M. H. GRAYSON

ELECTRICAL COMMUNICATION

Technical Journal of the
International Telephone and Telegraph Corporation
and Associate Companies

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ADJACENT-CHANNEL LOW-POWER MOBILE TRANSMITTER-RECEIVER

LONG-RANGE-NAVIGATION INSTRUMENTATION

TRIODE AMPLIFIERS FOR OPERATION FROM 100 TO 420 MEGACYCLES

TWINPLEX AND TWINMODE RADIOTELEGRAPH SYSTEMS

DESIGN CONSIDERATIONS FOR A RADIOTELEGRAPH RECEIVING SYSTEM

PRECISION CALIBRATOR FOR LOW-FREQUENCY PHASE-METERS

TUNABLE WAVEGUIDE FILTERS

CURRENT FLUCTUATIONS IN DIRECT-CURRENT GAS DISCHARGE PLASMA

NOTE ON REACTIVE ELEMENTS FOR BROAD-BAND IMPEDANCE MATCHING

TELEPHONE STATISTICS OF THE WORLD

INTERNATIONAL TELEPHONE AND TELEGRAPH

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Adjacent-Channel Low-Power Mobile Transmitter–Receiver

By E. B. MOORE

Federal Telephone and Radio Corporation; Clifton, New Jersey

INCREASED demand in recent years for very-high-frequency mobile service in the United States has made it necessary for the Federal Communications Commission to assign channels in the 148–174-megacycle range at 60-kilocycle intervals in the same area. This action, which was required to provide service to the maximum number of mobile-equipment users, has made necessary the development of vehicular transmitters and receivers having special performance features to permit satisfactory operation with such close channel spacing. To meet the requirements of the majority of mobile-service subscribers, a 10-watt low-cost equipment, which has been designated as *FT–153–10*, is available in both 6- and 12-volt-powered versions.

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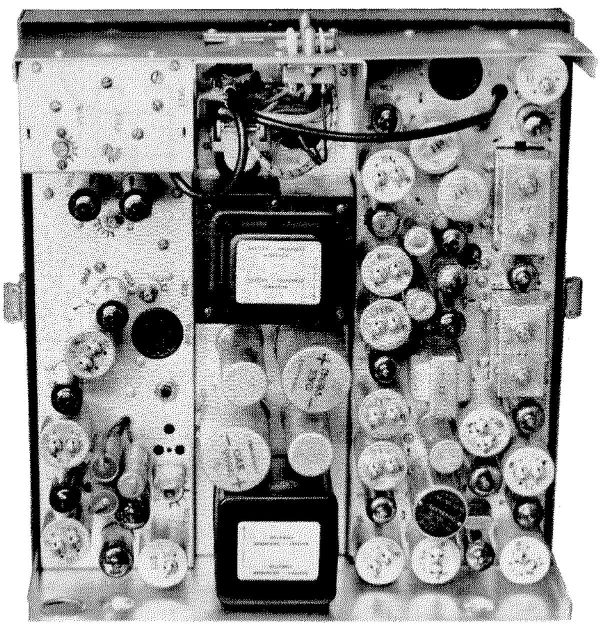
The standards established by the Federal Communications Commission are based around a frequency- or phase-modulated type of transmission having a maximum carrier deviation of 15 kilocycles for modulation frequencies up to 3000 cycles. The Commission also requires that all vehicular and fixed-station transmitting equipments be provided with deviation-limiting systems to prevent undesired radiation of sideband frequencies into the adjacent channels.

Transmitter characteristics of importance for use in a vehicular system include: high degree of frequency stability, high noise attenuation, low harmonic and spurious radiation power, and low battery drain.

The receiver selectivity must be sufficient to attenuate greatly the sideband frequencies of the adjacent-channel transmissions and still have a nose bandwidth great enough to accept the sidebands of the desired "on-channel" transmission. Consideration must be given to obtaining the higest practicable immunity to interference of the carrier intermodulation and desensitization types, commensurate with sensitivity requirements. Image and other spurious responses must be minimized in like fashion.



Top chassis view of the *FT-153-10* equipment. The receiver is at the right, the transmitter at the left, and the power supplies are between them.

1. Transmitter

The phase-modulated transmitter has as an exciter, a crystal-controlled Miller-type oscillator, whose frequency is multiplied 36 times in 4 stages. The output of the last multiplier drives two *5812* beam power tubes in push-pull class-*C* operation. For convenience in stocking tubes and standardization, only two types are used, the *12AT7* and *5812*, in the 7-tube circuit. The *5812* tubes are used as the multiplier-drivers in the 3 stages preceding the output amplifiers. This arrangement affords a large saving in average current drain inasmuch as the five *5812* tubes have filamentary type emitters that may be energized almost instantaneously. The filament supply voltage for these stages is under the control of a microphone hookswitch, when the transmitter is operated in the complete mobile system. Therefore, the filaments of these stages are unenergized until the microphone is lifted from its receptacle. Plate supply voltage is applied when the operator depresses the microphone button.

Figure 1 shows a simplified schematic diagram of the oscillator-modulator. The operation of the phase modulator is as follows. The voltage appearing at the output of the oscillator at point *A* is advanced in phase approximately 60 degrees by means of capacitor *C2* and resistances *R1* and *R2*, and is applied at reduced level to the grid of the modulator at point *B*. The voltage at point *A* is also coupled to the plate of the modulator through *R3* and *C1* and appears in the same phase as the voltage at point *A*. When the modulator is in its quiescent unmodulated condition, its output is formed from two voltages, one at the phase of point *A*, and the other 120 degrees displaced from that voltage because of the 60-degree phase shift of the resistance–capacitance network and the 180-degree phase reversal of the modulator. These components are approximately equal in amplitude. As the modulating voltage is applied, the resulting gain variation at the audio modulation rate causes a variation in magnitude of that component of the voltage at point *C*, which is amplified by the modulator. As this component varies in amplitude, there is a corresponding

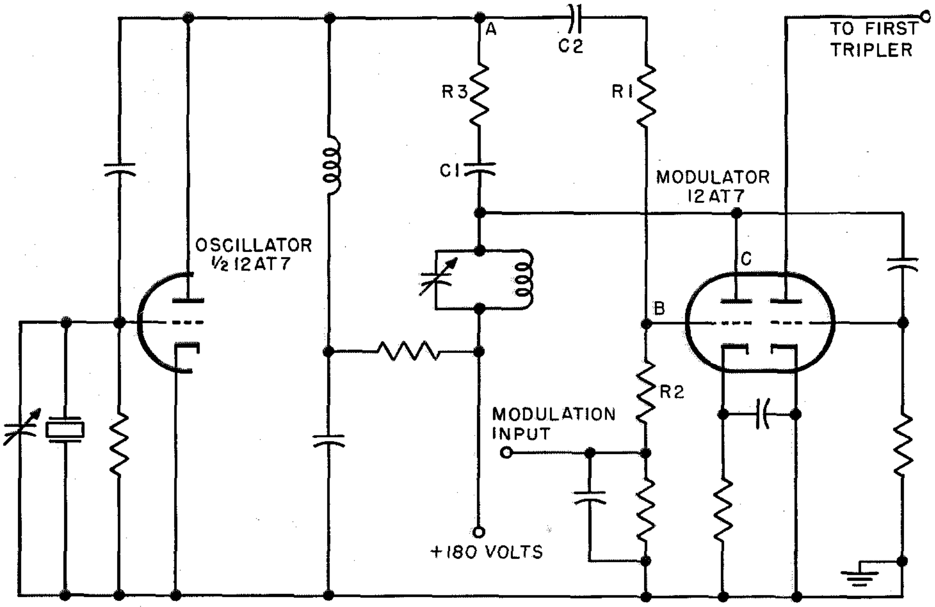


Figure 1—Circuit arrangement of the oscillator and modulator.

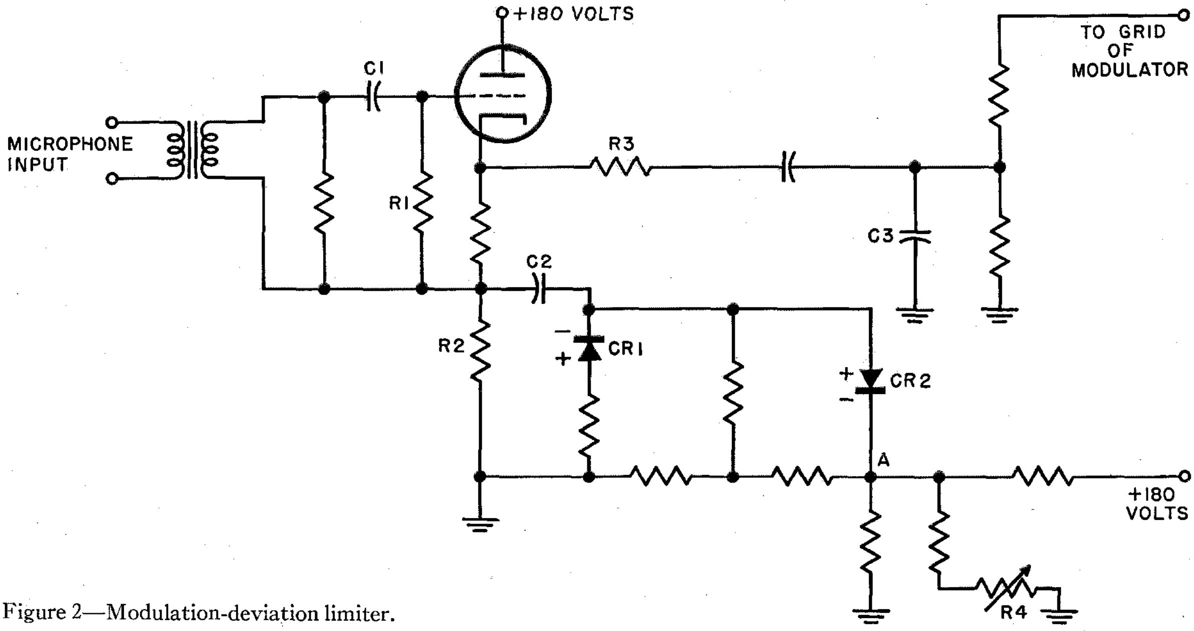


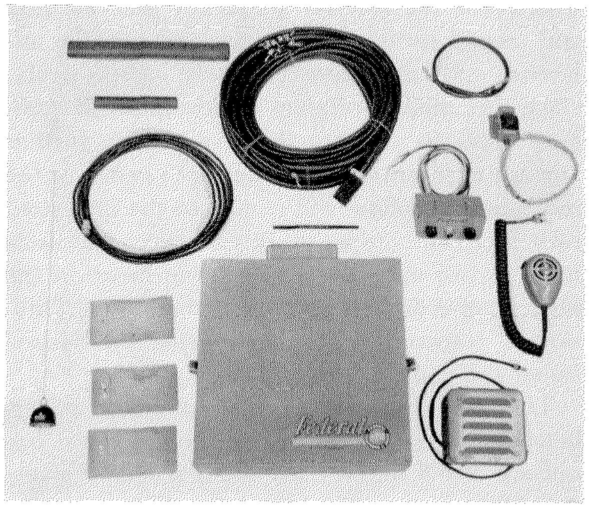
Figure 2—Modulation-deviation limiter.

change in the resulting phase, thereby providing the desired phase-modulation characteristic. There is also an accompanying amplitude variation, which is effectively eliminated due to the fact the following multiplier–driver stages are saturated.

The modulation-deviation-limiter circuit is shown in Figure 2. The audio signal appearing at the output of the microphone transformer passes through the differentiating network *C1–R1* to the grid of a cathode-loaded amplifier. The resistance–capacitance ratio of the differentiating network is such as to provide 3 decibels attenuation at 3000 cycles. Limiting is accomplished in the cathode load of the audio modulation amplifier by means of germanium diodes *CR1* and *CR2*. These rectifiers are biased at a level of about 20 volts at point A with positive direct voltage developed from the plate supply through a voltage divider. The diodes, which are in series, therefore are each biased with approximately 10 volts. This rectifier network is shunted across the cathode load *R2* through coupling capacitor *C2*. When the audio signal at the cathode exceeds 10 volts, zero to peak, each rectifier starts conduction on its half cycle, thereby presenting a low impedance to the cathode output of the amplifier, resulting in limiting action above this level. Control of the level at which limiting takes place is accomplished by bias control *R4*. The output of the cathode-loaded amplifier feeds the grid of the modulator tube through an integrating network consisting of *R3* and *C3* which has a 3-decibel attenuation at 300 cycles.

The maximum audio distortion produced by this limiter system is maintained at about 10 percent. The use of the symmetrical or balanced type of limiter reduces the second-harmonic content to a low level, leaving a predominant third harmonic in addition to the fundamental. If it is assumed that with the most severe conditions of limiting a square-wave output is obtained from the limiter, this third harmonic would be thirty percent of the fundamental. The integrating network, which provides a frequency attenuation of about 6 decibels per octave, further attenuates the third harmonic another 12 decibels, resulting in an over-all reduction of about 22 decibels. It will be seen readily that the fifth harmonic of the square wave will be attenuated to a point that it is small in comparison to the third harmonic.

This combination of differentiation, limiting, and integration, provides a system capable of audio-frequency modulation in the range of 300 to 3000 cycles, the standardized audio range for mobile service, with very effective deviation limiting at a minimum cost. The fact that the system does not depend on time-constant parameters, such as are required for volume compressors or other automatic-volume-control systems, makes its action instantaneous in nature, thereby preventing the possibility of momentary interference into the adjacent channel.



Complete equipment including installation material.

2. *Receiver*

To afford the highest practical degree of rejection to interfering responses due to adjacent-channel transmitters, the receiver must be equipped with selective filters having large attenuation to the carrier and phase-modulated sidebands of such transmissions that are only 60 kilocycles removed. It is further required that careful consideration be given to the distribution of gain and selectivity, without degradation of sensitivity, in order that desensitization does not occur in the early amplifier stages of the receiver before the full selectivity is developed.

The most severe adjacent-channel interference is the carrier intermodulation type that is produced by the mixing of the second harmonic of the interfering adjacent-channel signal and the fundamental of the alternate-channel signal. It can easily be seen that one product of this type of mixing produces an on-frequency signal as follows:

$$2(f_s + 60) - (f_s + 120) = f_s, \qquad (1)$$

where f_s is the receiver frequency and the first and second terms of the expression are the adjacent and alternate channels, respectively. Once this mixing product is developed in sufficient level to activate the receiver, no further amount of filtering in the selective circuits of the receiver that follow will eliminate the undesired response. This type of interference manifests itself in two ways; either by being heard in the absence of a desired signal and therefore a "nuisance" response or by "capturing" a desired signal and causing a call to be lost.

The receiver is a double-conversion superhetrodyne type having a first intermediate frequency of 3.8 megacycles and a second conversion frequency of 455 kilocycles. By converting to 3.8 megacycles from a carrier frequency in the range of 148–174 megacycles in the first mixer, it is possible to reject the adjacent channel in the band-pass filters of the first intermediate-frequency amplifier sufficiently to minimize desensitization and carrier intermodulation at this point. Converting to this relatively low frequency, however, imposes stringent requirements on the selective circuits of the radio-frequency amplifiers because of the proximity of the image and other spurious responses. To reject these image responses in the radio-frequency amplifiers, four high-Q delay-line type of tuned circuits are used in addition to a single-tuned antenna coil. The delay-line tank circuits are contained in silver-plated brass-shielded assemblies, fabricated from square tubing. These circuits provide attenuation at frequencies one percent off resonant frequency ranging from 56 to 42 decibels from the low end to the high end of the band, respectively. The resulting spurious-response rejection is greater than 90 decibels over the frequency range.

The major part of the receiver's selectivity is produced by the interstage filters of the second intermediate-frequency amplifier, which has a center frequency of 455 kilocycles. Three quadruple-tuned dissipative band-pass filters[1] having the over-coupled or "Chebishev" amplitude-response shape are used to obtain the desired selectivity. The quadruple-tuned resonant circuits are formed from low-cost double-tuned transformers coupled by mutual inductance within each assembly with the two double-tuned assemblies capacitively coupled to each other

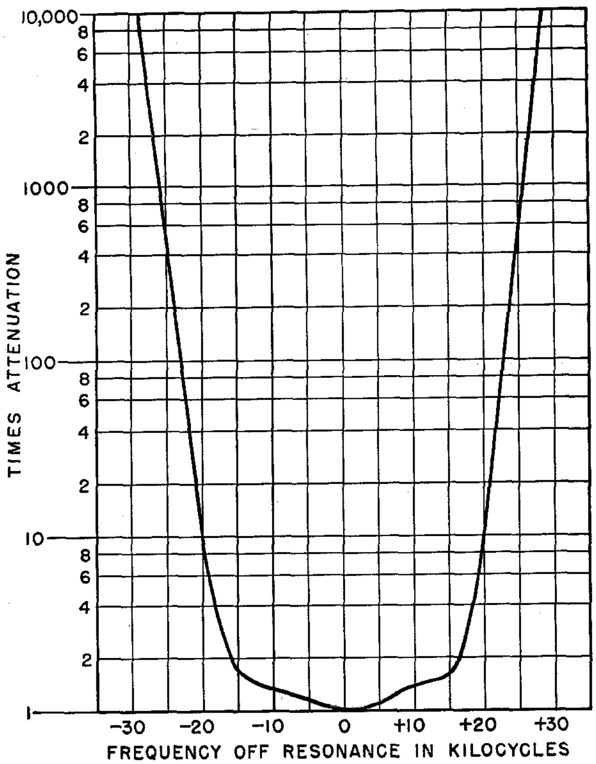


Figure 3—Selectivity of the 455-kilocycle intermediate-frequency amplifier. Three quadruple-tuned circuits are used.

to form the complete quadruple-tuned filter. The resulting selectivity of the second-conversion intermediate-frequency amplifier is shown in Figure 3. The schematic of a typical amplifier stage with its associated quadruple-tuned filter section is shown in Figure 4.

The coupled type of quadruple filter offers several advantages over a stagger-tuned or combination stagger-tuned and coupled filter. Some of these advantages are: less filter insertion

[1] M. Dishal, "Design of Dissipative Band-Pass Filters Producing Desired Exact Amplitude–Frequency Characteristics," *Electrical Communication*, v. 27, pp. 56–81; March, 1950; also *Proceedings of the I.R.E.*, v. 37, pp. 1050–1069; September, 1949.

loss, thereby permitting fewer amplifier stages; and the fact that only a single-frequency signal source is required for alignment rather than the four frequencies that are required for the stagger-tuned method. The same selectivity shape is obtainable with either a stagger-tuned or coupled design for an equal number of resonant tanks. It is necessary, however, to maintain the coupling tolerance in each transformer section to narrow limits when using the coupled type. It was also determined that control of physical spacing of the universal-wound coils was not sufficient and for this reason a test fixture was designed to permit the setting of the spacing to a specific coupling, measured as a function of the product KQ.

To be effective, a sensitive vehicular receiver must be equipped with a squelch system to eliminate undesired receiver noise when no signal is being received. The squelch system incorporated in this receiver operates on the noise-quieting phenomenon resulting when a carrier enters the input of the receiver. A signal of 0.3 microvolt at the antenna input is sufficient to open the receiver fully. The schematic diagram of the squelch circuit is shown in Figure 5 and operates as follows.

The output of the discriminator is coupled to resonant circuit *Z1*, which is a nonadjustable tank having a resonant frequency in the vicinity of 16 kilocycles. The spectrum of noise in the region of this frequency is then used to control the squelch action. *Z1* is coupled to a two-stage high-gain amplifier having a similar tank *Z2*, as the output load. When no signal is received, the receiver noise is amplified and coupled to diode *D1* of the *6AQ6* dual-diode triode. This diode acts as a direct-current restorer or clamper to the noise voltage and effectively doubles the negative peak voltage with respect to ground. In the absence of signal and with the squelch control *R1* set to the threshold of operation, the amplified receiver noise develops an average direct-current potential of about 70 volts, negative with respect to ground. The peak negative voltage will then be nearly twice this value. Under such conditions, the neon lamp *N* conducts on the negative half cycle causing about 20 volts negative to be developed at point *A*, which biases the triode section of the *6AQ6* amplifier and the output amplifier to plate-current cutoff. When a signal of about 0.3 microvolt enters the receiver, the receiver noise is captured in the amplitude-limiter stages, which precede the discriminator. This action causes the noise to decrease about 10 decibels and the developed average direct voltage at diode *D1* to reduce to about 23 volts negative. Because the neon lamp requires a 60-volt potential to conduct, this voltage does not transfer to point *A* and the cutoff bias voltage at the grids of the audio amplifiers is removed accordingly. Diode *D2* is used as a protective device to prevent the possibility of positive voltage from being developed in the system due to circuit failure.

Noise quieting for control and the application of the

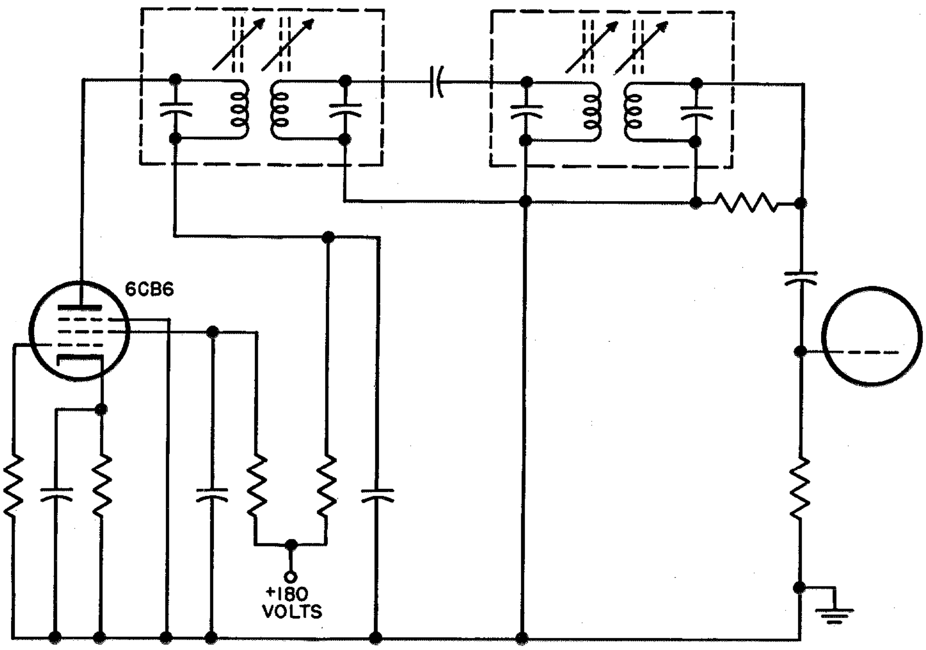


Figure 4—Quadruple-tuned intermediate-frequency amplifier stage. Three are used in the receiver.

neon lamp as a switching device, provide an effective fast-acting squelch operation with little cost. In addition to muting the speaker output when no signal is received, this squelch circuit arrangement also serves to reduce average current drain inasmuch as the audio power amplifier does not draw plate current until a call is received.

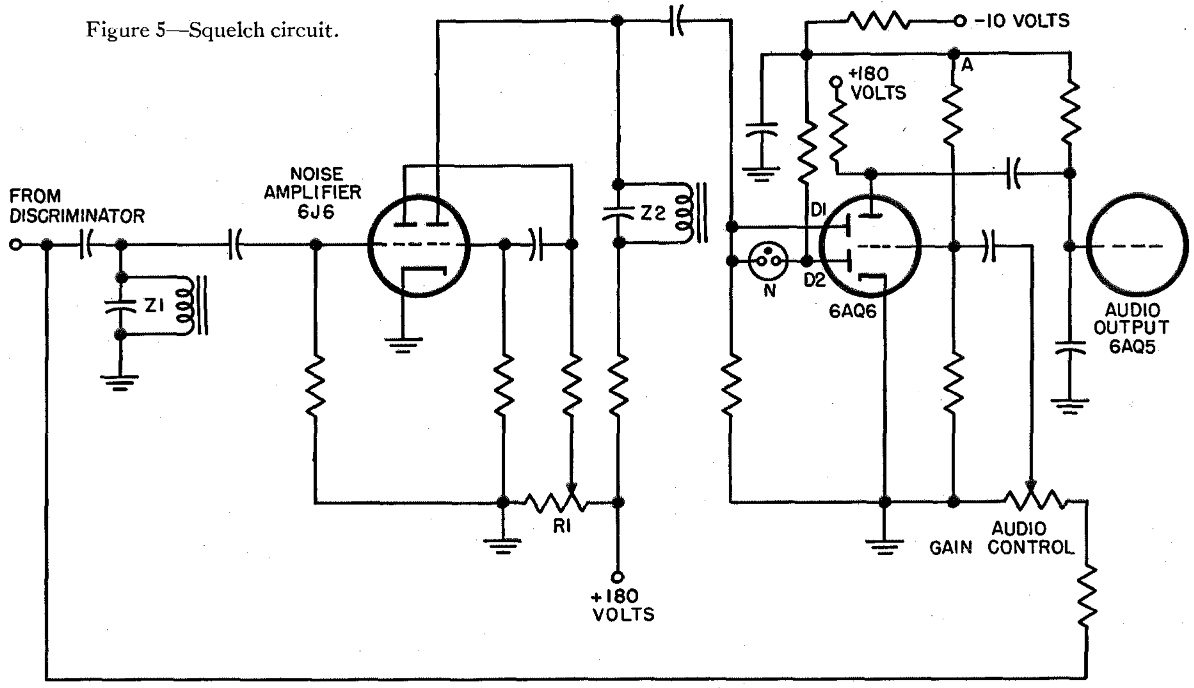


Figure 5—Squelch circuit.

3. Power Supply

Consistent with the policy of conserving battery drain and minimizing cost, synchronous-type vibrator supplies, which do not require the use of rectifier tubes, are used for both the receiver and transmitter. The vibrator units used in the receiver and transmitter sections of the power supply are identical. This flexibility simplifies the stocking of these components and permits interchanging of vibrators when emergency operation is desired on either the transmitter or receiver, when a vibrator failure occurs.

To obtain increased life expectancy of the transmitter vibrator, the vibrator reed is energized when the microphone is lifted from the hookswitch. Current is not applied to the interrupter contacts, however, until the microphone button is depressed. By starting the vibrator reed before applying current to the interrupter contacts, the possibility of the contacts freezing due to arcing is reduced.

4. Acknowledgment

The author makes grateful acknowledgment to Messrs. R. C. Davis, J. Zavoda, G. Menhennett, M. Macomber, and T. MacMillan for their work in the development of the equipment. Acknowlegment is made to the Federal Telecommunication Laboratories for their assistance in the design of the delay-line radio-frequency filters and quadruple-coupled filters used in the receiver.

Long-Range-Navigation Instrumentation

By BEN ALEXANDER

Federal Telecommunication Laboratories, Incorporated; Nutley, New Jersey

IF THE EARTH consisted of a solid land mass and if cost were no object, ground-navigation facilities for long-range navigation could be closely spaced and almost any desired degree of accuracy and reliability obtained. Because of the distances between usable land masses, however, it is generally agreed that the ground-based facilities of a world-wide long-range navigation system must be capable of service ranges in excess of 1500 miles.

Careful examination of the radio-frequency spectrum shows that only frequencies below a few hundred kilocycles per second can provide the order of reliability necessary for long-range navigation, and the optimum frequency band from a cost consideration seems to be near 100 kilocycles.[1]

The design of a system of long-range navigation initiated in 1945 has now reached the stage of field evaluation and has produced much information that would seem to apply, with some modifications, to the long-range navigation problem in general as well as to the Navaglobe[2] system.

The aperture of the transmitting antenna structure turns out to be the most characteristic feature of ground-based long-range navigation systems. Antenna spacings of many wavelengths and spacings of a small fraction of a wavelength both present difficulties which, somewhat surprisingly, are overcome when a spacing near one-half wavelength is used.

Large apertures, at the frequencies involved, require either large antenna arrays or multiple transmitting sites. This is expensive. More seriously, one of the propagation errors (height error) is negligible with small apertures but gets very large with apertures of more than one or two wavelengths. Finally it turns out that large-aperture systems tend to involve a bothersome amount of airborne complexity in order to obtain direct-reading instrumentation.

Very small apertures, on the other hand, introduce another type of difficulty. Since a considerable depth of modulation (with position) is necessary for the transmissions to be useful, apertures smaller than about a quarter wavelength almost of necessity involve antennas fed antiphaseally. This requires large antenna circuits for relatively small amounts of radiation. In consequence, small assymmetry of the terrain in the vicinity of the antenna structure results in large distortion of the pattern actually radiated. This seems to be the origin of the relatively large errors in Adcock-type direction finders and also of the large errors experienced with the low-frequency omnirange system.

Clearly, a compromise must be made. It has been found that an antenna aperture of about one-half wavelength balances the advantages, avoiding many of the errors characteristic of both larger and smaller apertures. It is interesting to note that some antenna configurations—notably regular polygons with an odd number of sides—show distinct advantages in further reducing the height error.

Examination of the experiences gained in all sorts of navigation systems and analysis of propagation measurements lead to the conclusion that half-wave antenna systems can be made so that the root-mean-square azimuth error resulting from propagation uncertainties and terrain irregularity near the antennas is about one-quarter degree. The question then arises: Can the system be given practical airborne instrumentation sufficiently accurate to keep the over-all error almost as small? There are two classes of errors involved in this question: first, errors due to noise on the signal, and second, errors due to imperfections of the airborne instrument itself.

Minimization of errors due to noise must make use of the fact that long-range navigation systems can always tolerate a reading time of as

[1] P. R. Adams and R. I. Colin, "Frequency, Power, and Modulation for a Long-Range Radio Navigation System," *Electrical Communication*, v. 23, pp. 144–158; June, 1946.

[2] H. Busignies, P. R. Adams, and R. I. Colin, "Aerial Navigation and Traffic Control with Navaglobe, Navar, Navaglide, and Navascreen," *Electrical Communication*, v. 23, pp. 113–143; June, 1946.

long as 30 seconds. By using this permissible delay, instrumentation can often be devised that reduces the effects of noise to a marked degree. This process is substantially one of filtering the information to discard apparent position changes that seem to occur more rapidly than the airplane itself can move. This filtering process can be effected by various techniques such as the use of filters, integration, cross correlation, etc.

Practical embodiments of these principles should produce a signal-to-noise ratio of better than one to one before the received signal is detected. This requires reducing noise in the vicinity of the signal before detection by cross correlation, narrow-band filtering, or an equivalent process. Once the one-to-one signal-to-noise ratio has been reached, it is desirable to detect and then make use of the full reading time available by means of relatively cheap post-detector integration.

There is one further matter related to noise that was not apparent early in the Navaglobe program. A linear detector has the property of causing a bearing "push" when the signal is noisy. This is an error in the average position of the bearing indicator that no amount of subsequent averaging or post-detector integration can reduce. A square-law detector, however, does not have this property. This phenomenon, on further examination, appears to be common to many navigation systems operating in the presence of high noise levels.

By incorporating these techniques, noise errors can usually be made negligible, even at thousands of miles from the transmitter.

In considering instrumental errors, the magnitude of the problem can be realized from a simple example. Suppose it were necessary (in the process of instrumenting the system) to obtain bearings by measuring the phase difference between two audio-frequency tones. This is usually done by shifting the phase of one of the tones until it is cophaseal with the other and measuring the amount of phase shift required. This phase shifting is best performed with the kinds of highly accurate two-phase resolvers now available to the industry. These devices, and their associated circuitry, however, have errors up to one-half degree under service conditions and their root-mean-square error is near one-quarter degree. So it can be seen that one element of the instrument can in itself have errors as large as the expected propagation errors unless it is arranged to have one electrical degree represent less than one degree of azimuth. Fortunately, the half-wave aperture, which appeared desirable for entirely different reasons, naturally tends to provide the equivalent of two electrical degrees for each azimuth degree. Consequently, a one-quarter-degree error of measurement causes only one-eighth-degree azimuth error.

In every navigational instrumentation, there are two or more signals to compare. In the omnirange example, these are the reference and the bearing waveforms, the phases of which must be compared. In Navaglobe, it is the amplitude of the signals received from three sequentially keyed antenna pairs that is compared. Consequently, a sound conclusion is that accurate noncritical instrumentation can be attained conveniently if a common channel carries all the signals up to the comparison point. Then, the problem of constantly calibrating the system with pilot signals is eliminated, as are most tendencies of the indicator to wander out of calibration with tube aging, voltage and temperature variations, etc.

This leads naturally to a servo type of instrumentation. By arranging to compare signals present in a single channel in such a way as to bring the output of the comparison device to zero when the indicator is properly oriented, the post-detector averaging or integration can be performed on an error signal, rather than on the unprocessed data. Since accurate storage for a long time is a complex operation, a technique of error storage (which needs to be only roughly proportional) is very much to be desired. It turns out, of course, that any desired amount of post-detector integration can be obtained merely by suitably choosing the gear train that couples the output shaft of the system with a motor energized by the error signal coming from the comparison device.

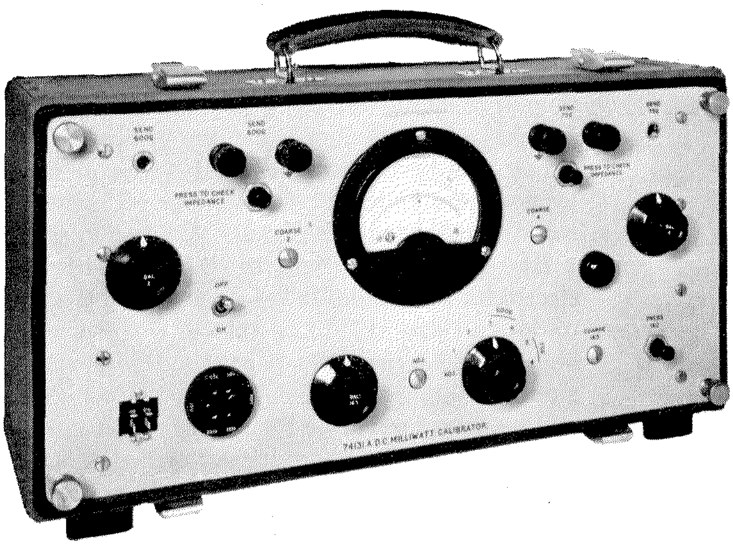
There has been considerable difference of opinion in the past on the relative merits of "simultaneous-transmission" systems like omniranges and "sequential-transmission" systems like Navaglobe. Actually, there is nothing to justify selection of one approach over the other from the standpoint of information theory, provided there is equivalent power, modulation depth, antenna aperture, etc. However, from considerations of equipment accuracy and simplicity, the possibility of completely common circuits offered by sequential transmission provides a fundamental advantage that should not be discarded lightly. The problems of receiver design bear this out dramatically. Designing a receiver with distortion down 60 decibels requires careful engineering, but designing a receiver with small differences in phase shift in separate channels is a major undertaking. Navaglobe instrumentation accuracy of within 5 minutes seems to bear out this point.

The author acknowledges the cooperation of Mr. Harry H. Davis, Mr. Fred Moskowitz, and Captain Oscar D. Goldberg of the Air Force's Rome Development Center and their associates for their participation in the project to which this paper relates.

Recent Telecommunication Development

Sub-Standard Milliwatt Calibrator

THE ACCURACY of milliwatt test sets is normally checked by reference to a sub-standard direct-current milliammeter or a standard direct-current potentiometer, neither of which is suitable for general use in laboratory or field if its absolute accuracy is to be guaranteed. This lack of a suitable instrument has been removed by the introduction of the *74131-A* Direct-Current Milliwatt Calibrator by Standard Telephones and Cables, Limited.



The calibrator will send a direct-current power of 1 milliwatt into a 600-ohm or 75-ohm load and is, therefore, suitable for checking 600-ohm and 75-ohm milliwatt test sets. The ultimate standard voltage is a Weston miniature standard cell (manufactured by Messrs. Muirhead Limited), which has an absolute accuracy within ±0.01 per cent, its temperature coefficient being negligible in this application. This enables an accuracy of better than ±0.02 decibel to be achieved in the calibrator. The cell is extremely robust both electrically and mechanically and can be stood upside-down for long periods without detriment.

An electronic bridge network enables the source of direct-current power to be accurately adjusted by reference to the voltage of the standard cell, as it is extremely sensitive to small differences in the potentials applied to the balancing arms. In fact, a change of 0.05 decibel in the power fed to the milliwatt test set being calibrated causes a change of deflection of over $\frac{1}{4}$ inch (6 millimetres) on the meter.

The calibrator is housed in a robust aluminium box capable of withstanding severe field use and operates from alternating-current mains. Its weight is 16 pounds (7.3 kilograms), and its dimensions are $17\frac{1}{2} \times 9\frac{1}{8} \times 8\frac{7}{8}$ inches ($445 \times 232 \times 225$ millimetres).

Triode Amplifiers for Operation from 100 to 420 Megacycles*

By D. C. ROGERS

Standard Telephones and Cables, Limited; Ilminster, England

CONCENTRATION during the war years on the design of valves suitable for the shorter decimetric wavelengths resulted in there being very few valves available for use in the frequency range from 100 to 420 megacycles per second other than expensive all-glass disc-seal or lighthouse types. This state of affairs was particularly noticeable in the case of small transmitting amplifiers.

By careful attention to the geometry of electrodes and leads, it has been found possible to design triode valves capable of operation at frequencies up to 420 megacycles per second in grounded-grid circuits, of being mounted on conventional pressed-glass bases, and of using only the recognized techniques of receiving-valve manufacture. Design features requiring extra consideration and some of the results achieved with various special valve types are outlined.

• • •

The necessity during the war years to extend the frequency range of all types of radio valves as far as possible gave rise to the introduction of a range of triodes able to operate at decimetre and even centimetre wavelengths. This was achieved by using very small interelectrode spacings to reduce transit-time effects and by employing disc-seal connections to the electrodes to reduce lead inductances and facilitate incorporation of the valve into coaxial-line circuits. The success achieved with these valves resulted in relatively little effort being applied to extending the frequency range of more conventional types, particularly of small transmitting valves.

The designer of equipment working at wavelengths in the region of one or two metres frequently finds himself compelled to incorporate in his equipment a disc-seal valve capable of operation at very much shorter wavelengths. The special techniques involved in the manufacture of these valves inevitably bring increased cost, and in view of the greatly extended use of these wavelengths since the war, it is clearly desirable to search for a more economical alternative. Furthermore, the circuit equipment associated with disc-seal tubes is also in most cases more expensive and is frequently less convenient in use, particularly from the point of view of replacement in event of failure.

The success of the disc-seal triode at very-high frequencies is due not only to the special form of construction but also to the grounded-grid circuit in which this type of tube is almost invariably used. It seems natural to enquire whether the grounded-grid circuit can be used to advantage with tubes of more orthodox construction. Strangely enough, this possibility does not appear to have been extensively investigated from the point of view of small transmitting valves, although receiving valves for grounded-grid operation have been designed and the grounded-grid circuit has been used for many years in large transmitters. Indeed, it was to large transmitters that the grounded-grid circuits were first applied.

By careful attention to the geometry of electrodes and leads, it has been found possible to design tubes able to operate at frequencies as high as 420 megacycles per second, and yet be mounted on conventional pressed-glass bases and be manufacturable by employing only those techniques used in producing orthodox small tubes. By this means, their cost has been reduced to but a small fraction of that of their disc-seal counterparts, with a considerable increase in convenience and only a small sacrifice in performance.

1. *Lead Inductance and Interelectrode Capacitances*

It will be of interest to examine the significance in the grounded-grid circuit, which differs from

* Reprinted from *British Institution of Radio Engineers Journal*, v. 11, pp. 569–575; December, 1951.

the grounded-cathode case, of the various lead inductances and interelectrode capacitances.

The first essential is, of course, that it must be possible to design an output circuit of satisfactory efficiency resonant at the desired frequency; for convenience, it is normally desirable to use a coil rather than a transmission line where possible. The anode-to-grid capacitance must, therefore, be kept to a minimum to reduce circulating currents and to enable a practical coil to be used. Reducing the anode lead inductance will also ensure that as much as possible of the inductive part of the circuit is external to the valve.

At the input circuit, the cathode–grid capacitance will inevitably be considerably greater than the anode–grid capacitance due to the small cathode–grid distance needed to obtain a high mutual conductance and adequately short transit time. In comparison with the anode circuit, losses in the cathode circuit are of much less significance, owing to the very low impedance and Q of the circuit. In any case, input circuit losses will introduce loss in gain, but not loss in efficiency, which in most cases is more important.

At the highest frequencies, it is quite possible to operate beyond the frequency at which the cathode lead inductance resonates with the cathode–grid capacitance, since this will merely result in the input impedance, measured at the grid pins, becoming inductive; this can be allowed for in the design of the input impedance-transforming network, and corresponds to the case of a disc-seal tube with a transmission-line input circuit in which the first voltage node is within the valve envelope. Such a condition would not be acceptable at the anode, as the losses would probably be too high; thus it is usually found that the maximum usable frequency of a given valve is limited by the behaviour of the anode circuit.

There is, however, further reason for reduction of the cathode-lead inductance, peculiar to the use of grounded-grid amplifiers under class-C conditions. The low impedance presented by the cathode to the driving source is non-linear and unless the driving source has negligible impedance at harmonics of the driving frequency, distortion of the input waveform will take place with consequent loss in efficiency. A low cathode-lead inductance assists in obtaining the required condition, in contrast to the linear class-A case, where a reactance in the cathode lead can usually be accommodated as part of the impedance-transformation network from the previous stage. The falling reactance-versus-frequency characteristic of the cathode–grid capacitance is also of advantage in reducing distortion of the input waveform.

Excessive inductance in series with the grid of a grounded-grid amplifier can cause feedback and instability. In general, it is found necessary to connect the grid to as many base pins as possible, using extensive conductors within the valve envelope. By suitable choice of pin arrangement, these connecting pieces can also be used to shield the cathode leads from the anode and thus simulate the effect of the disc of a disc-seal valve. Returning input and output circuit by-pass capacitors to different grid pins is another device to reduce the impedance common to the two circuits.

By comparison with a similar valve in a grid-driven circuit, the grounded-grid amplifier has a much lower power amplification. Hence, to produce a valve acceptable to the circuit designer, the valve designer must be prepared to use large values of amplification factor and mutual conductance. However, if these are possible in disc-seal valves, there is no reason why there should be any unsurmountable difficulty in more orthodox valves, and in some of the designs now to be described mutual conductances between 20 and 30 milliamperes per volt are obtained.

2. 24-Watt Valve for Frequencies Up to 200 Megacycles

One of the first applications to arise for a high-slope triode for grounded-grid use occurred in the design of a transmitter–receiver for the civil airlines. This equipment operated in the band from 118 to 132 megacycles and was required to deliver an output of at least 10 watts; economy in space and weight demanded that the valve should operate from the same anode supply voltage as the remainder of the equipment, namely, 300 volts. Assuming the instantaneous anode potential at the moment of maximum anode current to be about 100 volts, a peak anode

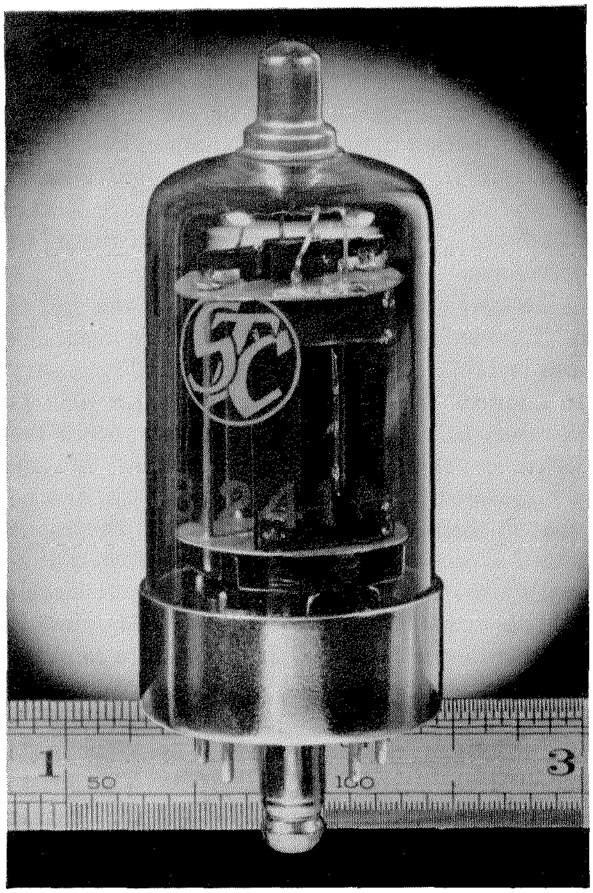


Figure 1—Type *3B/241M* grounded-grid triode for frequencies up to 200 megacycles.

current of about 300 milliamperes is required to produce the desired output power. Since in the grounded-grid circuit the anode current flows also through the input circuit together with the grid current, that is, the input current exceeds the output current, it is readily apparent that a high power gain can be obtained only by keeping the driving voltage small, and that this involves a very-high value for the mutual conductance. More detailed consideration will show a mutual conductance of at least 20 milliamperes per volt to be desirable.

One of the obstacles encountered in obtaining a large mutual conductance is that the necessarily small cathode–grid clearance gives rise to overheating of the grid and consequent grid emission. In valves for grounded-grid use, however, all internal screening is connected to the grid, and thus a large area of metal becomes available for grid cooling.

Figure 1 is a photograph of the valve evolved to meet this airborne application and manufactured commercially under the code *3B/241M*. Figure 2 shows some details of the construction and illustrates the means by which the requirements relating to capacitances, lead inductances, etc. have been met. It will be seen that to reduce the anode-to-grid capacitance, the anode does not encircle the cathode and grid assembly, but is constructed in two completely separate portions facing the active surfaces of the cathode; this avoids the capacitance that would otherwise exist between the grid support rods and parts of the anode that serve no purpose in collecting the beam. The two halves of the anode are joined by two nickel tapes to the top cap connection. A top cap is used to reduce the anode-to-grid

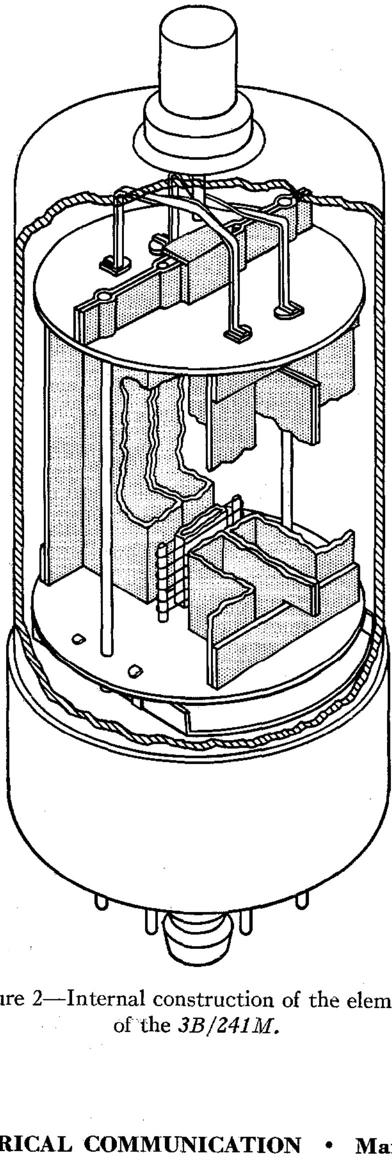


Figure 2—Internal construction of the elements of the *3B/241M*.

capacitance and to avoid voltage breakdown that would otherwise occur in the base in airborne application at high altitudes. The permissible anode dissipation without external cooling is 15 watts. The anode itself, however, is capable of greater dissipation without overheating, and this limit is set by the temperature of the bulb. If forced air-cooling of the bulb is employed, the dissipation may be increased to 24 watts.

The *3B/241M* is mounted on a loctal pressed-glass 8-pin base, and low grid-lead inductance is obtained by direct connection in the base of the grid to the screen, which serves to shield the cathode connections from the anode. This screen is itself welded directly to 4 of the base pins and thus serves both as a low-inductance connection for the grid and as a rigid mounting for the electrode assembly; it also provides cooling for the grid. Two pins are used for the cathode connections. Internally they are both welded to a flat nickel plate that extends to

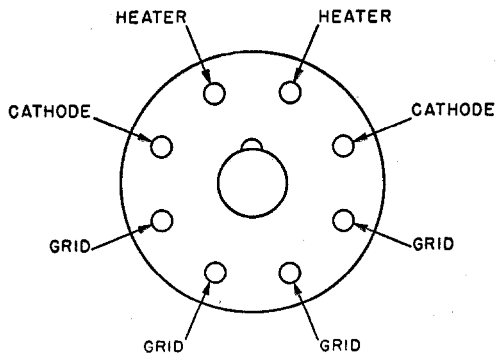


Figure 3—Pin arrangement and connections to provide multiple low-impedance connections to the cathode and grid. An 8-pin loctal base is used.

within ⅛ inch (3.2 millimetres) of the end of the cathode; two short nickel tapes provide connection between this plate and the cathode. The arrangement of the pins, shown in Figure 3, and disposition of the internal connections is such that at the highest frequencies a low-impedance transmission-line can be used to drive the cathode. The valve has a mutual conductance of 28 milliamperes per volt and an amplification factor of 100; typical operating conditions are given in Table 1.

TABLE 1

3B/241M, Typical Operating Conditions

Anode Voltage	300 Volts
Anode Current	80 Milliamperes
Load Impedance	1600 Ohms
Power Output	15 Watts
Efficiency	62.5 Per Cent
Input Impedance	120 Ohms, Approximately
Driving Power	2.0 Watts, Approximately
Grid Bias	−10 Volts
Grid Current	25 Milliamperes

Note: These operating conditions are based on the power delivered by the valve into the anode circuit. No account is taken of losses in the anode circuit but in view of the low load impedance these will normally be small.

It has been found that with careful attention to layout satisfactory circuits can be devised without the use of resonant lines at frequencies as high as 200 megacycles. The interelectrode spacings are such that transit-time effects at this frequency are small. Considerable care is necessary in keeping the inductance of the external grid connections low enough to avoid instability; it is desirable to use separate capacitors soldered

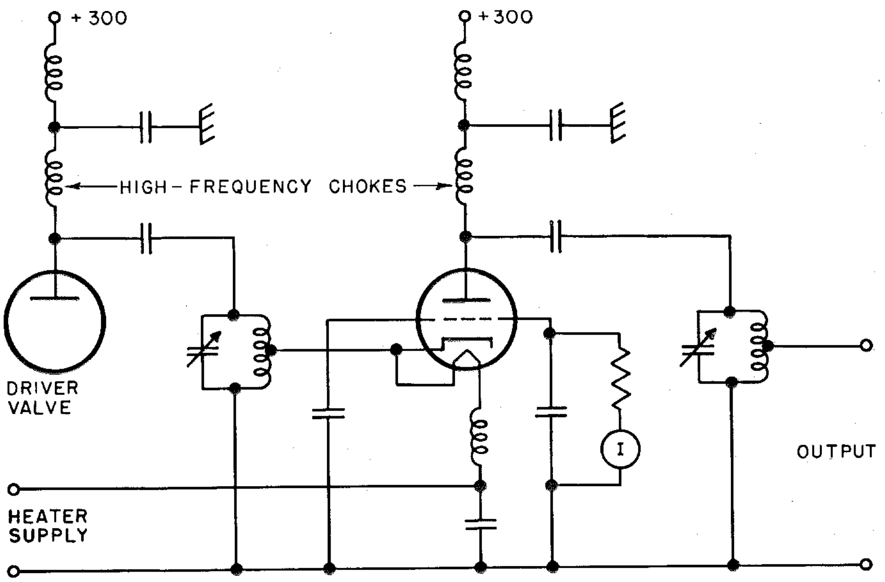


Figure 4—Grounded-grid circuit for working the valve as an amplifier.

to each of the 4 grid connectors. In some cases, a special capacitor constructed in the form of a ring surrounding the base has been used successfully.

At frequencies up to about 150 megacycles, a satisfactory low-impedance source for the cathode drive can be obtained by various means, for example, by connecting the cathode to a tap on the output-circuit coil of the driver stage, as in Figure 4. Above this frequency, the necessary inductance to be provided between cathode and earth becomes inconveniently small, and it is helpful to use the arrangement of Figure 5. A bridge of metal is soldered to the two cathode pins and the ends of the bridge are soldered to the chassis. Thus a very-low-inductance path to earth is formed, dependent on the areas of the two loops A in the figure. The bridge can conveniently be a pressed sheet-metal part, and values of inductance can then be reproduced within close limits. It is used as part of the resonant output circuit of the driving stage, the remainder of the coil being soldered to the centre of the bridge.

A significant advantage of this type of valve over the lighthouse type of disc-seal valve at these frequencies is that the effective anode capacitance of the lighthouse valve, when used with lumped circuits, is much higher than its catalogue value, due to the capacitance of the anode disc to the surrounding metalwork. In the case of the airborne equipment already mentioned, it was found possible to operate in any channel between 118 and 132 megacycles without retuning. This would not have been possible with a disc-seal valve.

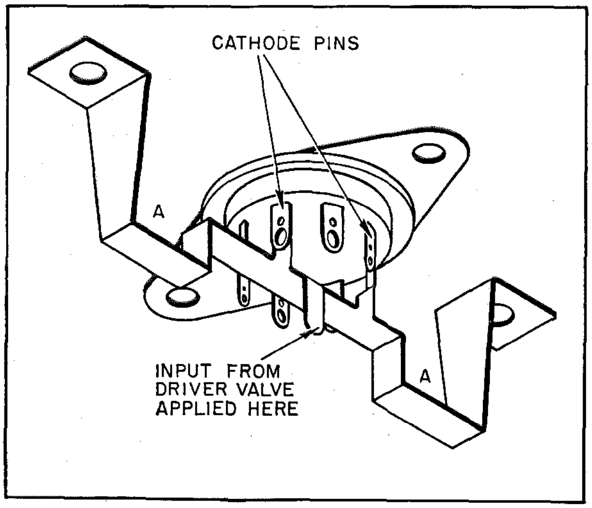


Figure 5—Cathode inductance for frequencies above 150 megacycles.

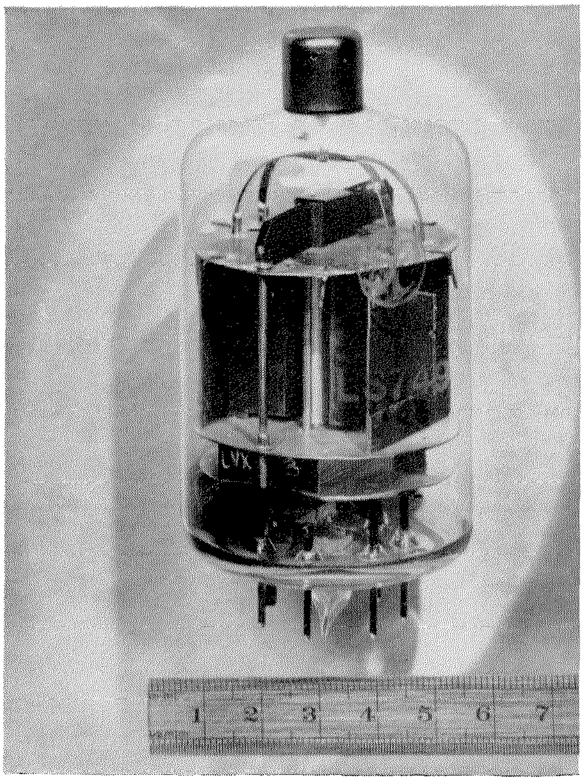


Figure 6—Experimental 50-watt valve for frequencies up to 200 megacycles.

3. *50-Watt Valve for Frequencies Up to 200 Megacycles*

A photograph of an experimental valve with an anode dissipation of 50 watts is shown in Figure 6. Capable of a total anode input of a little over 100 watts at 600 volts, this valve will deliver output powers in excess of 50 watts over the same frequency range as the valve just described. This larger valve will not be described in detail, as in most respects it is similar to the *3B/241M;* it is mentioned to illustrate the fact that larger output powers are available from valves having this simple form of construction.

4. *24-Watt Tube for Frequencies Up to 420 Megacycles*

It is possible that the frequency range of the type of valve just described could be extended by the use of resonant-line circuits; a more convenient alternative, however, exists in the double valve, already in use in the form of the

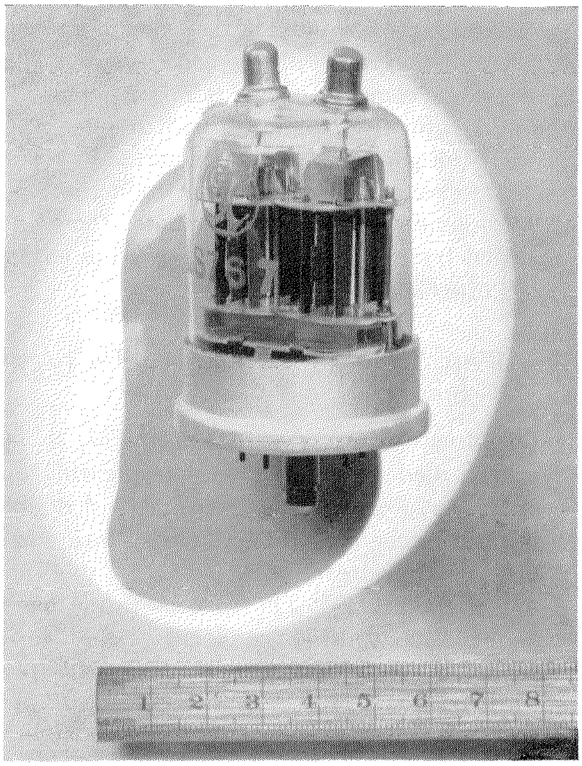


Figure 7—Type *LS767* experimental valve for frequencies up to 420 megacycles.

double beam tetrode, such as the type *832*. If a given design of triode is re-arranged to form two triodes each having half the mutual conductance, anode current, etc., and connected in a balanced or push-pull circuit, the effective capacitance across the tuned circuit is reduced by a factor of 4, and the maximum operating frequency is raised by a factor of 2 on this account alone. At the same time, the grids of the two triodes can be linked by a low-impedance connection within the envelope, thereby almost eliminating the high-frequency current in the external grid connection, and the inductance of this connection is of much less importance. Of course, in practice, a 4-to-1 reduction in capacitance cannot quite be achieved, due to the additional end-effects that obtain and the capacitance of the additional leads. Furthermore, the importance of losses in the anode circuit is greater, since the impedance is raised 4 times. Nevertheless, this type of valve has shown considerable promise, and a description will be given of an experimental valve, coded *LS767* and shown in Figure 7, which has operated successfully at 420 megacycles. Figure 8 shows some of the details of construction, and it will be seen that the same general form of electrode assembly as in the *3B/241M* has been retained, in view of its low inherent output capacitance. Two such units have been mounted as far apart as allowed by the bulb dimensions to reduce the capacitance between the respective anodes. Anode-lead inductance has been reduced slightly by the use of a pressed metal part to form the connection to the top cap, in place of the thin tapes of the *3B/241M*.

Arrangements for interconnection of the two grids are of interest. A low-impedance high-frequency path between them is needed, but owing to the difficulty in manufacture of balancing accurately the characteristics of two valve assemblies of the necessary high slope, it was felt that separate biasing arrangements for the two sections would be desirable. A by-pass capacitor has therefore been incorporated in the

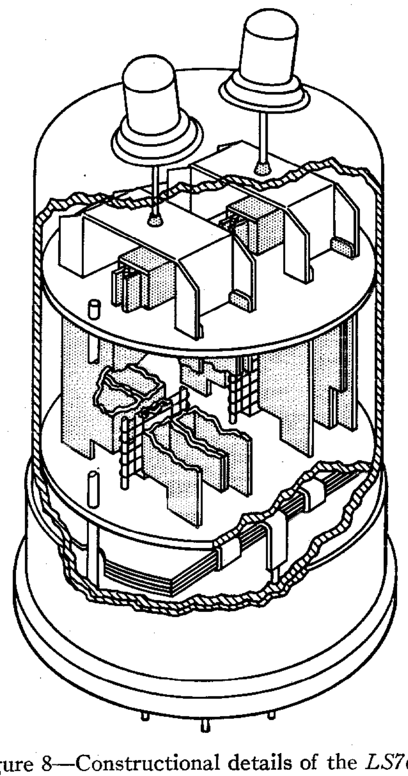


Figure 8—Constructional details of the *LS767*.

base screen. One grid is connected directly to this screen, and is provided with 2 base-pin connections. These pins provide the main high-frequency connection for both grids and the direct-current connection for grid *1* only, whilst grid *2* is capacitively coupled to this screen and is provided with a separate base pin for direct-current connection. Each triode section of this valve has a mutual conductance of 17 milliamperes per volt, an amplification factor of 90, and an anode dissipation of 12 watts. Typical operating conditions are given in Table 2.

TABLE 2

LS767, TYPICAL OPERATING CONDITIONS AT 420 MEGACYCLES

Anode Voltage	300 Volts
Anode Current (Per Section)	70 Milliamperes
Grid Current (Per Section)	15 Milliamperes
Grid Bias	−7.5 Volts
Output Power (Both Sections)	17 Watts
Efficiency	40 Per Cent
Driving Power (Both Sections)	3 to 4 Watts

Anode circuits for this valve can readily be constructed for frequencies up to 420 megacycles; for example, a parallel-conductor transmission line of strips 0.75 inch (19 millimetres) wide spaced 0.75 inch apart has a length of 1.75 inches (45 millimetres) external to the valve when resonant at 420 megacycles. Tuning these circuits over a wide range has presented some problems. In general, eddy-current tuning methods, in which the inductance of the coil is varied by means of a movable copper slug acting as a short-circuited turn, and circuits of the so-called butterfly type appear to be most suitable.

The resonant frequency of the cathode–grid capacitances together with the inductances of the cathode leads is lower than 420 megacycles so the drive circuit must be designed to present a capacitive impedance to the valve. A method that has been found to be satisfactory is to couple the cathodes to the anodes of the driving stage by means of small capacitors, whose reactances are large in comparison with the reactance of the lead inductance, which thus becomes of no consequence. This arrangement depends on the falling reactance of the grid–cathode capacitance with frequency to prevent the generation of harmonic voltages at the cathodes, as set out in the introduction.

A direct-current path for the cathodes is provided by small chokes; the complete circuit is as shown in Figure 9.

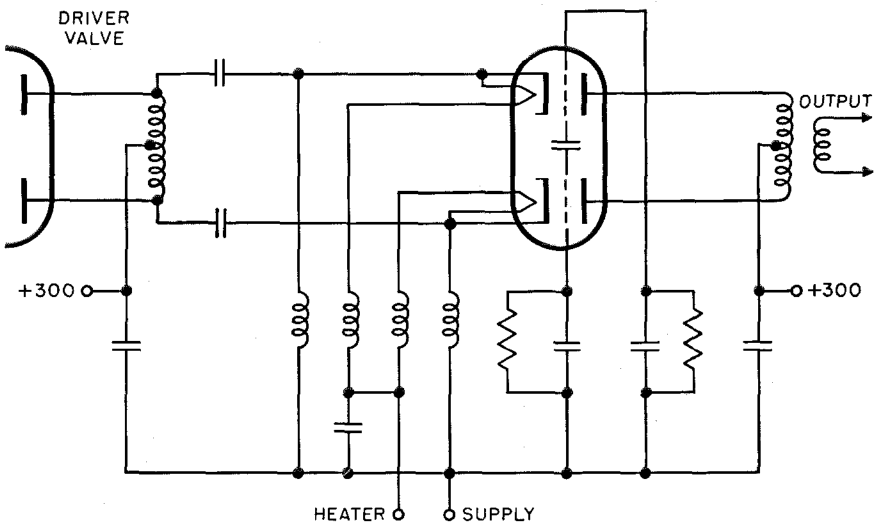


Figure 9—Circuit diagram of 420-megacycle amplifier using an *LS767* valve. A capacitor built into the valve permits the grids to be maintained at the same high-frequency potential whilst permitting separate direct-current biases to be applied for balancing purposes.

5. 5-Watt Tube for Frequencies Up to 420 Megacycles

Encouraging results with the *LS767* so far described made it worth while investigating the possible usefulness of the grounded-grid double triode in receiving applications. Two small double triodes already exist, namely, the type *6J6* and the *12AT7*. The *6J6*, however, has a common cathode and cannot be used in a balanced cathode-input circuit; the disposition of electrodes and pin connections of the *12AT7* is not very well suited to balanced grounded-grid use, although the author understands from an unpublished report that some success has been obtained with it as an amplifier at 400 megacycles. Clearly some improvement should be possible by the use of a more convenient arrangement of electrodes and connections, and some preliminary work has been carried out on an experimental valve, coded *LS774*, in which the electrode structure is as shown in Figure 10.

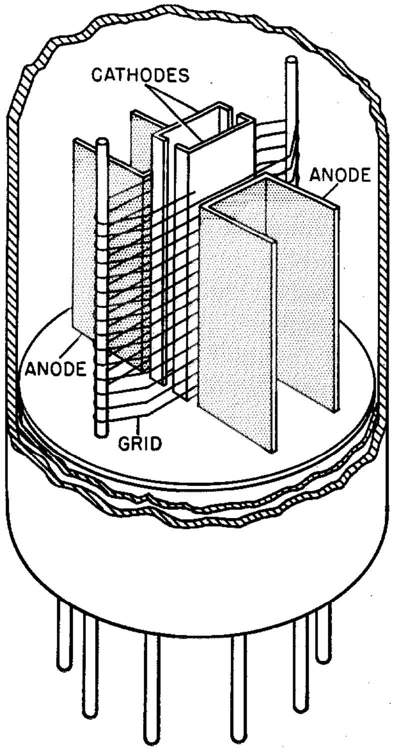


Figure 10—Constructional details of the *LS774* experimental valve.

A common grid is used to control the current from two cathodes, and thus the impedance between the two triode grids is effectively reduced to zero. To reduce heater–cathode capacitance and to conserve heater power, the cathodes themselves are single-sided and have a common heater, as shown in the figure. Static characteristics and interelectrode capacitances of this tube are given in Table 3.

The valve is single-ended and uses a miniature 9-pin glass base, so arranged that input and output connections are symmetrically disposed and separated by earthed connections for screening purposes.

Early experiments with this valve have shown that gains of more than 10 decibels per stage can be obtained at frequencies up to about 430 megacycles, and that at this frequency it is still possible to use a coil as the anode resonant circuit. At the time of writing no noise-factor measurements had been made, but it is expected that the noise-factor will be rather critically dependent on the design of the input circuit.

The single-ended arrangement makes it easy to use this valve as an oscillator, merely by connecting small feedback capacitors from the anodes to their respective cathodes, to which direct-current connection is made by chokes. Experiments on its use as a mixer are also in hand. Thus it appears possible that this valve may enable receivers to be designed for frequencies up to 420 megacycles or more with little greater complexity than for, say, 100 megacycles. Furthermore, its 2·5-watt dissipation per anode is adequate for it to form a driver stage for the *LS767* transmitting tube described in the previous section.

TABLE 3

LS774 Experimental Grounded-Grid Double Triode

Amplification Factor (Per Section)	100
Mutual Conductance (Per Section)	8 Milliamperes Per Volt
Anode Dissipation (Per Section)	2.5 Watts
Interelectrode Capacitances	
Either Cathode to Grid and Heater	6.0 Micromicrofarads
Cathode to Anode (Per Section)	0.1 Micromicrofarad
Either Anode to Grid and Heater	1.6 Micromicrofarads
Cathode to Cathode	1.0 Micromicrofarad
Anode to Anode	0.1 Micromicrofarad

6. Conclusion

It is hoped that these valves may assist in filling the gap that has developed in the radio spectrum at wavelengths of about a metre. The author acknowledges the valuable suggestions regarding the mechanical design of these tubes received from Mr. W. W. Marsh, who has been in charge of their manufacture.

Twinplex and Twinmode Radiotelegraph Systems

By CHRISTOPHER BUFF

Mackay Radio and Telegraph Company; New York, New York

DURING the past ten years, the use of frequency-shift transmission has been steadily increasing to the point where, today, we find a majority of the international radio circuits using this system for either Morse code or automatic printer operation, with a predominant trend toward the latter. The reliability afforded by frequency-shift transmission has permitted the use of several types of time- and frequency-division multiplex systems to handle increasing traffic loads. On circuits where the requirements are for two channels, the basic system herein described will provide a high grade of service combined with a maximum of flexibility and at a minimum of cost.

This system, retaining the advantage derived from frequency shift, has been developed to the stage where it is in daily operation on several long-haul radio circuits. The basic principle of the system, which is called Twinplex, may be extended to provide 4 or more channels, using single-sideband or subcarrier modulation methods. At present, its greatest utility is in converting existing single-channel frequency-shift radio circuits to two-channel, nonsynchronous, Morse, printer, or mixed-code operation. Alternatively, the Twinplex system may be used to transmit one channel of 3-element cable code. This latter technique has found a rather-important application in the direct connection of radio circuits to ocean cable systems.

The Twinplex system is based on the proposition that two 2-element mark–space channels may be combined to form on the frequency scale a single 4-element channel wherein each of 4 frequencies represents one of the 4 possible mark–space combinations shown in Table 1.

TABLE 1

MARK–SPACE COMBINATIONS FOR TWO CHANNELS

Channel *A*	Channel *B*	Frequency
Mark	Mark	*F1*
Mark	Space	*F2*
Space	Mark	*F3*
Space	Space	*F4*

At the present time, 400-cycle-per-second separation is used between each of the 4 frequencies, making a total frequency shift of 1200 cycles. Total bandwidth, including keying sidebands, is confined to 1700 cycles with 60-word-per-minute printers keying on both channels. The rate of transition between the 4 signaling conditions has been held to that prevailing for standard operation on a single-channel basis so that the only increase in occupied bandwidth is due to the increase in over-all prime frequency shift. Compared to the standard 850-cycle shift for a single channel, the additional bandwidth for 2-channel Twinplex will be only 350 cycles.

Extensive field testing and daily operation of many long-distance circuits have shown on the basis of error counts that, generally speaking, 2-channel Twinplex is equally effective on any radio circuit that normally supported a single-channel frequency-shift printer circuit.

In order to convert to Twinplex operation, one small combiner unit is added at the transmitting station and three pieces of conversion equipment at the receiving station. The conversion equipment has been designed to fit into the space vacated by single-channel equipment. This results in great economy of space and permits existing installations to be easily modified for 2-channel operation.

Finally, the great flexibility of operation cannot be overemphasized. Full freedom is allowed the traffic department in utilizing the available channels. For example, a 60-word-per-minute printer may be on one channel and a 75-word-per-minute printer on the second or a printer and Morse combination may be used. Branch-office extensions offer no problem and forked circuits may be efficiently handled. In international radio traffic operations, differences in printer speeds, codes, and operating procedures are likely to be with us for some time to come. The basic Twinplex system would seem to offer the most economical means of doubling traffic-handling capacity while retaining the inherent advantages of single-channel frequency-shift transmission.

1. Theory of Twinplex Signaling

The Twinplex concept of 2-channel signaling involves certain transitional phenomena that must be considered in the design of effective equipment using this principle of operation.

When two 2-element channels are combined, there are only 4 combinations possible and these may be represented as a 4-element channel on the frequency scale. Such a set of signaling conditions is shown in Table 2.

It is interesting to note at this point that to extend this type of grouping beyond two channels would involve transitions between so many frequencies as to make the design of the equipment highly impractical. The relationship which holds is

$$F_N = 2^N, \qquad (1)$$

where F_N is the number of frequencies and N is the number of channels.

Three channels would involve 8 frequencies, four channels 16, and so on. However, multiples of the 2-channel Twinplex grouping applied, for example, as a multiplex group on a single-sideband transmitter would be quite economical of frequency spectrum and offer a 3-decibel over-all power gain due to the fact that the available power is distributed among half the number of frequencies that would be required for conventional mark–space keying of each channel.

In the case of nonsynchronized keying on the present 2-channel system, the carrier frequency is shifted among 4 values at completely random intervals. For example, 10 percent of a mark pulse on channel A may take place on $F1$ and 90 percent on $F2$, or any other division of the mark may occur. The receiving circuit arrangements are such that whenever a transition occurs on one channel, the signaling condition on the other channel is undisturbed except during actual transition from mark to space on the first channel. This holds true for all the possible signal combinations that may occur. However, when keying takes place between two frequencies such as $F1$ and $F3$ and the output of a bandpass filter centered on $F2$ is observed on an oscilloscope, it will be seen that a "pip" of energy comes through during each transition between $F1$ and $F3$ (Figure 1). The amplitude of these pips is dependent, mainly, on the rate of transition between $F1$ and $F3$ and the attenuation characteristics of the $F2$ filter. During signaling, the transition rate and the bandpass-filter characteristics are fixed and so the amplitude of the pip is fixed in relation to the main signaling pulse. In the design of a practical system, the pip amplitude has been held to about $\frac{1}{4}$ of the main pulse amplitude. To bring it down further would

TABLE 2

TWINPLEX SIGNALING CONDITIONS

Input Keying Channels		Relative Combined Output in Volts	Transmitted Carrier Frequency in Cycles	Detected Receiver Frequency in Cycles	Output Tone-Keyed Channels	
A	*B*				*A*	*B*
Mark	Mark	0	$F1 = F_c + 600$	1950	Mark	Mark
Mark	Space	1	$F2 = F_c + 200$	2350	Mark	Space
Space	Mark	2	$F3 = F_c - 200$	2750	Space	Mark
Space	Space	3	$F4 = F_c - 600$	3150	Space	Space

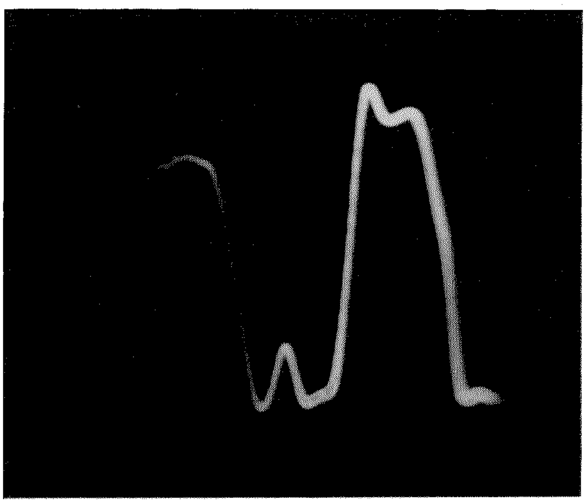


Figure 1—Transitional pip through $F2$ filter at bottom.

require a much faster transition rate and thereby increase the keying sidebands. In present Twinplex equipment, the mark-to-space transition rate that prevailed with single-channel frequency-shift keying has been maintained.

A second effect called "splitting," takes place when a mark–space transition occurs within a mark pulse causing part of the pulse to occur on one frequency and the remainder on a second frequency. If we look at the recombined mark at the output of a low-pass filter, a split will be observed at the point within the mark where the transition occurred. The amplitude of the split is also about $\frac{1}{4}$ of the main pulse amplitude and is governed by the same factors as the pip. Figure 2 shows a recombined mark pulse on channel-B low-pass filter output. The split at the top is caused by a mark-to-space transition on A.

The circuits following the receiving filters and signal rectifiers are designed to work in the area that lies between the split from the top and the pip from the bottom of the pulse.

To obtain a wide operating margin, the tone pulses are amplified to a rather high voltage level. The pips and splits will be amplified proportionately but, in terms of volts lying in between, we gain operating margin. The filtered tone pulses are then rectified and passed through suitable low-pass filters before being applied to the grid of a direct-current amplifier that is adjusted to key to mark several volts above the half-amplitude point of the rectified pulse and to key to space several volts below this point. By careful design and adjustment, the direct-current amplifiers may be made to key within less than one-half the margin available between the pip and split regions.

The mark–space signaling combinations of the system are such that both splits and pips may occur on channel B, but splits only can occur on channel A. This might at first seem to indicate that the failure point of channel B would occur at a considerably higher signal-to-noise ratio than for channel A. Measurements of the signal-to-noise ratio show that channels A and B fail at very nearly the same level, regardless of whether the opposite channel is keying or not. When the signal-to-random-noise ratio becomes unfavorable in the limiter, the energy distribution of the limiter output is spread more and more over the entire input bandpass region. Consequently, there is less output available on the desired signaling frequencies. This causes the familiar depression towards zero of the desired signal at the output of the mark-channeling filter and, if deep enough, subsequent failure of the signal. It has been observed that this depression of the signal due to noise tends to mask

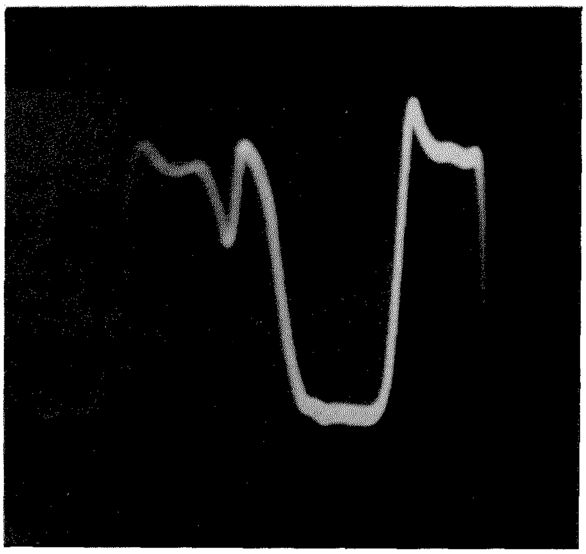


Figure 2—Recombined mark pulse of B channel showing split.

the inherent pips and splits somewhat before the actual failure point occurs. The breaking point is very sharp and seems to occur on single-channel keying within less than 2 decibels after keying failure with both channels has occurred. Measurements of the signal-to-noise ratio have been made using receiver-tube hiss and thermal agitation as a noise source. These measurements indicate that failure of the tone output of A and B channels occurs only after the noise level at the converter input is raised to 6 to 8 decibels above the signal level.

2. *Bandwidth and Keying Characteristics*

The total frequency shift of 1200 cycles is centered about the assigned carrier frequency so that mark–mark is 600 cycles plus, space–space 600 cycles minus and the two intermediates 200 cycles plus (mark–space) and 200 cycles minus (space–mark). Keying sidebands are determined by the transition rate from marking to spacing intervals, sometimes called the slope of the keying. In frequency-shift transmission, the transition rate may be closely controlled by choice of the proper time-constant in the grid circuit of the reactance-tube modulator, through which the keying pulses must pass. For standard single-channel operation, a resistance–capacitance filter providing a time-constant of about 200 microseconds has been used to handle keying speeds up to 240 dot-cycles per second without undue biasing of the pulses. For Twinplex operation, the same filtering has been retained so that the bandwidth will increase only by the amount of additional frequency shift, or 350 cycles, when compared to standard 850-cycle single-channel transmission. The over-all Twinplex system allows for a maximum keying speed of about 50 dot-cycles per second per channel with the limitation being placed on the receiving-end filters in order to circumvent the pips and splits that are inherent in this method of transmission. When two 60-word-per-minute 5-unit printers are used, operating margins of 65 to 75 points are normally obtained and regeneration is not required. This is equivalent to the performance obtained on standard single-channel circuits. It is expected that the present system will handle two 100-word-per-minute printers, when these become available.

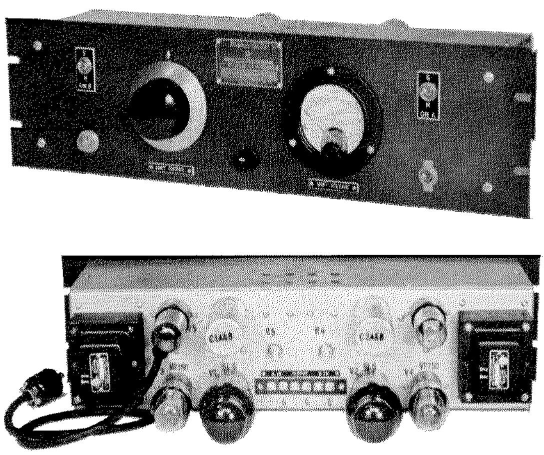


Figure 3—Front and rear views of the Twinplex combiner.

In ocean-cable-code transmission, a 3-element code consisting of dot, dash, and neutral elements is used, corresponding to positive current, negative current, and no current through the cable. On a frequency basis, three signal frequencies are required for its transmission. Before the development of the Twinplex system, it was necessary to resort to manual transcription or mechanical reperforation equipment for interconnecting ocean cable and radio networks. For cable-code transmission using Twinplex, only three frequencies are used namely, mark–mark for neutral, mark–space for dot, and space–mark for dash, making a total frequency shift of 800 cycles. Cable-code keying speed in dot-cycles per second is slow compared to Morse or 5-unit printer code. Maximum operating speed of the cables connecting with one of the Twinplex radio circuits, is about 1200 center holes per minute on the perforated tape. This is equivalent to 20 bauds or 10 dot-cycles per second. However, the information or message content of cable code, requiring about 23 center holes per word, is roughly twice that of international Morse so that traffic flowing at a speed of 10 dot-cycles per second is equivalent to about 53 words per minute in Morse.

In an intermixed network of cable and radio circuits, the use of Twinplex has greatly simplified the speedy handling of traffic by eliminating mechanical reperforation and attendant maintenance problems, while at the same time improving operating efficiency.

3. Transmitting Apparatus

One small unit, the Twinplex combiner, is required in addition to that already at hand for standard single-channel frequency-shift transmission. Front and rear views are shown in Figure 3. The combiner has two direct-current channel inputs and one closely regulated direct-current output, which is used to bias the reactance-tube grid in a frequency-shift exciter. The two inputs are keyed from standard tone rectifiers and ignore any voltage fluctuations therein above the threshold level required to key the combiner unit, thus isolating the frequency-shift-exciter reactance circuit from any variable conditions existing on the keying channels.

Essentially, the combiner consists of two small well-regulated power supplies in which fractions of the individual voltage sources are triggered off across individual load resistors by two keyer tubes. One source is adjusted to produce twice the output voltage of the other. These resistors are connected in series across a third resistor, which forms the output load impedance. When the latter resistor is very large compared to the individual load resistors, the voltages will add up

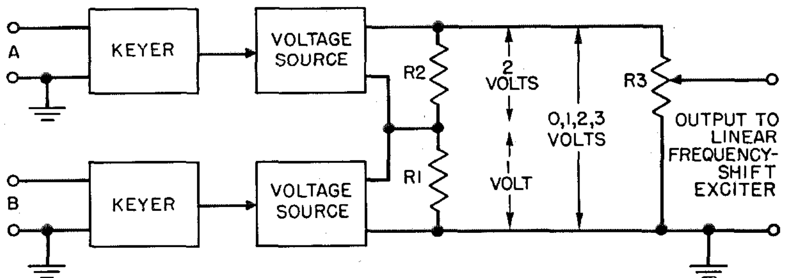


Figure 4—Principle of combining two keying channels. $R1 = R2$ and $R3 = 200$ $R1$. The output voltages and transmitted frequencies are given in Table 2.

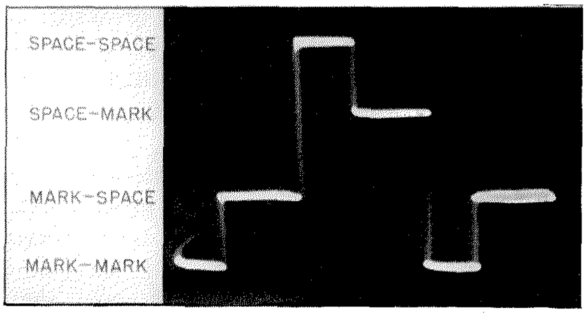


Figure 5—The 4-step keying pattern.

to a sufficiently accurate degree for all practical purposes. Figure 4 shows the principle of operation used in the combiner. When the voltage across $R1$ is, say, 1 volt and that across $R2$ is 2 volts, the voltage across $R3$ is very close to 3 volts. With a 200:1 ratio of $R3$ to $(R1+R2)$, it is actually 2.98 volts. Negligible current is drawn by the reactance-tube input circuit so the 1, 2, 3 relationship may be closely maintained.

The combiner operates as follows. With keyer tubes A and B both cut off, zero volts output (mark–mark); A cut off and B conducting, 1 volt output (mark–space); A conducting and B cut off, 2 volts output (space–mark); A and B both conducting, 3 volts output (space–space). Over-all frequency shift is controlled by the voltage divider $R3$, with intermediate shifts remaining proportional at any setting.

The frequency-shift exciter associated with the combiner must shift linearily to at least 1200 cycles in order that the transmitted frequencies will be exactly proportional to the combiner output voltages. Also, the stability must be high so each of the 4 frequencies may be held to within ±50 cycles at a carrier of 20 megacycles over operational periods of 8 hours or longer. This degree of frequency stability has been achieved by improvements in the 200-kilocycle oscillator circuit and in the characteristics of the temperature-control oven of existing frequency-shift exciters. Figure 5 shows the 4-step keying pattern at the combiner output produced when the two channels are keyed at random.

4. *Receiving Apparatus*

The receiving apparatus consists of the Twinplex converter, tuning monitor, and dual tone keyer, which together occupy a total panel space of 17½ inches, making it possible to mount these units in the same rack space required by certain types of single-channel dual-diversity frequency-shift conversion equipments (Figure 6). The

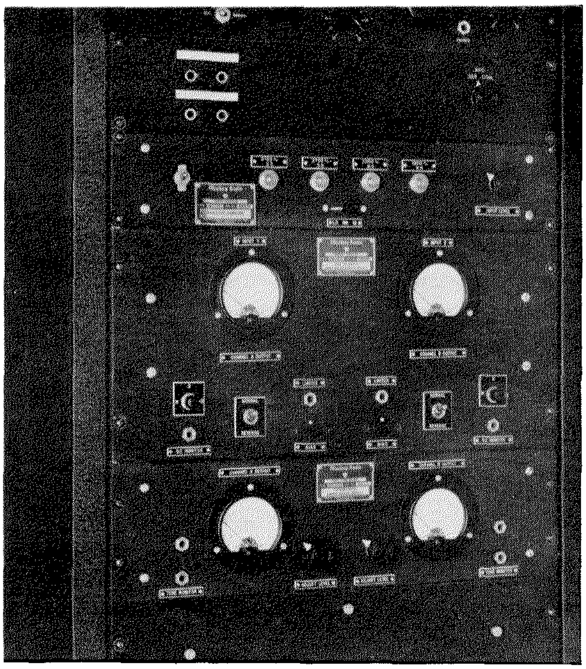


Figure 6—Twinplex receiving terminal apparatus.

Twinplex converter is a dual-diversity unit that accepts the 4 audio frequencies from the two radio receivers and, after amplitude limiting, separates them into the A and B channels. The frequency separation is accomplished by 3 bandpass filters having bandwidths of ± 100 cycles about center frequencies of 1950 (mark–mark),

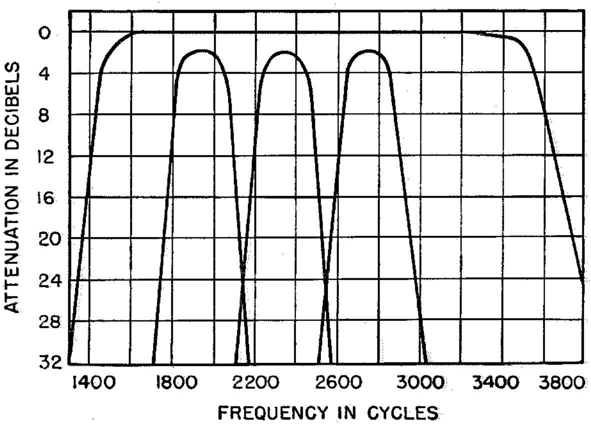


Figure 7—Characteristics of the receiving filters and of the input bandpass filter.

2350 (mark–space), and 2750 (space–mark) cycles as may be seen in Figure 7. The fourth frequency, 3150 cycles (space–space) is not passed through any filter. This frequency holds down the limiter output to the exclusion of noise, and since it does not pass on to the diode signal rectifiers, both channels A and B go to the space condition. The rectified output of the 1950-cycle (mark–mark) filter keys both channels A and B to mark, 2350 cycles (mark–space) keys A to mark and B to space, and 2750 cycles (space–mark) keys A to space and B to mark. For dual-diversity operation, two identical transient-free limiter channels, saturating at an input of -40 decibels, are used to drive the two sets of three filters. These filters are of a high-impedance type each working from the plate circuit of a triode and terminating in 50,000 ohms. The voltage gain obtained in the filter simplifies the equipment considerably by eliminating additional amplifier stages. By rectifying at a relatively high signal voltage level in the diversity-connected diodes, a wider operating range for the following direct-current amplifiers is obtained.

Differential rectification is not used with this system but due to careful consideration of the basic factors affecting the failure point, the system is on a par with several types of single-channel equipment that use differential rectification. The output of the signal-rectifying diodes is passed through a separate low-pass filter, the characteristics of which are shown in Figure 8, for the A and B channels and then to the first grid of a 2-stage direct-current amplifier, whose operating threshold may be adjusted so that keying from mark to space takes place in one-half the margin available between the pip and split regions of the pulse. The direct-current output of the converter keys a dual tone keyer for transmission of the signals to the central office over regional facilities. Switches are provided on the converter for reversing the mark–space condition on either channel and for placing static test signals on the line. Jacks are provided for monitoring the limiter, bandpass filter, low-pass filter, and tone channel outputs so that step-by-step maintenance is simplified.

A tuning monitor on a $3\frac{1}{2}$-inch panel is provided as an aid in adjusting the common-oscillator injection frequency to produce the 4 proper audio frequencies for driving the converter. This unit consists essentially of an input amplifier and 4 high-Q tuned circuits, each operating in the plate circuit of a triode tube. Across each circuit, a neon indicating lamp is connected, which glows

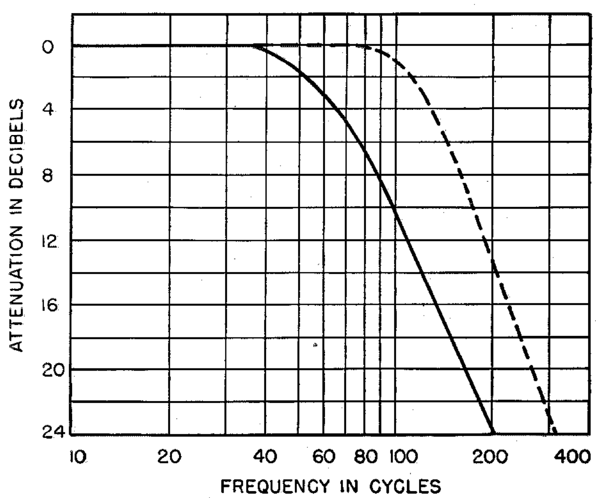


Figure 8—Either of the two characteristics shown may be selected for the low-pass filter through which the rectified signal currents flow.

only when the correct frequency, ±50 cycles, is present. Thus, when both channels are keying, all 4 lights will glow on and off, intermittently. When channel *A* keys and *B* is on mark, the mark–mark (1950-cycle) and space–mark (2750-cycle) lights, only, will glow and similarly the lights will follow all other keying combinations. In addition, the mark–mark (1950-cycle) circuit is provided with a high–low switch that cuts in capacitors for tuning the circuit to 75 cycles higher and 75 cycles lower than normal. By manipulation of this switch while tuning the common-oscillator frequency control, the audio frequencies may be adjusted within ±10 cycles of their proper values. Frequency drift of the transmitter or receiver oscillator may be detected and corrected before keying failures occur in the output of the converter.

The dual tone keyer occupies $5\frac{1}{4}$ inches of panel space. It uses two resistance–capacitance oscillators of the phase-shift type, two phase inverters, and two push-pull amplifiers that are keyed in the cathode circuit by the converter direct-current output channels. Tone output of each channel is adjustable to 0 decibels, which is 6 milliwatts into 600 ohms impedance.

These three units comprise the Twinplex receiver conversion equipment. They may be mounted in a dual-diversity receiving bay together with two radio receivers, a common oscillator unit, and a terminal power supply as may be seen in Figures 9 and 10.

5. *Four-Channel Operation —Twinmode*

An extension of the Twinplex system has been carried out to provide 4-channel operation on a radio transmitter equipped with a high-level amplitude modulator. This system, called Twinmode, utilizes two groups of Twinplex keying, one group on the carrier-shift side to provide channels *A* and *B* and the second group on a frequency-shifted subcarrier applied through an amplitude modulator to provide channels *C* and *D*. Standard Twinplex equipment already described is used with one additional unit, an audio-shift keyer, being required to translate

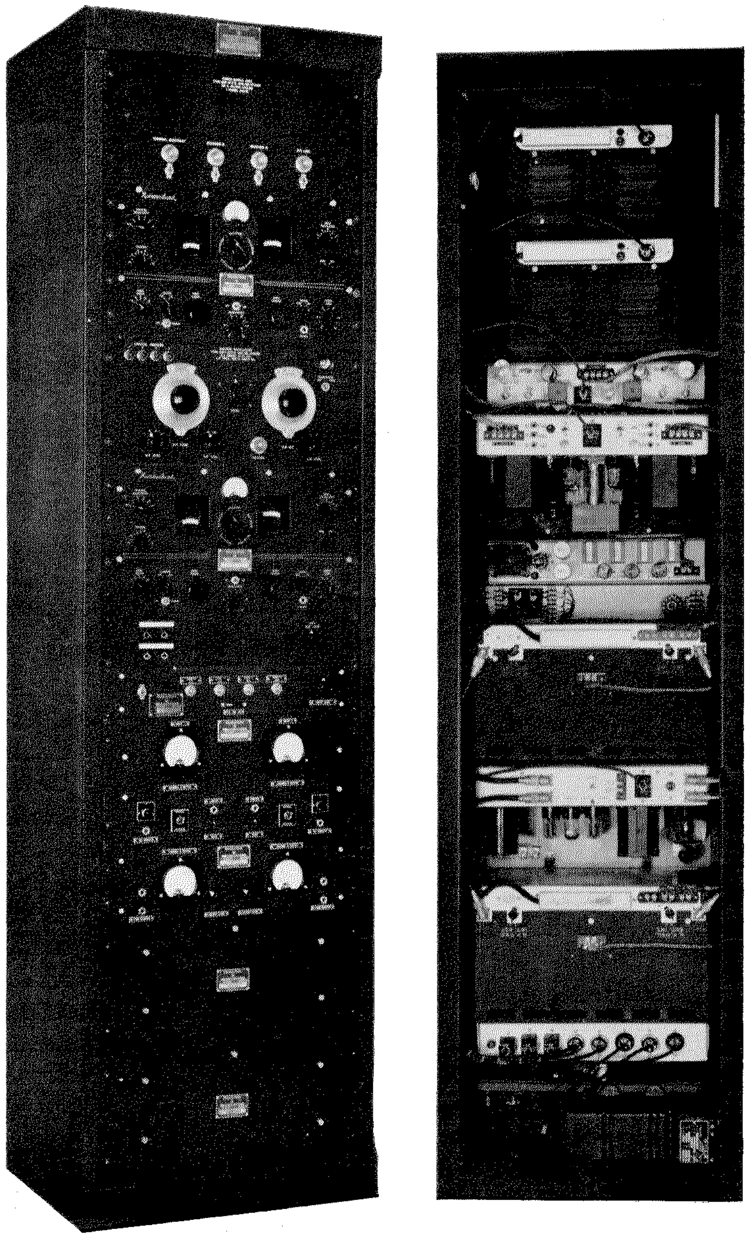


Figure 9—Front and rear views of Twinplex receiving bay.

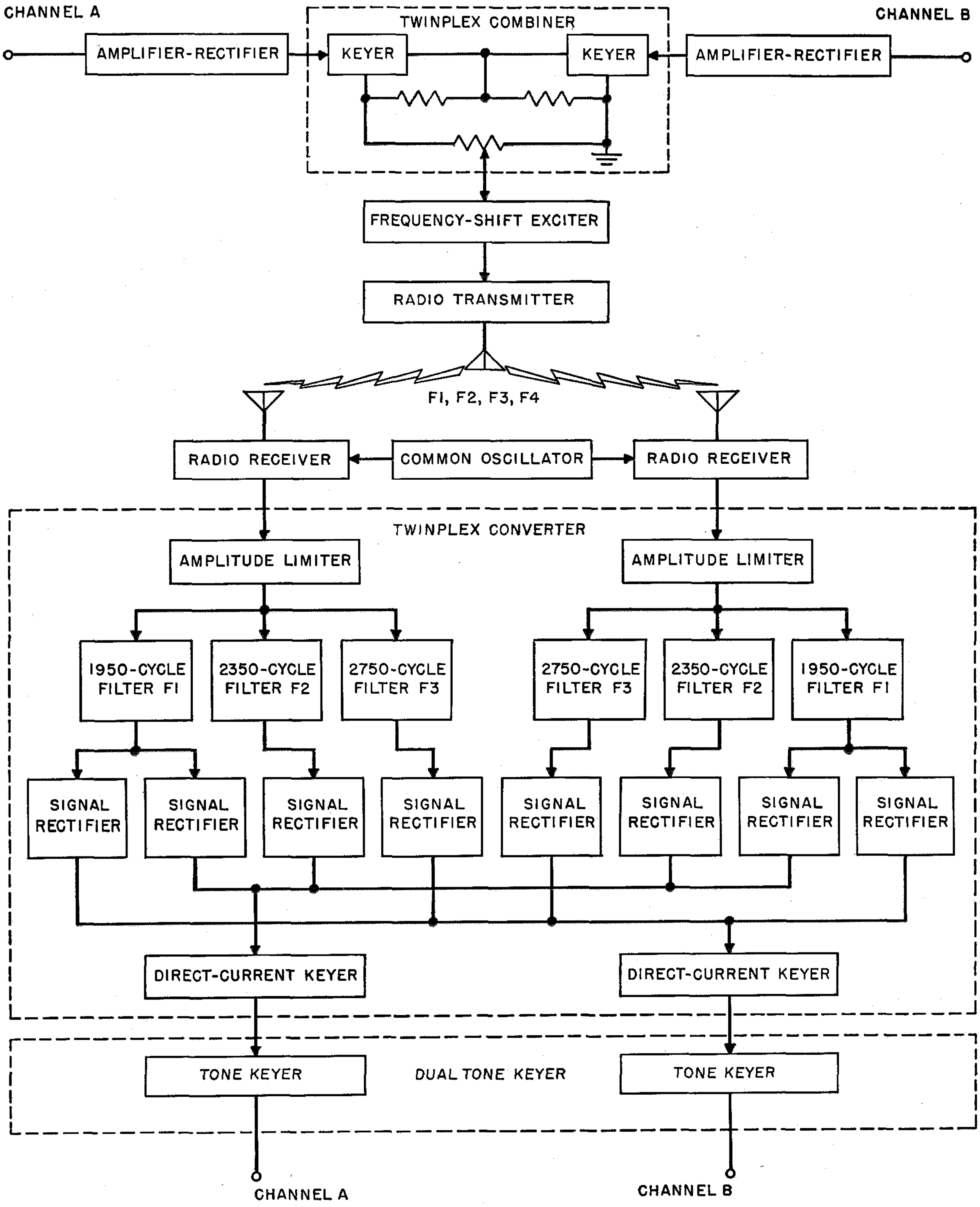


Figure 10—Block diagram of a dual-diversity Twinplex system. The keying voltages and frequencies are given in Table 2.

the combiner direct-current output pulses into any one of 4 audio frequencies for keying on the subcarrier side. The 4 audio frequencies used to modulate the carrier are the same as those normally derived from a Twinplex receiver by beat-frequency detection for carrier-shift reception, namely, 1950, 2350, 2750, and 3150 cycles. This permits the conversion equipment for the subcarrier frequency-modulated channels to be identical with that already described for receiving two channels on the carrier-shift side. The only difference in reception technique is that the subcarrier frequency-modulated channels, C and D, are received with the beating intermediate-frequency oscillator turned off. Figure 11 shows a Twinmode radio circuit.

The audio-shift keyer designed for use with this system consists of two stable oscillators operating near 150 kilocycles, which are adjusted so that the frequency difference between them produces the required audio frequency at the output of a mixer stage. The frequency of one of the oscillators is shifted by means of a reactance-tube modulator, whose grid is biased by the direct-current keying pulses from a Twinplex combiner unit. The mixer stage is followed by a cathode-follower output tube and then a 1500-to-3500-cycle bandpass filter to remove any

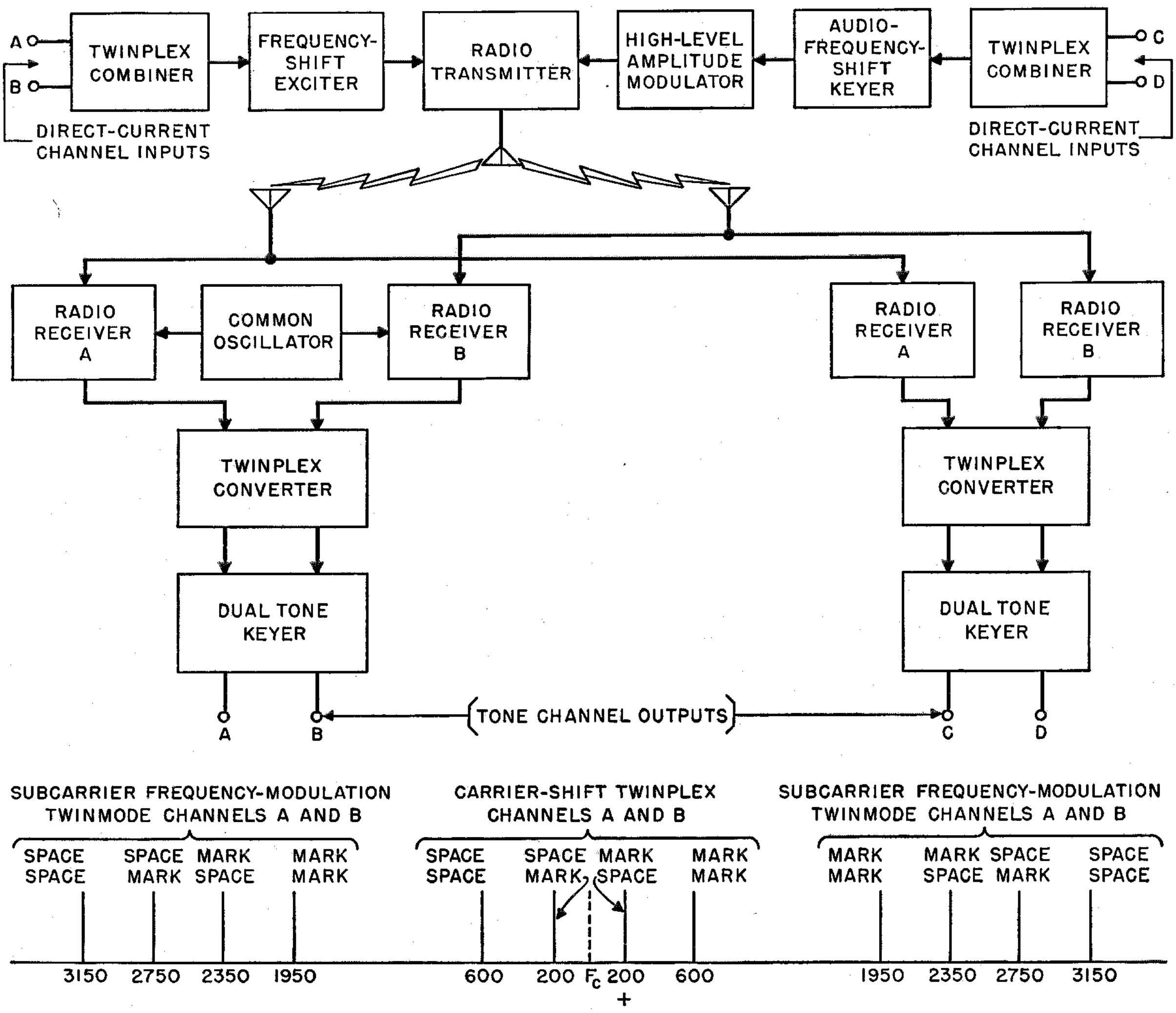


Figure 11—Block diagram of Twinmode system. The signaling frequencies in cycles relative to the transmitted center frequency F_c are indicated for a radio channel 7500 cycles wide.

existing distortion in the signal. For a linear frequency shift of 1200 cycles, from 1950 to 3150 cycles, the attendant amplitude change in output level is less than 5 percent. This unit delivers about reference power (6 milliwatts into 600 ohms) and may be connected directly to the speech-amplifier input of the modulated radio transmitter. Panel space required is $5\frac{1}{4}$ inches. Figure 12 shows the audio-shift keyer.

The transmitter is adjusted to 80-percent amplitude modulation for Twinmode operation. Full modulator power is always concentrated in one pair of prime side frequencies plus some keying sidebands. This gives about a 3-decibel improvement over the use of a pair of mark and space tones for each channel on the frequency-modulated subcarrier. Total bandwidth for four channels keying, two on carrier shift, and two on subcarrier frequency modulation is about 8 kilocycles with double-sideband modulation. With the audio modulating frequencies chosen, there is no direct interference caused by the *C* and *D* channels on subcarrier frequency modulation to the *A* and *B* channels on carrier shift, or vice versa. A compilation of the possible carrier beat frequencies shows that the nearest interfering tones at the receiver output will be 1200 cycles, on the low side, and 3900 cycles, on the high side, of the converter input bandpass filter. At 1200 cycles, the level is 50 decibels down and at 3900 cycles it is 40 decibels down from the passband level.

Under multipath propagation conditions, however, the differences in successive mark and space carrier signal amplitudes may be very great. These variations in carrier mark and space levels may cause clicks to be produced in the output of the linear diode detector used for receiving the subcarrier frequency-modulation channels. The clicks have a high harmonic content and can, under certain propagation conditions, cause mutilation of the *C* and *D* channels. It has been found that high-pass filtering immediately after the diode detector is beneficial in reducing these harmful effects. A 1400-cycle high-pass filter having a steep attenuation characteristic has been used with good results for suppressing the click fundamental and all harmonics up to the desired pass-band region starting at 1500 cycles. Channels *C* and *D* suffer a power disadvantage of around 4 decibels compared to the carrier-shift channels, *A* and *B*. Also, the receiver intermediate-frequency bandwidth acceptance must be 8 kilocycles as compared to 3 kilocycles required for the carrier-shift side,

Figure 12—Audio-shift keyer.

thereby further decreasing the signal-to-noise ratio of the received subcarrier frequency-modulation signal. Repeated observations show that due to the above deficiencies the subcarrier frequency-modulation channels will fail somewhat earlier, on a normal propagation day compared to the carrier-shift channels. During transitional periods, two-channel operation may, therefore, be employed. The two carrier-shift channels are not degraded by the presence of the subcarrier frequency modulation at any time.

The Twinmode system has been tested over a 3500-mile path with satisfactory results on 8-, 13-, and 19-megacycle frequencies with transmitter carrier power inputs of 5 to 12 kilowatts. The flexibility of channel usage and simplicity of operation, as well as the standardization of equipment used, make this system attractive for use in mixed-code communications where more than the two channels provided by Twinplex carrier-shift operation are required.

More recently the New York-to-Lima, Peru, radio circuit shown in Figure 13 was changed over from Twinplex to Twinmode operation. The Twinplex circuit formerly carried a 3-element cable-code channel, which was used at Lima to key the cable chain running down the west coast of South America to Valparaiso, across the Andes to Buenos Aires, and up the east coast to Rio de Janeiro. The Twinmode circuit now in operation provides this same cable-code channel on the subcarrier transmission, occupying channels *C* and *D*. At the same time, two 60-word-per-minute printer circuits are provided on channels A and B. The A printer channel is also taken off at Lima and forwarded to a customer's office on a leased basis. The New York Twinmode transmissions will also be received at Bogota, Colombia, where the B-channel signals will be selected to operate a printer, thereby releasing the two assigned carrier frequencies, two antennas, and the transmitter formerly used to provide this service on a straight single-channel frequency-shift basis. To accomplish this, it is only necessary to provide Bogota with a Twinplex converter unit to replace a single-channel frequency-shift converter.

The commercial operation of a circuit of the type just described clearly demonstrates the advantages of versatility and flexibility that accrue through the use of a system employing frequency-division principles.

6. Future Developments

Frequency-shift transmission is basically single-sideband technique as applied to telegraphy wherein the intermediate-frequency beat oscillator in the receiver acts as the reinserted carrier and the information is conveyed by the beat tones between this local carrier and the mark and space sidebands that are transmitted. When multichannel telegraphy on a single-sideband basis is practiced, it is customary to use the same type of transmitter and receiver as is required for telephony—with the exception that pairs of mark–space tones are substituted for voice or program modulation and the necessary filters for their separation are applied at the receiving end. It is suggested that a single-sideband system designed specifically for telegraph service would result in equipment of considerably less circuit complexity and cost while at the same time providing increased effectiveness and flexibility in circuit operations. This type of system may be built up from existing frequency-shift and Twinplex techniques. A proposed single-sideband telegraph system having this general form and using certain equipment already described as building blocks will be briefly described.

The single-sideband exciter at the transmitting end would consist of four stable low-distortion frequency-shifted oscillators symmetrically spaced around 200 kilocycles. A reactance modulator is associated with each oscillator and is keyed from a standard Twinplex combiner unit, described in Section 3, thus providing four 2-channel groups for a total of 8 channels. The oscillator outputs are combined and applied to a fixed balanced modulator from which the high-frequency sidebands produced by a 1.8-megacycle crystal and the 200-kilocycle oscillator

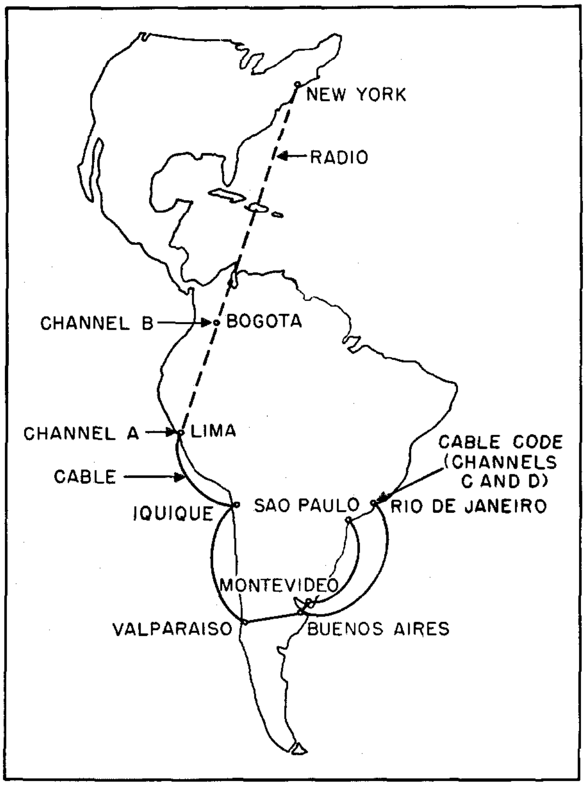


Figure 13—A Twinmode radio circuit from New York to Lima also keys on a subcarrier channel the cable system that extends to Rio de Janeiro. Bogota is served as well by one of the Twinmode channels.

frequencies are taken off. The four groups of shifted 2-megacycle frequencies are applied to a second balanced modulator and are mixed with 6-to-26-megacycle frequencies from a crystal oscillator and multiplier chain to produce low-frequency sidebands in the 4-to-24-megacycle range. This is followed by a straight linear amplifier to provide a single-sideband output of several watts suitable, after passing through a coaxial line, for driving the next stage in the transmitter. Alternatively, it may be desirable to make the exciter part of the transmitter proper.

Certain types of transmitters now employed

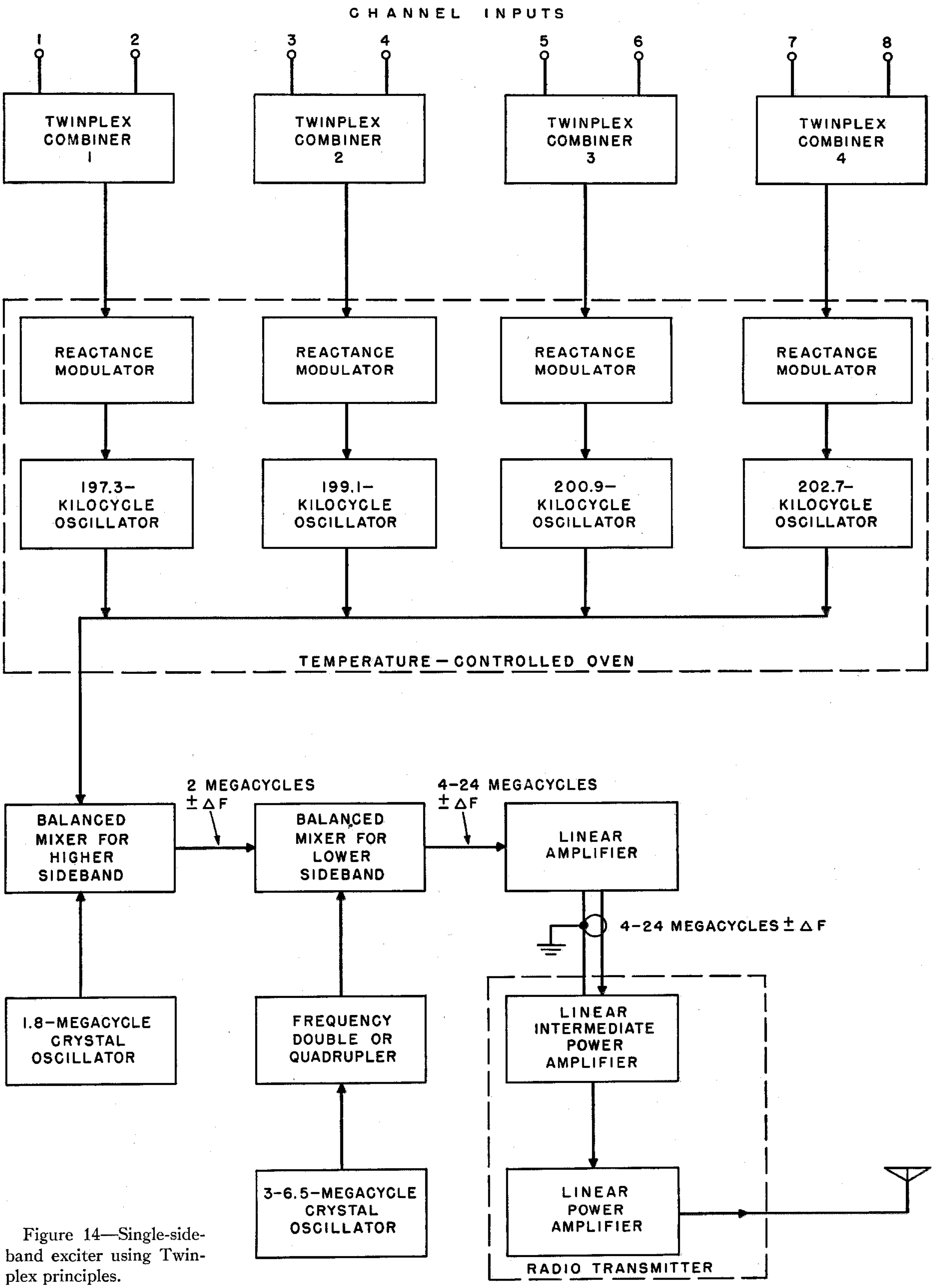


Figure 14—Single-side-band exciter using Twinplex principles.

for telegraphy using class-C amplifiers may be suitably modified for linear operation of the intermediate and power amplifiers so they may be driven with several watts from the single-sideband exciter output. There is, of course, a specific problem with each different transmitter.

are required as compared to eight for conventional multitone practice. This results in an over-all power gain of 3 decibels since the available power output is distributed among half the number of frequencies at any time. Also, the intermodulation possibilities are reduced.

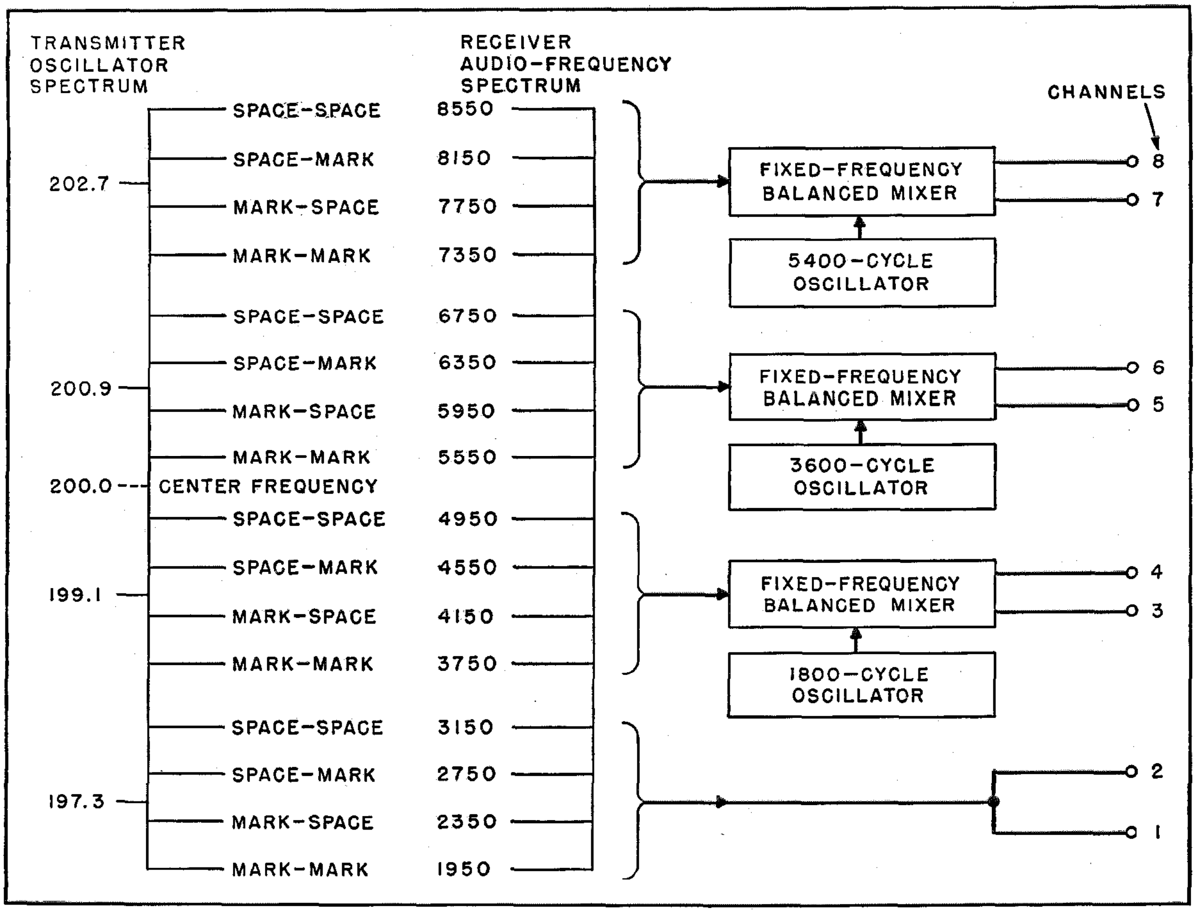


Figure 15—Single-sideband signaling frequencies. There are four groups of four frequencies each. The spacing between groups is 600 cycles and between adjacent frequencies in a group, 400 cycles, making a total spectrum of 6600 cycles. The center frequency of the oscillator is 200 kilocycles, which is multiplied to the assigned transmission frequency from which the receiver audio-frequency spectrum is recovered. A 470.25-kilocycle oscillator beats against the 465-kilocycle intermediate frequency to produce the above spectrum.

One type of transmitter presently operated in class-C uses two *4-250A* tubes in parallel in the intermediate power amplifier and an *880* in the power amplifier. The problem of operating these two stages as linear class-B amplifiers does not appear difficult. A functional block diagram of the proposed single-sideband telegraph exciter is shown in Figure 14. By utilizing the Twinplex combination scheme, only four shifted oscillators

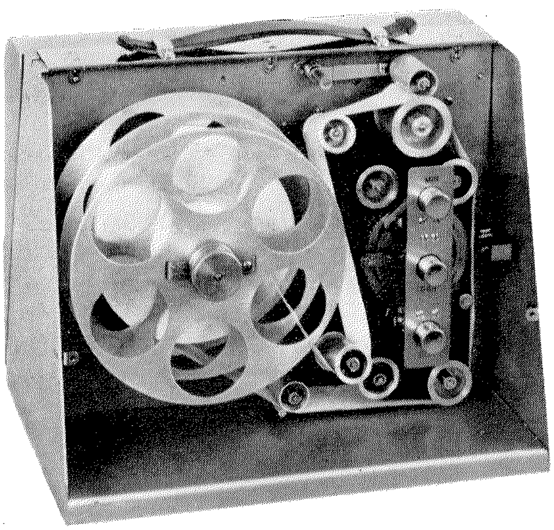
A standard dual-diversity Twinplex receiving bay of the type already described and shown in Figure 9 would form the basis of reception. For 8 channels, three additional Twinplex conversion equipments will be required. These are the standard units illustrated in Figure 6 and now used for 2-channel operation. In order to utilize the same filter frequencies in all converters, thereby effecting a valuable standardization in manufacture, the receiver audio frequencies produced by the single-sideband transmission, extending to 8550 cycles, will be changed to the basic four filter frequencies in each case, namely 1950, 2350, 2750, and 3150 cycles. This may be done quite simply by means of a fixed oscillator and mixer stage in front of each standard converter input. Figure 15 shows the actual frequency situation that would exist at the transmitter and at the output of the receiver.

The system described herein is highly compatible with certain existing single- and two-channel frequency-shift equipments thereby minimizing obsolescence when channel requirements increase on a given radio circuit. True single-sideband telegraphy results, with a 3-decibel power gain over conventional mark–space tone keying of each channel. A new multichannel radiotelegraph system utilizing these principles is now being developed for commercial use.

Recent Telecommunication Development

Printing Register

SUBSCRIBER-LINE activities, malicious calls, and toll calls in automatic telephone systems may be recorded on paper tape with a printing register designed by Bell Telephone Manufacturing Company.



The equipment is mounted in a self-contained unit of either portable or fixed form and can be connected to the exchange apparatus by jacks or patch cords.

Five printing wheels are actuated by electromagnetic clutches and for subscriber-line observations may record the date, hour, and minute of start and termination of each call, the number of the called line, and a metering indication. As many as 99 different tariffs or other metering data may be indicated.

Modifications may be made to adapt the register for observation on toll junctions in systems using ticket printers. Three of the printing wheels record time in hours, minutes, and seconds. The fourth wheel records toll prefix, selection pulses, called number, calling number, tariff, and release of connection. The fifth wheel provides for start, translation of selection, end of selection, answer of called subscriber, and end of connection. The last two wheels indicate 20 and 10 numbers, respectively. This record is very complete and covers all phases of the establishment and release of a connection.

A master clock provides the required signals for controlling the time wheels.

Design Considerations for a Radiotelegraph Receiving System*

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THE PAPER discusses some of the difficulties encountered in the reception of frequency-shift and amplitude-modulated telegraph signals over radio links. Comparison between both methods of signalling leads to the conclusion that greater transmission efficiency can be obtained by the use of frequency-shift signalling. Fading and interference affect both systems and give rise to distortion characteristics that depend not only on the type of modulation used but to some extent on the particular shape of the modulating waveform. The three main components of distortion are defined, and bandwidth requirements before and after demodulation are discussed. The use of direct printing circuits introduces difficulties during long marking periods and some methods of overcoming this problem are given. The need for exceptional frequency stability in relation to the reduction of errors for both methods of signalling is stressed, and methods of minimizing the effects due to frequency drift by the use of automatic frequency control and other means are discussed. The means of obtaining the greatest benefit from a diversity system are considered, and mention is made of an effective method of diversity working using amplitude-modulation signalling. Some observations are made on the effects of noise and on the main features of a satisfactory system for the reception of high-speed signals using either method of signalling.

• • •

1. Comparison Between Frequency-Shift and Amplitude-Modulated Systems

The first experiments with frequency-shift signalling over a radio link were made in about 1928, but no notable advance in technique appeared before 1939. Single-channel and multichannel frequency-shift signalling has been used increasingly since 1939 in preference to the on–off-keying or amplitude-modulated system. The advantage of frequency-shift signalling has been shown by comparisons over radio links and by laboratory tests that simulated frequency-selective fading. Typical results obtained in the field[1, 2] have shown the transmitted power needed to give equal average rates of error to be 10 decibels less, and the lost capacity to be 39 per cent less, for frequency-shift signalling than for the other system. Laboratory tests have confirmed these results.[3, 4] For example, telegraph distortion of 40 per cent, corresponding to the selector-failure point of a teleprinter, has been obtained with frequency-shift signals $2\cdot5-4\cdot5$ decibels lower than corresponding amplitude-modulation signals: for 15-per-cent distortion the difference was 4–6 decibels, and for still lower distortion the difference was about 10 decibels. Calculations based on comparison of the signal-to-noise ratios of the systems agree closely with observed results.[5]

It is difficult to obtain reliable printing with an amplitude-modulation system during periods of rapid fading. If, for example, the printer were set to operate at half the normal amplitude of the demodulated wave, there would be complete failure with a drop in level of 6 decibels. The operating point is usually set at a much lower level, near the spacing noise level, so that greater changes in signal level can be tolerated. This expedient, used with automatic gain control, is usually effective provided the signal level does not change too quickly. It is advantageous to have an automatic-gain-control system with adjustable time-constant. A long time-constant prevents the receiver noise from rising during short intervals between signals, and improves

* Reprinted from *Journal of the Institution of Electrical Engineers*, Part 3, v. 98, pp. 253–262; July, 1951. The unnumbered photographs and their legends did not appear in the original printing.

[1] Numbered references appear in Section 10 on page 48.

the signal-to-noise ratio if the input signal is sufficient to give full automatic-gain-control action. On the other hand, a short time-constant is needed to follow rapid fades.

A narrow-band amplitude-modulation system may fail in conditions of rapid fading, whatever the transmitter power, because of the generation of transients in the selective circuits. A fairly wide pass-band, maintaining a substantially square waveform at the demodulator, is needed to avoid this trouble.

A frequency-shift system is almost unaffected by fading if amplitude limiters are used.[6] Limiter circuits can be made almost completely free from transient response during rapid fading, and with signal-to-noise ratios above 6 decibels at the input a favourable amplitude discrimination occurs, giving output signal-to-noise ratios greater than 10 decibels. This means that faultless printing is obtainable if the signal-to-noise ratio is not less than 6 decibels at the input terminals.[4]

Another advantage of the frequency-shift system is that it is essentially a double-current system, whereas it is difficult to obtain the advantages of polar or double-current working with an amplitude-modulation system. Amplitude-modulation systems using tone rectifiers to give polar currents corresponding to on and off periods are not strictly double-current systems. In the frequency-shift system, the marking and spacing signals both control the output, but only noise is received in the spacing periods of an amplitude-modulation system.

The bandwidth required for frequency-shift signalling need be no greater than for amplitude-modulation signalling, and for equal bandwidths the frequency-shift system has about a 6-decibel advantage in signal-to-noise ratio.

2. *Operation Over a Radio Link*

2.1 Fading and Interference

Selective fading is the main cause of unstable operation of a long-distance radiotelegraph system.[7] The distorting effect of selective fading depends on the type of modulation used and to some extent on the modulation waveform.

A direct-printing system fails if the signal falls below noise level for about half the duration of an elemental dot, but even if the signal never falls below noise level the system may fail through the generation of transients in the receiver caused by very rapid changes in signal amplitude: 10 fades a second of more than 40-decibel depth are not uncommon. The receiver must therefore be designed to deal with very large changes in signal level at various rates. A receiver so designed will give printing errors that fall progressively in number as the transmitter power is raised to the level above which fades below noise level do not occur.

If selective fading occurs through the relatively simple condition of transmission by two paths, and if the lengths of the two paths differ by a distance of the same order as the wavelength, fading is uniform over the whole spectral width of the signal, and simple variation of signal amplitude results. If, on the other hand, the difference in path length corresponds to a time lag of the order of a period of modulation, the modulation suffers severe distortion.

Figure 1 shows how the changes in amplitude and phase of the resultant of a single-frequency signal transmitted by two paths vary with the phase and amplitude differences between the components at the receiver. It will be seen that the maximum variation occurs when the components are of equal amplitude. Thus maximum distortion occurs when the path difference is great and the components are of equal amplitude.

In an amplitude-modulation system, with a time lag between component signals of the order of a dot length, the successive arrival of a dot by the two paths causes an elongation of the dot when the components are in phase, and double marking when they are out of phase. The telegraph distortion due to elongation is then given by

$$D = \Delta T W / (12 \cdot 5), \text{ per cent} \qquad (1)$$

where W is the speed of signalling in words per minute, and ΔT is the delay between the signal components in milliseconds.

When in a frequency-shift system there is two-path transmission, there are periods when the marking signal arriving by one path overlaps at the receiver the spacing signal arriving by the other. Maximum distortion occurs, as for amplitude-modulation systems, when the amplitudes of the two components are equal.[8] Moreover, the transients produced in these circumstances may cause distortion of much longer duration than the period of overlap of the spacing and marking signals. This is one of the factors that determines the choice of bandwidth of the receiver circuits and the degree of frequency shift used.

The pass-band of the receiver may be wide enough to pass dots with substantially square waveforms. It may be reduced, however, until the bandwidth in cycles per second approaches the signalling speed in bauds (dot-periods per second), when a dot is passed with substantially sinusoidal waveform. This may be done to reduce noise, or it may be done at the transmitter to conserve space in the frequency spectrum. When the bandwidth is substantially narrower than the frequency separation of simultaneous selective fades, the fading has little effect on reception until a fade crosses the pass-band. Diversity reception must then be relied on to prevent breakdown. As the pass-band is reduced, so the frequency shift must be reduced also, which increases the risk of errors due to frequency drift of the apparatus at either terminal or to small changes in frequency during transmission resulting from Doppler effect caused by rapid changes in path length. In a system with a pass-band of 100 cycles or less, the systematic distortion can be kept below 4 per cent at 80 bauds, dots being passed as sine waves with about half the steady-signal amplitude.

2.2 Distortion

The main reason for using direct-printing systems is to obtain speed and accuracy. For that reason, the reduction of distortion is of major importance. The mutilation of a letter or even a word may be tolerated if the text is in plain language so that the fault is easily recognized, but mutilation of text in code is intolerable. There are two methods of safeguarding accuracy: first by the use of error-detecting codes; and secondly by the use of signal-shape regeneration. Both methods may be used together. Error-detecting codes are not infallible, but failures are rare. (The five-unit code, which is not error-detecting, appears to be generally used because the specialized equipment is easily added to existing apparatus.[2]) There remains, however, the risk of undetectable errors and the printing of random characters caused by noise or interference.

Regeneration at a radio receiver enables the waveshape of the signals to be corrected before being applied to a further transmission link such as a land line. Without regeneration, the dis-

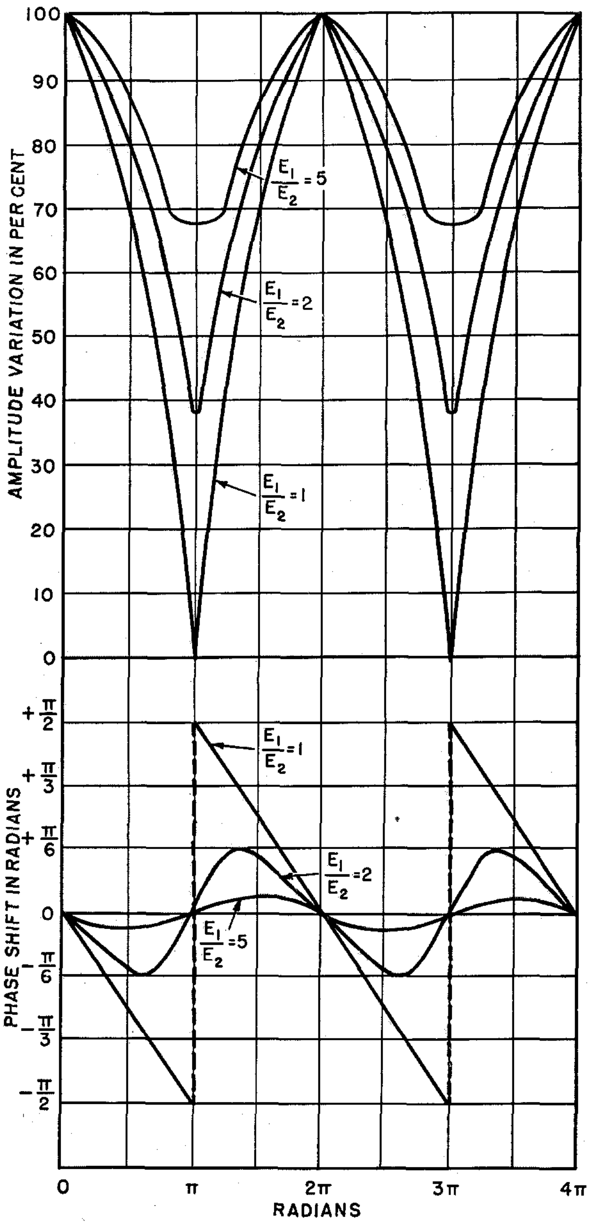


Figure 1—Variations in amplitude and phase during selective fading, as a function of $2\pi F\Delta T$, where F=frequency in cycles, ΔT=difference in arrival time in seconds, and E_1, E_2=amplitudes of delayed and undelayed signals. Amplitude variation is shown at top, and phase variation below.

tortion might, in such circumstances, exceed the limit permissible for satisfactory operation. Signal regeneration depends on sampling the polarity of the demodulated signal for a very small fraction of each unit period. The sampling is usually most effective if it occurs at the middle of each unit period. Even so, errors may occur if noise or interference coincides with the sampling.

Distortion[9] is usually regarded as having three components; bias, characteristic, and radio distortion, which are illustrated in Figure 2. Figure 2A shows a perfect signal of six unit-time elements, each of 20 milliseconds duration. The dotted portion of each element is of six milliseconds duration and is the signal-selecting or sampling period. Figure 2B shows bias distortion, which is a lengthening of the mark or space signals at the expense of the others. Figure 2C shows characteristic distortion, which is caused by the electrical characteristics of the transmission circuits. Bias and characteristic distortion are collectively called systematic distortion and are associated with the systematic features of the apparatus. Radio distortion, illustrated by Figure 2D, is characterized by random time-shifts of the edges of the signal elements and by breaks and indentations of the signal shapes caused by irregularity of the radio transmission and other fortuitous fluctuations in the communication system.

3. *Bandwidth*

The radio-frequency bandwidth required to transmit substantially square-wave modulation is about the same for amplitude-modulation and frequency-shift systems if the modulation index of the frequency-shift system (the frequency shift divided by the baud speed*) is less than unity. This bandwidth may be written

$$B \approx 1 \cdot 5 \text{ (speed in bauds)} + C = \frac{1500}{D} + C$$
$$= 1 \cdot 2W + C \qquad (2)$$

where B = bandwidth between half-power points in cycles; C = a variable term, allowing for frequency drift; D = dot length in milliseconds; W = speed in words per minute. If only the fundamental component of the modulation is transmitted, the bandwidth may be restricted to about two thirds of the value given in (2). As the modulation index of a frequency-shift system is raised

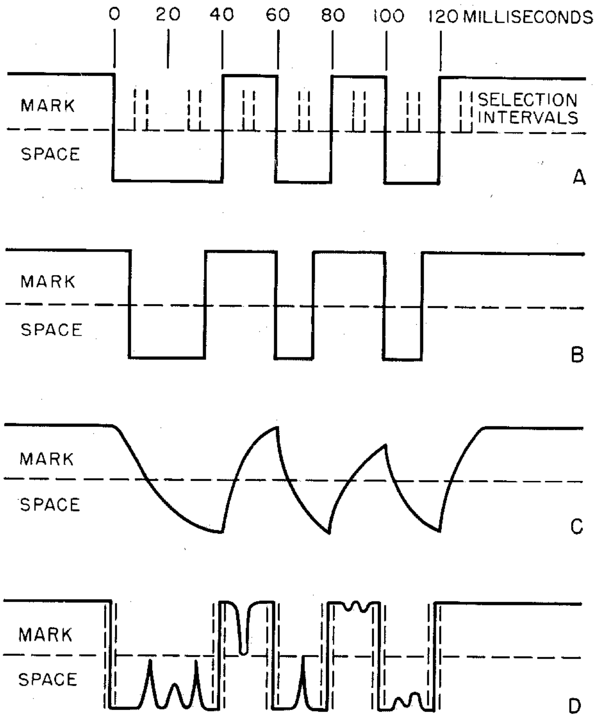


Figure 2—Illustration of the three main components of distortion. *A*—perfect signal, *B*—bias distortion, *C*—characteristic distortion, and *D*—radio distortion.

beyond 1, the bandwidth required approximates to twice the frequency shift, or Nf, whichever is the greater (where N is the number of significant sidebands and f is the keying frequency). It is important, however, to determine the optimum bandwidth for each system, having regard to prevailing conditions of noise and selective fading. Some of the main conclusions obtained from laboratory tests [3,4,10] are as follows.

Increase of bandwidth and frequency shift in a frequency-shift system improves the signal-to-noise ratio at low distortion levels provided the ratio exceeds 6 decibels. No such improvement is obtained by increasing the bandwidth for an amplitude-modulation system beyond that given by (2). Faultless operation of a frequency-shift

* The frequency shift is the total excursion between the marking and spacing frequencies. The keying frequency is half the baud speed, because one keying cycle comprises one dot and one space. Thus the modulation index equals the total frequency excursion divided by the baud speed.

system is obtained if the signal-to-noise ratio never falls below 6 decibels. The corresponding figure for an amplitude-modulation system is 10 decibels. In conditions of selective fading, whereas very few errors are recorded by a frequency-shift system provided the signal-to-noise ratio exceeds 6 decibels at the minimum of a fade, an amplitude-modulation system can give many errors even though the signal-to-noise ratio is high.

The bandwidth required after demodulation depends, for both systems, mainly on the baud speed. The degree of low-pass filtering does not depend on the frequency shift used in a frequency-shift system.

Reduction of distortion in the low-pass filter is important particularly if the filter is not followed by signal-shaping circuits. The input to the filter usually consists of a series of pulses of non-uniform length, and it is necessary that the slope of the leading edges of the pulses should be uniform. This requirement may be met by making the cut-off frequency greater than twice the keying frequency.

Some degree of low-pass filtering is necessary in both systems to attenuate the higher-frequency components of noise and so improve the signal-to-noise ratio. It has been shown[11] that the addition of a low-pass filter after a linear discriminator improves the signal-to-noise ratio of a frequency-shift system as compared with an amplitude-modulation system by

$$20 \log_{10}(3^{1/2} \Delta F/\Delta f), \text{ decibels,}$$

where ΔF = frequency shift, and Δf = filter pass-band, for high signal-to-noise ratio.

If the noise momentarily exceeds the level of the signal in a frequency-shift system, the signal will be suppressed, the output from the limiter will consist mainly of noise, and the demodulated output will momentarily fall to zero. As the noise increases relative to the signal, so the frequency and duration of these "holes" increase. Low-pass filtering will not remove this effect, but it reduces distortion substantially until the "hole" occurs, by reducing the high-frequency components of noise. Distortion caused by multi-path propagation contains many high-frequency components, which are also removed by low-pass filtering.

4. Direct-Printing Operation

In printing-telegraph systems, the sending and receiving machines are synchronous, and the synchronous driving power is derived either from a separate source of constant frequency or directly from the sender. In the start-stop system, synchronism is obtained by the transmission of two synchronizing impulses for each character. These impulses are the start (spacing) and stop (marking) signals. The length of the impulses depends on the signalling speed, but during non-signalling conditions, i.e. during short intervals between characters or words, or during longer pauses, the transmitter is on mark. The retention of the marking signal is important because a holding current is provided for the magnet of the receiving machine. If the unidirectional current furnished by the marking signal is lost because of the use, after demodulation, of circuits with relatively short time-constant, the onset of the next signalling impulse may result in the recording of an error. It is therefore desirable to use some form of direct coupling between the output of the demodulator and the final stage of the unit feeding the receiving machine.

In a frequency-shift system using a linear discriminator, there is no appreciable change in amplitude or waveshape of the output signal with frequency drift, but a direct-current displacement occurs, and it is therefore desirable to use some form of direct-current elimination. This requirement is at variance with the requirement during a long marking interval. A number of solutions of this difficulty have been tried with different success. One system[3] uses a coupling network that blocks the direct-current component and passes only the useful signalling component: the output of the network contains a feedback loop that passes only the direct-current component and low frequencies to compensate for the loss in the coupling network. Similar action is obtained with the circuits given by Hargreaves.[12] Direct-current clamping circuits[13] are useful. The output from the discriminator can be alternating current coupled to a pair of diodes in such a manner that part of the waveform is held at a fairly uniform level (within $\frac{1}{4}$ volt) without subsequent amplitude changes.

The disadvantage of this type of circuit appears during very long pauses between signals, when the driving voltage decays in the alternating-current network, leaving the clamping circuits responsive to noise.

An alternative method[14] is used in the equipment described in Section 8. Direct-current elimination is obtained by capacitive coupling after the linear discriminator. This avoids the difficulties associated with frequency drift: for example, a drift rate equivalent to changing from mark to space in one minute corresponds to a frequency of 0·008 cycle, and with a capacitance of 1 microfarad and a resistance of 1 megohm the attenuation at this frequency is about 26 decibels. During non-signalling conditions, the steady component due to the marking signal is lost, but the receiving machine does not respond, because a holding current is provided by a separate circuit that is quiescent during signalling periods.

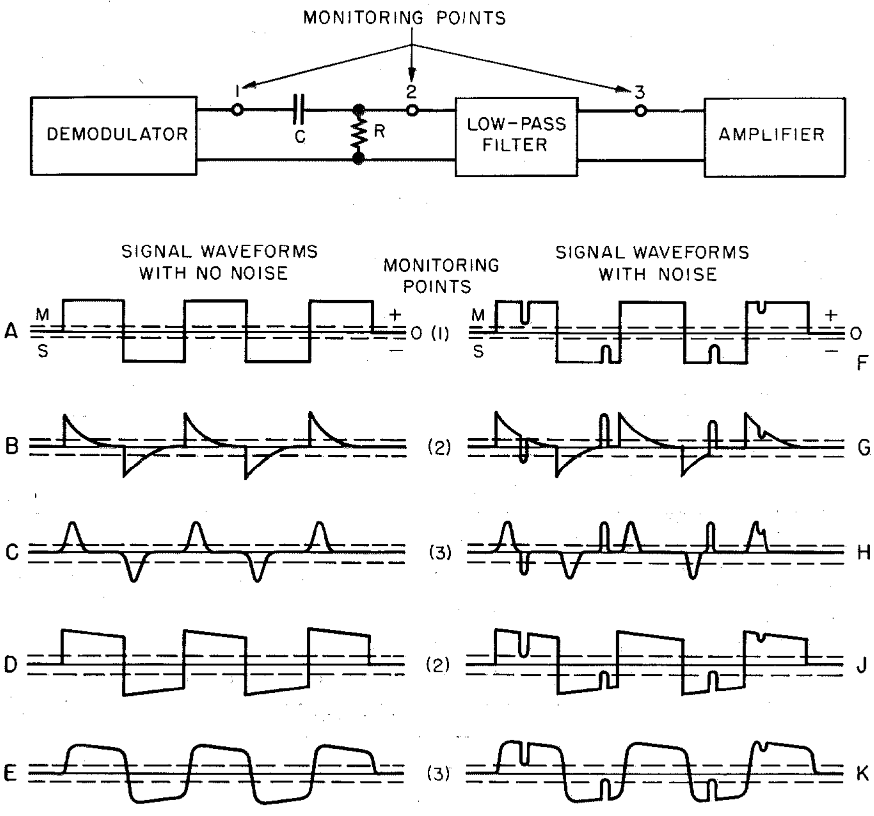


Figure 3—Reconstruction of signal shapes after demodulation.

The time-constant of the capacitive network following the discriminator should be greater than the duration of the longest unbroken sequence of mark or space elements. If the time-constant is of such a value that the demodulated waveform is effectively differentiated, the system becomes much more susceptible to noise. Figure 3 illustrates the type of action obtained. The waveforms shown at *B* and *C* correspond to the use of a time-constant considerably less than the duration of a signalling element, and those at *D* and *E* correspond to a time-constant some 50 times greater than an elemental period. Either mode of operation will give a clean output in the absence of noise. The waveform at *C* can be converted readily to a form suitable for recording if the amplifier in Figure 3 is capable of precipitate amplification only. Amplifiers of this type, sometimes known as "flip-flops" or dead-beat multivibrators, possess the property of having only two stable conditions corresponding respectively to the polarity of the signalling impulses, and the output consists of a series of amplified rectangular pulses. There will be no bias distortion provided the mark-to-space and space-to-mark transition delays are equal.

The effects obtained with differentiation of the marking and spacing impulses in the presence of noise are shown in Figure 3 at *G* and *H*. In these cases, spurious impulses are produced because the noise peaks extend below zero level and trigger the amplifier in the opposite sense to that of the signal element. The effect obtained with the longer time-constant under similar conditions is shown at J and K. In this case the output is unaffected by the noise, and the waveform at the output of the amplifier will be similar to that shown in A. (The noise impulses shown in Figure 3 are represented, for clarity, as single peaks extending towards the voltage zero.)

Figure 3 illustrates an advantage secured by the use of precipitate amplifiers or similar triggered circuits, in contrast to the use of direct-current amplification, namely that the final output is unaffected by noise provided the spurious impulses do not extend beyond the voltage zero in the sense opposite to that of the signal element.

5. *Frequency Stability and Automatic Frequency Control*

Radiotelegraph systems require a high degree of frequency stability if narrow bandwidths are used. Now the operation of equipment under pan-climatic conditions is becoming of increasing importance: for example, the operating temperature range of equipment for the armed services is -40 to $+70$ degrees centigrade. The advent of direct-printing installations in aircraft is not far distant, and equipment may well undergo rapid changes of temperature exceeding 60 degrees centigrade in range. Measures to prevent changes of frequency with temperature are therefore imperative.

Frequency drift of the carrier input to the receiving terminal causes bias distortion and, if severe enough, results in failure of the system. In multi-channel systems, all the channels may be put out of action.

In an amplitude-modulation system, bias distortion rises rapidly as the carrier frequency approaches the cut-off region of the filters, due to the reduction in amplitude of the demodulated signal. In a frequency-shift system, the amplitude of the demodulated signal is constant, but the direct-current component of the signal changes as the carrier moves away from the nominal frequency, with a steady rise in distortion. If the direct-current component is eliminated, the limiting condition before failure is again determined by the cut-off frequency of the filters preceding the demodulator, with one important difference from the amplitude-modulation case: the signal-to-noise ratio decreases as the carrier moves away from the centre of the pass-band.

Long-term stability of the order of 10 parts in a million is obtainable with oven-controlled crystal oscillators. This figure represents a summation of the errors due to cutting and grinding, secular effects, and variation of ambient temperature.

The requirements for short-term frequency stability are far more stringent for frequency-shift than for amplitude-modulation systems. For the former, an instantaneous frequency deviation of only five parts in a million can be sufficient to cause errors through approximating to the frequency shift used for signalling. In an amplitude-modulation system, sudden changes of frequency of that order are unlikely to cause errors unless accompanied by transient changes in amplitude.

With a frequency-shift system, the summation of the frequency errors occurring over the link, and the degree of shift used, set a limit to the use of systems based on filters centred on the mark and space frequencies.[15] The cross-over point corresponding to the nominal carrier frequency is invariable, and the operating margin is rapidly reduced if the deviation of frequency from nominal exceeds 300 cycles. On the other hand, a system using a linear discriminator followed by a network that is alternating-current-coupled or gives direct-current coupling without a direct-current displacement will continue to print without error even if the mark and space frequencies are displaced.

For some conditions of service, in which a range of frequencies has to be covered at the receiving terminal, it may be desirable in the interests of flexibility and economy in crystals to use variable-frequency oscillators. Long-term stability of the order of 10 parts in a million per degree centigrade is obtainable with existing techniques. In one system,[23] this performance is approached by the use of a commercial variable capacitor of rugged design with temperature compensation at two settings corresponding respectively to the minimum and maximum frequencies of the oscillator.

There are two common methods of covering the required frequency range. First, by frequency multiplication of a variable-frequency oscillator; and secondly, by using a small number of crystals which, after frequency multiplication, cover the frequency range, the gap between each crystal frequency being covered by heterodyne combination with a tunable oscillator operating at a relatively low frequency. The required stability is more easily obtained by the second method, but careful choice of the fixed and variable frequencies is required if freedom from unwanted responses is to be achieved. These responses can be minimized by the use of balanced mixers, and the ratio of the final frequency to the variable-oscillator frequency should not exceed[17] about 20:1.

Satisfactory automatic frequency control is difficult with frequency-shift working, but it is usually possible to obtain long periods of unattended operation without frequency correction if the oscillators are crystal-controlled and if the receiver bandwidth is sufficient to accommodate the shift and the frequency error. If these conditions are not possible, some form of automatic frequency control is usually needed. One method of providing automatic frequency control consists of suppressing the spacing signals and using the marking signals to provide a control voltage. A disadvantage of this method is that the marking pulses have to be integrated by circuits of long time-constant to provide a uniform control voltage during signalling intervals, resulting in a rate of correction that may be too low. Alternatively, both the marking and spacing signals can be used by converting them to the same frequency by a heterodyne process (i.e. by beating them with a signal having a frequency equal to the frequency shift) yielding a composite and continuous control voltage. The advantage of this method is that the rate of correction can be controlled more easily than in the method previously mentioned.

Unless the automatic frequency control is entirely satisfactory, it is better to dispense with it and secure frequency stability of the various oscillators in the link. If all the oscillators in the receiving terminal are crystal-controlled, the system is inflexible. Some degree of frequency drift can, however, be compensated by using a free-running oscillator in the final conversion process. Manual variation of about one-fifth of the bandwidth is useful if the frequencies of all the other oscillators in the link are fixed.

It is difficult to ensure that an automatic-frequency-control system will be unresponsive to noise or adjacent-channel interference unless some type of sampling is used. Such a process might be used to apply frequency control during periods of high signal-to-noise ratio and freedom from interference only.

6. *Diversity Reception*

Frequency diversity requires more frequency space and radiated power than space diversity, but is useful when space diversity is difficult through limitation of site area, e.g. on ships. With frequency diversity using amplitude or frequency modulation, two-, four-, or six-tone working is commonly used with single-sideband operation. An alternative method of frequency diversity has been suggested for frequency-shift systems,[3] which may be effective if the necessary bandwidth can be used. The frequency-shifted signal is phase-modulated, which spreads the energy of the signal over a wider frequency band. For example, with a phase swing of $\pm 1\cdot 4$ radians at a frequency some three times the frequency shift, three major components of the signal are available and can be separately demodulated and combined to form, in effect, a triple-diversity system.

The main object of diversity working is to reduce the operating time lost through unfavourable conditions in the medium. If, on the other hand, a certain proportion of lost time can be tolerated, diversity working will allow a reduction in the transmitted power. The effective gain due to a diversity system, taking into account time lost, is given by[18]:

$$\text{Diversity gain} \approx \frac{20}{m}\left(1-\frac{1}{n}\right)\log_{10}\frac{1}{K}\text{, decibels} \quad (3)$$

where K = proportion of time-loss allowed; n = number of constituent signals; m = a constant. If the received signal is considered as the resultant of a number of signals of equal amplitude and random phase relationship, the value of m for more than two signals[19] is approximately 2. With this value, putting $K = 0\cdot 01$, the diversity gain is shown in Figure 4 as a function of n.

The improvement offered by the use of diversity reception depends to a great extent on the method of selecting and combining the separate signals. Direct addition of the outputs from the diversity paths is unsatisfactory because a phase delay of the incoming signal produces an equal phase delay of the audio-frequency signal

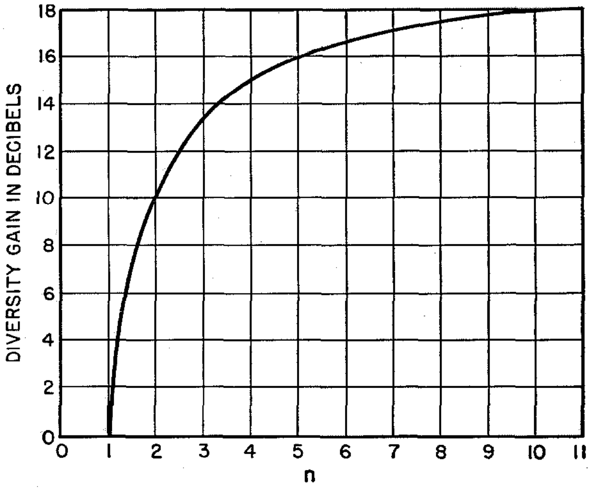


Figure 4—Diversity characteristics. n = number of constituent signals.

when the output is obtained by heterodyning alone. Combination of the signals after demodulation is better because the delay in the modulation envelope is equal to the time delay between the arrivals of the signal components which, being usually not greater than two milliseconds, is short compared with a modulation period (about one tenth at 50 bauds): phase cancellation of the signals is therefore impossible.

Diversity circuits for amplitude-modulation systems are of two main types: A) those in which the automatic-gain-control circuits of the receivers are combined in a common load resistance; B) those in which the highest-level audio-frequency output is selected from the alternative receptions. Both types of circuit may be used together. With common automatic gain control, it is necessary that the automatic-gain-control characteristics of the different receivers should be similar to maintain a uniform output during changeover between paths. A useful feature of the second system is that the receivers require no modification for diversity working, for it is only necessary to feed their audio-frequency outputs into a combining unit.[20]

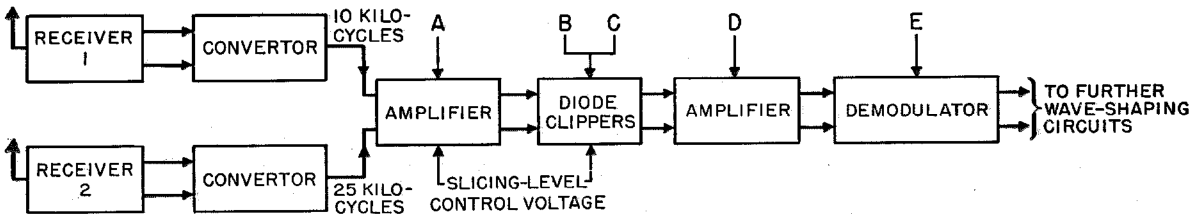
Another method used with amplitude-modulation reception,[16,21,22] which does not fall into categories *A* or *B* mentioned in the previous paragraph, is that used in the equipment described in Section 8. A block diagram and the waveforms at various points in the circuit are shown in Figure 5. The common-frequency signals from receivers *1* and *2* are converted to 10 and 25 kilocycles respectively. These two signals are combined and amplified in a common load. The output waveform at this point, which may also contain noise components, is shown at *A*. The diode clippers are arranged to cut a thin slice out of the marking impulses, as indicated by the dotted lines in *B*. The thickness of the slice is constant for all signal conditions, but the slicing level is automatically held more or less at the position relative to the signal amplitude shown in the diagram, by control voltages derived from the signal. The output waveform after this clipping is shown at *C*. At *D* is shown the resultant waveform after direct-current elimination and further amplification. Both components of the signal may be present, together with noise, but nevertheless the marking impulses have been completely separated from the spacing noise level. The waveform at *E* is obtained after further demodulation. The time-constant of the control voltage that determines the clipping levels is automatically related to the keying speed.

Very rapid switching between receivers is required in a diversity system. If, however, the delay between the alternative transmissions is relatively small compared to the length of an elemental dot, it may add momentarily to the distortion every time a change-over from one receiver to another takes place.

Switching is controlled by differences in level between the received signals, operation with differences of the order of 1 decibel or less being typical of many systems. The possibility of making full use of such characteristics depends to a large extent on how closely the gain of the circuits prior to the diversity-combining unit can be equalized. It is important that the gain characteristics should be identical at all levels of signal input. It is also necessary that the switching process should not be accompanied by transients at the fastest fading rates expected. On the other hand, operation with level differences approaching 3 decibels, while giving some deterioration in performance, appears to be satisfactory, and avoids the necessity of obtaining identical gain characteristics in each receiver.

In general, diversity-circuit arrangements for frequency-shift working can be divided into two main classes: A) those that use some form of gating circuit to select the higher-amplitude signal; B) those in which the common frequency signals from each receiver are reconverted to different frequencies and fed into a common limiter. Both classes of circuit automatically select the stronger signal and suppress the weaker. Circuits of the first type should be free from transients during rapid change-overs. Circuits of the second type have to be designed to be insensitive to amplitude modulations caused by spurious responses, close to resonance.

Combination of the automatic-gain-control circuits of the receivers may help to improve the diversity action. This applies particularly to amplitude-modulation systems with reversed on–off keying, because combination makes the output circuits less prone to disturbance by noise during spacing intervals. With frequency-shift working, effective automatic gain control is required if limiting occurs at any stage below the threshold level of the diversity combining-unit. Automatic gain control may also be of assistance during long standby periods.



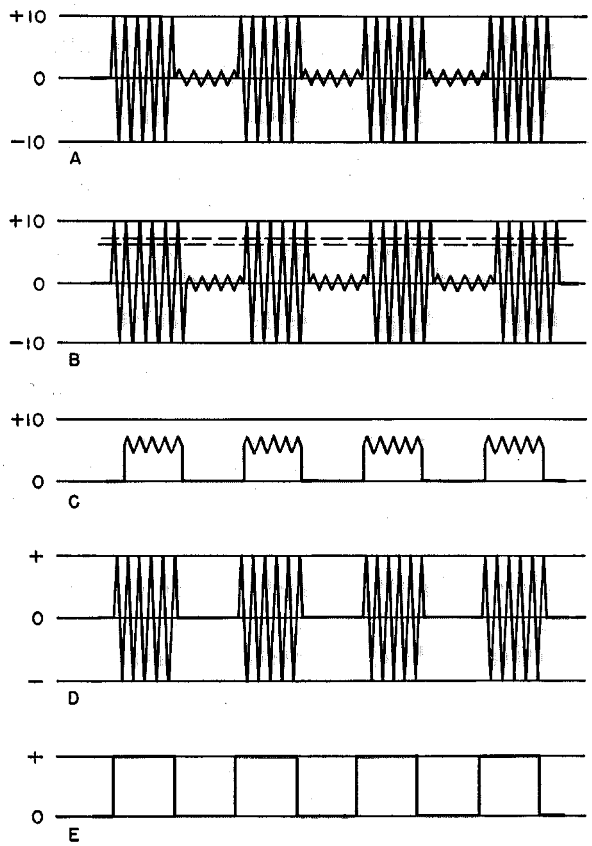


Figure 5—Reconstruction of signal shape in an amplitude-modulation system arranged for dual diversity, with 50-baud reversals. *A*—original signal, *B*—clipping process, *C*—signal after clipping, *D*—further amplification and direct-current elimination, and *E*—demodulation.

7. *Noise Suppression*

The presence of noise gives rise to radio distortion which, with teleprinter working, may be sufficient to cause overprinting, erratic operation of the line-feed mechanism, and the recording of spurious impulses. Noise induced in the receiving aerial comprises: A) atmospherics, B) disturbance of local origin.

Many methods of minimizing the effects of noise are available which, in conjunction with suitable gating circuits, can be arranged to mute the output circuits of the receiving system to give protection against random operation of the printing machines.

The efficacy of these methods depends to a great extent on their ability to discriminate between the prevailing signal and noise levels. Muting circuits that operate at a predetermined level of the incoming carrier may not be always satisfactory because it is often impossible to avoid muting before the signal-to-noise ratio is poor enough to justify such action.

The ability of limiters in obtaining improved conditions against the effects of impulsive and fluctuation noise has been established.[21, 22] In a frequency-shift system provided with an effective limiter, there is a marked increase in output signal-to-noise ratios above the improvement threshold. Below this point, which may be defined as equality of peak-noise-to-peak-carrier input levels, the signal-to-noise ratio decreases rapidly. Between the limits of noise only and conditions corresponding to signal-to-noise ratios greater than unity, the output waveform consists of, either, random noise components or a series of rectangular pulses due to the signalling elements. When the signal level is below the improvement threshold, the output of the limiter becomes amplitude modulated by the noise and it should, theoretically, be possible to rectify the noise currents in order to obtain a control voltage for muting purposes.

The most difficult case, in considering the problem of noise suppression, is that due to interference from impulsive noise. In this case, the effectiveness of the limiter action depends to a large extent on the bandwidth preceding the limiter. With narrow bandwidths, the duration of the noise pulses becomes excessively lengthened and the signalling pulses may be subjected to random time shifts with consequent printing errors.

With increasing use of narrow bandwidths, the problem of interference by impulsive noise becomes of outstanding importance, particularly in locations such as on board ships. In this case, due to the proximity of noise to the receiving system, the probability of breakdown due to this cause is high.

With receiving terminals on land, local sources of impulsive noise should be kept more than a quarter of a mile from the receiving aerials. Mechanical senders are often a source of noise, and they require good shielding and filtering. Measurements on such devices indicate that the peak noise field may be as high as 1200 microvolts per metre at a distance of one foot, falling to 50 microvolts per metre at 10 feet.

8. System for Receiving Amplitude-Modulation or Frequency-Shift Signals

8.1 Description and Details of Performance

The system is built up of a number of units that can be combined to provide non-diversity, dual-, or triple-diversity working. A block diagram of the system is shown in Figure 6. (Three commercial equipments corresponding to these types are shown in the photographs on page 46.)

With frequency-shift working, the intermediate frequency of two of the receivers is changed from 10 to 25 and 45 kilocycles respectively; that of the third remains at 10 kilocycles. The three frequencies are presented to a common limiter followed by selective circuits at each frequency. The 25- and 45-kilocycle channels are reconverted to 10 kilocycles and fed to discriminators. This process is followed by low-pass filtering and signal-shaping circuits.

Similar action is obtained with amplitude-modulation diversity working, except that the output of the amplitude-modulation demodulator is fed directly to the signal-shaping circuits.

The characteristics of the system are as follows:

Frequency Range: 4–25 megacycles.

Output (Amplitude Modulation or Frequency Shift): 60–0–60 milliamperes polar direct current and 800-cycle tone; 100 milliwatts in 600 ohms.

Range of Shifts: The permissible frequency errors plus the maximum value of shift used must not exceed ±1000 cycles. The lower limit of shift is 80 cycles.

Telegraph Distortion: Less than 5 per cent at keying speeds of 12 to 150 bauds for a non-recurrent waveform with signals removed not more than ±500 cycles from the nominal value.

Keying Speed: 600 words per minute maximum.

Sensitivity: The rated output is obtained with an input signal of 2 microvolts.

Frequency Stability: Within 1–2 parts in a million per degree centigrade with crystal control; 10 parts in a million per degree centigrade without crystal control.

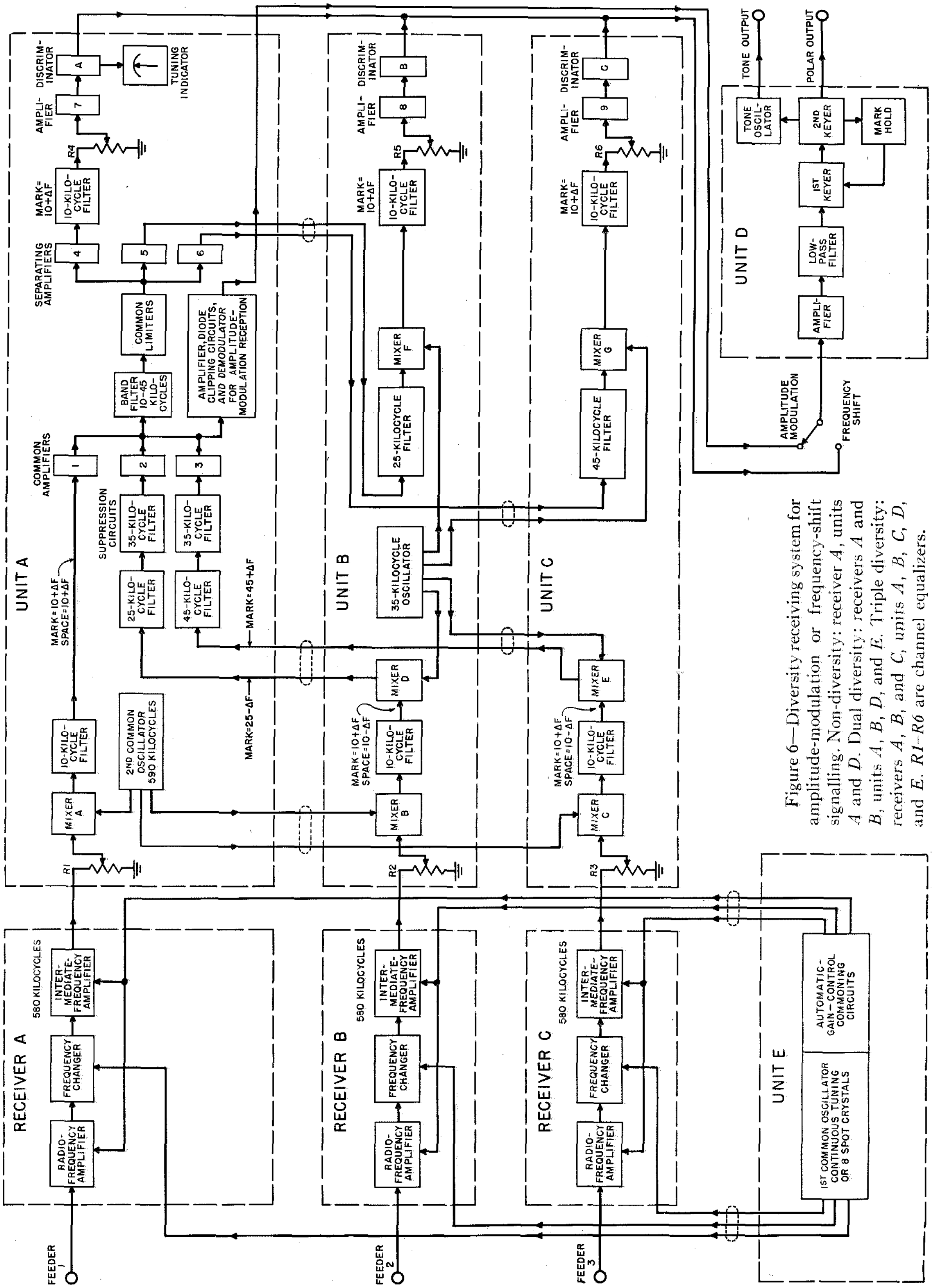


Figure 6—Diversity receiving system for amplitude-modulation or frequency-shift signalling. Non-diversity: receiver *A*, units *A* and *D*. Dual diversity: receivers *A* and *B*, units *A*, *B*, *D*, and *E*. Triple diversity: receivers *A*, *B*, and *C*, units *A*, *B*, *C*, *D*, and *E*. *R1–R6* are channel equalizers.

8.2 Diversity Arrangements

With frequency-shift working, the selection of the stronger signal is obtained by the use of a common limiter aided by an effective method of interconnecting the channel discriminators. Only the discriminator with the highest input level contributes to the output. Typical diversity characteristics are shown in Figures 7 and 8. The interconnection of three balanced discriminators to a common load [24] is shown in Figure 9, and the corresponding diversity characteristics are shown in Figure 10. The amplitude-modulation diversity system has been described in Section 6.

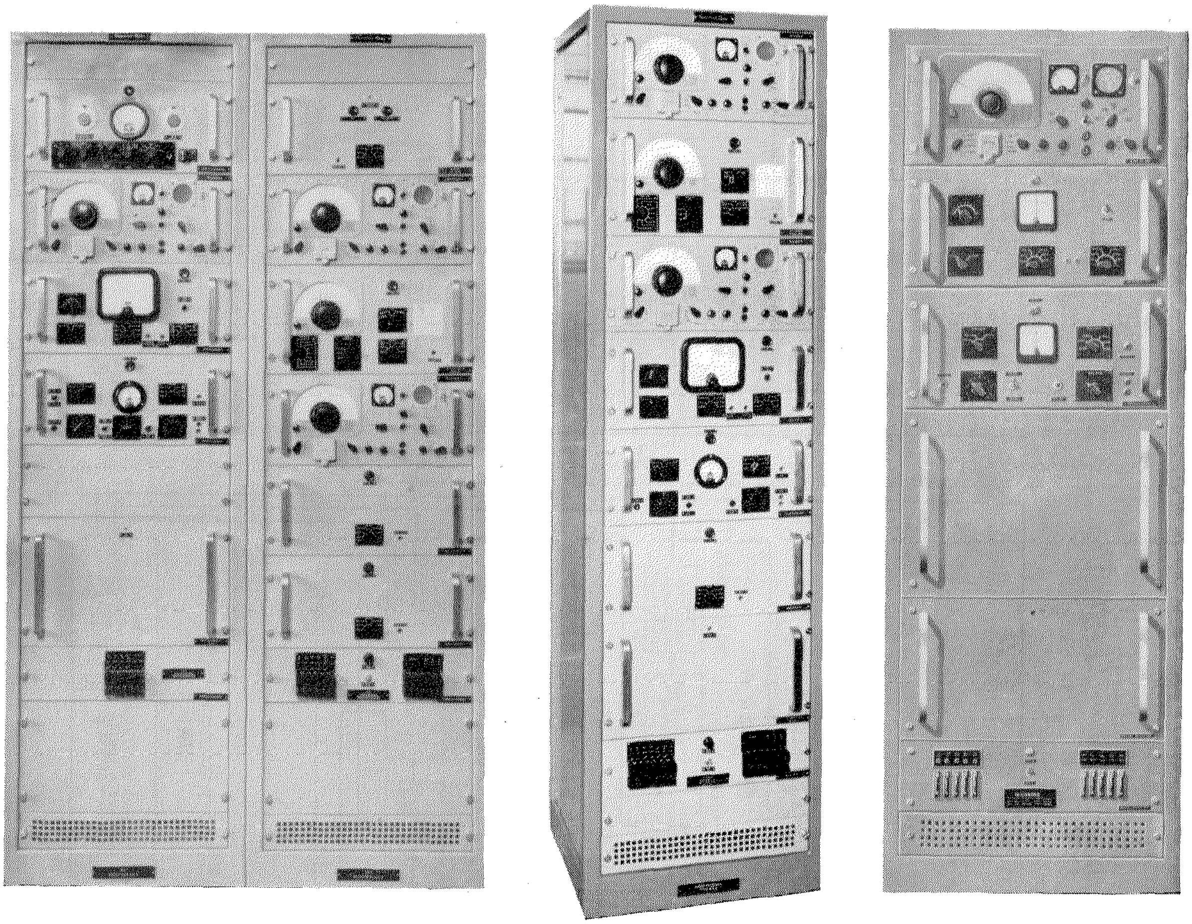
8.3 Keying and Signal-Shaping Circuits

These circuits are in unit *D* (Figure 6), and are shown schematically in Figure 11.

The square signal-pulses from the discriminators are amplified by *V1* and filtered. Alternative filters are provided to cover the range of keying speeds.

Two trigger circuits, variants of the flip-flop type, are used and are keyed by the signal transitions at the grid of *V3*. The characteristic point at which the triggering takes place is determined by the setting of *R1*. This control is effective in removing bias distortion, but the action is not automatic because no correction is applied for random variations of bias due to changes in the medium.

During prolonged marking intervals the steady component of the marking signal is lost owing to the capacitive coupling used after the demodulator *C1* in Figure 9. A restoring action has been obtained as follows. During signalling intervals, the polar output developed across *R2* is passed through *C0* to two diodes (*V9* and *V10*



From left to right are the *RX4*, *RX5*, and *RX6* radiotelegraph receivers produced by Standard Telephones and Cables, Limited. They are respectively triple-, dual-, and non-diversity equipments.

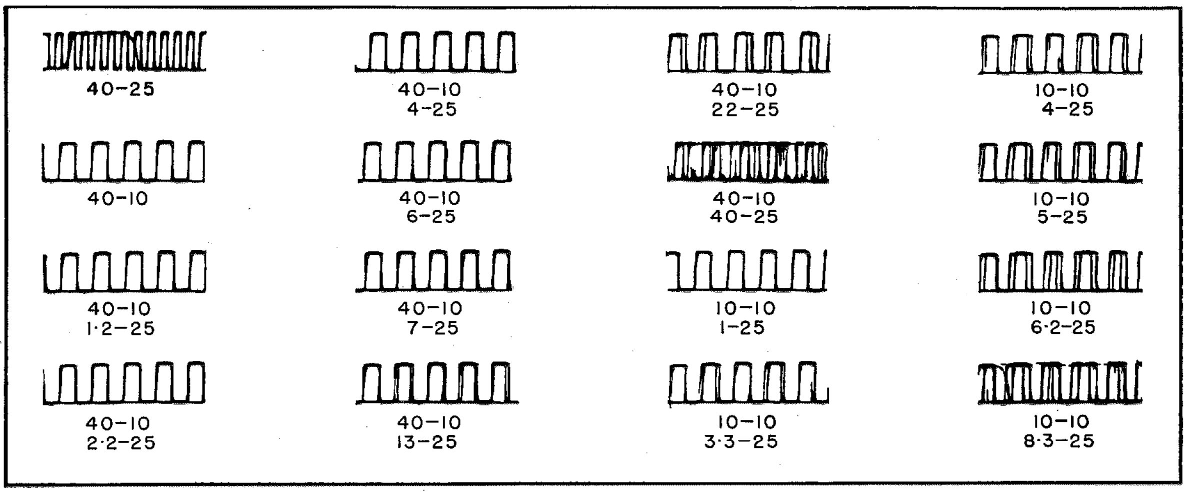
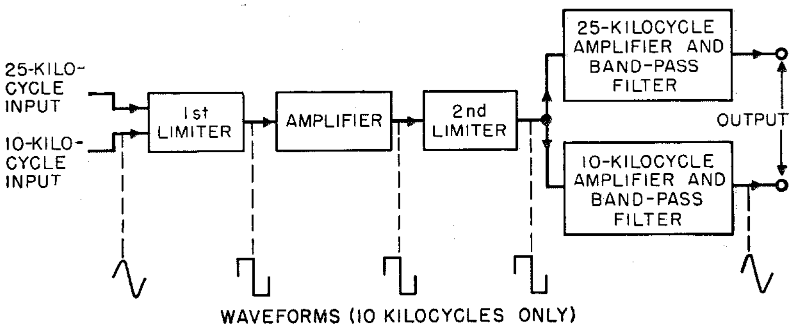


Figure 7—Signal shapes at the output terminals of common limiters for different input levels and frequencies. The first figure under the waveform in each case is the input level in volts, and the figure following the dash is the frequency in kilocycles.



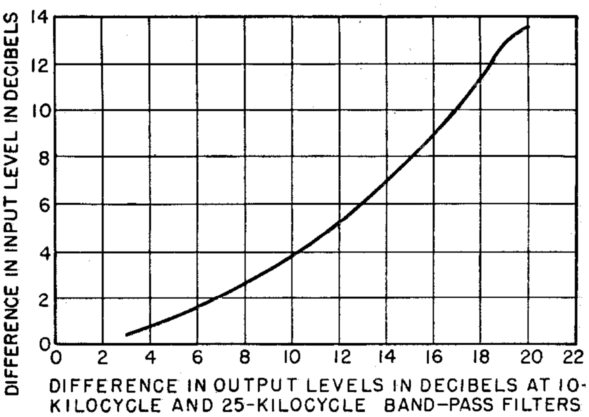


Figure 8—Diversity characteristics due to common-limiter action.

in Figure 11). A negative-going pulse drives *V2* to cut-off, and the resulting positive pulse developed at the anode of this valve is suppressed by the conducting diode *V11*. The valve *V3* is therefore triggered by the incoming signal transitions only. *V2* is also driven to cut-off during a positive pulse arriving through *C0* due to the conducting diode *V10*, and the mark holding valve *V2* becomes inoperative. At the end of a signalling period, *C2* discharges with a consequent holding-off bias at the grid of *V2*. This action produces a negative-going pulse at the suppressor grid of *V3*, which is fed back to the grid of *V2* through *C4*. Hence *V3* is cut off and *V4* is left in a conducting state. Since the ouput of *V4* controls the second trigger circuit, the output polarity at the terminals *3* and *4* is determined.

If a noise pulse is received during the idling period and is of sufficient level to actuate the first trigger circuit, the cycle of operations is repeated. The mark-holding pulses developed at the grid of *V3* are of approximately 0·4-millisecond duration, which ensures that the first signalling element received after a long standby period is practically undistorted.

9. *Acknowledgments*

The author wishes to thank Mr. C. E. Strong, chief radio engineer of Standard Telephones and Cables, Limited, for granting permission for publication, and to acknowledge the valuable assistance he has received from discussions with

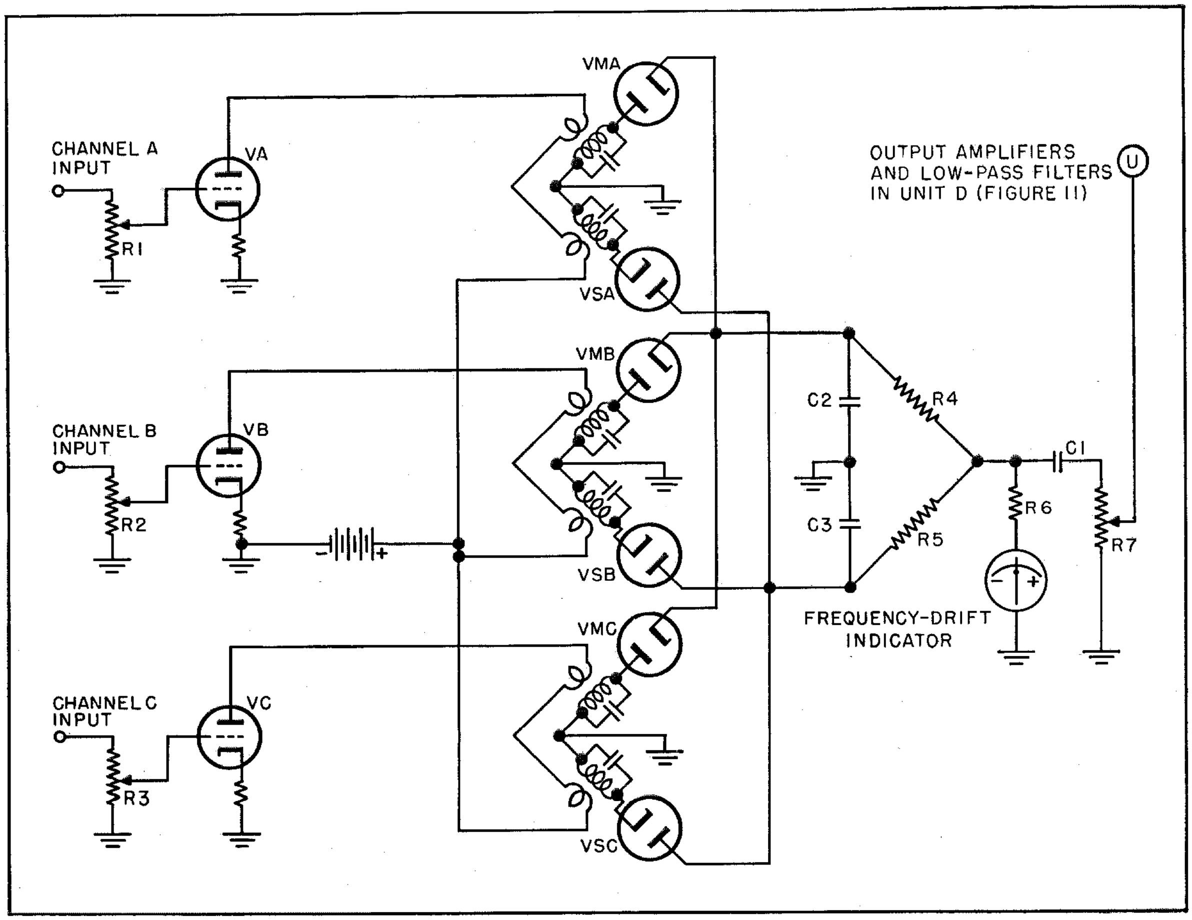


Figure 9—Interconnections for three balanced discriminators. *VMA*, *VMB*, *VMC*—marking diodes. *VSA*, *VSB*, *VSC*—spacing diodes. *VA*, *VB*, *VC*—channel amplifiers. *R1*, *R2*, *R3*—channel equalizing controls.

Messrs. D. Hamilton, T. F. Hargreaves, J. M. Kirk, H. T. Prior, and the team that developed the equipment described in Section 8.

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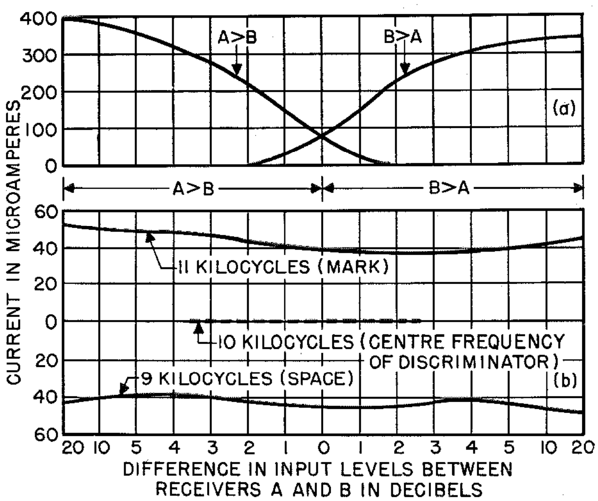

Figure 10—Overall diversity characteristics (units *A* and *B*). At top—discriminator diode currents, and below—output current from two balanced discriminators (monitored by meter shown in Figure 9).

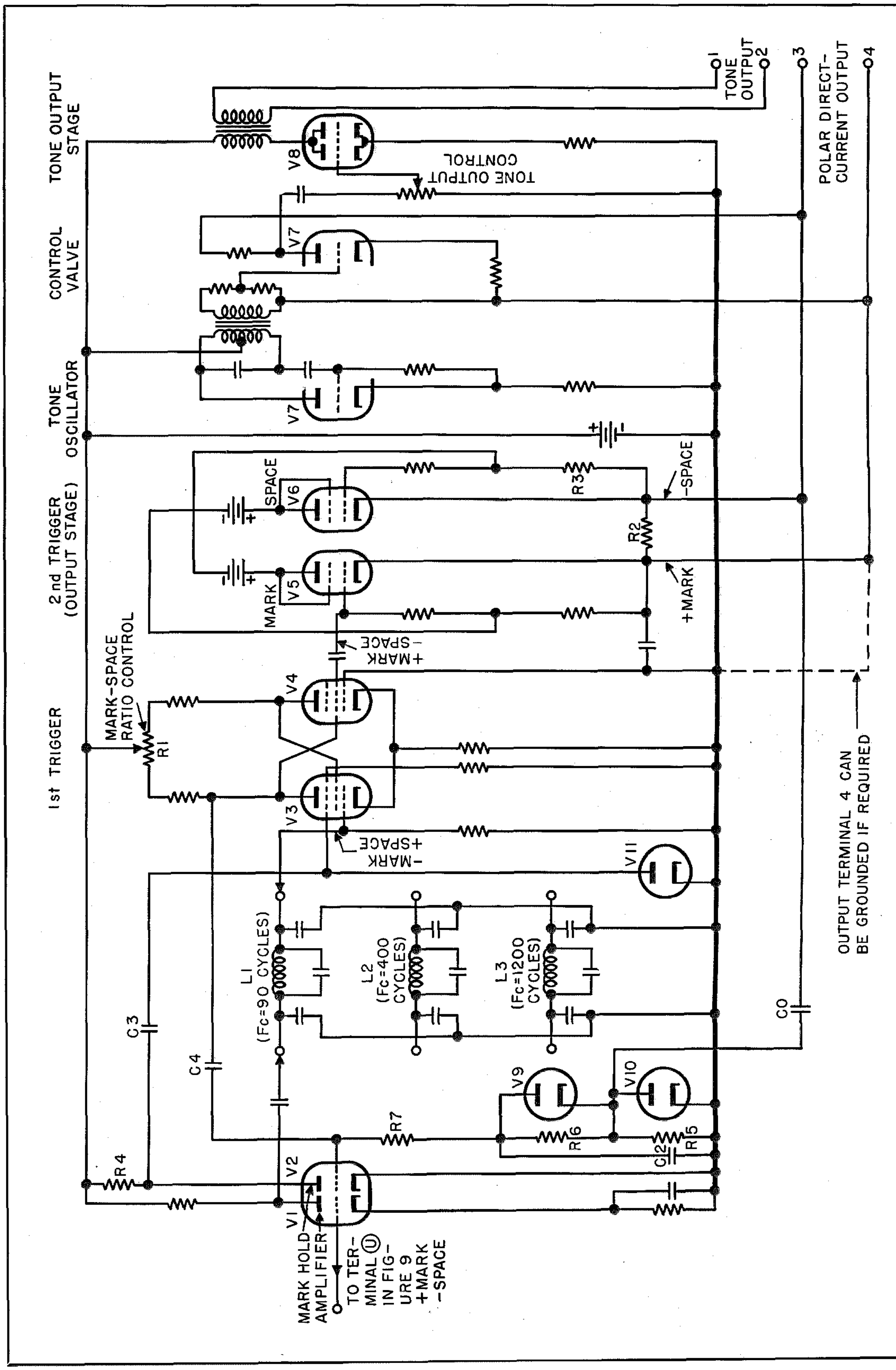

Figure 11—Signal-shaping, mark-holding, and output circuits (unit *D*). 800-cycle tone output = 50 milliwatts in 600 ohms. Polar output = 50-0-50 milliamperes in 1200 ohms.

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Precision Calibrator for Low-Frequency Phase-Meters*

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IN RECENT YEARS several navigational aids have been developed, each involving the measurement of the phase relationship between two alternating voltages of the same frequency. These aids include certain types of automatic direction finder (e.g., the phase-modulation type[1]) and some omni-directional beacons (e.g., Luck beacon,[2] the Civil Aeronautics Administration beacon,[3] etc.). Examples requiring a phase measurement exist in many other fields of electrical and radio engineering.

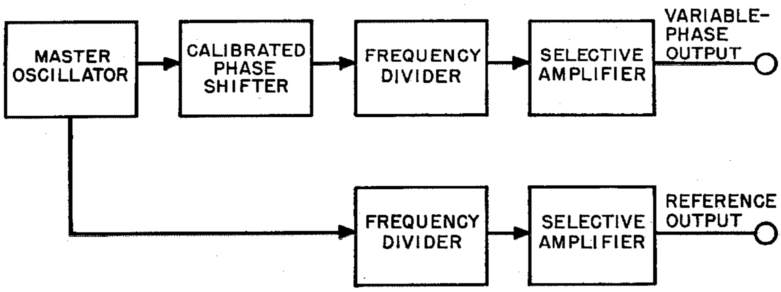


Figure 1—Block schematic diagram of the phase calibrator.

An accurate phase calibrator is therefore required, and it is the purpose of this article to describe such a calibrator. Other methods of calibration already exist but all have a strictly limited accuracy. All depend on accurate mechanical and electrical construction. Furthermore, the accuracy is not easily checked to better than ±1 degree due to the lack of a reference having sufficient accuracy and due to the inherent difficulty of phase measurement. First, there has been no absolute standard of phase difference constructed to date with one known exception.[4] Secondly, even when a standard has been set up and it is required to check the calibrator against it, a measurement of phase is involved and this cannot be performed with sufficient accuracy.

Thus it is necessary to build a calibrator that is fundamentally accurate, so that it does not require to be used with reference to a standard. A simple calibrator of this type has already been constructed independently in America,[4] but that to be described is superior in several respects. It will be appreciated from the following description that its short-term accuracy is unlimited and its long-term accuracy is excellent.

1. Principle of Operation

The calibrator provides two outputs at the required frequency. The phase of one output is fixed and this is termed "the reference output." The other output is termed "the variable-phase output" and the channel producing it is provided with the necessary phase shifter. This is calibrated so that the two outputs can be arranged to have any desired phase difference.

In simple calibrators, the phase shifting is performed at the output frequency, so that the accuracy of any phase shift cannot be greater than the accuracy of the phase shifter itself. However, in the calibrator to be described, the phase shifting is performed at a harmonic of the output frequency, and the output is obtained by frequency division. The reference output is obtained from a second frequency divider for which the input is not shifted in phase. The block schematic diagram is shown in Figure 1.

The first published reference to the use of harmonics in a related manner[5] referred to the measurement of the phase relationship between two voltages by performing the measurement with the 8th harmonic of one of them. This harmonic was produced by frequency multiplication of one input.

Referring to the phase calibrator, suppose that the required output frequency is f and the nth harmonic nf is supplied to the phase shifter. An oscillator is set up giving an output at the frequency nf. This is applied to two separate channels. One channel is operated at fixed phase. The other is the variable-phase channel.

* Reprinted from *Wireless Engineer*, v. 28, pp. 197–208; July, 1951.

† Mr. Wintle is now with the Admiralty Signal and Radar Establishment.

[1] Numbered references appear in Section 10 on page 61.

The fixed-phase or reference channel consists of one or more frequency dividers having an overall frequency division factor n and produces an output at the required frequency f at constant phase. The output circuits may be selective, if required to give a good waveform.

The variable-phase channel has at its input end a phase shifter whose accuracy need not be great. The requirements of this phase shifter are that its output should have an approximately constant amplitude as the phase is varied, that the output should be continuous, and that the phase of the output should be correct to an accuracy to be decided later.

The phase shifter is followed by a frequency-divider circuit identical with the one in the complete reference channel.

Suppose that the division factor of each channel is 36. Then an error of phasing in the phase shifter of 3·6 degrees at any setting will give an error in the output phase of 0·1 degree only. It is clear that increasing the division factor increases the accuracy of the output phase. In fact,

$$\text{Output phasing error} = 1/n \text{ (phasing error at the output of the phase shifter).}$$

The most important feature of the method is this. If the shaft of the phase shifter is rotated by exactly 1 revolution from any initial position, so that the output of the phase shifter varies in phase through 360 degrees, then final output at frequency f varies through exactly 360 degrees/n whatever the division factor n. This is a fundamental property of the calibrator, so that it is self-calibrating every 360 degrees/n.

If n is 36, the phase of the output is exact every 10 degrees and this becomes 1 degree if n is increased to 360. Such a large division factor is unnecessary, however, because there is a certain initial accuracy in the phase shifter itself. If a square potentiometer[6] or an electromagnetic phase shifter is used and the two-phase input is accurately phased, then the output phase is also as accurate as the construction allows. If the two-phase input is incorrectly phased or the amplitudes are unequal, there is an additional and calculable error. The error need be known only to a first order because it is reduced in order by the frequency-division process. It is also possible, if a resistive phase shifter is to be used, to employ a construction somewhat simpler than that of the square potentiometer. This is described later when it will be shown that a maximum error of 4 degrees is produced. Division by 36 reduces this to 0·11 degree. However, the simplified phase-shifting potentiometer is still accurate at the four feed points if the input phasing and amplitudes are correct, so that the input phase after division by 36 is accurate at intervals of 2·5 degrees. This is generally sufficient for the calibration of a phase-meter.

The two divider channels are made identical, so that any change of high-tension voltage, etc., affects them equally.

2. *Master Oscillator*

The frequency of the master oscillator is decided by the division factor and the final frequency required. For omni-directional beacons, the frequency at the phase-meter is now generally 30 cycles per second[8] and for the phase-modulation direction finder the frequency may be up to 100 cycles approximately.[1] It will be considered as an example that the final frequency is to be 30 cycles and the division factor 36. This requires an original frequency of 1080 cycles and this may be produced by a resistance–capacitance oscillator, tuning-fork oscillator, or otherwise. In the apparatus used by the writer, the frequency was required to be very accurate and therefore a crystal master oscillator was used. Since crystals operating below 4 kilocycles are not readily available, one operating at 4320 cycles was obtained. The output of the crystal oscillator must be divided in frequency by a factor of 4 to give the required input to the calibrator proper. Alternatively the division factor $n = 36$ can be increased to 144, when one revolution of the phase shifter shifts the final phase by 2·5 degrees. However, such a large division factor makes the use of the calibrator laborious.

Quartz crystals in this frequency range are of the four-terminal type and as they are not well known some description is included. The crystals are cut in the form of a bar with four gold electrodes, and if an appropriate pair are joined, the three-terminal network remaining simulates a tapped tuned circuit.

The crystals can then be connected, for example, in the same way as the tuned circuit in the Hartley oscillator. A precaution is essential, however, in that some form of automatic gain control must be used. The crystals have a high Q and the build-up time is several seconds. Once the crystal begins to oscillate it can do so very violently, and is easily fractured. One maker of these crystals* recommends a 2-valve circuit but the writer has used a single pentode valve circuit of Hartley type with success (Figure 2). A circuit tuned approximately to the crystal frequency is used as anode load of *V1* but it may be replaced by a resistance. The alternating anode voltage is rectified by *V2* and used as automatic-gain-control bias on the suppressor grid. The inclusion of a high resistance *R2* in the screen-grid circuit lowers the screen voltage when the anode current is reduced by the action of the automatic gain control, and the latter is made more effective. In addition, the maximum dissipation of the screen grid is not exceeded. The anode current is cut off by a suppressor-grid voltage of about −50 volts in an average receiving-type pentode. This corresponds to an alternating anode voltage of about 34 root-mean-square volts. Therefore, in normal operation, the automatic gain control will cause a stable amplitude to be reached when a voltage of about 25 root-mean-square volts appears at the anode.

3. *Variable-Phase Channel*

3.1 Main Phase-Shifting Control

The continuous phase shifter, which follows the master oscillator, may be of any convenient type such as the Drysdale type phase-shifting transformer (e.g., the Magslip), or a resistive polygonal potentiometer,[6] of which the 'square' potentiometer is a particular case.

However, as already pointed out, the process of frequency division reduces any error in the phase shifter. Thus, depending on the final accuracy required, some inaccuracy can be allowed in the phase shifter. The manufacture of a highly accurate square or polygonal potentiometer is, therefore, not justified and a simpler construction is allowable. The square potentiometer becomes a 'round potentiometer' which is easier to make.[7]

The round potentiometer consists of a circular winding similar to those used in good quality wire-wound potentiometers with the exception that it is an endless toroidal winding. Four taps

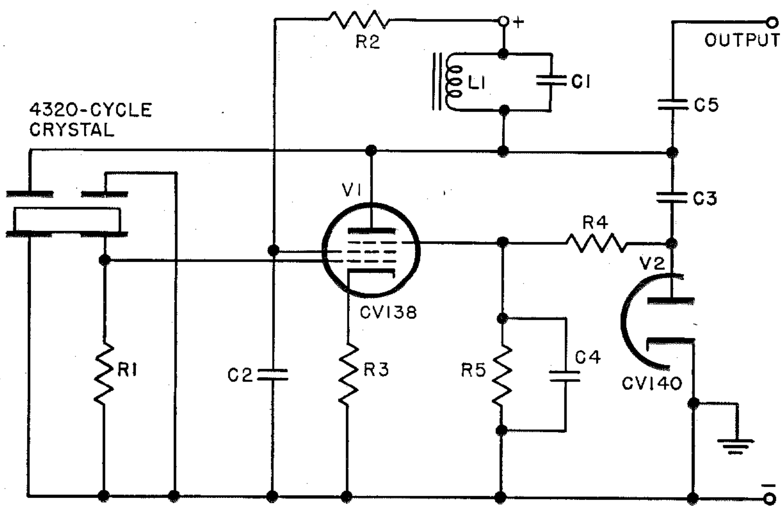


Figure 2—Crystal oscillator circuit.

are made at intervals of 90 degrees. The wiper can be rotated continuously and it is important that the contact should not be intermittent, as explained later. The wiper is associated with a simple revolution counter which is an extension of the Maltese-cross mechanism. This counter divides by the same number as the overall division factor of the electrical frequency dividers, so that the output shaft of the mechanical revolution counter indicates roughly the output phase of the output wave after frequency division.

The revolution counter indicates the number of complete revolutions of the shaft of the wiper and it is marked in 10-degree intervals from zero through 360 degrees back to the same zero. The position of the shaft of the potentiometer subdivides these 10-degree intervals and the potentiometer scale is divided into 10 equal divisions, each of which represents 1 degree change in the final phase. The phase-shifting potentiometer is, therefore, provided with coarse and fine scales.

Whereas the square potentiometer is theoretically perfectly accurate, the round potentiometer is fundamentally inaccurate if a uniform circular winding is used. The error is repetitive and is zero when the position of the wiper corresponds with the four axes of symmetry, if the input

* Standard Telephones and Cables, Limited.

phases and amplitudes are correct. There are thus eight zeros in the error curve and the error is termed octantal in the terminology of radio direction finding. The error curve is plotted in Figure 3, and it will be seen that its amplitude is

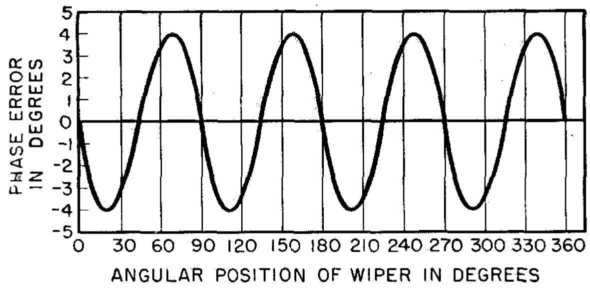


Figure 3—Error curve of the round potentiometer phase shifter.

± 4 degrees. If a phase-shifter having this error is followed by a frequency divider with a division factor of 36, the maximum error in the final phase due to this cause is ≈ 0.11 degree.

This is normally satisfactory for the calibration of most phase-meters, but it can be reduced if desired in several ways.

A. By increasing the division factor of the frequency divider.

B. By feeding the potentiometer at more than four points from a polyphase source.

C. By using a non-uniform winding.

D. By using a mechanical correcting device.[7]

E. By using a square potentiometer or a good phase-shifting transformer.

Photographs of the round potentiometer are shown in Figure 4.

The output of the master oscillator is divided into four symmetrical phases, of equal amplitude, and these outputs are applied in order of phase to the taps of the potentiometer, taken in order. As the wiper is rotated, the voltage appearing thereon progresses or retards in phase by an amount equal to the angle of rotation of the wiper. There is a repetitive error of phase as previously mentioned, but a complete revolution from any initial position shifts the output phase by exactly 360 degrees.

The amplitude of the output varies by 3 decibels as the wiper rotates, as a simple consideration will show, but since the output is to be squared and limited, this is of no consequence.

It has been stated that the output must be continuous. This arises because the output is to be applied to frequency dividers or counters, and an unintentional transient in the applied voltage can cause a counting error. It may be argued that a counting error can cause the final phase to shift only by exact multiples of 360 degrees/n and that unless the phase-meter under test is very inaccurate, it will indicate what phase slip has occurred. However, this effect is to be avoided, since the coarse scale of the phase-shifting potentiometer will give a false indication, and the calibration of a given phase-meter will take longer than it need.

Any slight discontinuity of the output of the phase shifter can be removed by feeding this output to a ringing circuit via a resistance that is

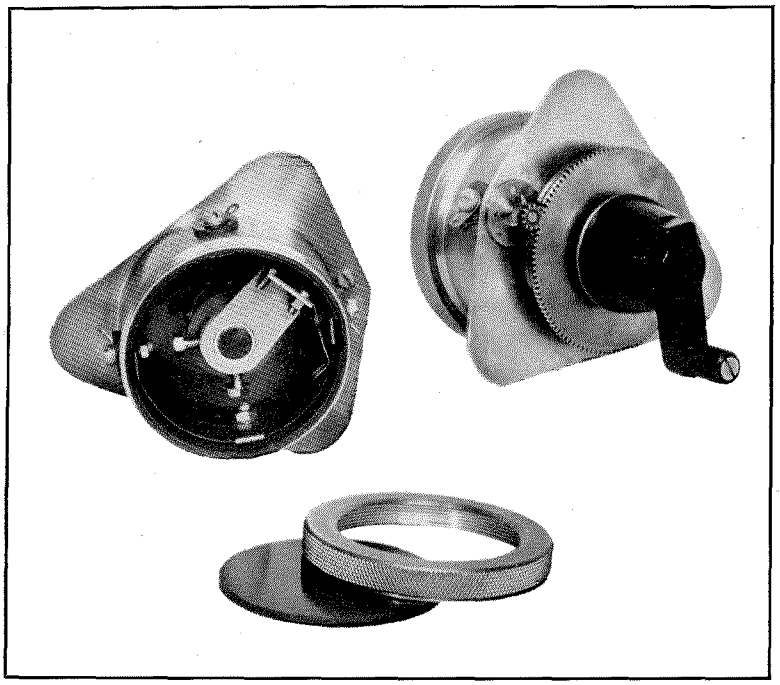


Figure 4—The round potentiometer phase shifter.

sufficiently large to cause negligible load on the potentiometer.

The complete circuit associated with the above description is shown in Figure 5.

The crystal master oscillator is followed by the frequency divider, which provides the correct frequency at the input of the phase-shifting circuit. This frequency divider is not shown, but it can be of any type. It need not be phase-stable since it supplies both output channels. Its output is applied to the transformer *T1*, tuned to 1080 cycles by capacitor *C6* to remove harmonics.

The output phase is shifted by ± 45 degrees, by the phase-shifting bridge *R6*, *R7*, *C7*, *C8*, in which the resistances and reactances are all numerically equal. The voltages at the output points are applied to the grids of two cathode-followers *V3* and *V4*. The cathode loads are two transformers *T2*, *T3* having the secondary centre-taps earthed. The four output leads then represent a symmetrical 4-phase system, and the leads are joined in phase-order to the taps of the main phase-shifting potentiometer *P1*.

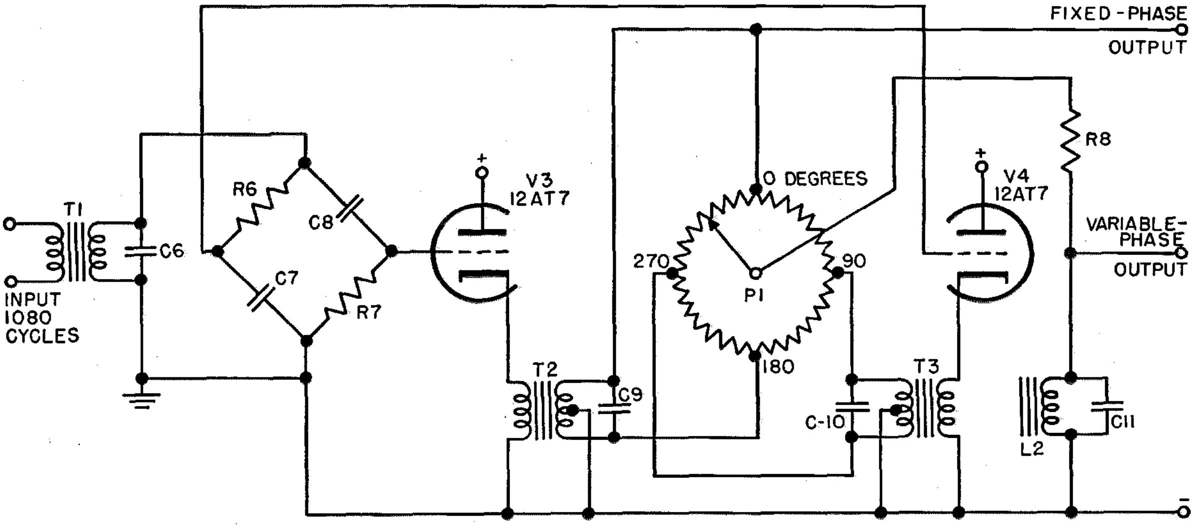


Figure 5—The round potentiometer phase-shifting circuit.

Two outputs are taken to high-impedance circuits, one from a feed point of *P1*, and one from the wiper. The output from the wiper is applied via *R8* to the ringing circuit *L2*, *C11*, tuned to 1080 cycles.

R8 is large compared with the component resistances of *P1*. Also the impedance of the ringing circuit is comparable with *R8*. The variable phase output is taken from the ringing circuit.

A changing phase-shift in the ringing circuit can cause an error of the output phasing. However, since the *Q* of the ringing circuit need not be greater than 5 when in circuit, such accidental phase-shifts will be small. It is clearly preferable to eliminate the need for the ringing circuit by making the output of the phase-shifter perfectly continuous.

3.2 Amplifer and Squarer

The crystal oscillator is followed by a two-stage amplifier and squarer (*V5*, *V6*, Figure 6). This produces at the anode of the second stage a square wave whose peak-to-peak voltage is a large fraction of the high-tension voltage.

3.3 First Frequency Divider (1080 to 180 Cycles)

A modification of the familiar stepping counter[9] has been found to be a very stable frequency divider, but it is not as economical in valves as some other types. In the calibrator that gave rise to this article, stepping counters are used throughout.

The original stepping counter operates as follows, with reference to Figure 6. A fall in voltage at the anode of *V6* causes *C13* to charge through the left-hand diode of *V7*. When the voltage rises, *C13* loses some of its charge through the right-hand diode to the reservoir *C14*. The voltage across *C14* is therefore of 'staircase' form as time progresses, and the envelope is exponential, as shown in Figure 7. The anode voltage of the thyratron is therefore raised and at some point, depending on the grid bias, and, assuming anode triggering of the thyratron in the simple circuit, the operation is clearly more certain.

Since both the peak voltage of the input square wave and the diode bias voltage are proportional to the high-tension voltage, the operation of the

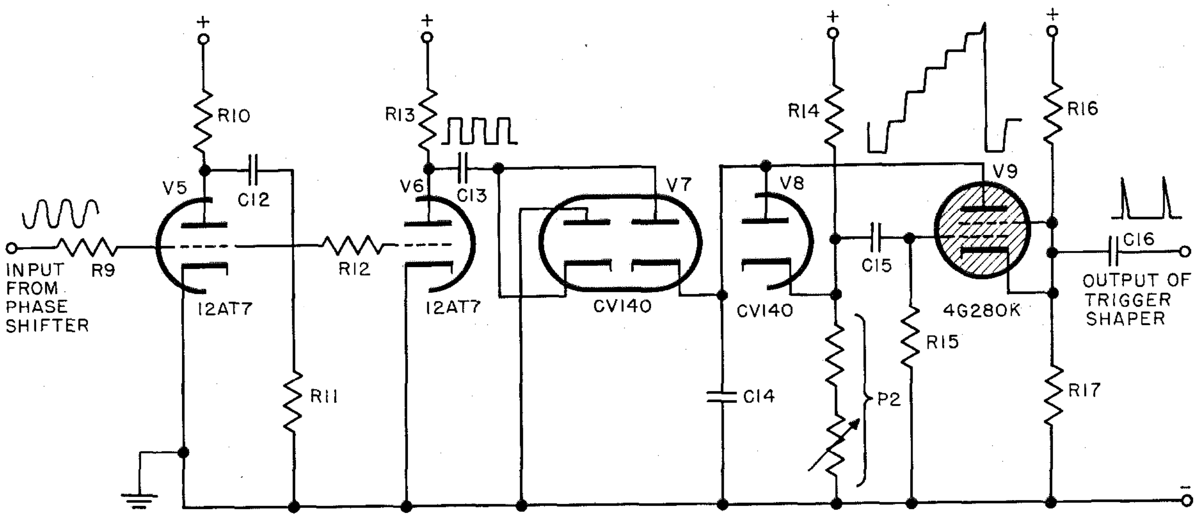


Figure 6—Amplifier squarer and first frequency divider circuit (1080–180 cycles).

C15 disconnected, it will fire. The reservoir capacitor is then discharged and the process repeats.

The division factor is decided by the grid bias of the thyratron and by the ratio *C14/C13*. Thyratrons are not perfectly stable, however, and the voltage required at the anode to cause firing will vary with the high-tension voltage and the ambient temperature, etc. This causes variations of the division factor, except when this is small.

By adding a simple diode gate circuit, the counter can be made very stable for any ambient temperature and for a 2:1 ratio of the high-tension voltage. The components to be added are *V8*, *R14*, *P2*, and *C15*.

The thyratron is biased at the cathode by means of *R16*, *R17*, so that it cannot fire within the expected range of its anode voltage. The voltage across the reservoir *C14* is applied to the anode of diode *V8*. The cathode of *V8* is biased positively by means of *R14* and *P2*. When the voltage across *C14* exceeds this bias voltage, the diode conducts and its cathode voltage is raised. This rising voltage is applied via *C15* to the thyratron grid and causes it to fire. Since this voltage is of the same order as that which causes circuit is unaffected by changes of high-tension voltage over a wide range. In addition, changes of ambient temperature cannot affect the operation since the thyratron does not fire naturally, but is triggered by pulses from an independent circuit.

The division factor can be controlled over a small range by adjusting *P2*. Wider variation may require alteration of the values of *C13* and *C14*. The calculation of the actual values required for a given circuit is covered in detail in the Appendix. The values required for this circuit are deduced as an example.

When the thyratron is triggered, a large posi-

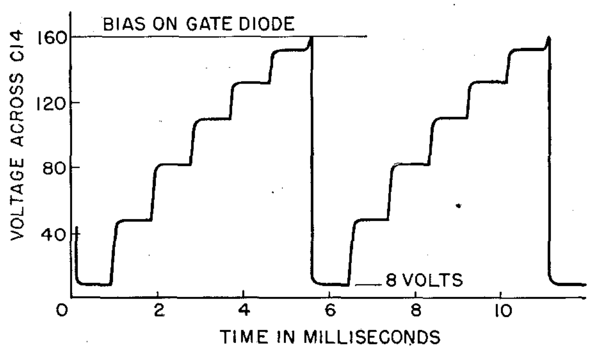


Figure 7—The graph of the voltage across *C14*.

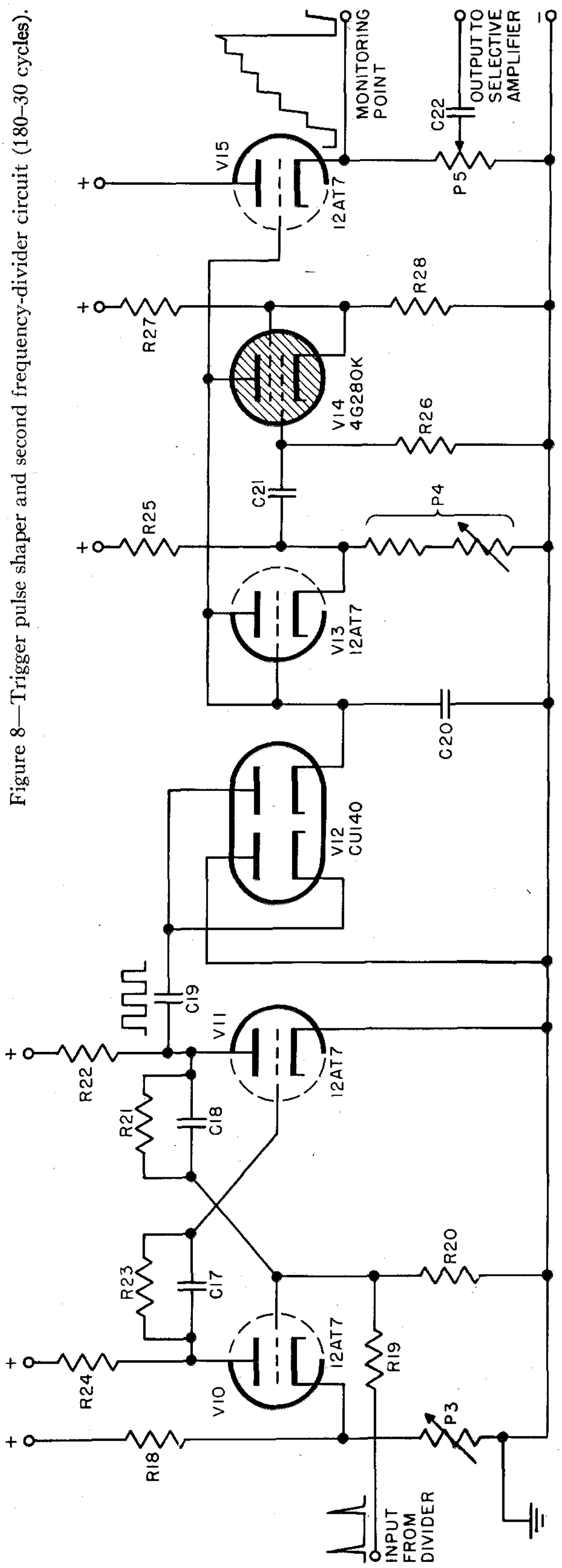


Figure 8—Trigger pulse shaper and second frequency-divider circuit (180–30 cycles).

tive pulse of short duration appears across the cathode bias resistor. This pulse is used as input to the following circuits.

3.4 Pulse–Shaping Trigger Circuit

Two frequency dividers are to be used in cascade to give a division factor of 36. The second divider also requires a square-wave input. At no point in the circuit of the first divider does such a waveform exist, so that a shaping circuit is required. A single-stability multivibrator circuit consisting of *V10*, *V11* is used and is shown in Figure 8. It is designed to convert the short positive pulses at the cathode of the thyratron into a square wave of approximately unity on/off ratio and having the largest possible amplitude. The repetition frequency is 180 cycles. The circuit requires no adjustment when it is once correctly set up. The output is taken from the anode of *V11*.

3.5 Second Frequency Divider (180 to 30 Cycles)

The output of the trigger shaper is applied to the second frequency divider (Figure 8). This is identical with the first except that the charging and reservoir capacitances *C19* and *C20* may be increased by a factor not exceeding 6, since the input frequency is less in this ratio.

3.6 Output Cathode-Follower and Gain Control

Most phase-meters of the type under consideration require the two applied voltages to be sinusoidal. The output from the second divider must therefore be tuned up to remove harmonics and the output waveform of the divider should therefore contain a large fundamental component. The 'staircase' waveform across the reservoir capacitor *C20* is most suitable for this purpose. Direct connection to *C20* is undesirable and connection is made via a cathode-follower *V15* (Figure 8). This requires no grid leak resistor and it has an almost infinite input impedance. A large cathode resistance *P5* must be used to eliminate grid current, since the voltage between the grid and ground rises to half the high-tension voltage approximately.

The presence of the cathode-follower also provides the divider with a low-impedance monitoring point. In addition, it is an advantage when setting up the calibrator to provide the first frequency divider with a similar monitoring facility.

The cathode load of the output cathode-follower is a potentiometer *P5*, which acts as a gain control.

3.7 Selective Amplifier

Harmonics may be removed from the output of the cascaded divider circuits by means of normal tuned circuits of inductance and capacitance. However, at these low frequencies, physically large iron-cored inductors with large inductances are required, and, in any case, these do not make satisfactory tuned circuits. This is because the presence of iron in the circuit produces harmonics, due to hysteresis, and a change of any steady current carried by the windings produces a change of inductance. The inductance varies rapidly with change of level. Also mechanical shocks can move the core and alter the inductance, particularly if the core is gapped.

Due to these disadvantages experience has shown that iron-cored inductors are to be avoided where constancy of phasing is required, except where the inductor is used as the cathode load of a cathode-follower or is fed by any other low-impedance source.

The calibrator under discussion uses an amplifier that is made selective by the inclusion of a frequency-selective resistance–capacitance circuit in the feedback loop. The amplifier is a simplification of published circuits.[10,11]

The voltage appearing at the output of cathode-follower *V15* is applied to a potentiometer pad consisting of *R29* and *R30* (Figure 9). The voltage appearing across *R30* is applied to a cathode-follower *V17*, whose output is applied to a twin-T selective circuit consisting of *C25*, *C26*, *C27*, *R33*, *R34*, and *R35*. This is very accurately adjusted to give zero output at the fundamental frequency. The output of this circuit is applied to the grid of a pentode valve *V16* operating as a high-gain voltage amplifier and having *R30* as anode load. *V16* does not modify the voltage of fundamental frequency appearing across *R30*. However, harmonic voltages appearing across *R30* produce heavy negative feedback, since the twin-T network gives an output at all frequencies except the fundamental. Thus the whole circuit is selective and the selectivity depends on the voltage loss of the cathode-follower, the loss of the twin-T network, and the gain of the feedback amplifier.

Comprehensive information about this type of selective amplifier is given in the original articles, but some design notes will be given. The twin-T circuit must be fed from an impedance that is low compared with its component impedances, and it must be terminated with a very high load impedance. The impedances in the series arms should be double those in the shunt arm. The circuit must be constructed with stable resistors and capacitors and must be very accurately tuned by varying, for example, *R33* and *C27*. To obtain good selectivity, the voltage gain of *V16*

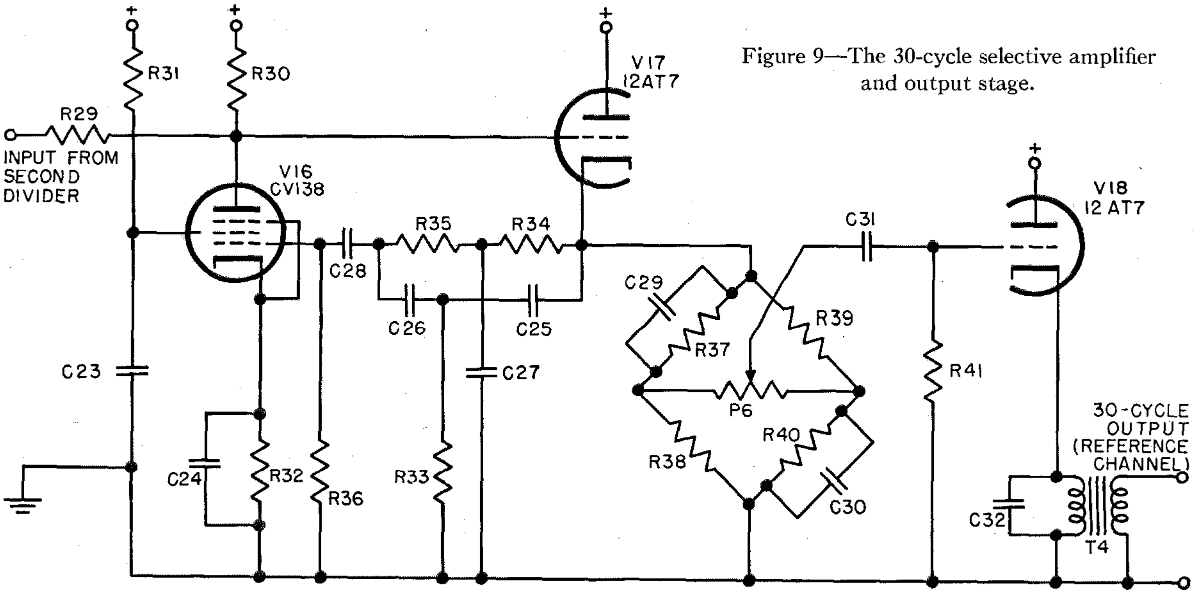


Figure 9—The 30-cycle selective amplifier and output stage.

must be at least 100 times, a requirement which is easily satisfied. The selectivity may be varied by varying this gain.

3.8 Phase–Setting Control

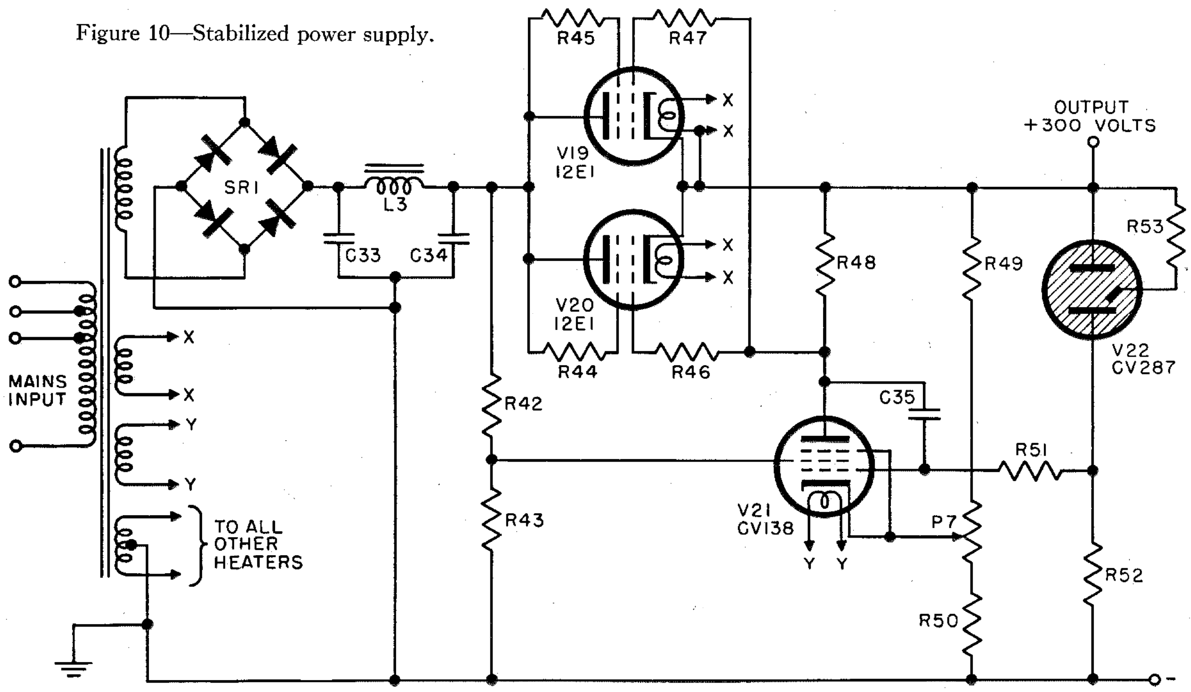
The sinusoidal output of the cathode-follower *V17* is applied to a bridge circuit consisting of *C29*, *R37*, *C30*, *R40*, etc. The potentiometer *P6* allows small phase adjustments to be made. This is a useful facility in setting up the calibrator.

The input is obtained at constant phase, as indicated in Figures 1 and 5; i.e., is not obtained via the round-potentiometer phase shifter. The phase-setting control *P6* and the associated bridge circuit are omitted.

5. Power Supply

Although the frequency-divider circuits described previously are not affected by change of high-tension voltage, a stabilized power supply

Figure 10—Stabilized power supply.



3.9 Output Stage

The output of the phase-shifting bridge is applied to a cathode-follower *V18*, which acts as an output stage. The cathode load is a transformer *T4* tuned to the fundamental frequency. It will be appreciated that this tuned circuit is not selective owing to the low output impedance of the cathode-follower. The reason for tuning is to raise the impedance of the transformer at the fundamental frequency and make negligible the phase shift between grid and cathode voltages. The transformer has an appropriate secondary winding to give a convenient output voltage.

4. Fixed-Phase Channel

The fixed-phase channel is identical with the variable-phase channel, with two exceptions.

is desirable. The final divider and output circuits produce, in the high-tension supply, currents at 40 cycles. Any normal power-pack with filter circuits consisting of inductors and capacitors has an appreciable internal impedance at 40 cycles. This causes interaction between the two channels and a semicircular repetitive error is produced in the output phase difference. A power supply that is stabilized electronically can be made to have negligible or even zero internal impedance at low frequencies. In this case, there is no coupling between the two channels due to this cause, and a suitable circuit is shown in Figure 10. An alternative solution is to use two separate power supplies for the two channels.

It should be mentioned that a semi-circular repetitive error in the output phase difference can also be produced if there is any magnetic coupling between the two output transformers of the two channels, and these transformers must be placed accordingly.

6. Panel Controls

In the calibrator under discussion, it has been found desirable to have the following controls brought out to knobs.

A. The 'round-potentiometer' phase shifter.

B. The gain control for each channel.

C. The subsidiary phase shifter giving a few degrees phase shift for setting the zero of the calibrator.

D. An output switch with the following positions:—

- a. Zero output from both channels.
- b. Zero output from the reference channel.
- c. Zero output from the variable-phase channel.
- d. Both outputs connected normally.
- e. The reference channel connected to both pairs of output terminals.

It is also desirable to have five preset controls brought out for screw-driver adjustment. Each control is associated with a frequency-divider circuit and the monitoring point of each divider is connected to a small terminal on the panel of preset controls. A lead is also taken from the master oscillator to a terminal on this panel.

The five preset controls and the six monitoring points are concealed behind a small cover which is removable. These controls are required only infrequently.

7. Operation of Calibrator

It will be appreciated from the preceding description that the calibrator does not provide an accurate absolute phase difference, but an accurate change of phase difference. This defect is usually unimportant, for most phase-meters are provided with a built-in zero check. It is then only necessary to check the scale for linearity.

Alternatively, using the output switch described in Section 6, the following procedure may be adopted. Using switch position *e*, a common input is applied to both channels of the phase-meter under test. The phase-meter reading is noted. The main phase control of the calibrator is set to zero, and the output switch is set to position *d* for normal operation. The phase-setting control (Section 3.8) is then adjusted until the phase-meter gives the same reading as before. The phase-zero of the calibrator is then set, and the phase-setting control is locked in this position. In these two tests, the amplitude of the input to a given channel of the phase-meter must be the same. Then, the phase-meter acts only as a transfer indicator and introduces no error, since it is working under identical conditions in the two tests.

It may be found convenient to build into the calibrator a simple phase discriminator that can be used in the zero-setting procedure. This has not been done in the equipment described.

A slight modification of the round potentiometer provides for more rapid checking of phase-meters at a number of discrete readings. The four tapping points of the potentiometer are internally connected to four small metal plates. These are arranged so that even if the wiper is not positioned accurately to line up with a tapping point, it still makes electrical contact with the tapping point; i.e., the tapping point is extended along the line of motion of the wiper. By this means, it is not necessary to ensure that the wiper is set exactly to correspond with a tapping point, and the speed of operation is increased. It will be obvious that the phasing error of the potentiometer at intermediate points will be increased. However, if spot checks at intervals of 10 degrees or 2·5 degrees prove sufficient, then this modification is allowable, since these spot checks are obtained with the wiper set to correspond with the four tapping points only.

This modification may be carried a stage further, with the latter proviso. The phase-shifting potentiometer may be replaced by an 8-position rotary switch, in which the rotor can rotate continuously. The distributed resistive winding of the round potentiometer is replaced by a ring of 8 lumped equal resistances connected between the eight contact points of the switch. This crude potentiometer is fed at four symmetrical points as before. The wiper must make continuous contact with the studs, so it is necessary for it to short-circuit adjacent studs as it crosses from one to the next. It is for this reason that 8, and not 4, resistances and studs are used. As the wiper rotates, the output phase advances in steps. However, the phase is correct at the 4 feed points and this is all that is required if the calibrator is to be used to provide spot checks only.

8. Conclusions

A calibrator that was constructed in the manner described has proved highly successful and has been in frequent use for one year. It has proved invaluable in the investigation of a phasemeter of the differential-detector type.

The principle of operation at a harmonic frequency is general and has a number of important applications. As an illustration, the method of phase measurement using a harmonic frequency may be mentioned again.[5] Suppose a system involves a phase measurement. The phase measurement may be performed at the fundamental frequency, in which case the accuracy is not large. Harmonic frequencies may be generated and these may be used in the phase measurement. However, whether this will give any improvement in accuracy is doubtful, for, although the accuracy of the phase measurement itself is increased, the frequency-multiplication process may introduce serious errors. The remedy is to design the system as a whole so that it generates the harmonic frequencies inherently. This principle has already been applied successfully, in the design of omni-directional beacons.

9. Acknowledgments

Acknowledgments are due to Mr. C. W. Earp for the original idea, to Mr. H. W. Hawkes for assistance with the experimental work, and to Mr. T. J. Cox who was responsible for the design and construction of the complete phase-shifting potentiometer.

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11. Appendix—Calculation of the Component Values for Frequency Dividers[12, 13]

The relevant section of Figure 6 is reproduced in Figure 11. The charge and discharge sections of *V7* will be called *V7C* and *V7D*.

Since the square-wave input is applied via capacitor *C13*, the direct-current level of the input voltage is unimportant. It will be assumed, for simplicity, that the input square wave has zero voltage initially, and that it reaches a steady voltage E on alternate half cycles.

The discharge of *C14* by the thyratron is incomplete, and *C14* remains charged to a voltage v. Initially, *C13* will not be charged, but at the commencement of subsequent cycles of the 'staircase' waveform, it will have some charge. This latter case will be considered.

Initial state: Input voltage $= 0$

$$\text{Charge on } C_{13} < C_{13}.E$$
$$\text{Charge on } C_{14} = C_{14}.v = q.$$

1st negative half cycle, Figure 11B.

The input voltage falls to $-E$. V7*C* conducts and acts as a low resistance. *C*13 therefore acquires a charge $C_{13}.E = Q$, in the sense shown.

1st positive half cycle, Figure 11C.

The input voltage rises to zero and *V7D* conducts. The input circuit and *V7D* act as resistances only. *C13* and *C14* therefore share their charges until they reach equal voltages; i.e., the total charge divides in the ratio of the capacitances.

Charge on $C_{14} = (Q + nQ + n^2Q + n^2q)n$.
Voltage across $C_{14} = e_3$

$$= (Q + nQ + n^2Q + n^2q)n/C_{14}.$$

pth positive half cycle

It is now clear that during the pth positive half cycle, C_{14} will have a charge:

$$(nQ + n^2Q + n^3Q + \ldots \text{ to } p \text{ terms}, + n^p.q).$$

Voltage across $C_{14} = e_p$

$$= \frac{(nQ + n^2Q + n^3Q + \ldots \text{ to } p \text{ terms} + n^pq)}{C_{14}}.$$

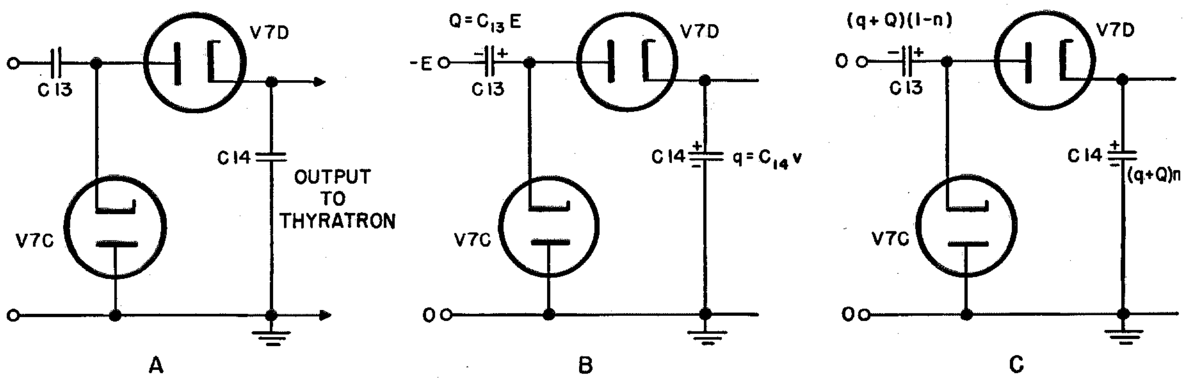

Figure 11.

Total charge $= Q + q$

Charge on $C_{14} = (Q + q)\dfrac{C_{14}}{C_{13} + C_{14}} = (Q + q).n,$

where $n = \dfrac{C_{14}}{C_{13} + C_{14}}.$

Voltage across $C_{14} = e_1 = \dfrac{(Q + q)n}{C_{14}}.$

2nd negative half cycle

The charge on *C13* is made up to Q when V7*C* conducts.

2nd positive half cycle

The charges are shared.

Total charge $= Q + nQ + nq$.
Charge on $C_{14} = (Q + nQ + nq)n$.
Voltage across $C_{14} = e_2 = (Q + nQ + nq)n/C_{14}$.

3rd negative half cycle

The charge on C_{13} is made up to Q.

3rd positive half cycle

The charges are shared.
Total charge $= Q + nQ + n^2Q + n^2q$.

Completion of the Cycle

The thyratron is triggered during the rise of a step (i.e., during a positive half cycle of the input square wave) when *C13* has already lost to *C14* some of its charge $C_{13}.E$.

The thyratron discharges *C14* to a voltage v, so that the initial conditions are regained. The cycle repeats.

If *C14* is not discharged, but continues to be charged, its final voltage can be found thus:

Voltage across *C14* after p steps $= e_p$

$$= \frac{nQ + n^2Q + n^3Q \ldots \text{ to } p \text{ terms} + n^p.q}{C_{14}}.$$

Since $n < 1$, $n^p \to 0$ as $p \to \infty$

$$\therefore \text{ sum to infinity} = \frac{Q}{C_{14}}\frac{n}{1 - n}$$

$$= \frac{C_{13}E}{C_{14}}.\frac{C_{14}}{C_{13} + C_{14}}.\frac{C_{13} + C_{14}}{C_{13}} = E,$$

i.e., the voltage across *C14* rises to the peak-to-peak voltage of the input square wave.

Height of the pth Step

$$e_p = \frac{(nQ+n^2Q+n^3Q+\ldots\text{to } p \text{ terms}, +n^pq)}{C_{14}}$$

$$e_{p-1} = \frac{(nQ+n^2Q+n^3Q+\ldots\text{to } p-1 \text{ terms}, +n^{p-1}.q)}{C_{14}}$$

$$\therefore e_p - e_{p-1} = \frac{1}{C_{14}}.\{Qn^p - qn^{p-1}(1-n)\}.$$

Substituting for Q and q in terms of E and v

$$\begin{aligned} e_p - e_{p-1} &= \frac{C_{13}En^p}{C_{14}} - \frac{C_{14}vn^{p-1}.(1-n)}{C_{14}} \\ &= E(1-n)n^{p-1} - v(1-n)n^{p-1} \\ &= (E-v)(1-n)n^{p-1}. \end{aligned}$$

Therefore, the heights of successive steps are in geometric progression, so that the envelope of the charging curve is exponential.

Height of the first step $= (E - v)(1 - n)$.

This shows the physical significance of n. Initially, the voltage across *C14* is v, and after an infinite number of steps, the voltage is E. Therefore, the complete staircase waveform extends over a voltage range $(E - v)$. Thus $(1 - n)$ is the height of the first step expressed as a fraction of $(E - v)$, and n is the fractional height remaining, to be occupied by the charging curve after an infinite number of steps.

Condition for the Height of the pth Step to be a Maximum

If the thyratron is to fire during the pth step, greatest stability of operation is obtained when this step has maximum height.

$$\begin{aligned} \therefore \frac{d}{dn}(e_p - e_{p-1}) &= 0 \\ &= (E-v)\{(1-n)(p-1)n^{p-2} - n^{p-1}\} \\ \therefore p - 1 - np + n - n &= 0 \\ \therefore n &= \frac{p-1}{p} \\ \therefore 1 - n &= \frac{1}{p} = \frac{C_{13}}{C_{13}+C_{14}}. \end{aligned}$$

Therefore, to make the height of the pth step a maximum, the fractional height of the first step must be $1/p$. Then—

Maximum height of the pth step

$$\begin{aligned} &= (E-v)(1-n)n^{p-1} \\ &= (E-v)\frac{1}{p}\left(\frac{p-1}{p}\right)p-1 \\ &= (E-v)\frac{(p-1)^{p-1}}{p^p}. \end{aligned}$$

Bias Required on the Gate Diode

When the circuit is correctly adjusted to have a division factor p, the thyratron is triggered when the mid-point of the pth step is reached. To find the voltage across *C14* corresponding to this point, it is necessary to know the voltage across *C14* after the $(p - 1)$th and pth steps.

Voltage across *C14* after p steps $= e_p$

$$\begin{aligned} &= \frac{(nQ+n^2Q+n^3Q\ldots\text{to } p \text{ terms}+n^pq)}{C_{14}} \\ &= \frac{nC_{13}E}{C_{14}}(1+n+n^2+n^3+\ldots\text{to } p \text{ terms})+\frac{n^pC_{14}v}{C_{14}} \\ &= nE\frac{(C_{13}+C_{14}-C_{14})}{C_{14}}\,\frac{(1-n^p)}{1-n}+n^pv \\ &= nE\left(\frac{1}{n}-1\right).\frac{(1-n^p)}{1-n}+n^pv \\ &= E(1-n^p)+n^pv \\ &= (E-v)(1-n^p)+v. \end{aligned}$$

The pth step is to have maximum height

$$\therefore n = \frac{p-1}{p}$$

$$\therefore e_p = (E-v)\left\{1 - \left(\frac{p-1}{p}\right)^p\right\} + v.$$

Similarly

$$e_{p-1} = (E-v)\left\{1 - \left(\frac{p-1}{p}\right)^{p-1}\right\} + v.$$

At the mid-point of the pth step

Voltage across $C_{14} \times \frac{1}{2}(e_p + e_{p-1})$

$$\begin{aligned} &= \frac{E-v}{2}.\left\{2-\left(\frac{p-1}{p}\right)^p-\left(\frac{p-1}{p}\right)^{p-1}\right\}+v \\ &= \frac{E-v}{2}.\left\{2-\frac{(p-1)^{p-1}.(p-1+p)}{p^p}\right\}+v. \end{aligned}$$

If a modern thyratron, having a large control ratio, is used then the triggering pulse required is very small and can be neglected. With this approximation, the above equation also represents the bias required on the gate diode.

$\therefore$ Bias required on the gate diode

$$= v + (E - v)\left\{1 - \frac{(p-1)^{p-1}(2p-1)}{2p^p}\right\}.$$

The first frequency divider of the calibrator will be considered as an example.

C13 must be capable of charging and discharging fully in a time equal to the half-period of the input square wave. Neglecting the differential anode resistances of the diodes *V7C* and *V7D*, the larger resistance involved in the time constant is the anode load *R13* of the second squaring valve *V6*, and is obtained when *V6* is non-conducting.

Source resistance *R13* = 68 kilohms.
Input frequency = 1080 cycles.
Half-period of input square wave = 463 microseconds.

Suppose the time-constant $C_{13}R_{13}$ is 10 per cent of this half-period

$$C_{13}R_{13} = 0\cdot1 \times 463 \times 10^{-6} = C_{13} \times 68 \times 10^3 \text{ seconds}$$

$\therefore C_{13} = 680$ micromicrofarads.

The division factor $p = 6$

$$\frac{1}{p} = \frac{C_{13}}{C_{13} + C_{14}}$$

$$\therefore p = 1 + \frac{C_{14}}{C_{13}}$$

$$\therefore \frac{C_{14}}{C_{13}} = 5$$

$\therefore C_{14} = 3400$ micromicrofarads.

In practice, it may be preferable to use the round values 700 and 3500 micromicrofarads for *C13* and *C14*, and stable capacitors with mica dielectrics should be used.

In the circuit under consideration, the high-tension voltage is 300 volts, and the anode load of the second squaring valve is 68 kilohms. From the characteristics of the valve type *12AT7* used, it can be found that the square wave obtained will have a peak-to-peak voltage of 250 volts, approximately. Also, the thyratron *4G280K* becomes non-conducting when its anode voltage is less than 8 volts.

$$\therefore E = 250 \text{ volts}$$
$$v = 8 \text{ volts.}$$

$$\text{Height of 6th step} = (E - v)\frac{(p-1)^{p-1}}{p^p}$$
$$= 242 \times \frac{5^5}{6^6} = 16\cdot4 \text{ volts.}$$

Bias on the gate diode $= v + (E - v)$.

$$\left\{1 - \frac{(p-1)^{p-1}(2p-1)}{2p^p}\right\}$$
$$= 9 + 242\left\{1 - \frac{3150 \times 11}{2 \times 46500}\right\} = 160 \text{ volts.}$$

This bias is obtained by means of a resistive potentiometer across the high-tension supply. If the resistance *R14* to positive high tension is 100 kilohms,

$$\text{Resistance to earth} = \frac{160}{300 - 160} \times 100 \text{ kilohms}$$
$$= 114 \text{ kilohms.}$$

This is shown as *P2* in Figure 6, and can best be made up with a fixed resistance of 90 kilohms, in series with a 50-kilohm potentiometer connected as a variable resistance. All these components should be wire wound.

The grid bias on the thyratron is $-2\cdot5$ volts corresponding to a critical anode voltage of 250 volts. For a critical anode voltage of 160 volts, a grid bias of $-2\cdot1$ volts is required. Thus, when the anode–cathode voltage reaches 160 volts, triggering is caused by a grid pulse of $+0\cdot4$ volts only. This is a very small fraction of the height of the 6th step, so that the circuit is reliable.

The second frequency divider works under identical conditions, but the input frequency is lower by a factor of 6. Thus, the charging and reservoir capacitances *C19* and *C20* may be increased by this factor, although the circuit will operate satisfactorily at the lower frequency with no change.

Monitoring with Cathode-Ray Oscilloscope

As mentioned in Section 3.6, connection to *C14* or *C20* should only be made via a cathode-follower. The picture obtained on the cathode-ray oscilloscope screen will include a small spike, as shown in Figure 7. When the circuit is in correct adjustment, this spike should be half the height of the 6th step. However, the sixth step does not appear in full, so that an estimate must be made from the height of the preceding step.

Tunable Waveguide Filters*

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HERETOFORE, tunable waveguide filters have consisted of cavities connected with quarter-wavelength coupling lines and tuned by means of a lumped reactance in the cavity. Such filters exhibit asymmetrical frequency response and a bandwidth proportional to the cube of frequency when tuned.

A different method of tuning is to maintain constant guide wavelength as the filter is tuned. A filter so tuned will exhibit symmetrical pass-band response and essentially constant bandwidth.

Various methods of varying guide wavelength are investigated, including A) changing the guide width, B) inserting a dielectric strip in the broad face of the guide, and C) inserting a metal strip in the broad face of the guide.

Measurements on a three-section filter tuned by a dielectric strip show that almost constant bandwidth with a straight-line-frequency tuning curve can be obtained over at least a 12-percent frequency range.

• • •

Microwave filters are used to minimize interference from transmissions on unwanted frequencies. The design of microwave filters is covered by Fano and Lawson.[1] A good exposition of the design methods applied to maximally flat band-pass filters and references to other work are given by Mumford.[2] Band-pass filters can be designed and built to give the desired characteristics at one frequency. Mumford,[2] for example, has built a fifteen-cavity fixed-tuned filter that worked very well. Tunable filters, however, that preserve their design characteristics have not been built. This paper gives methods of designing satisfactory tunable filters.

1. Discussion

Two fundamental methods of changing the resonant frequency of a waveguide cavity have been investigated: A) The insertion of a lumped susceptance in the cavity, and B) varying the parameters of the waveguide in such a manner that the guide wavelength at resonance is constant.

1.1 Lumped Susceptance

The insertion of a variable lumped susceptance in a cavity will alter the resonant frequency of the cavity.

This method of tuning is subject to several severe limitations. The bandwidth of such a cavity constructed from capacitive or inductive irises is

$$\left.\begin{aligned} BW_{cap} &\doteq \frac{2c}{\pi l}\frac{\lambda_0\lambda_{g0}}{X^2}\left(1-\frac{2\lambda_{g0}^2}{X^2}\right)\doteq K/f^2 \\ BW_{ind} &\doteq \frac{2c}{\pi l}\frac{\lambda_0}{Y^2\lambda_{g0}^3}\left(1-\frac{2}{Y^2\lambda_g^2}\right)\doteq Kf^4 \end{aligned}\right\}. \quad (1)$$

(Symbols are defined in Section 7.) The superiority of capacitive irises for tunable filters has been demonstrated by Smullin.[3]

A further limitation is that the coupling lines between cavities are not tuned. As a result, the pass-band response becomes more and more asymmetrical as the filter is tuned further from its design frequency.[4]

The limitations on the tuning-post method are such that the filter can only be employed over a narrow frequency range. A three-section inductive-iris filter operating in the 5000-megacycle region had a bandwidth that increased 26

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† Mr. Augenblick is now with Microlab, South Orange, New Jersey.

[1] G. L. Ragan, "Microwave Transmission Circuits," McGraw-Hill Book Company, New York, New York; 1948: chapters 9 and 10.

[2] W. W. Mumford, "Maximally-flat Filters in Waveguide," *Bell System Technical Journal*, v. 27, pp. 684–713; October, 1948.

[3] L. D. Smullin, "Design of Tunable Resonant Cavities with Constant Bandwidth," Technical Report 106, Massachusetts Institute of Technology, April 5, 1949. (For abstract, see *Proceedings of the I. R. E.*, v. 37, p. 1442; December, 1949.)

[4] See page 694 of footnote reference 1.

percent for a 4-percent change in resonant frequency. The percentage change for a filter constructed from capacitive irises is somewhat less but nevertheless sufficient to restrict the operating range of the filter.

deviation from constant bandwidth is generally not too serious, however.

A further advantage of the constant-guide-wavelength filter is the fact that the coupling lines are exactly tuned over the entire frequency range. Thus, the pass-band response will not suffer from the asymmetrical response of the tuning-post filter.

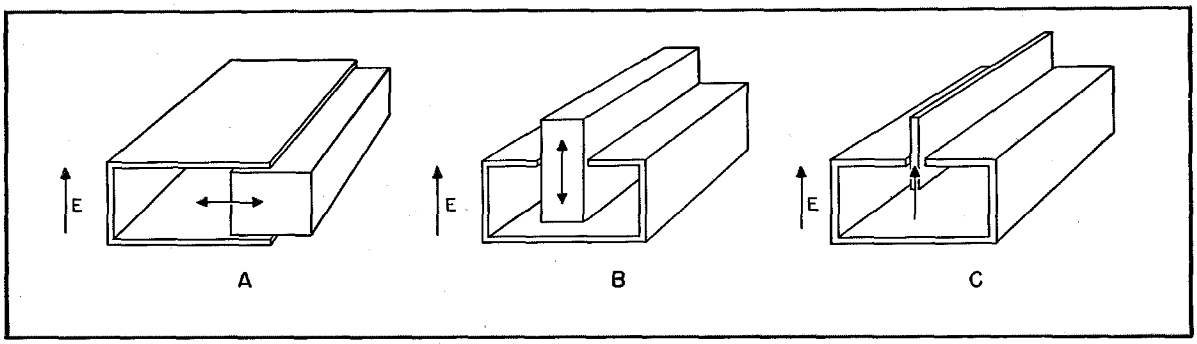

Figure 1—Methods of changing guide wavelength. *A*—Movable wall. *B*—Variable-depth dielectric strip. *C*—Ridge waveguide.

1.2 Constant Guide Wavelength

Most properties of a dispersive waveguide system depend on the guide wavelength and not on frequency directly. For example, the electrical length of a waveguide line is $2\pi l/\lambda_g$ and not $2\pi l/\lambda$; the susceptance of a capacitive iris varies as $1/\lambda_g$ and not as frequency. This principle can be used in the design of many tunable waveguide systems, but will be applied here only to the design of filters.

The doubly loaded Q_{λ_g} of a simple cavity is

$$Q_{\lambda_g}=\frac{\lambda_{g0}}{\Delta\lambda_g}=\left(\frac{\pi l}{2\lambda_g}\right)(B^4+4B^2)^{1/2}. \tag{2}$$

This equation shows that if a cavity has a certain Q_{λ_g} at one guide wavelength and one frequency it will have the same Q_{λ_g} at a different frequency if the guide wavelength is the same, provided that B is not changed.

The fact that Q_{λ_g} is constant over the band unfortunately does not mean that Q on a frequency basis is constant.

$$Q_{\lambda_g}=\frac{\lambda_{g0}}{\Delta\lambda_g}=\frac{\lambda_0}{\Delta\lambda}\left(\frac{\lambda}{\lambda_g}\right)^2=\frac{f_0}{\Delta f}\left(\frac{\lambda}{\lambda_g}\right)^2=Q_f\left(\frac{\lambda}{\lambda_g}\right)^2. \tag{3}$$

Thus, a system with a constant Q_{λ_g} has a Q that varies as the square of the frequency. Usually a constant bandwidth is desired, which means a Q_f proportional to frequency. This

2. *Methods of Changing Guide Wavelength*

2.1 Variable-Width Waveguide

The guide wavelength for the dominant mode in a rectangular waveguide is given by

$$\lambda_g=\frac{\lambda}{[1-(\lambda/2a)^2]^{1/2}}. \tag{4}$$

Thus, the guide wavelength may be held constant by adjusting the guide width for each resonant frequency. One of the narrow walls of the waveguide is removed and replaced with a movable metal strip that makes good contact with the walls and the irises, as in Figure 1A.

The susceptance of a capacitive iris is not a function of the width of the waveguide. Hence, this particular iris may be employed in the variable-width filter. This is not true of other geometries, such as the inductive iris. The end susceptance and impedance mismatch of the tuning section may be minimized by designing the tuning section to have normal guide dimensions in the center of the desired band.

The major limitation of such a filter is the difficulty of making good and uniform contact between the movable wall and the rest of the system.

2.2 Dielectric Strip

Another method of changing the guide wavelength is to insert a dielectric strip into a slot cut in the broad face of the guide, Figure 1B. The guide wavelength at a fixed frequency is a minimum when the strip extends all the way across the guide, the wavelength increasing as the strip is retracted. The filter is designed to operate properly at the high-frequency end of the band when the strip is completely out of the guide. To tune the filter to a lower frequency, the dielectric strip is inserted to such a depth that the guide wavelength is the same as the design guide wavelength. The dielectric strip is made wide enough so that the guide wavelength with the strip almost all the way across the guide at the lowest desired frequency is the same as the design guide wavelength. The proper width can be determined from an equation derived by Frank.[5]

The end susceptance and impedance mismatch of the tuning section is low, and matching transformers are quite simple. The filter may suffer from comparatively high losses unless the slot is shielded properly. This particular filter appears to have the most merit of all systems tested.

2.3 Ridge Waveguide

A ridge waveguide[6] can be used to change the guide wavelength in the filter, Figure 1C. The principle is the same as that discussed above for the dielectric strip. While any width of ridge can be used, a very thin wedge is most suitable because radiation from a narrow slot is small. The characteristic-impedance mismatch between the tuning section and the normal waveguide is high and matching transformers are difficult. Otherwise, the method of tuning is satisfactory.

3. *Constant-Bandwidth Filters*

A good approximation to a constant bandwidth can be obtained by using half- or quarter-wavelength transformers between the constant-guide-wavelength filter and the rectangular waveguide.

Figure 2A shows schematically a filter with matching sections between the filter proper and the normal waveguide. Figure 2B is the equivalent circuit of a filter made of three identical sections. B is the equivalent susceptance of each filter section and is given by

$$B \cong 4Q\frac{\Delta\lambda_g}{\lambda_{g0}} \cong 4Q\left(\frac{\lambda_g}{\lambda}\right)^2\frac{df}{f_0}, \quad (5)$$

where

Q = guide wavelength Q of one section
df = deviation from resonant frequency.

For three sections, this becomes

$$B = 1.526\left(\frac{f}{f_0}\right)\left(\frac{2df}{w}\right), \quad (6)$$

where

f_0 = design frequency
w = bandwidth of whole filter at design frequency.

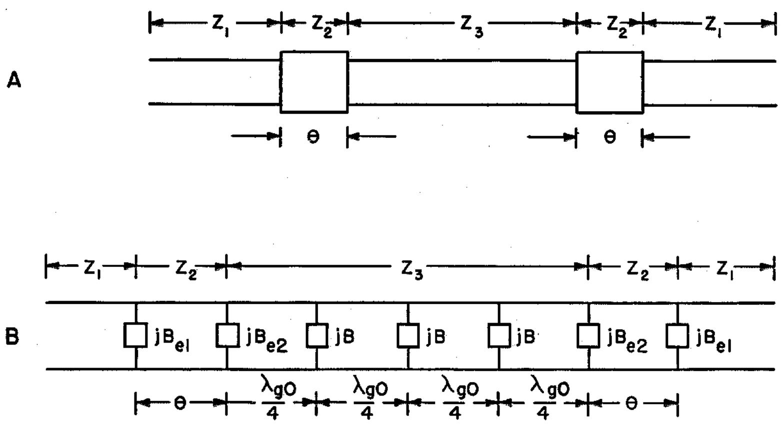

Figure 2—Filter with matching sections. *A*—Schematic of filter. *B*—Equivalent circuit.

The characteristic impedance of a waveguide is given by

$$Z_0 = K\frac{b}{a}\frac{\lambda_g}{\lambda}. \quad (7)$$

[5] C. G. Montgomery, R. H. Dicke, and E. N. Purcell, "Principles of Microwave Circuits," McGraw-Hill Book Company, New York, New York, 1948: page 387.
[6] S. B. Cohn, "Properties of Ridge Waveguide," *Proceedings of the I. R. E.*, v. 35, pp. 783–788; August, 1947.

The insertion-loss ratio (LR) at frequencies near the resonant frequency is given in (8), assuming that B_{e1} and B_{e2} are equal to zero, that the $\lambda_{g0}/4$ coupling lines and the matching sections are a constant electrical length, and that the susceptances of the irises and the characteristic impedances are constant.

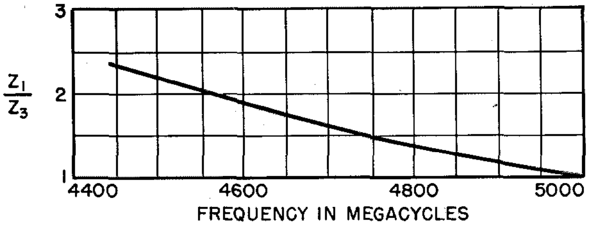

Figure 3—Mismatch between ridge guide and normal guide. $Z_1/Z_3=\lambda_{g1}/\lambda_{g3}$.

$$LR=1+\frac{B^2}{4}\left[\left(\frac{Z_1}{Z_3}\right)^2(B^2-2)+\left(\frac{Z_3}{Z_1}\right)^2\right]^2 \tag{8A}$$

$$LR=1+\frac{B^2}{4}\left[\left(\frac{Z_2}{Z_1}\right)^2\left(\frac{Z_2}{Z_3}\right)^2(B^2-2) + \left(\frac{Z_1}{Z_2}\right)^2\left(\frac{Z_3}{Z_2}\right)^2\right]^2. \tag{8B}$$

Equation (8A) is for matching sections a half-wavelength long and (8B) is for matching sections a quarter-wavelength long. If (Z_3/Z_1) equals the ratio of the guide wavelength in the filter to the guide wavelength in the rectangular guide (as is the case when the filter is tuned with a dielectric strip), (6) and (8A) show that for a 12-percent shift in the resonant frequency the bandwidth increases 16 percent. If in addition $Z_2=(Z_1Z_3)^{1/2}$, (6) and (8B) show that the bandwidth is directly proportional to the ratio of the design frequency to the resonant frequency. If (Z_3/Z_1) equals the ratio of the guide wavelength in the filter to the guide wavelength in the rectangular guide and Z_2 equals Z_3, (6) and (8B) show that almost constant bandwidth can be obtained.

4. Experimental Results

4.1 Movable-Wall Filter

A cavity made of two capacitive irises with half-wavelength transformers on each end was built in a 2-inch by 1-inch by 0.064-inch-wall waveguide (*RWR 187* or *RG-49/U*) and tested. The wall of the half-wavelength transformers, as well as the wall of the cavity, was made movable. When the width of the waveguide was 1.872 inches, the center frequency was 4325 megacycles and the bandwidth between 3-decibel points was 91 megacycles. When the width was 1.42 inches, the center frequency was 5073 megacycles and the bandwidth was 125 megacycles. The loaded Q is

$$Q_f=Q_{\lambda g}\frac{a_3}{a_1}\left(\frac{\lambda_{g3}}{\lambda}\right)\left(\frac{\lambda_{g1}}{\lambda}\right). \tag{9}$$

This equation predicts a ratio of bandwidths equal to 1.46, whereas a ratio of 1.37 was obtained experimentally.

4.2 Ridge Waveguide

Calculations were made to determine to what depth a metal strip $\frac{1}{32}$-inch wide must be inserted to obtain a constant guide wavelength in the ridge guide at frequencies between 4400 and 5000 megacycles. The constant guide wavelength was the same as the guide wavelength in a normal 2-inch by 1-inch guide at 5050 megacycles. The relative characteristic impedance Z_1/Z_3 of the ridge waveguide, determined by measuring the relative impedance (in normal 2-inch by 1-inch guide) at the face of the ridge, is shown in Figure 3. Since the large change in characteristic impedance would make the design of matching sections difficult, no filters were constructed of ridge waveguides.

4.3 Dielectric-Strip Filter

Measurements were made to determine to what depth a $\frac{1}{4}$-inch-thick polystyrene strip must be inserted to maintain the guide wavelength constant by placing them on the bottom of a slotted section of waveguide, the load end of which was short-circuited. The guide wavelength was determined by measuring the distance between successive minima. The depth required to maintain the guide wavelength constant at

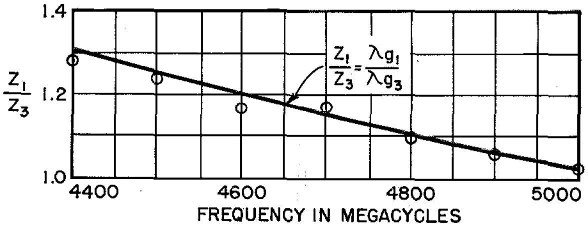

Figure 4—Mismatch between dielectric-strip guide and normal guide.

7.48 centimeters versus frequency was determined (7.48 centimeters is the guide wavelength at 5100 megacycles in a rectangular waveguide 1.872 inches wide). The depth-versus-frequency

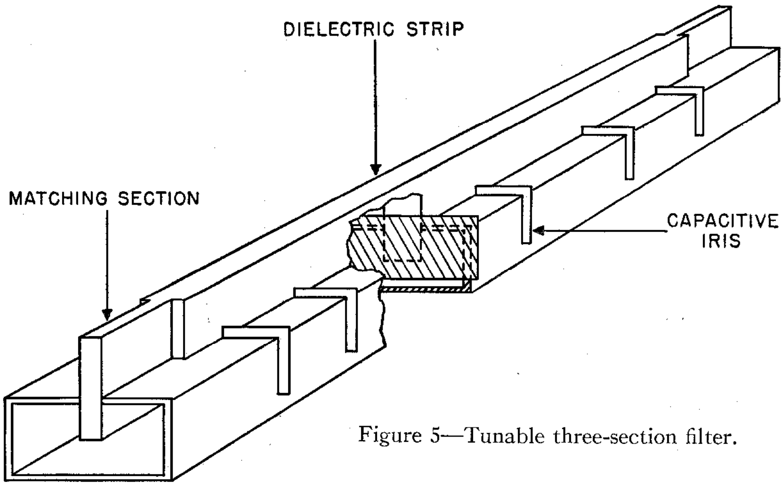

Figure 5—Tunable three-section filter.

curve was essentially a straight line between 4400 and 5000 megacycles.

The relative characteristic impedance Z_1/Z_3 of the partially filled waveguide, determined by measuring the relative impedance at the face of the strip, is shown by circles on Figure 4. Also shown on Figure 4 is the variation of Z_1/Z_3 with frequency, assuming that (7) holds.

A three-element filter, shown in Figure 5, was designed to resonate at 5050 megacycles, with a cavity bandwidth of 84 megacycles so that the bandwidth of the whole filter would be 65 megacycles.[7] The measured standing-wave ratio in the pass band with the filter tuned to three different frequencies is shown in Figure 6. Also shown there is the resonant frequency versus depth of dielectric. For this measurement, a quarter-wavelength transformer with $Z_2=(Z_1Z_3)^{1/2}$ was used.

At 4700 megacycles, the midband insertion loss was 2 decibels due to radiation from the poorly shielded slot. At 50 and 100 megacycles off the resonant frequency, the insertion loss was 21.5 and 42.5 decibels, respectively. The theoretical insertion losses at these frequencies are 22.5 and 41.5, respectively. These measurements were made with a relatively crude model constructed to test the principle. Separate pieces of dielectric were used in the cavities, quarter-wavelength coupling lines, and matching sections to allow tuning each section to compensate for constructional errors. The shift-of-minimum method[8] was used to align the filter sections at each resonant frequency. The depth of dielectric in the three cavities varied slightly. The greatest variation, ±0.016 inch from the average, was at 4375 megacycles. The depth of dielectric in the quarter-wave coupling lines was consistently less than the depth in the cavities. The greatest variation, 0.083 inch, was at 4950 megacycles. It is believed that these variations are due to constructional errors and the

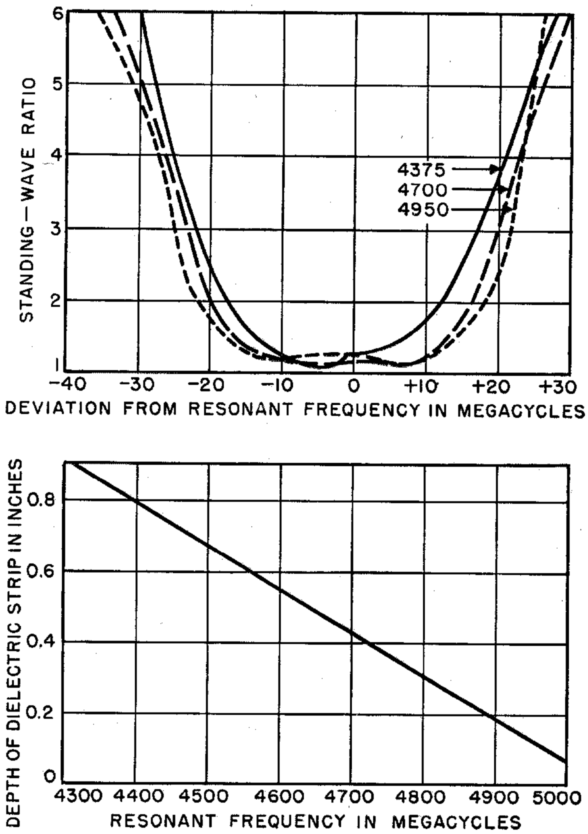

Figure 6—Response of three-element filter. At top, standing-wave ratio is plotted against deviation from the frequencies indicated in megacycles. Below, the slope of the linear curve is 0.85 megacycle per 0.001 inch.

[7] See page 682 of footnote reference 1.

[8] See page 714 of footnote reference 1.

method of alignment and are not inherent in the method of tuning.

5. *Conclusions*

The theory and measurements show that a tunable waveguide filter with almost constant bandwidth can be built by using a variable-depth dielectric strip to keep the guide wavelength in the filter constant as the resonant frequency is changed. Single-knob tuning with a linear frequency scale is possible. These features, not possessed by other types, are highly desirable in some systems.

6. *Acknowledgment*

Acknowledgment is due to D. J. LeVine and H. Seidel for making some of the measurements reported in this paper.

7. *Glossary*

a = waveguide width
a_3 = waveguide width in filter section
b = waveguide height
B = normalized susceptance of iris
B_{e1} = normalized end susceptance between transformer and normal guide
B_{e2} = normalized end susceptance between transformer and filter section
BW = frequency bandwidth
C = velocity of light
K = a constant
l = iris separation
Q_f = loaded Q on a frequency basis
$Q_{\lambda g}$ = loaded Q on a guide-wavelength basis
Z_1 = characteristic impedance of normal guide
Z_2 = characteristic impedance of matching section
Z_3 = characteristic impedance of filter section
λ = air wavelength
λ_0 = air wavelength at resonance
λ_g = guide wavelength
λ_{g0} = guide wavelength at resonance
λ_{g1} = guide wavelength in normal guide
λ_{g3} = guide wavelength in filter section
$\Delta\lambda_{g0}$ = guide wavelength bandwidth
θ = electrical length $= 2\pi l/\lambda_g$.

Recent Telecommunication Development

Etchings of M. I. Pupin

Michael I. Pupin (1858–1935) is the subject of the latest etching in the series published by the International Telecommunications Union.

Pupin was born in 1858 in Serbia. He emigrated to the United States and served for many years as professor of electromechanics at Columbia University. He is perhaps best known for the inductive loading of telephone lines, which was a necessary element in the initiating of long-distance telephony. He also experimented in the propagation of electrical waves, resonance phenomena, and multiplex telegraphy. In 1896, he discovered secondary X-ray radiation and developed the use of a fluorescent screen for the photographing of X rays, thus reducing importantly the exposure time required for examination by photography.

This Pupin portrait is the seventeenth in the series that started in 1935. The etchings are on a good grade of paper and with margins measure 9 by $6\frac{5}{8}$ inches (23 by 17 centimeters). These etchings are available at 3 Swiss francs each from Secrétariat général de l'Union internationale des Télécommunications, Palais Wilson, 52, rue des Pâquis, Genève, Suisse. The etchings are of Ampère, Baudot, Bell, Erlang, Faraday, Ferrié, Gauss and Weber, Heaviside, Hertz, Hughes, Marconi, Maxwell, Morse, Popov, Pupin, Siemens, and Tesla.

Current Fluctuations in the Direct-Current Gas Discharge Plasma*

By PHILIP PARZEN and LADISLAS GOLDSTEIN†

Federal Telecommunication Laboratories, Incorporated; Nutley, New Jersey

THE NOISE POWER from a gas discharge plasma may be ascribed to the electron current fluctuations in the plasma due to collisions of electrons with atoms or ions. The noise power, in general, is derived from both the thermal velocities, which are characterized by the electron temperature, and from the direct-current power, which is characterized by the average current.

• • •

1. *Introduction*

Radio-frequency energy produced in a direct-current discharge plasma in different gases has been measured by Goldstein.[1] Mumford[2] has recently indicated the use of such tubes as noise sources in the microwave region. This noise power can be accounted for by a study of the electron-current fluctuations in a gas discharge plasma, and, in general, in any electron gas. It will be seen that the electron-current fluctuations can be separated into two parts. One can be ascribed to the electron temperature and the other to the direct current in the gas discharge tube.

2. *Calculation of Current Fluctuations in a Gas Discharge Tube*

Let us consider the electrons in the gas discharge (Figure 1) whose thermal speeds lie between v and $v+dv$ with a mean collision frequency Z_v. This means that on the average an electron in this range will suffer Z_vT collisions in a time T. Actually, there will be fluctuations in this number, and the probability $p(K)$ that an electron will experience K collisions in a time T is[3]

$$p(K)=\frac{e^{-Z_vT}(Z_vT)^K}{K!}. \tag{1}$$

The probability that the time between consecutive collisions of an electron lies between θ and $\theta+d\theta$ is

$$q(\theta)=Z_ve^{-Z_v\theta}d\theta. \tag{2}$$

The convection current measured between the electrodes due to an electron that has collided at time t_k with a subsequent free time θ_k is

$$\left.\begin{aligned} i_x(t-t_k;\,\theta_k)&=(e/d)[v_x+a(t-t_k)]\\ i_y(t-t_k;\,\theta_k)&=(e/d_y)v_y\\ i_z(t-t_k;\,\theta_k)&=(e/d_z)v_z \end{aligned}\right\}t_k\leqslant t\leqslant t_k+\theta_k$$
$$=0 \text{ everywhere outside this time interval,} \tag{3}$$

where d is the length of the tube parallel to the direction of applied electric field E; $a=eE/m$;

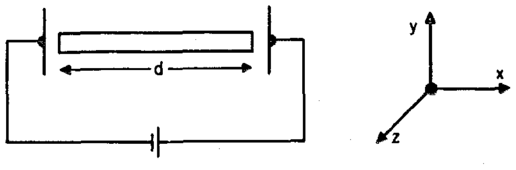

Figure 1.

d_y and d_z are transverse dimensions; v_x, v_y, v_z are components of the thermal velocity; and, similarly, i_x, i_y, i_z are components of the convection current.

We shall now restrict ourselves to the calculation of i_x as the calculations for the others are quite similar. Now let

$$\phi_k=t-t_k. \tag{4}$$

* Reprinted from the *Physical Review*, v. 82, pp. 724–726; June 1, 1951. This work was sponsored by the Signal Corps Engineering Laboratories of the United States Army.
† Dr. Goldstein is now with the University of Illinois.
[1] L. Goldstein and N. L. Cohen, "Radiofrequency Conductivity of Gas-Discharge Plasmas in the Microwave Region," *Physical Review*, v. 73, p. 83; January 1, 1948.
[2] W. W. Mumford, "A Broad-Band Microwave Noise Source," *Bell System Technical Journal*, v. 28, pp. 608–618; October, 1949.
[3] S. O. Rice, "Mathematical Analysis of Random Noise," *Bell System Technical Journal*, v. 23, pp. 282–332; July, 1944.

Hence

$$i_x(\phi_k;\, \theta_k) = (e/d)(v_x + a\phi_k), \quad 0 \leqslant \phi_k \leqslant \theta_k \quad \text{(3A)}$$
$$= 0 \text{ elsewhere.}$$

The current $I(t)$ will be a random function of time, depending on the values of ϕ_k and θ_k, which may be looked on as random values. Thus the methods described by Rice[3] for the study of shot-effect processes may be used to compute the average value and the spectrum of $I(t)$.

2.1 Computation of $\langle I(t)\rangle_{\text{Av}}$

Following Rice,[3] the time average of the current $I(t)$ due to a single electron is found by averaging $I(t')$ for a given t' over M intervals of duration T; M being very large. Thus

$$\langle I(t)\rangle_{\text{Av}} = \lim \frac{1}{M} \sum_{i=1}^{M} I_i(t'). \quad (5)$$

$I_i(t')$ is the current at time t' in the ith interval. In $Mp(K)$ of these intervals, an electron will experience K collisions that occur at times $t_1,\, t_2 \ldots t_K$. The t_i will vary in a random manner over these $Mp(K)$ intervals. For these intervals, the contribution to the time average is

$$p(K)\langle I_k(t - t_i;\, \theta_i)\rangle_{\text{Av}}.$$

The latter average is now over all the random-varying t_i in these $Mp(K)$ intervals. Now let us consider those intervals for which $t_i < t' < t_{i+1}$. The average contribution for such intervals would be

$$\int_0^T \int_0^{\theta_i} i_x(\phi_i) \frac{d\phi_i}{T} Z_v\, e^{-Z_v\theta_i}\, d\theta_i \quad (7)$$

$$\theta_i = t_{i+1} - t_i,$$

which equals

$$\frac{Z_v}{T}\frac{e}{d}\left(\frac{v_x}{Z^2_v} + \frac{a}{Z^3_v}\right) \text{ as } T \to \infty. \quad (8)$$

Hence summing over all t_i

$$\langle I_x(t)\rangle_{\text{Av}} = \sum_{K=0}^{\infty} \frac{Kp(K)}{T} \frac{e}{d}\left(\frac{v_x}{Z_v} + \frac{a}{Z^2_v}\right) \quad (9)$$

$$\langle I_x(t)\rangle_{\text{Av}} = \frac{ev_x}{d} + \frac{ea}{d}\frac{1}{Z_v}, \quad (10)$$

since

$$\sum_{K=0}^{\infty} Kp(K) = Z_v T.$$

Similarly,

$$\left.\begin{aligned} \langle I_y(t)\rangle_{\text{Av}} &= ev_y/d_y \\ \langle I_z(t)\rangle_{\text{Av}} &= ev_z/d_z. \end{aligned}\right\} \quad (11)$$

3. Calculation of Correlation Function

Following Rice, the correlation function $\psi_{\text{Av}}(\tau)$ is given by

$$\langle\psi(\tau)\rangle_{\text{Av}} = \lim_{T\to\infty} \frac{1}{T}\int_0^T \langle I(t)I(t+\tau)\rangle_{\text{Av}}\, dt, \quad (12)$$

where the average is, as previously, over many intervals of duration T. Thus,

$$\langle I(t)I(t+\tau)\rangle_{\text{Av}}$$
$$= \sum_{K=0}^{\infty} p(K)\langle I_K(t - t_i;\, \theta_i) I_K(t+\tau-t_j;\, \theta_j)\rangle_{\text{Av}}, \quad (13)$$

where the average is now over the random variables t_i and θ_i. Now for those intervals (t_i, t_{i+1}) for which t and $t+\tau$, $\tau > 0$, both lie in this interval or $0 < \phi_i < \theta_i - \tau$, the contribution is

$$\langle\psi_1(\tau)\rangle_{\text{Av}} = \sum_{K=0}^{\infty} Kp(K)\frac{Z_v}{T}$$
$$\int_\tau^T \int_0^{\theta_i - \tau} i_x(\phi_i) i_x(\phi_i + \tau) e^{-Z_v\theta_i} d\phi_i\, d\theta_i. \quad (14)$$

For those intervals (t_i, t_{i+1}) and (t_j, t_{j+1}) for which t and $t+\tau$ lie in different intervals, the contribution is

$$\langle\psi_0(\tau)\rangle_{\text{Av}} = \sum_{K=0}^{\infty} K(K-1)p(K)\frac{Z^2_v}{T^2}$$
$$\times\left[\int_0^T \int_0^{\theta_i} i_x(\phi_i) e^{-Z_v\theta_i} d\phi_i\, d\theta_i\right]^2. \quad (15)$$

Substituting for i_x from (3a) and neglecting those terms that are proportional to v_x inasmuch as they will not contribute to the average over-all velocity classes

$$\psi_{\text{Av}}(\tau) = \langle\psi_1(\tau)\rangle_{\text{Av}} + \langle\psi_0(\tau)\rangle_{\text{Av}}$$

$$\langle\psi_1(\tau)\rangle_{\text{Av}} = \frac{e^2}{d^2}\left[v^2_x + \frac{2a^2}{Z^2_v} + \frac{a^2\tau}{Z_v}\right] e^{-Z_v\tau} \quad (16)$$

$$\langle\psi_0(\tau)\rangle_{\text{Av}} = \frac{e^2}{d^2}\left[v^2_x + \frac{a^2}{Z^2_v}\right]. \quad (17)$$

4. Calculation of the Spectrum $\langle w(f)\rangle_{\text{Av}}$ of Current Fluctuations

$$\langle w(f)\rangle_{\text{Av}} = 4\int_0^\infty \psi(\tau)\cos 2\pi f\tau\, d\tau. \quad (18)$$

The spectrum resulting from $\psi_0(\tau)$ does not represent a fluctuation as it is due to $\langle I(t)\rangle_{Av}$. Thus

$$\langle[I(t)-\langle I(t)\rangle_{Av}]^2\rangle_{Av}=\int\langle w_1(f)\rangle_{Av}df$$
$$=\langle\psi_1(0)\rangle_{Av}=\frac{e^2}{d^2}\left[v^2{}_x+\frac{2a^2}{Z^2{}_v}\right] \quad (19)$$

and

$$w_1(f)=4\int_0^\infty\langle\psi_1(\tau)\rangle_{Av}\cos 2\pi f\tau d\tau$$
$$=\frac{4e^2}{d^2}\left[\left(v^2{}_x+\frac{2a^2}{Z^2{}_v}\right)\frac{Z_v}{Z^2{}_v+4\pi^2f^2}+\frac{a^2}{Z_v}\frac{Z^2{}_v-4\pi^2f^2}{(Z^2{}_v+4\pi^2f^2)^2}\right]. \quad (20)$$

A similar derivation for metals with no direct current has been carried through by Bakker and Heller.[4] In their case, the current fluctuations, assuming classical statistics, is given by (20) with $a=0$.

5. *Summation Over All Electrons*

It now remains to sum over all electrons. It is assumed that all electrons act independently of each other. Naturally, some assumptions will have to be made about the number of electrons per unit velocity range F and the variation of Z_v with v.

We shall assume that F in the gas discharge plasma is Maxwellian,[5]

$$\left.\begin{aligned}F(v_x,v_y,v_z)&=A\exp\left[-\beta^2(v^2{}_x+v^2{}_y+v^2{}_z)\right]\\A&=N(m/2\pi kT_e)^{3/2}\\\beta^2&=m/2kT_e,\end{aligned}\right\} \quad (21)$$

where N is the total number of electrons $=n_0Sd$, n_0 is the number of electrons per cubic centimeter, S is the cross-sectional area of the tube, and T_e is the electron temperature. It is also assumed that Z_v is independent of v and equal to Z (22). This is true for some gases; e.g., argon, krypton, xenon for velocities less than 1 volt and mercury for still higher velocities.[6]

With these assumptions, denoting the total quantities (sum over all electrons) by a subscript N, the average total current is given by

$$\langle I_N(t)\rangle_{Av}=\frac{eaN}{d}\frac{1}{Z}. \quad (23)$$

Hence the direct-current resistance R_0 is given by

$$R_0=\frac{1}{N}\frac{m}{e^2}d^2Z. \quad (24)$$

Furthermore, in accordance with the theory of conductivity of electrons in gases,[7] the alternating-current admittance is given by

$$\left.\begin{aligned}Y_\omega&=\frac{Ne^2}{md^2}\left[\frac{Z+j2\pi f}{Z^2+4\pi^2f^2}\right].\\&=G(\omega)+jB(\omega)\end{aligned}\right\} \quad (25)$$

$$\langle[I_N(t)-\langle I_N(t)\rangle_{Av}]\rangle^2{}_{Av}=\frac{Ne^2}{md^2}kT_e+2\left[\frac{\langle I_N(t)\rangle_{Av}}{N}\right]^2. \quad (26)$$

Except for a numerical factor of 2, this checks with a result obtained by other considerations by Brillouin[8] for electrons in metals

$$\langle w_1(f)_N\rangle_{Av}=4kT_eG(\omega)+4\left[\frac{\langle I_N(t)\rangle_{Av}}{N}\right]^2\times\left[\frac{2Z}{Z^2+4\pi^2f^2}+\frac{Z(Z^2-4\pi^2f^2)}{(Z^2+4\pi^2f^2)^2}\right]. \quad (27)$$

For I_y and I_z, the spectrum will be given by just the first term of (27).

6. *Calculation of Available Noise Power in Waveguides*

We shall now apply these results to calculate the available noise power from a gas discharge plasma placed in the transverse plane of a rectangular waveguide propagating only in its lowest mode.

The available noise power

$$P_\omega=\frac{|I_E|^2}{4}\frac{1}{G_\omega}, \quad (28)$$

[4] C. J. Bakker and G. Heller, "On the Brownian Motion in Electric Resistances," *Physica*, v. 6, pp. 262–274; March, 1939: p. 282.

[5] Langmuir and Mott-Smith, "Theory of Collectors in Gaseous Discharges," *Physical Review*, v. 28, pp. 727–763; October, 1926.

[6] R. B. Brode, "Quantitative Study of the Collisions of Electrons with Atoms," *Reviews of Modern Physics*, v. 5, pp. 257–279; October, 1933.

[7] H. Margenau, "Conduction and Dispersion of Ionized Gases at High Frequencies," *Physical Review*, v. 69, pp. 508–513; May 1 and 15, 1946.

[8] L. Brillouin, "Fluctuations of Current in a Conductor," *Helvetica Physica Acta*, v. 7, Supplement 2, pp. 47–67; 1934. This is due to the fact that in Brillouin's model the average current is not produced by an accelerating direct-current electric field but is assumed to be given. Brillouin's answer may be obtained by putting $a=0$ and increasing the thermal velocity by a constant drift velocity.

where I_E is the current in the direction of the E vector. Hence, in this case

$$P_\omega = \left\{ kT_e + \frac{P_0}{NZ} \cos^2 \theta \left[2 + \frac{Z^2 - 4\pi^2 f^2}{Z^2 + 4\pi^2 f^2} \right] \right\} df, \quad (29)$$

θ is the angle between the E vector and axis of tube, P_0 is the direct-current power dissipated in the tube, and at $\theta = \pi/2$ this reduces to the case of Mumford[2] for which

$$P_\omega = kT_e df. \quad (30)$$

7. *Conclusions*

The noise power from gas discharge plasma may be ascribed to the electron-current fluctuations in the plasma due to collisions of electrons with atoms or ions. The noise power, in general, is derived from both the thermal velocities, which are characterized by the electron temperature, and from the direct-current power, which is characterized by the average current. The fluctuations due to the positive-ion current may be neglected since the ion gas temperature is usually much lower than that of the electron gas, and the ionic current is much less than the electronic current. It has been assumed that the electron velocity-distribution function is Maxwellian and that the collision frequency of an electron is independent of speed. This has been verified by experiment in certain cases. However, for those discharges where these assumptions are not experimentally verified, the current fluctuations and noise power may still be computed by a similar method from the experimentally obtained distribution function and variation of collision frequency with speed. The current fluctuations due to the fluctuations in the various electron removal and production processes in the plasma have not been accounted for by this method.

In general, the frequency-sensitive portion of the available noise power is only a few percent of the KT_e term. However, it may become important in low-electron-temperature gas discharge plasma, such as occur in caesium vapor.

Recent Telecommunication Development

Advanced Theory of Waveguides

Mr. L. Lewin, a member of the technical staff of Standard Telecommunication Laboratories, Limited, is the author of a new book on "Advanced Theory of Waveguides", published by Iliffe and Sons, Limited, Dorset House, Stamford Street, London, S.E. 1, England.

The seven chapters are titled as follows: Electromagnetic Theory and Its Application to Waveguides; Cylindrical Posts in Waveguides; Diaphragms in Waveguides; The Tuned Post and the Tuned Window; Waveguide Steps, T-Junctions, and Tapers; Radiation from Waveguides; and Propagation in Loaded and Corrugated Guides. These chapters together with a 24-page bibliography and a short index and preface make up the 192-page $5\frac{1}{2}$-by-$8\frac{3}{4}$-inch (14-by-22-centimeter) hard-covered book. Price, 30 shillings plus postage.

Note on Reactive Elements for Broad-Band Impedance Matching

By LEONARD LEWIN

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BROAD-BAND reactance compensation can be achieved by the use of a series element in combination with a parallel element and load. Three applications are given in which broad-band matching of waveguide components is achieved by this means. The need for the development of convenient series elements is stressed.

• • •

The use of an inductance to tune out a capacitance is, of course, a well-known and much-used device. The tuning can be done at one frequency only: broad-band tuning would require negative circuit units, which cannot be constructed from passive elements. The position is modified in the presence of a load, and many constant-R networks are known. How negative elements may be simulated over a band, using only passive reactances, will be shown with particular reference to waveguides.

1. Series and Parallel Elements

Figure 1 shows a line terminated by a reactance jx_1 in series with a parallel arrangement of a reactance jx_2 and a load 1 (all quantities referred to the line impedance). The input impedance Z is given by

$$Z = jx_1 + 1/(1 + 1/jx_2). \tag{1}$$

If the load is approximately matched, jx_2 will be large and jx_1 small. Expanding the denominator of the second term, we get

$$Z \approx jx_1 + 1 - \frac{1}{jx_2} = 1 + j\left(x_1 + \frac{1}{x_2}\right). \tag{2}$$

If $x_2 = -1/\omega C$ and $x_1 = \omega L$, then the total reactance in (2) is zero when $L = C$. This cancellation holds not only at a chosen frequency, but over the whole range for which the expansion used in obtaining (2) is valid. Similarly, if $x_2 = \omega L$, then we take $x_1 = -1/\omega C$ with $C = L$.

If x_2 consists of a parallel combination of L and C, then it is readily seen that x_1 should be a series combination with the values of L and C interchanged.

An identical analysis holds for the arrangement of Figure 2, in which jx_1 is in parallel with a load 1 in series with jx_2.

This scheme of reactance compensation is valuable only when the load is already approximately matched. For example, if $x_2 = 5$, so that the initial mismatch produces 10-per-cent voltage reflection, then the combination of a series element jx_1 chosen as above provides a first-order matching, the residual being the neglected terms in the expansion, approximately $1/x_2^2 = 1/25$, or 2-per-cent reflection. Since the error goes up as the square of the reactance x_2, it is important to start with, wherever possible, an initial device that is fairly well matched. In

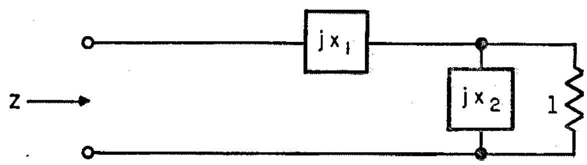


Figure 1.

practice, however, it is unlikely that the load will be exactly unity over the band, and deviations may sometimes be used, wholly or partially, to correct the second-order effects.

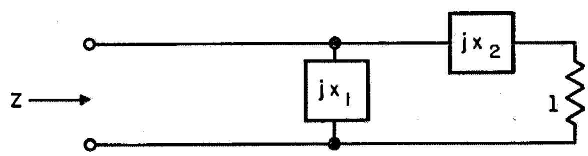


Figure 2.

The representation of a load as a series or parallel arrangement is usually a matter of convenience only: in most waveguide applications, it is desirable to work in susceptances since the commoner waveguide structures such as diaphragms, tuned windows, etc., are shunted across the guide. In practice, this may mean that an impedance that is most naturally represented as a series combination has to be put in the parallel form, and this may result in the appearance of what seem to be negative elements. Actually, this is an approximation over a finite band only, and corresponds to the expansion whereby (2) was obtained from (1). However, when this phenomenon was first noted, this interpretation had not been considered, and the appearance of negative elements was accepted as a fortunate occurrence that permitted easy broad-band matching. With the realisation that nothing more than a series to parallel transformation was involved, the scope of the method was immediately widened, since ordinary shunt elements could be compensated by means of series elements, an uncommon practice in waveguides since such elements are not so readily constructed as shunt susceptances. A waveguide gas seal, described below, is one such application.

2. *Some Waveguide Applications*

The discontinuity at the beginning of a taper in the narrow side of a waveguide can be represented as a series capacitance or, in the parallel form, as a shunt negative inductance.[1] It is obvious that the first form is the significant one and the latter only an approximation. However, it points to the use of an inductive diaphragm for broad-band compensation, and this is in fact achieved without difficulty.

A second example is provided by the H-bend corner. Measurements showed that the equivalent circuit could be approximated to by a parallel tuned circuit across the corner, the circuit elements being negative. This fact shows that the more natural equivalent circuit would have been a series tuned circuit in series with the guide at the corner. A tuned window across the corner provides broad-band compensation, provided the window dimensions are correctly chosen. In practice, it was found that the inductive element could be conveniently provided by a further cut-back of the corner, thereby simplifying the matching structure considerably.

[1] L. Lewin, "Reflection Cancellation in Waveguides," *Wireless Engineer*, v. 10, pp. 258–264; August, 1949: also *Electrical Communication*, v. 27, pp. 48–55; March, 1950.

A waveguide gas seal consists of a mica window clamped across the guide. The reflection is about 3 per cent, varying somewhat over the waveband. The usual practice is to tune out the capacitance at mid-band with an inductive diaphragm. This leaves a residual of 1 to 1·5 per cent at the edges of the band. Broad-band compensation is achieved by providing at the clamp a series inductance, which can conveniently take the form of a narrow cavity milled out of the clamping flanges, which must be specially designed for this purpose. Virtually complete reactance compensation can be obtained in this way.

3. *Limitation of the Method*

Apart from the obvious limitations, discussed in Section 1, that the load must be fairly well matched to start with, a further limitation appears through the need to provide series inductances and capacitances. A series inductance is an excrescence on the outside of the waveguide, and this is undesirable particularly when the unit to be matched is a radiating horn or other element for which the outside guide surface is in an electromagnetic field. The capacitive element is even more troublesome, since it would seem to involve a break in the guide surface, which would therefore radiate. If the break were enclosed in a box, resonances could occur, whilst if the box were made lossy a resistive component would appear along with the capacitance. No doubt, these limitations can be overcome in due course. An investigation into convenient forms of series elements seems highly desirable.

Telephone Statistics of the World*

TELEPHONES in service throughout the world numbered approximately 74.8 million as of January 1, 1951, reflecting an increase of 4.5 million telephones, or 6.4 percent, over the previous year's total.

Only ten countries were served by more than one million telephones. They were: United States, United Kingdom, Canada, Germany, France, Japan, Sweden, Russia, Italy, and Australia. Among large countries, six reported more than 15 telephones for each 100 of their populations; namely, United States (28.1), Sweden (23.9), Canada (20.8), Switzerland (19.1), New Zealand (19.1), and Denmark (16.9).

More than one-half the world's telephones are in the United States, where the entire industry is operated under private ownership. While some 82 percent of this country's 43 million telephones were operated by the Bell System, comprising the American Telephone and Telegraph Company and its 20 principal telephone subsidiaries, there were more than 5500 other privately owned telephone companies operating within the United States at the beginning of 1951.

The telephone user in the United States may be connected with most of the world's telephones by land-lines, submarine cables, or radiotelephone facilities. Because of restrictions in certain areas resulting from war conditions, it is difficult to estimate accurately the number of telephones that may be connected at the present time. However, physical facilities exist for the potential connection of telephone users in the United States with about 96 percent of the world's telephones.

Only those telephones that are available to the general public are included in this compilation. Private intercommunicating sets and private line telephones used exclusively by railroads and other agencies that do not have connection with a commercial system are excluded. Statistics are presented in so far as possible as of January 1, 1951. Where current official data could not be obtained, estimates have been based on the latest data available.

TELEPHONES IN CONTINENTAL AREAS

January 1, 1951 (a)

Continental Area	Total Telephones			Privately Owned		Automatic (Dial)	
	Number	Percent of Total World	Per 100 Population	Number	Percent of Total	Number	Percent of Total
North America	45,933,800	61.4	27.5	45,554,400	99.2	31,923,300	69.5
Middle America	555,000	0.7	1.1	511,100	92.1	385,600	69.5
South America	1,815,000	2.4	1.6	903,900	49.8	1,355,900	74.7
Europe	21,300,000	28.5	3.5	3,149,900	14.8	14,772,600	69.4
Africa	895,200	1.2	0.5	15,100	1.7	579,400	64.7
Asia	2,655,000	3.6	0.2	197,700	7.4	1,152,000	43.4
Oceania	1,646,000	2.2	11.9	109,600	6.7	1,040,800	63.2
World	74,800,000	100.0	3.1	50,441,700	67.4	51,209,600	68.5
United States	43,003,832	57.5	28.1	43,003,832	100.0	30,170,000	70.2

(a) Partly estimated; data reported as of other dates have been adjusted to January 1, 1951.

* Abridgement from a booklet issued by The American Telephone and Telegraph Company, New York, New York.

Telephones in Countries of the World

January 1, 1951

Country		Total Telephones		Ownership		Automatic (Dial)	
		Number	Per 100 Population	Government	Private	Number	Percent of Total Telephones
NORTH AMERICA							
Alaska		17,919	13.8	78	17,841	6,206	34.6
Canada	(1)	2,911,900	20.8	379,200	2,532,700	1,747,100	60.0
Greenland		0	—	—	—	—	—
St. Pierre and Miquelon		137	2.9	137	—	0	—
United States		43,003,832	28.1	—	43,003,832	30,170,000	70.2
MIDDLE AMERICA							
Bahamas		4,234	5.3	4,234	—	4,198	99.1
Barbados		4,272	2.0	—	4,272	4,166	97.5
Bermuda	(2)	6,300	17.0	—	6,300	6,300	100
British Honduras		678	1.1	678	—	0	—
Canal Zone	(3) (5)	6,112	7.6	6,112	—	6,112	100
Costa Rica		9,500	1.2	200	9,300	0	—
Cuba		120,668	2.2	518	120,150	100,938	83.6
Dominican Republic		6,446	0.3	146	6,300	4,537	70.4
El Salvador	(1)	6,800	0.3	2,400	4,400	0	—
Guadeloupe	(1)	850	0.3	850	—	0	—
Guatemala	(1)	4,200	0.2	4,200	—	3,800	90.5
Haiti		3,809	0.1	3,809	—	3,355	88.1
Honduras	(1)	2,850	0.2	2,850	—	1,450	50.9
Jamaica		13,700	1.0	—	13,700	13,087	95.5
Leeward Islands		871	0.8	871	—	0	—
Martinique		2,539	1.0	2,539	—	0	—
Mexico	(1)	285,600	1.1	1,000	284,600	191,738	67.1
Netherlands' Antilles		4,613	2.8	4,613	—	4,463	96.7
Nicaragua		3,439	0.3	3,439	—	0	—
Panama		11,403	1.4	—	11,403	9,289	81.5
Puerto Rico		38,490	1.7	2,215	36,275	20,154	52.4
Trinidad and Tobago		14,794	2.3	—	14,794	12,397	83.8
Virgin Islands		1,039	2.9	1,039	—	0	—
Windward Islands		2,150	0.8	2,150	—	0	—
SOUTH AMERICA							
Argentina		798,391	4.6	719,088	79,303	613,783	76.9
Bolivia	(1)	7,950	0.2	—	7,950	7,600	95.6
Brazil		549,700	1.0	1,500	548,200	424,000	77.1
British Guiana		3,337	0.8	3,337	—	167	5.0
Chile		134,869	2.3	—	134,869	90,230	66.9
Colombia		100,850	0.9	83,250	17,600	48,000	47.6
Ecuador		8,800	0.2	7,600	1,200	1,000	11.4
Falkland Islands		305	13.3	305	—	0	—
French Guiana		218	0.8	218	—	0	—
Paraguay		5,273	0.4	5,273	—	4,529	85.9
Peru		46,733	0.5	—	46,733	37,729	80.7
Surinam		1,710	0.8	1,710	—	0	—
Uruguay		89,871	3.8	88,791	1,080	66,233	73.7
Venezuela		67,000	1.3	—	67,000	62,617	93.5
EUROPE							
Albania	(1)	1,300	0.1	1,300	—	0	—
Andorra		60	1.2	60	—	0	—
Austria		412,394	5.8	412,394	—	312,777	75.8
Belgium		687,012	7.9	687,012	—	514,599	74.9
Bulgaria	(1)	59,000	0.8	59,000	—	26,000	44.1
Czechoslovakia	(1)	385,000	3.1	385,000	—	230,000	59.7
Denmark		723,443	16.9	35,468	687,975	296,914	41.0
Finland		328,394	8.0	42,695	285,699	179,626	54.7
France		2,405,802	5.7	2,405,802	—	1,474,509	61.3
Germany, Eastern, and Berlin	(1)	332,200	1.6	332,200	—	200,000	60.2
Germany, Western		2,393,013	5.0	2,393,013	—	2,050,637	85.7
Gibraltar		1,348	6.1	1,348	—	1,205	89.4
Greece		81,905	1.0	—	81,905	77,020	94.0
Hungary		116,000	1.2	116,000	—	85,800	74.0
Iceland	(1)	20,801	14.4	20,801	—	13,787	66.3
Ireland		82,031	2.7	82,031	—	52,634	64.2
Italy		1,244,152	2.7	—	1,244,152	1,147,179	92.2
Liechtenstein		1,805	12.9	1,805	—	0	—
Luxemburg		23,412	8.0	23,412	—	16,796	71.7
Malta and Gozo		7,030	2.3	7,030	—	0	—

(1) Data partly estimated.
(2) June 30, 1951.
(3) Excluding telephone systems of the U.S. military forces.
(4) March 31, 1951.
(5) June 30, 1950.
(6) Including all Asiatic territory of the U.S.S.R.
+ Less than 0.1.

Telephones in Countries of the World—*Continued*

Country		Total Telephones		Ownership		Automatic (Dial)	
		Number	Per 100 Population	Government	Private	Number	Percent of Total Telephones
EUROPE—*Continued*							
Monaco		2,800	14.7	2,800	—	2,500	89.3
Netherlands		781,678	7.7	781,678	—	702,896	89.9
Norway	(5)	451,727	13.8	376,714	75,013	263,084	58.2
Poland	(1)	230,000	0.9	230,000	—	155,000	67.4
Portugal		152,580	1.8	52,529	100,051	75,045	49.2
Romania		136,000	0.9	136,000	—	102,000	75.0
San Marino	(1)	100	0.7	100	—	0	—
Spain		667,639	2.4	19,558	648,081	510,700	76.5
Sweden		1,685,200	23.9	1,683,230	1,970	1,083,774	64.3
Switzerland		896,398	19.1	896,398	—	863,324	96.3
Trieste		23,654	6.6	500	23,154	23,154	97.9
Russia	(1) (6)	1,500,000	0.7	1,500,000	—	400,000	26.7
United Kingdom	(4)	5,433,614	10.7	5,433,614	—	3,893,242	71.7
Yugoslavia	(1)	85,000	0.5	85,000	—	55,000	64.7
AFRICA							
Algeria	(1)	96,400	1.1	96,400	—	68,500	71.1
Anglo-Egyptian Sudan		8,300	0.1	8,300	—	6,520	78.6
Belgian Congo		5,432	+	5,432	—	3,379	62.2
British East Africa		19,771	0.1	19,771	—	13,863	70.1
British South Africa		29,500	0.4	29,500	—	21,700	73.6
British Southwest Africa		4,571	1.5	4,571	—	2,374	51.9
British West Africa		18,858	+	18,858	—	1,390	7.4
Comoro Archipelago		0	—	—	—	—	—
Egypt	(1)	115,500	0.6	115,500	—	66,700	57.7
Eritrea	(1)	2,600	0.2	2,600	—	1,050	40.4
Ethiopia		2,050	+	2,050	—	1,600	78.0
French Cameroon		1,096	+	1,096	—	0	—
French Equatorial Africa		1,169	+	1,169	—	0	—
French Somaliland		350	0.7	350	—	0	—
French Togo		470	+	470	—	0	—
French West Africa		11,740	+	11,740	—	5,208	44.4
Libya		4,434	0.4	4,434	—	3,102	70.0
Liberia		0	—	—	—	—	—
Madagascar	(1)	5,600	0.1	5,600	—	0	—
Mauritius		4,749	1.0	4,749	—	277	5.8
Morocco		68,484	0.6	53,694	14,790	45,979	67.1
Portuguese Africa		6,705	+	6,705	—	1,459	21.8
Reunion	(1)	2,050	0.8	2,050	—	0	—
St. Helena		87	1.7	87	—	0	—
Seychelles	(1)	75	0.2	75	—	0	—
Somalia	(1)	350	+	350	—	0	—
Somaliland Protectorate		165	+	165	—	0	—
Spanish Guinea		271	0.1	—	271	0	—
Spanish Sahara		0	—	—	—	—	—
Tunisia		25,207	0.8	25,207	—	14,625	58.0
Union of South Africa		458,851	3.7	458,851	—	321,682	70.1
Zanzibar and Pemba		400	0.1	400	—	0	—
ASIA							
Aden, Colony of		1,338	1.3	1,338	—	1,338	100
Afghanistan	(1)	4,500	+	4,500	—	300	6.7
Bahrain		364	0.3	364	—	364	100
Brunei		65	0.2	65	—	0	—
Burma	(1)	5,200	+	5,200	—	1,040	20.0
Ceylon		16,860	0.2	15,811	1,049	14,734	87.4
China	(1)	255,000	+	160,000	95,000	178,000	69.8
Cyprus		5,002	1.1	—	5,002	0	—
French India		76	+	76	—	76	100
French Indo-China		11,531	+	11,531	—	4,200	36.4
Hong Kong	(5)	28,705	1.3	—	28,705	28,705	100
India	(4)	168,397	+	166,559	1,838	70,800	42.0
Indonesia	(1)	43,000	+	43,000	—	1,000	2.3
Iran		28,620	0.2	3,126	25,494	10,725	37.5
Iraq		17,630	0.3	17,630	—	11,455	65.0
Israel		28,956	2.1	28,956	—	24,826	85.7
Japan	(4)	1,802,558	2.1	1,802,558	—	667,667	37.0
Jordan		4,245	0.3	4,245	—	1,800	42.4
Lebanon	(1)	13,500	1.1	13,500	—	0	—
Malaya		23,694	0.4	23,694	—	8,569	36.2
North Borneo		578	0.2	578	—	110	19.0
Pakistan		18,771	+	18,771	—	9,328	50.0
Philippine Republic		20,386	0.1	—	20,386	15,425	75.7

Telephones in Countries of the World—*Continued*

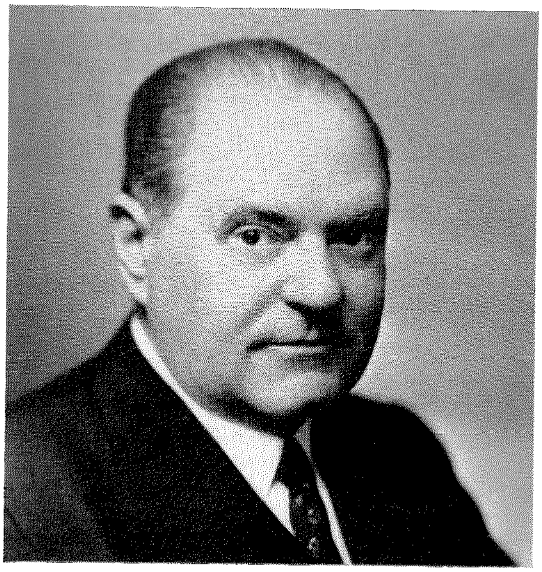
Country		Total Telephones: Number	Total Telephones: Per 100 Population	Ownership: Government	Ownership: Private	Automatic (Dial): Number	Automatic (Dial): Percent of Total Telephones
ASIA—*Continued*							
Portuguese Asia		1,630	0.1	1,630	—	1,410	86.5
Sarawak		713	0.1	713	—	0	—
Saudi Arabia		5,600	+	5,600	—	0	—
Singapore		18,601	1.9	—	18,601	18,601	100
Syria		11,516	0.4	11,516	—	7,376	64.1
Thailand	(1)	6,100	+	6,100	—	6,100	100
Turkey		65,150	0.3	65,150	—	56,410	86.6
Other Places	(1)	80,300	0.2	80,300	—	24,000	29.9
OCEANIA							
American Samoa		313	1.7	313	—	313	100
Australia		1,158,202	13.9	1,158,202	—	720,981	62.3
Fiji	(1)	2,500	0.9	2,500	—	0	—
French Oceania		2,198	1.9	2,198	—	0	—
Hawaii		109,579	21.9	—	109,579	103,177	94.2
New Zealand	(4)	369,989	19.1	369,989	—	216,749	58.6
Pacific Islands (British)	(1)	410	0.2	410	—	0	—
Papua-New Guinea		1,614	0.1	1,614	—	0	—
Portuguese Timor	(1)	280	+	280	—	0	—
South Seas Mandate (U.S.)		5,432	3.9	5,432	—	2,705	49.8
Western New Guinea	(1)	525	0.1	525	—	0	—
Western Samoa		351	0.4	351	—	0	—

Telephone Conversations for the Year 1950

Country		Number of Conversations (1) in Thousands: Local	Toll	Total	Per Capita
Alaska		41,900	300	42,200	327.1
Argentina		2,743,400	36,500	2,779,900	161.7
Australia		900,200	71,200	971,400	118.8
Belgium		420,500	58,100	478,600	55.4
Brazil		1,863,000	24,000	1,887,000	36.2
Canada		4,889,100	117,900	5,007,000	361.6
Ceylon		36,000	2,500	38,500	5.1
Chile		443,600	19,500	463,100	79.7
Cuba		670,000	4,000	674,000	126.0
Denmark		923,200	185,400	1,108,600	259.6
Finland		458,200	74,900	533,100	131.2
France		1,110,500	426,200	1,536,700	36.7
Germany, Western		1,705,500	329,800	2,035,300	42.8
Greece		257,700	3,800	261,500	32.9
Hawaii		238,100	4,500	242,600	473.8
Iceland		50,400	1,200	51,600	360.8
Ireland		69,300	10,300	79,600	26.5
Israel		61,500	2,800	64,300	56.1
Italy		1,996,000	106,100*	2,102,100	45.4
Jamaica		40,000	500	40,500	28.9
Japan	(2)	5,300,000	330,500	5,630,500	67.2
Malaya		67,100	9,000	76,100	14.6
Mexico		656,000	8,500	664,500	26.2
Netherlands		556,000	159,100	715,100	70.7
Norway	(3)	461,500	47,800	509,300	156.7
Paraguay		16,000	200	16,200	11.5
Peru		166,400	2,100	168,500	20.0
Philippine Republic		165,900	200	166,100	8.5
Portugal		143,800	27,700	171,500	19.9
Puerto Rico		82,800	2,500	85,300	38.9
Spain		1,360,000	59,000	1,419,000	50.2
Sweden		2,090,000	118,000	2,208,000	313.3
Switzerland		381,800	295,400*	677,200	144.9
Tunisia		14,200	5,900	20,100	6.0
Turkey		73,000	8,700	81,700	4.0
Union of South Africa		547,500	38,000	585,500	47.5
United Kingdom	(2)	3,117,400	251,700	3,369,100	66.6
Uruguay		273,300	4,000	277,300	116.5
Venezuela		278,200	1,100	279,300	57.7
United States		54,195,000	2,045,000	56,240,000	370.6

* Three-minute units.
(1) Telephone conversation data were not available for all countries.
(2) Year ended March 31, 1951
(3) Year ended June 30, 1950

In Memoriam



FREDERICK TURNER CALDWELL

FREDERICK TURNER CALDWELL was associated with the International Telephone and Telegraph Corporation in positions of responsibility from the early period of its formation and made a host of loyal friends by whom he will continue to be remembered for his energy and devotion to his work, his integrity, his kindly manner and unfailing consideration for others, his pervasive humor, and his great personal courage.

Mr. Caldwell was born at Bloomfield, Iowa in 1883. Graduating from Iowa State College in 1903, he worked for various independent telephone companies in the United States until 1911, when he became general manager of the Canadian municipal telephone system of Edmonton, Alberta. In 1915, he was appointed superintendent of telegraphs and time service for the Grand Trunk Pacific Railroad and the Canadian Government Railway, with headquarters at Winnipeg.

During the first World War, Mr. Caldwell served as a captain of the United States Army Signal Corps, seeing action at Chateau Thierry, San Mihiel, and the Argonne.

Mr. Caldwell's long association with the International System began in the year of its foundation, 1920, when he became chief engineer of the Cuban Telephone Company. He was transferred to Spain in 1924 as chief engineer of the Spanish Telephone Company.

After serving as technical supervisor of all telephone operating companies of the International System in South America from 1931 to 1934, Mr. Caldwell returned to Spain as vice president and technical director of the Spanish Telephone Company, becoming its executive vice president on the eve of the Spanish civil war.

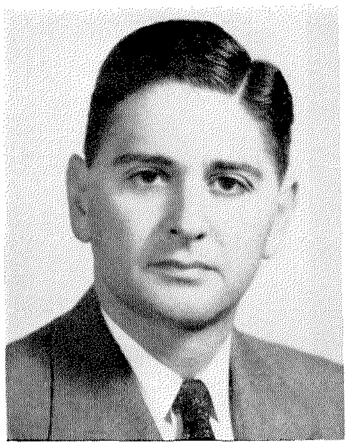
Mr. Caldwell was made a vice president in 1937 and a director in 1943 of the International Telephone and Telegraph Corporation. Early in 1945, he received similar appointments in the International Standard Electric Corporation. That same year, at the request of the War Department, he was assigned to Germany as deputy commander of the communications division of the United States Military Government.

Returning to the United States late in 1945, Mr. Caldwell assumed the presidency of International Standard Electric Corporation and early in 1948 of Federal Telephone and Radio Corporation. The strain of these manifold activities took heavy toll of his health and in 1949 Mr. Caldwell relinquished his positions in the United States and returned to Spain as vice president for the Iberian peninsula.

At the time of his death, Mr. Caldwell held the following offices in the International System: director of International Telephone and Telegraph Corporation and of International Standard Electric Corporation; divisional representative for the Iberian peninsula; president and director of International Telephone and Telegraph Corporation (España); vice president and director of Standard Eléctrica, Madrid; and Sociedad Anónima Radio Argentina, Buenos Aires; and director of Standard Eléctrica, Lisbon; and Compañía Radio Aérea Marítima Española, Madrid.

Mr. Caldwell died on December 21, 1951, at Memorial Hospital in New York City.

Contributors to This Issue



Ben Alexander

Ben Alexander was born at New York City on October 21, 1919. He received a B.A. degree from Cornell University in 1940.

From 1941 to 1945, Mr. Alexander served in the United States Navy, becoming communications officer of the U.S.S. Phoenix.

In 1945, he joined Federal Telecommunication Laboratories, where he is now a department head in the aerial navigation development laboratories.

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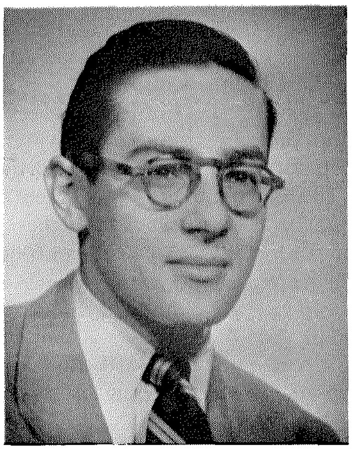
Harry A. Augenblick, Jr. was born on July 21, 1926, at South Orange, New Jersey. He received the B.S. degree in electrical engineering from Massachusetts Institute of Technology in 1946.



Harry A. Augenblick, Jr.

On graduation, he joined Federal Telephone and Radio Corporation, transferring later to Federal Telecommunication Laboratories, where he worked on microwave antennas and components. In 1950, he founded Microlab, where he is presently engaged in the design and production of microwave components and test equipment.

Mr. Augenblick is a member of Sigma Xi and an Associate of the Institute of Radio Engineers.

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Christopher Buff was born on January 6, 1917 at Camden, New Jersey. He took engineering courses at Union County (New Jersey) Junior College and at Hofstra College. He has been an active licensed radio amateur since 1931.

He was employed from 1941 to 1946 in various engineering capacities by Press Wireless, Incorporated, and was engaged in the development and design of some of the first carrier-frequency-shift keying equipment.

Since 1946, he has been with Mackay Radio and Telegraph Company working on frequency-shift keying and associated terminal equipment.

Mr. Buff is a Member of the Institute of Radio Engineers.

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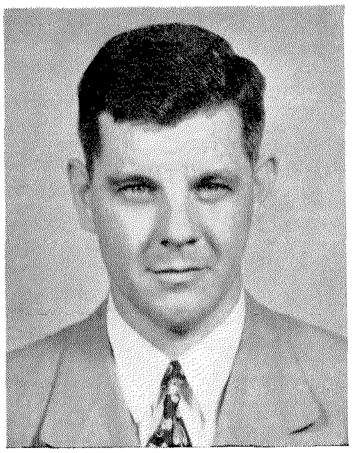
Ladislas Goldstein was born in 1906 in Hungary. He received from the University of Paris a bachelor's degree in 1924, master's in 1928, and doctor of science in 1937.

He was employed as a research physicist in the Curie Laboratory of the Institute of Radium of the University of Paris from 1928 to 1940. During the following year, he was with the Institute of Atomic Physics of the University of Lyon.

In 1941, he came to New York City where from 1942 to 1944 he was employed as a research physicist by the Canadian Radium and Uranium Corporation. Since 1945, he was with Federal Telecommunication Laboratories as a senior project engineer. He recently joined the electrical engineering department of the University of Illinois.

Dr. Goldstein is a member of the American Physical Society.

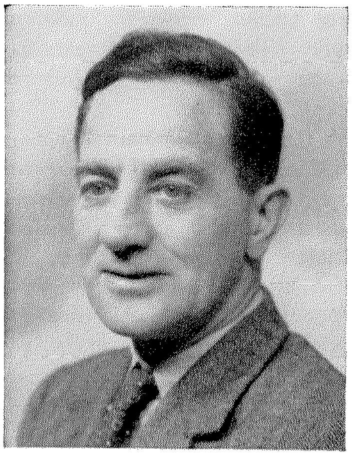


Christopher Buff

J. D. Holland was born at London in 1902. He was educated at the Salisian School in Battersea and completed a three-year apprenticeship at the Park Royal Engineering Works Limited in 1922. In 1939, he was awarded the National Certificate and a first-class pass in the Final of the City and Guilds examination in radio communication.

From 1922 to 1928, he served with the Royal Signals and spent three of these years in North China assisting in establishing a radio link between Peking and Hong Kong. He joined Standard Telephones and Cables, Limited, in 1929 as an electrical fitter and was transferred to development work in 1935. Recently, as leader of a development group, he has been con-

Ladislas Goldstein



J. D. Holland

cerned with the problems of frequency- and amplitude-modulation multichannel communication in the ultra-high-frequency band.

Mr. Holland is an Associate Member of the Institution of Electrical Engineers, having submitted a thesis on the design of high-stability oscillators in partial fulfillment of the requirements for that grade.

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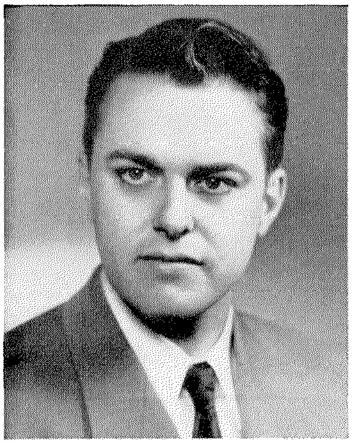
Leonard Lewin was born in 1919 at Southend-on-Sea, England. He studied mathematics with particular reference to transcendental functions and the electromagnetic theory of radiation.

During the war, he did research work on antennas and mirrors at the British Admiralty and in 1945 served as chairman of the Inter-Service Committee on Radar Camouflage.



E. B. Moore

Mr. Lewin joined the engineering staff of Standard Telecommunication Laboratories at London in 1946.

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E. B. Moore was born at Ho-Ho-Kus, New Jersey, on October 15, 1922. He received the M.E. degree from Stevens Institute of Technology in 1944.

From 1943 to 1946, he served in the U.S. Navy, becoming an electronics officer. In 1946, he joined Federal Telephone and Radio Corporation, where he has been working on radio receiving equipment for airborne navigation, direction finding, and communications.

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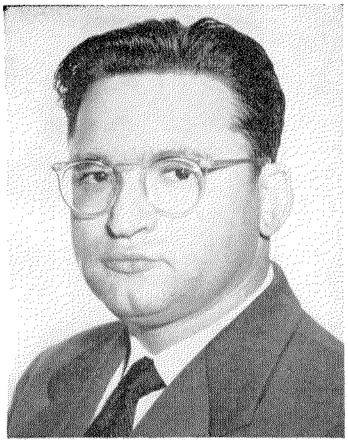

Philip Parzen

Philip Parzen was born on June 28, 1916, in Poland. He received the B.S. degree in physics from the College of the City of New York in 1939 and the M.S. degree in physics from New York University in 1946.

During the second world war, he was employed at the Westinghouse Research Laboratories. Since 1947, he has been with Federal Telecommunication Laboratories and has worked on microwave tubes and electromagnetic-wave propagation.

Mr. Parzen is a member of the American Physical Society.

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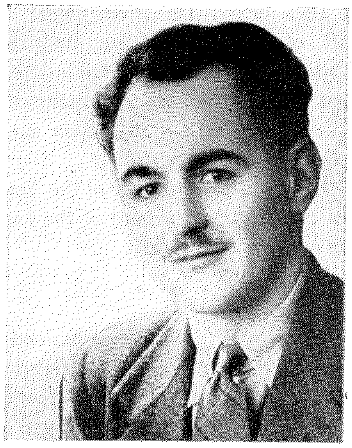

Douglas C. Rogers

Douglas C. Rogers was born in 1920 at Richmond, Surrey.

He joined the staff of Standard Telephones and Cables, Limited, in 1939. During the war, he was assigned to the ultra-high-frequency laboratories at Eltham. Since then, he has been concerned with the development of ultra-high-frequency valves in the valve laboratories at Ilminster.

Mr. Rogers is an Associate Member of the Institution of Electrical Engineers.

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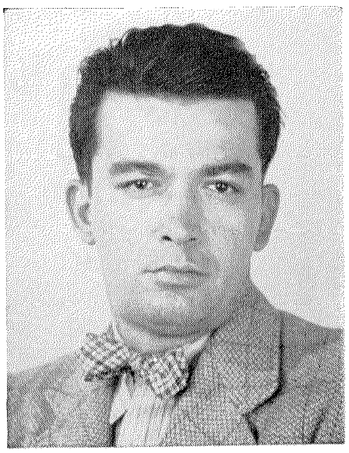
William Sichak was born on January 7, 1916, at Lyndora, Pennsylvania. He received the B.S. degree in physics from Allegheny College in 1942.

From graduation until near the end of 1945, he was at the Radiation Laboratory at Massachusetts Institute



William Sichak

of Technology. He then joined the staff of Federal Telecommunication Laboratories, where he is now a department head in the radio and radar components division and has been active in the microwave field.

Mr. Sichak is a member of the American Physical Society and of the Institute of Radio Engineers.

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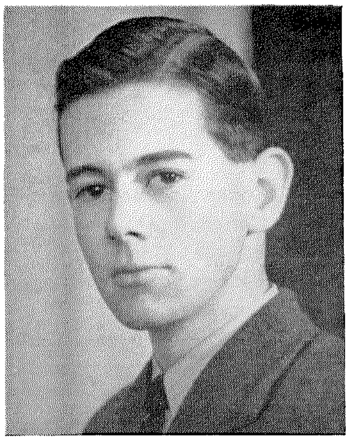

Malcolm F. Wintle

Malcolm F. Wintle obtained a degree in natural science at Cambridge University in 1941. He then spent a year as an instructor at a Royal Air Force training center for radio mechanics followed by four years as a lecturer in electronics at University College, Exeter.

He then joined Standard Telephones and Cables, Limited, where he carried out advanced development work on automatic direction finders, frequency-modulation systems, and omnidirectional beacons. During this period, he obtained a degree in electrical engineering from London University.

In 1950, Mr. Wintle became Senior Scientific Officer at the Admiralty Signal and Radar Establishment.

INTERNATIONAL TELEPHONE AND TELEGRAPH CORPORATION

Associate Manufacturing and Sales Companies

United States of America

International Standard Electric Corporation, New York, New York

Federal Telephone and Radio Corporation, Clifton, New Jersey

International Standard Trading Corporation, New York, New York

Capehart-Farnsworth Corporation, Fort Wayne, Indiana

The Coolerator Company, Duluth, Minnesota

Flora Cabinet Company, Inc., Flora, Indiana

Thomasville Furniture Corporation, Thomasville, North Carolina

Great Britain and Dominions

Standard Telephones and Cables, Limited, London, England

Creed and Company, Limited, Croydon, England

International Marine Radio Company Limited, Croydon, England

Kolster-Brandes Limited, Sidcup, England

Standard Telephones and Cables Pty. Limited, Sydney, Australia

Silovac Electrical Products Pty. Limited, Sydney, Australia

Austral Standard Cables Pty. Limited, Melbourne, Australia

New Zealand Electric Totalisators Limited, Wellington, New Zealand

Federal Electric Manufacturing Company, Ltd., Montreal, Canada

South America

Compañía Standard Electric Argentina, Sociedad Anónima, Industrial y Comercial, Buenos Aires, Argentina

Standard Electrica, S.A., Rio de Janeiro, Brazil

Compañía Standard Electric, S.A.C., Santiago, Chile

Europe

Vereinigte Telephon- und Telegraphenfabriks Aktiengesellschaft Czeija, Nissl & Co., Vienna, Austria

Bell Telephone Manufacturing Company, Antwerp, Belgium

Standard Electric Aktieselskab, Copenhagen, Denmark

Compagnie Générale de Constructions Téléphoniques, Paris, France

Le Matériel Téléphonique, Paris, France

Les Téléimprimeurs, Paris, France

C. Lorenz, A.G. and Subsidiaries, Stuttgart, Germany

Mix & Genest Aktiengesellschaft and Subsidiaries, Stuttgart, Germany

Süddeutsche Apparatefabrik Gesellschaft m.b.H., Nuremberg, Germany

Nederlandsche Standard Electric Maatschappij N.V., The Hague, Netherlands

Fabbrica Apparecchiature per Comunicazioni Elettriche, Milan, Italy

Standard Telefon og Kabelfabrik A/S, Oslo, Norway

Standard Electrica, S.A.R.L., Lisbon, Portugal

Compañía Radio Aérea Marítima Española, Madrid, Spain

Standard Eléctrica, S.A., Madrid, Spain

Aktiebolaget Standard Radiofabrik, Stockholm, Sweden

Standard Telephone et Radio S.A., Zurich, Switzerland

Telephone Operating Systems

Companhia Telefônica Nacional, Porto Alegre, Brazil

Compañía de Teléfonos de Chile, Santiago, Chile

Compañia Telefónica de Magallanes S.A., Punta Arenas, Chile

Cuban American Telephone and Telegraph Company, Havana, Cuba

Cuban Telephone Company, Havana, Cuba

Compañía Peruana de Teléfonos Limitada, Lima, Peru

Porto Rico Telephone Company, San Juan, Puerto Rico

Radiotelephone and Radiotelegraph Operating Companies

Compañía Internacional de Radio, Buenos Aires, Argentina

Compañía Internacional de Radio Boliviana, La Paz, Bolivia

Companhia Radio Internacional do Brasil, Rio de Janeiro, Brazil

Compañía Internacional de Radio, S.A., Santiago, Chile

Radio Corporation of Cuba, Havana, Cuba

Radio Corporation of Porto Rico, San Juan, Puerto Rico

Cable and Radiotelegraph Operating Companies

(Controlled by American Cable & Radio Corporation, New York, New York)

The Commercial Cable Company, New York, New York[1]

Mackay Radio and Telegraph Company, New York, New York[2]

All America Cables and Radio, Inc., New York, New York[3]

Sociedad Anónima Radio Argentina, Buenos Aires, Argentina[4]

[1] Cable service. [2] International and marine radiotelegraph services. [3] Cable and radiotelegraph services. [4] Radiotelegraph service.

Laboratories

Federal Telecommunication Laboratories, Inc., Nutley, New Jersey

International Telecommunication Laboratories, Inc., New York, New York

Laboratoire Central de Télécommunications, Paris, France

Standard Telecommunication Laboratories, Limited, London, England