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Volume 41 **1966** **Number 3**

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Maximum-Likelihood Smoothing—The available data about a time-varying quantity are considered to be a set of measurements. Each measurement is expressed as a likelihood function. These likelihood functions can be manipulated so that the time-varying quantity having the maximum likelihood of occurrence is obtained. The manipulations consist of multiplication, convolution, scaling, and partitioning. If the likelihood functions are Gaussian they can be represented by two parameters, mean and weight, and the manipulations reduce the simple algebraic operations.

The maximum-likelihood approach facilitates the optimum combination of data from diverse sources; for example, radio measurements and inertial measurements of the position of a vehicle. The data required will often be easier to obtain and more intuitively meaningful than that for the alternative smoothing approach utilizing time series. It also turns out that the mathematics for the likelihood approach are relatively simple, partly because the elements of the process are more readily comprehensible, thus making it easier to make reasonable approximations.

On the Relation Between Time, Space, and Holding-Time Distribution Functions—In applying the probability theory to telephone switching problems, care should be taken to avoid two principal errors: the incorrect interpretations of the mathematical meaning of the traffic values that appear in the various probability equations and the unwarranted application of probability equations to cases for which they are not valid.

The present paper deals with the latter problem and contains a theoretical investigation of the conditions under which the various well-known equations are valid. For this purpose a general equation is used that is based on the principle of statistical equilibrium but for a measurable lapse of time. This equation was first published in volume 38, number 2, 1963 of this periodical.

The solution of problems by the aid of this equation gives rise to a number of algebraic lemmas. Further, the paper contains a new delay equation comprising Bernoulli terms.

Planning of Telephone Systems Using Small-Diameter Coaxial Cable—Crosstalk among symmetric twisted pairs in cable increases with frequency and limits them to about 120 channels. For coaxial pairs in cable, crosstalk decreases with frequency and is significant below about 60 kilohertz.

Logically, the early coaxial-cable systems were designed to accommodate thousands of channels to meet heavy trunk needs. The 10:1 gap in channel capacity between the twisted-pair and coaxial-pair systems has now been filled with a coaxial cable having an inner-conductor diameter of 1.2 millimeters and an outer-conductor diameter of 4.4 millimeters. Transistor repeaters supplied with operating power over the inner conductor of the coaxial pair are essential elements of the new system. Initial designs are for 300, 960, and 1260 telephone channels with unattended repeaters.

To fit into existing systems, the coaxial pair should have a characteristic impedance of about 75 ohms, which for polythene dielectric and copper requires a ratio of 3.6 for the diameters of the facing surfaces of the coaxial conductors. The minimum spacing between conductors is limited by the total of the operating voltages of all the repeaters in the feed link in series and the voltage drops along the inner-conductor sections between repeaters. Under fault conditions, this total voltage may be applied across the coaxial pair. A 30-mile (50-kilometer) link for 2700 channels would have 25 amplifiers at 1.25-mile (2-kilometer) intervals. Suitable choice of repeater locations for a 300-channel system will permit them to be retained if additional repeaters are interspersed later to provide for an increase in channel capacity.

By isolation from inducing power systems, sheathing of the cable containing the coaxial pairs, insulation of the inner and outer coaxial conductors from earth, shortening of the links fed from a single source, and use of discharge devices, induced voltages may be reduced and thus protect the transistor amplifiers from damage.

Multichannel Telephone Equipment of Standard Elektrik Lorenz for Small-Diameter Coaxial Cable—Line equipment to provide 300 and 960 channels over the new small-diameter coaxial cable with transistor repeaters was designed to meet the requirements of the International Telegraph and Telephone Consultative Committee and the German Post Administration. The use of transistors permits underground installation of unattended dependent repeaters and the economical application of direct-current power over the coaxial cable for long distances.

Two types of line equipment form one system family, using compatible repeater section lengths and identical mechanical design. This allows for straightforward conversions from one type to the other if the line capacity is to be increased from 300 to 960 speech channels.

Multichannel Telephone Equipment of Standard Telephones and Cables for Small-Diameter Coaxial Cable—Two carrier systems are described capable of transmitting 300 and 960 telephone channels over two 0.174-inch (4.4-millimeter) diameter coaxial pairs of conductors employed as a 4-wire circuit. Both systems meet the recommendations of the International Telegraph and Telephone Consultative Committee and the specifications of the British Post Office. The repeater spacings are 2.5 miles (4 kilometers) for 960 channels and 3.75 miles (6 kilometers) for 300 channels.

Both equipments have a considerable number of common parts and circuits to ease the development, manufacturing, and maintenance problems. They can also share a common repeater housing intended for installation in an underground box or manhole, the whole repeater installation being designed around the British Post Office decision to employ a 4-tube coaxial cable as standard. The repeaters are power-fed over the center conductors of the 4-wire coaxial circuit and a comprehensive fault location and alarm scheme is built into the terminal repeaters.

Multichannel Telephone Equipment of Standard Téléphone et Radio for Small-Diameter Coaxial Cable—This contribution describes the 300-channel system designed by Standard Téléphone et Radio and constructed in the new Swiss equipment practice 62. The facilities mentioned are mainly those that differ from the other two systems.

High-Power Varactor Frequency-Doubler Chains—For frequency multiplication, about 10 times the power can be handled by varactors driven for a fraction of the cycle into the forward-bias region than if they are limited to the reverse-bias region as has normally been the case.

The varactor capacitance is a function of the drive power and exceeds the zero-bias value. The conversion resistance, which represents the ability to convert power at frequency f to $2f$, varies inversely as the input frequency. Best efficiency occurs when the load resistance equals the conversion resistance. The power that may be used is limited by either the varactor heat dissipation or by the peak voltage developed across the varactor. Circuit designs for both lumped and distributed coupling networks are outlined.

A chain of 6 doublers produces 1 watt at 4288 megahertz from an input of 10 watts at 67 megahertz, a multiplication of 64 times. Outputs of 5 watts have been obtained in the 4-gigahertz region.

Trends in Radio and Television Receiver Components in Germany—About 15 years ago the selenium rectifier started to replace the vacuum-tube power rectifier in radio receivers. The later invention of the transistor expanded the application of semiconductors and about 1959 the all-transistor portable radio receiver operating in the very-high-frequency region was marketed in Germany.

The introduction of semiconductors into mains-operated receivers is taking much longer. In the television receiver the limiter stages were the first to use transistors, followed by the ultra-high-frequency circuits. With the exception of the picture tube, high-voltage rectifier, and indicator tubes, all vacuum tubes can now be replaced by semiconductors in home television receivers. Price has been a major consideration.

Development of passive components has emphasized great reduction in physical size, often resulting from new materials that in some cases improve operating functions as well. Automatic assembly on printed-wiring boards has required mechanical redesign of previously acceptable components.

Goudet Elected President of Société française des Électroniciens et des Radioélectriciens for 1966

Georges Goudet, Managing Director of Compagnie Générale de Constructions Téléphoniques, President of Laboratoire Central de Télécommunications, and President of Société des Produits Industriels ITT (all in Paris, France), has been elected President of Société française des Électroniciens et des Radioélectriciens for 1966.

Among his predecessors in this position were: Duke de Broglie, Louis Lumière, Edouard Belin, Louis Bréguet, and Prince Louis de Broglie.

Georges Goudet was born in Dijon, France, on 2 June 1912. He received the titles of Agrégé and then Docteur in physical sciences from École Normale Supérieure of Paris. He is an officer of the Legion of Honor.

McNitt Honored by Veteran Wireless Operators Association

At its 41st annual banquet in New York City, the Veteran Wireless Operators Association presented its Marconi Memorial Gold Medal to General James R. McNitt, president of ITT World Communications, in recognition of a "quarter century of valuable contributions to global communications."

Other International System personnel who previously received this award were the late Lee deForest, Admiral Ellery W. Stone, and Haraden Pratt.

Maximum-Likelihood Smoothing

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1. Introduction

The smoothing of a time-varying quantity can be approached in two ways. The more-common way is to consider the time-varying quantity to be composed of a true value contaminated with noise. Statistical properties of the uncontaminated variable and of the noise are then determined. An error criterion is next established, and an operator (often implemented by means of a filter) is established that, within the constraints of the system, minimizes the statistical error in accordance with the error criterion [1].

The other approach, taken in this paper, is to consider the available data to be a set of measurements about the time-varying quantity. Each measurement yields a likelihood function for the measured quantity; that is, each measurement, by itself, allows statistical inference about the measured quantity, which of course depends on the assumed precision of the measurement process. These likelihood functions can then be manipulated so that the time-varying quantity having the maximum likelihood of occurrence is obtained. If the measurements are not all independent, this must be taken into account in the manipulation of the likelihood functions.

There are several important advantages in the maximum-likelihood approach. It facilitates the optimum combination of data from diverse sources, for example, radio measurements and inertial measurements of the position of a vehicle. The data required, namely the likelihood (or distribution) functions for the various types of measurements employed, will often be easier to obtain and more-intuitively meaningful than would the statistical information of the time series involved required for the alternative smoothing technique. And it turns out that the mathematical manipulations involved in the use of the likelihood approach are relatively simple. This is so partly because the elements of this process are more-readily comprehensible, hence it is easier to make reasonable approximations. To illustrate this, a smoothing program implemented for radio-guided vehicles is described as an example.

2. Manipulation of Likelihood Functions

A likelihood function $L(x)$ expresses the likelihood of the occurrence of x; that is, the probability (or probability density) of the occurrence of x is $kL(x)$, where k is a constant. For example, $L(x)$ might be the result of an imperfectly accurate measurement of x. The following properties of likelihood functions follow immediately from the rules for probability.

(**A**) If likelihood functions $L_1(x)$ and $L_2(x)$ are obtained from independent measurements, then the likelihood function of x is $L(x) = kL_1(x)L_2(x)$.

(**B**) If likelihood functions $L_1(x)$ and $L_2(y)$ are obtained from independent measurements, then the likelihood function of z is

$$L(z) = k\int_{-\infty}^{\infty} L_1(z \mp y)L_2(y)\mathrm{d}y \equiv L_1(x)^*L_2(\pm y)$$

where $z = x \pm y$.

Suppose independent measurements are made of quantities x, y, z giving $L(x)$, $L(y)$, $L(z)$. If $z = x - y$, then this information can be used to give combined likelihood functions for x, y, and z as follows

$$L'(x) = [L(x)][L(z)^*L(y)]$$
$$L'(y) = [L(y)][L(x)^*L(-z)]$$
$$L'(z) = [L(z)][L(x)^*L(-y)].$$

If, for example, $L(x)$ is the result of a position measurement at time 1, $L(y)$ of a position measurement at time 0, and $L(z)$ of a measurement of average velocity, then the L' functions give improved estimates of position at times 1 and 0 and of average velocity.

It should be noted that a very-accurate measurement results in a very-narrow likelihood function approaching a delta function, while a dubious measurement results in a broad likelihood function approaching uniform distribution. The convolution of a broad and a narrow likelihood function results in a broad one; the multiplication of a broad and a narrow function results in a narrow one. In the calculation of $L'(x)$, $L'(y)$, and $L'(z)$ above, if two of the original measurements are highly accurate and the third is not, the result will be very-little influenced by the inaccurate measurement. If one of the measurements is not made, this is equivalent to a measurement of zero accuracy, and the results will then be determined solely by the two remaining measurements.

Suppose a particle is known to be traveling with constant velocity, and position measurements are made at times 0, 1, 2, resulting in $L_1(x)$, $L_2(y)$, and $L_3(z)$. There are three simple ways of calculating velocity likelihood, namely

$$\begin{aligned} L_1(v) &= L_2(y)*L_1(-x) \\ L_2(v) &= L_3(z)*L_2(-y) \\ L_3(v) &= L_3(z/2)*L_1(-x/2). \end{aligned}$$

However, the $L_i(v)$ cannot be multiplied together because the $L_i(v)$ are not independent measurements.

The most-general expression for $L(v)$ is

$$\begin{aligned} L(v) = {} & [a(y)L_2(y)*b(x)L_1(-x)] \\ & \times [c(z)L_3(z)*d(y)L_2(-y)] \\ & \times [e(z/2)L_3(z/2)*f(x/2)L_1(-x/2)]. \end{aligned}$$

The following constraints must be satisfied: $a(y)d(y) = 1$; $b(x)f(x) = 1$; $c(z)e(z) = 1$. In other words, a likelihood function $L(w)$ can be split into components whose product equals $L(w)$, and these components can be considered to be independent measurements. The remaining arbitrary functions can be selected to give the best $L(v)$.

The indicated procedure can be applied to more-general problems. However, if the measurement data are in the form of Gaussian likelihood functions, the computations are greatly simplified. The remainder of this paper will be limited to such Gaussian distributions, except where otherwise noted.

3. Gaussian Likelihood Functions

A Gaussian likelihood function can be expressed as $L(x) = ke^{-W(x-m)^2}$ and is thus completely specified by two parameters, namely mean m and weight W. The weight W is related to standard deviation σ by $W = 1/2\sigma^2$. The Gaussian likelihood function will be abbreviated as $L(x; W; m)$.

The following properties can easily be shown for Gaussian likelihood functions by carrying out the indicated operations (remembering that multiplicative constants are arbitrary).

$$\begin{aligned} & L(x; W_1; m_1)L(x; W_2; m_2) \\ & \qquad = L\left(x; W_1 + W_2; \frac{m_1W_1 + m_2W_2}{W_1 + W_2}\right) \end{aligned} \tag{1}$$

that is, if Gaussian likelihood functions are multiplied, the product is Gaussian with weight equal to the sum of the weights of the factors and mean equal to the weighted mean of the factors.

$$\begin{aligned} & L(x; W_1; m_1)*L(\pm x; W_2; m_2) \\ & \qquad = L\left(x; \frac{W_1W_2}{W_1 + W_2}; m_1 \pm m_2\right) \end{aligned} \tag{2}$$

that is, if Gaussian likelihood functions are convoluted to obtain sum or difference likelihood functions, the result has reciprocal weight equal to the sum of the reciprocal weights of the factors, and mean equal to the sum or difference, respectively, of the mean of the factors.

$$L(cx; W; cm) = L(x; W/c^2; m) \tag{3}$$

that is, multiplication of the variable of a Gaussian likelihood function by c multiplies the weight by $1/c^2$. (In other words, if a measurement is rescaled, the expected error must be rescaled correspondingly.)

$$\begin{aligned} & L(x; W; m) \\ & \qquad = L(x; aW; m)L(x; (1 - a)W; m), \\ & \qquad\qquad 0 \leq a \leq 1 \end{aligned} \tag{4}$$

that is, a Gaussian likelihood function can be split into independent components by merely splitting the weight.

4. Example of Maximum-Likelihood Smoothing

The theory developed thus far is adequate for the solution of some practical problems. Given below is an implementation of smoothing of the location of a moving vehicle in real time. A digital computer was available in the system under consideration; hence it was convenient to perform the smoothing numerically. However, the smoothing could readily have been implemented by analog means, had this been a requirement. For simplicity, the problem will be presented here in one dimension only; no essential features of the smoothing process will be lost thereby.

The position of the vehicle under consideration is determined by means of a radio location system. The measurements are repeated at regular intervals T, where T is sufficiently long compared with measurement system time constants so that the measurements at different times can be considered independent. The radio measurement system has an error, the predicted magnitude of which depends in a known way on the location of the vehicle relative to the fixed stations.

If nothing were known about the motion of the vehicle, then the position indicated by the radio location system would be the best estimate of the position of the vehicle at the instants of measurement. If, however, some information is available about the motion of the vehicle from one instant to the next, then this information can be used to improve the position location estimates. In the case of the vehicle under consideration, the vehicle travels with an acceleration (often zero) that is known within some tolerance.

An important consideration is how this tolerance in acceleration, which can be considered an error in a "measurement" of acceleration, is correlated between measurements. In the problem under consideration, the error in acceleration "measurement," which is largely caused by wind effects, was assumed independent from interval to interval. This assumption was made for the following reasons.

(**A**) It leads to very simple implementation. (Note that for any other assumption, the more-complicated procedure of the following section must be employed.)

(**B**) In the absence of detailed data, this assumption is as plausible as any.

(**C**) The equation obtained using this assumption can be modified to guard against the other extreme assumption; namely, that the acceleration error is constant, as will be shown.

It was also assumed that the measurement errors are a Gaussian distribution and for similar reasons:

(**A**) This assumption leads to simple implementation.

(**B**) Measurement errors are usually very close to a Gaussian distribution.

(**C**) One cause of departure from Gaussian distribution is gross errors that are due to some failure. Such an error can be expected to be more nearly uniformly distributed over the limits of the variable than a Gaussian distribution. The results obtained using Gaussian distribution can be modified to guard against gross error due to failure, as will be shown.

Assume now that, at some instant of time, the best estimates of position X_1 and of first difference in X, $V_1 = X - X_1$ and the corresponding weights W_{X1} and W_{V1} are available. At an instant of time T later, a measurement X_M is obtained with weight W_M. The second difference in X, $A_1 = V - V_1$ is also obtained from the intended maneuver (or measured) with weight W_A. There are now two ways of obtaining X at the second instant; namely

$$X = X_M \tag{E1}$$

$$X = X_1 + V_1 + A_1 \tag{E2}$$

and two ways of obtaining V, namely

$$V = X_M - X_1 \tag{E3}$$

$$V = V_1 + A_1. \tag{E4}$$

It should be noted that each of the two means of obtaining X or V involves different information and hence can be considered independent. There is a slight approximation in this statement, since X_1 and V_1 are not completely independent because they are obtained from the same basic data. However, the manipulation of the data to obtain these two quantities is completely different, they are hence very weakly coupled, and the error made in considering them independent therefore is insignificant. The alternative would be calculations involving all the past data rather than simple recursive equations. Making the assumption of independence, the best estimates of X, V, and their respective weights are obtained, from the rules of manipulating Gaussian likelihood functions, as follows.

$$X = \frac{W_{E1}X_M + W_{E2}(X_1 + V_1 + A_1)}{W_{E1} + W_{E2}} \tag{E5}$$

$$W_X = W_{E1} + W_{E2} \tag{E6}$$

$$V = \frac{W_{E3}(X_M - X_1) + W_{E4}(V_1 + A_1)}{W_{E3} + W_{E4}} \tag{E7}$$

$$W_V = W_{E3} + W_{E4} \tag{E8}$$

where W_{E1}, W_{E2}, W_{E3}, and W_{E4} are, respectively, the weights of the quantities in (E1), (E2), (E3), and (E4), and are obtained from

$$W_{E1} = W_M \tag{E9}$$

$$W_{E2} = (W_{X1}^{-1} + W_{V1}^{-1} + W_A^{-1})^{-1} \tag{E10}$$

$$\simeq \frac{W_{X1}W_{V1}}{W_{X1} + W_{V1}} \tag{E10A}$$

for relatively large W_A.

$$W_{E3} = \frac{W_M W_{X1}}{W_M + W_{X1}} \tag{E11}$$

$$W_{E4} = \frac{W_{V1}W_A}{W_{V1} + W_A}. \tag{E12}$$

The quantities X, V, W_X, and W_V are thus recursive and become X_1, V_1, W_{X1}, W_{V1} in the following computation. The weight W_M is time varying and is obtained from the known relation of the geometry between vehicle-measurement stations and measurement error, using unsmoothed values to make this computation. The weight W_A depends both on the maneuver being attempted and on the turbulence of the medium in which the vehicle is operating. In the example under consideration, only two values of W_A were employed, a high value for zero acceleration conditions and a lower value during maneuver.

It should be noted that the weights for position and velocity and hence the accuracy will (in the absence of rapid change in W_M and for relatively high W_A) increase with the number of measurements taken and eventually approach asymptotic values, approximately $W_V \simeq (W_A{}^2W_M)^{1/3}$ and $W_X \simeq (W_M{}^2W_A)^{1/3}$. Of course, the initial values for W_{V1} and W_{X1} are zero.

To deal with a steady error in acceleration, let it be required that an uncompensated steady velocity difference of magnitude a result in a position error that is no greater in magnitude than e.

Let $A_1 = 0$ in (E5) and (E7) while the actual acceleration is a, let there be no other measurement errors, and assume steady-state conditions have been reached (that is, constant e). Then

$$e = X_M - X \tag{E13}$$

$$a = V - V_1. \tag{E14}$$

Substituting (E13) and (E14) in (E7)

$$V = \frac{W_{E3}(X - X_1 + e) + W_{E4}(V - a)}{W_{E3} + W_{E4}} \tag{E15}$$

$$= \frac{W_{E3}(V + e) + W_{E4}(V - a)}{W_{E3} + W_{E4}} \tag{E16}$$

which reduces to

$$W_{E4} = \frac{W_{E3}e}{a}. \tag{E17}$$

For large W_A, a limit on W_{E4} is equivalent to a limit on W_V. If $W_X \gg W_M$, which is the case under steady-state conditions, then $W_{E3} \simeq W_M$. Hence, the position error can be limited approximately to e as a result of constant velocity difference a by limiting W_V to

$$W_V \leq W_M \left(1 + \frac{e}{a}\right). \qquad \text{(E18)}$$

To guard against gross error, two checks can be made. The velocity V (position difference) can be compared with a known maximum value of velocity L. The velocity difference $V - V_1$ can be compared with another limit, which is a function of W_V (in the example under consideration such that the 5σ limit was exceeded). If one of these limits is exceeded, then the assumption is made that a gross error has occurred and that the measurement distributions can no longer be assumed to be Gaussian. Instead of assuming that both ways of computing position (E1 and E2) or velocity (E3 and E4) contribute to the gross error, which is a consequence of Gaussian distribution, a more-likely explanation now is that the error is primarily or completely due to one or the other, but not to both. Which one is at fault is not known at this point, but the first time such a gross error occurs, the more-likely candidate for blame is the current measurement. However, a gross error in stored data, if not removed, is more serious since it will affect many subsequent points. Many schemes for dealing with gross error are possible; the following was employed in the example under consideration.

Calculate

$$C = \frac{K}{W_V[(V - V_1)^2 + (|V| - L)^2]}. \qquad \text{(E19)}$$

If $C \geq 1$, no gross error exists.

If $C < 1$ with frequency below a predetermined value, ignore the implicated X_M (by setting corresponding $W_M = 0$).

If $C < 1$ with greater frequency, mistrust stored X and V and multiply W_X and W_V by C for each such occurrence.

In summary, of the numbered equations, the ones computed in the implementation of the smoothing of the vehicle trajectory are (E5), (E6), (E7), (E8), (E9), (E10A), (E11), (E12), (E18), and (E19).

5. Methods of Treating Nonindependent Measurements

Assume N independent measurements are made of quantities $x_1, x_2, \cdots, x_N$ with nonzero weights $W_1, W_2, \cdots, W_N$. The measured values are $m_1, m_2, \cdots, m_N$. From these measurements it is desired to determine M parameters $y_1, y_2, \cdots, y_M$, where $N > M$. The relation between the x and y is given as

$$x_i = \sum_{j=1}^{M} a_{ij} y_j. \qquad (5)$$

One approach to obtain $L(y_j; W_j; m)$ is to eliminate all y_j, $j \neq J$ from (5) and obtain $N + 1 - M$ expressions for y of the form

$$y_{Jk} = \sum_{i=1}^{N} b_{iJk} x_i, \quad k = 1, 2, \cdots, N + 1 - M. \qquad (6)$$

The best estimate of y_j will be some weighted average of y_{Jk}, with the optimum weights to be determined. This could be done by considering each measurement to be split into $N + 1 - M$ independent measurements with the same measured m values, and weights W_{iK} constrained to have no negative weights and

$$W_i = \sum_{k=1}^{N+1-M} W_{iJk}, \quad i = 1, 2, \cdots, N. \qquad (7)$$

Then

$$L(y_J; W_J; m) = \prod_{k=1}^{N+1-M} \left[L\left(b_{1Jk}x_1; \frac{W_{1Jk}}{b^2_{1Jk}}; b_{1Jk}m_1\right) * L\left(b_{2Jk}x_2; \frac{W_{2Jk}}{b^2_{2Jk}}; b_{2Jk}m_2\right) * \cdots * L\left(b_{iJk}x_i; \frac{W_{iJk}}{b^2_{iJk}}; b_{iJk}m_i\right) * \cdots * L\left(b_{NJk}x_N; \frac{W_{NJk}}{b^2_{NJk}}; b_{NJk}m_N\right)\right] \quad (8)$$

where the W_{iJk} are chosen consistent with the constraints to maximize W_J. Although it is possible to perform this maximization explicitly, it is generally simpler to follow the procedure indicated below.

It can be shown that for linear operations with Gaussian distributions, minimum mean square error implies maximum likelihood and vice versa. Hence, the most-likely value of y can be found by minimizing the weighted mean square measurement error

$$\sum_{i=1}^{N} W_i(m_i - \sum_{j=1}^{M} a_{ij}y_j)^2 \quad (9)$$

with respect to y_j, $j = 1, 2, \cdots, M$. The result will be of the form

$$y_J = \sum_{i=1}^{N} c_{iJ}m_i. \quad (10)$$

The corresponding W_J is obtained from

$$\frac{1}{W_J} = \sum_{i=1}^{N} \frac{c_{iJ}^2}{W_i}. \quad (11)$$

As an example, let it be desired to find the best estimate of the speed of a particle known to travel at constant speed and whose position is measured independently at time 0, 1, 2, 3, 4, 5, giving measured values m_0, m_1, m_2, m_3, m_4, m_5. Suppose the weights of the Gaussian likelihood functions of the measurements are, respectively, W_0, 1, 1, 1, 1, 1. If y_0 is the position of the particle at time 0 and y_1 is the speed, then

$$x_i = y_0 + iy_1, \quad i = 0, 1, \cdots, 5 \quad (5A)$$

and mean square error E is

$$E = W_0(m_0 - y_0)^2 + \sum_{i=1}^{5} (m_i - y_0 - iy_1)^2. \quad (9A)$$

Solving $\partial E/\partial y_0 = \partial E/\partial y_1 = 0$ for y_0 and y_1 yields

$$y_1 = \frac{-15W_0m_0 + \sum_{i=1}^{5} (iW_0 + 5i - 15)m_i}{55W_0 + 50} \quad (10A)$$

and from (11)

$$\frac{1}{W_1} = \frac{225W_0}{(55W_0+50)^2} + \sum_{i=1}^{5} \frac{(iW_0+5i-15)^2}{(55W_0+50)^2} = \frac{W_0+5}{55W_0+50}. \quad (11A)$$

It is interesting to note that for $W_0 = \infty$, the velocity likelihood function can be obtained as

$$L(y_1; W_1; m) = L(x_1 - x_0; 1; m_1 - m_0) \times L\left(\frac{x_2-x_0}{2}; 4; \frac{m_2-m_0}{2}\right) \times L\left(\frac{x_3-x_0}{3}; 9; \frac{m_3-m_0}{3}\right) \times L\left(\frac{x_4-x_0}{4}; 16; \frac{m_4-m_0}{4}\right) \times L\left(\frac{x_5-x_0}{5}; 25; \frac{m_5-m_0}{5}\right)$$

in other words, by using only the differences between x_i and x_0.

For $W_0 = 1$, the velocity likelihood function can be obtained as

$$L(y_1; W_1; m) = \left[L\left(\frac{x_5}{5}; 25; \frac{m_5}{5}\right) * L\left(\frac{-x_0}{5}; 25; \frac{m_0}{5}\right)\right] \times \left[L\left(\frac{x_4}{3}; 9; \frac{m_4}{3}\right) * L\left(\frac{-x_1}{3}; 9; \frac{m_1}{3}\right)\right] \times [L(x_3; 1; m_3) * L(-x_2; 1; m_2)]$$

in other words, only differences between symmetrical data points are utilized.

As another example, consider that the same position measurements are made as in the previous example, but that the acceleration of the particle rather than the velocity is assumed constant, and two additional measurements giving values m_6 and m_7 are made: one of initial velocity x_6 with weight W_6, and one of acceleration x_7 with weight W_7. Let y_0 be the terminal position (corresponding to x_5), y_1 the terminal velocity, and y_2 the acceleration, then

$$x_i = y_0 + (i-5)y_1 + \frac{(i-5)^2}{2} y_2, \qquad i = 0, 1, \cdots, 5$$

$$x_6 = y_1$$

$$x_7 = y_2 \tag{5B}$$

and mean square error E is

$$E = W_0(m_0 - y_0)^2 + \sum_{i=1}^{5} \left[y_0 + (i-5)y_1 + \frac{(i-5)^2}{2} y_2 - m_i \right]^2 + W_6(m_6 - y_6)^2 + W_7(m_7 - y_7)^2. \tag{9B}$$

Numerical solutions for this problem have been obtained for $W_0 = 2.3$, $W_6 = 18$, and $W_7 = 240$. The c_{ij} coefficients and weights W_j are given in the following chart.

j \ i	0	1	2	3	4	5	6	7	W_j
0	−0.072	0.029	0.106	0.198	0.307	0.431	0.771	3.90	2.3
1	−0.142	−0.045	−0.018	−0.020	0.067	0.124	0.135	2.38	18

It should be noted that the weights for initial position and velocity have been so selected that they equal those for the terminal position and velocity. If the acceleration can be assumed to be correlated only over the span of this computation, then the above set of c_{ij} coefficients can be used recursively.

If measurements are made with equipments having memory networks, care must be taken that successive samples assumed to be independent are truly independent. Thus, if a linear filter is used the output of which decays to α in time interval T, and an output x_2 is observed T time units after an x_1 was measured and utilized, then the independent component of the second measurement is $x_2 - \alpha x_1$.

6. References

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On the Relation Between Time, Space, and Holding-Time Distribution Functions

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1. Introduction

The object of this paper is to investigate the conditions under which the principal distribution functions used for the calculations of telephone switching equipment, such as those originated by Bernoulli, Poisson, Erlang, and Engset, are valid. This discussion complements an earlier paper* in this periodical that dealt with the characteristic features of telephone traffic.

Such discussion seems appropriate, not only from a purely theoretical point of view but especially in connection with the calculation of link switching equipment, in which these functions are often used without discrimination and without regard to the traffic conditions implied by the use of a particular distribution function.

In the earlier paper it was explained that two distribution functions are needed for the definition of theoretical telephone traffic. One of these functions relates to the variation of the holding time of the connections, while for the second function one may choose between a time distribution function and a space distribution function.

The former type of function describes the distribution of the moments the calls arrive, and the latter relates to the variation of the number of connections in progress simultaneously.

Further, the earlier paper demonstrated that the following relation exists between the three describing functions, provided that no restrictions are imposed by some network configuration or by some hunting discipline.

$$ {}_2X_c(n, h) = \sum_{k=0}^{c} \sum_{j=k}^{\infty} B_k(j, d) \cdot {}_2X_j(n, h) \times \sum_{i=c-k}^{\infty} B_{c-k}(i, 1-d) \cdot {}_1X_{i,j}(n, h). \quad (1) $$

In this equation the time and space distribution functions are distinguished, respectively, by the suffixes 1 and 2, attached to the symbols X. The functions X are as yet undetermined. The symbol B is explained by the binomial law

$$ B_k(j, d) = \binom{j}{k} d^k (1-d)^{j-k}. \quad (2) $$

The symbol n stands for the expected number of calls arriving per unit of time and may either refer to every single busy hour or express the average value taken over an infinitely large number of hours; h is the average value of the holding time of the connections and ranges between the limits 0 and 1 hour; the variable c indicates the number of connections in progress simultaneously. (See Figure 1.)

The probability that a connection, started at any moment of a period h, does not last beyond this period is represented by d. Its value depends on the specific holding-time distribution function assumed and ranges between 0 and 1. When this function reads $p(> \tau) = f(\tau/h)$, then d is determined by

$$ d = 1 - \int_0^h f(\tau/h) d(\tau/h) \quad (3) $$

where τ is a variable within the period h. For example, if the holding-time distribution function of the connections follows an exponential law, the value of d amounts to e^{-1}.

In order not to burden the text of this paper, a number of lemmas have been grouped in Section 7 that are of general interest when solving the summations to which equation (1) gives rise.

2. General Conclusions

In equation (1) the holding time of the connections appears in two forms: in h, the average duration of the connections, and in d, the average of the holding-time distribution function between the limits h and ∞.

* J. Kruithof, "Theoretical and Practical Aspects of Telephone Traffic," *Electrical Communication*, volume 38, number 2, pages 252–263; 1963.

Consequently the variation of the holding time itself is of secondary importance. A great variety of distribution functions can be imagined that have the same h and d.

A second conclusion follows from the multiplication of both sides of equation (1) by c and by the subsequent summation between the limits $c=0$ and $c=\infty$. At the same time we shall demonstrate the application of lemmas contained in Section 7.

$$\sum_{c=0}^{\infty} c\cdot {}_2X_c(n,h)$$

$$= \sum_{c=0}^{\infty} c \sum_{k=0}^{\infty} \sum_{j=k}^{\infty} B_k(j,d)\cdot {}_2X_j(n,h)$$

$$\times \sum_{i=c-k}^{\infty} B_{c-k}(i,1-d)\cdot {}_1X_{i,j}(n,h)$$

$$= \sum_{j=0}^{\infty} {}_2X_j(n,h) \sum_{i=0}^{\infty} {}_1X_{i,j}(n,h)$$

$$\times \sum_{c=0}^{\infty} \sum_{k=0}^{c} c\cdot B_k(j,d) B_{c-k}(i,1-d).$$

By the application of lemma (5) to the last two summations we obtain

$$\sum_{c=0}^{\infty} c\cdot {}_2X_c(n,h) = \sum_{j=0}^{\infty} {}_2X_j(n,h)$$

$$\times \sum_{i=0}^{\infty} {}_1X_{i,j}(n,h)[jd + i(1-d)]$$

and finally

$$\sum_{i=0}^{\infty} i\cdot {}_1X_{i,j}(n,h) = \sum_{j=0}^{\infty} j\cdot {}_2X_j(n,h). \qquad (4)$$

This result is exemplified by lemmas (9) and (10) and signifies that the traffic quantities described by the time and space distribution functions are equal.

Thirdly, we may mention a check applicable to equation (1). From the consideration that X represents one of a sequence of probabilities, it follows that their sum must be equal to one. When we apply this condition, it should appear that the following relation is valid, without making any commitment concerning the functions X themselves.

We group together the factors comprising the parameter k.

$$\sum_{j=0}^{\infty} {}_2X_j(n,h) \sum_{i=0}^{\infty} {}_1X_{i,j}(n,h)$$

$$\times \sum_{c=0}^{\infty} \sum_{k=0}^{c} B_k(j,d) B_{c-k}(i,1-d) = 1.$$

If we reverse the sequence of summation of the last two summations and introduce the proper limits, we obtain

$$\sum_{j=0}^{\infty} {}_2X_j(n,h) \sum_{i=0}^{\infty} {}_1X_{i,j}(n,h)$$

$$\times \sum_{k=0}^{j} B_k(j,d) \sum_{c-k=0}^{i} B_{c-k}(i,1-d) = 1.$$

As each of these four summations is equal to one, equation (1) is verified.

3. Groups with *n*, *s*, or an Unlimited Number of Outlets

When dealing with the various problems that arise in conjunction with these groups of outlets, a variety of assumptions can be made

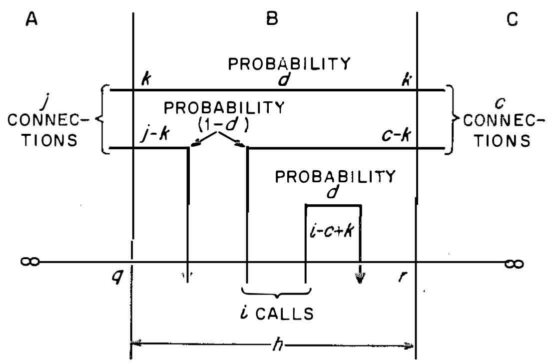

Figure 1—Events and probabilities.

and some additional notions must be introduced, which we believe are best stated in the sections where they appear.

The term "call" has been used exclusively to indicate the moment a "connection" commences.

To bring out the traffic conditions for which the well-known distribution functions are valid, we shall repeat summarily their derivation.

3.1 Bernoulli Law for n Calls

In the first place we shall deal with the straightforward problem, in which exactly n calls arrive during each of an infinitely large number of hours. Within each hour the calls are at liberty to choose any moment for their arrival, while they are independent of each other.

The probability that exactly c calls, neither more nor less, arrive during a period having a duration t, where $0 < t < 1$, equals

$$\binom{n}{c} t^c (1-t)^{n-c} = {}_1B_c(n,t).$$

This equation is not valid if the number of calls per hour varies and if n constitutes an average taken over an infinitely large number of hours, as in such instance the distribution function of n would have to play a role.

To pass from the time to the space distribution function, the duration of the period is made to equal the holding time of the connections. Under these conditions all connections and only those connections that began during the period h are still in existence at the moment the period terminates. Consequently the number of connections then existing will equal the number of calls that arrived during the period, so that their probabilities are the same. The describing time and space distribution functions are identical, so that we obtain

$$\begin{aligned} &{}_1B_c(n,t), \quad \text{for} \quad t = h \\ &= {}_2B_c(n,h) = \binom{n}{c} h^c (1-h)^{n-c}. \quad (5) \end{aligned}$$

This equation for the space distribution function is only valid if the holding time of the connections is constant for all connections; otherwise the distribution function of the holding time would have to enter into it.

We may check this result by means of equation (1). If we introduce equation (5) an identity should appear.

$$\begin{aligned} {}_2B_c(n,h) = \sum_{k=0}^{c} \sum_{j=0}^{n} B_k(j,d) B_j(n,h) \\ \times \sum_{i=c-k}^{n} B_{c-k}(i, 1-d) B_i(n,h). \end{aligned}$$

By the application of lemma $(1C)$, the above is transformed into

$${}_2B_c(n,h) = \sum_{k=0}^{c} B_k(n, dh) B_{c-k}[n, (1-d)h].$$

With constant holding time $d = 0$ and $k = 0$, the above equation is modified to

$${}_2B_c(n,h) = B_0(n,0) B_c(n,h)$$

which is an identity.

Summation of the above product of two Bernoulli functions is also possible if $d = (1-d)$, or if $d = \frac{1}{2}$. In this instance the right-hand part may be written, in accordance with lemma $(3B)$, as

$$\sum_{k=0}^{c} B_k(n, \tfrac{1}{2}h) B_{c-k}(n, \tfrac{1}{2}h) = B_c(2n, \tfrac{1}{2}h)$$

in which instance no identity is obtained. Consequently identity between the time and space distribution functions occurs only if the holding time of the connections has the constant value h.

However, it appears that the value $d = \frac{1}{2}$ has a significant meaning, as appears from the following reasoning.

If we introduce the following time and space distribution functions into equation (1)

$${}_1B_c\left(zn, \frac{h}{z}\right) \quad \text{and} \quad {}_2B_c\left(zn, \frac{h}{z}\right)$$

we obtain in the manner already demonstrated with the help of lemma ($3B$)

$$B_c\left(zn, \frac{h}{z}\right) = B_c\left(2zn, \frac{h}{2z}\right).$$

This equation is true only if $z = \infty$.

The value $d = \frac{1}{2}$, therefore, covers the transition from the Bernoulli to the Poisson distribution functions. The coefficient $\frac{1}{2}$ by which the holding time is multiplied is offset by a doubling of the number of calls, so that the traffic expressed in erlangs remains unaltered.

3.2 Bernoulli Law for s Sources

The reasoning followed in Section 3.1, which proves the identity between the time and space distribution functions of equation (5), holds good only if the connections are continuous, that is, if each connection consists of one continuous duration with no temporary interruptions.

The question may be raised whether some distribution function (of the Bernoulli type) exists that is based on the assumption that some or all of the connections are spaced out over the hour, so that each connection actually comprises groups of associated connections, each group having the same total holding time h.

Because of the relation thus introduced between connections of one group, the concepts of "source" and "occupation time" must be considered.

We assume that there are s sources each with the same occupancy h. Under these conditions the $\binom{s}{c}$ partial probabilities of the following space distribution function are equal and consequently additive.

$${}_2B_c(s, h) = \binom{s}{c} h^c(1 - h)^{s-c}. \qquad (6A)$$

The occupancy of the sources is constant, but the holding time of the actual connections may vary. It appears from equation (6A) that the distribution function of this holding time plays no role in the calculations, so that the scope of equation (6A) is much wider than that of equation (5).

Having abandoned the concepts of "connection" and "holding time," we can no longer indicate time and holding-time distribution functions. The concept of "call" is replaced by "source," and "holding time" by "source occupancy."

However, if we introduce equation (6A) into equation (1) in the manner already indicated, we find that

$${}_1B_c(s, t) = {}_2B_c(s, h), \quad \text{for} \quad t = h. \qquad (6B)$$

This time distribution function behaves, therefore, as if the number of calls is restricted to s connections with a constant holding time h.

We observe that the results in this section agree quite well with those in the preceding section. The relationship between the two cases lies in the fact that the former constitutes the boundary case of the latter. If s is made equal to n (the number of calls), the constant noncontinuous occupation time of the sources is modified to become the constant holding time of the connections.

The Bernoulli law based on sources may therefore be applied only to such practical examples in which the traffic is evenly spread over the inlets. This implies random hunting over the inlets. The Bernoulli equation may not be applied if the inlets, in their turn, are hunted over in a sequential order without slip or reversal.

3.3 Poisson Law

If we introduce the following time and space distribution functions

$$\left.\begin{aligned} {}_1P_c(n, h) &= (nh)^c e^{-nh}/c\,! \\ {}_2P_c(y) &= y^c e^{-y}/c\,! \end{aligned}\right\} \qquad (7)$$

into equation (1), we obtain

$$_2P_c(y) = \sum_{k=0}^{c} \sum_{j=0}^{\infty} B_k(j, d)P_j(y) \times \sum_{i=0}^{\infty} B_{c-k}(i, 1-d)P_i(y).$$

Application of lemma (2C) leads to

$$_2P_c(y) = \sum_{k=0}^{c} P_k(dy)P_{c-k}[(1-d)y].$$

In accordance with lemma (4B), this equation appears to be an identity.

Consequently, the time and space distribution functions for this case are identical if the period to which the time distribution function refers is equal to the average holding time of the connections. The distribution function of the holding time plays no part in this result.

In opposition to this result for the holding time, the distribution function of the average number of calls n may not be chosen in an arbitrary manner. This may readily be realized by the tentative introduction of some arbitrary distribution function, for example where H hours comprise $\frac{1}{2}n$ calls and a second group of H hours $1\frac{1}{2}n$ calls.

One permissible distribution function for the number of calls is if n constitutes a constant number. All hours contain exactly n calls.

Another permissible distribution function appears to be that of Poisson. In this instance $P_i(n)$ indicates the probability that any busy hour forming part of an infinitely large number of hours contains exactly i calls. Here n represents the average number of calls taken over the infinite number of hours considered.

If an hour comprises exactly i calls, the probability of finding c calls within any subperiod of the hour having a duration t amounts to $B_c(i, t)$. The total probability of finding c calls during any subperiod of the infinite number of busy hours equals

$$\sum_{i=0}^{\infty} B_c(i, t) \cdot P_i(n).$$

If lemma (2C) is applied, this probability appears to be equal to $P_c(nt)$.

Combining these two results (the variation of n in accordance with an exponential law, and the arbitrary distribution function for the holding time of the connections) it would seem that y (the traffic expressed in erlangs) may vary in an arbitrary manner.

4. Groups with Fewer Outlets than Inlets (Overflow)

The problems related to limited groups of outlets associated with overflow of traffic are characterized by the following.

(**A**) The traffic originated by the sources is directed to a group of outlets the size of which is—in general—equal to the number of inlets.

(**B**) A limited group of N outlets is divorced from this group.

(**C**) A hunting discipline is imposed on all calls while choosing a free outlet, in that they must give preference to free outlets of the limited group over free outlets of the remainder.

This presentation of the mathematical problem corresponds with the properties of the overflow of traffic, a problem commonly encountered in telephone practice. The agreement between the mathematical image and the practical phenomenon is complete, so that the results obtained by the theory are directly applicable to practical examples or cases.

However, the theory of limited groups associated with overflow is often applied to hunting problems, in which it is not known what happens to those calls that fail to find free outlets among the N outlets of the limited group. In daily telephone practice these calls are said to be lost, but this is merely a way of expression, as the temporary shortage of switching equipment will not annul the subscriber's need for these telephone connections.

Only the vain attempts to obtain a connection are lost. As these unsuccessful calls have practically no holding time, they do not develop into traffic that can be expressed in time or erlangs. Such traffic may obey some time distribution function but in no way can it affect any space distribution function (the functions that serve for the dimensioning of groups of outlets).

The application of equations pertaining to limited groups is therefore permissible only for the calculation of groups of outlets if the probability is of minor magnitude.

4.1 N Outlets Serving s Sources (Engset Equations)

Figure 2 shows by way of example a limited group of $N = 4$ outlets. No attempt has been made to introduce the s sources or inlets. Numbers have been placed with the connections to indicate that sequential hunting has been applied. The outlets are therefore seized in their numerical order. Random hunting within the limited group of N outlets might have been used instead, as the order of hunting within this group has no bearing on the efficiency of the group.

The intervals during which all N outlets are engaged have been shadowed for a purpose that will be explained presently. These intervals contribute to the blocking probability of the group and their durations contribute to the amount of overflow traffic.

As a general symbol ${}_2X_{i \text{ in } N}(s, h)$ is introduced, which indicates the probability of finding i out of N outlets of the limited group engaged, if the originating traffic is offered via s sources each with an occupation time of h hours or erlangs. This symbol is consistent with those already used in the foregoing.

Of the $N + 1$ equations needed to determine the probabilities ${}_2X_{0 \text{ in } N}(s, h) \cdots {}_2X_{N \text{ in } N}(s, h)$, two can be written without much comment. They are

$$\sum_{i=0}^{N} {}_2X_{i \text{ in } N}(s, h) = 1$$

and

$$sh = \sum_{i=1}^{N} i \cdot {}_2X_{i \text{ in } N}(s, h) + (s - N)h \cdot {}_2X_{N \text{ in } N}(s, h).$$

In the latter equation, the first term expresses the traffic offered to the main group of s outlets, the second the traffic carried by the N outlets, and the third the traffic overflowing the limited group.

In the following paragraphs we demonstrate that the second equation constitutes a boundary case of a more-general equation. To do this we shall follow a line of reasoning in which use is made of some peculiarities of sequences of traffic patterns, of which Figure 2 is an example. When reasoning about a specific sequence of traffic patterns, we shall draw some conclusions that hold good for any sequence and consequently for the pertinent probabilities.

Suppose that we eliminate from Figure 2 the shadowed intervals that contribute to the blocking probabilities. We then observe that of the N connections occupying the outlets, one connection causes the commencement of the blocking period, another connection the termination of this period. By the removal of such an interval contributing to the blocking period, the moment of arrival of the former call is postponed till the moment the period ends. But at that moment it will always find a free outlet within a group of $N - 1$ outlets.

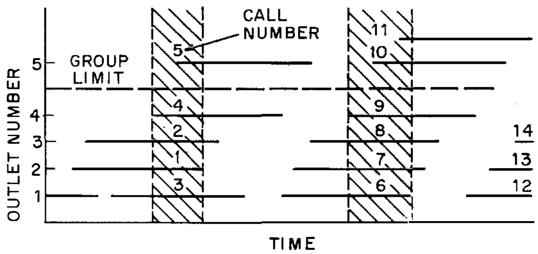

Figure 2—Arbitrary sequence of traffic patterns showing overflow.

This phenomenon occurs in the manner described, for every blocking interval. After the removal of the blocking intervals, we are therefore left with a new sequence of traffic patterns that pertain to a limited group of $N-1$ outlets. The sequence obtained in this manner constitutes a valid contribution to the probabilities pertaining to the $N-1$ group, as we have not disturbed in any way the imposed hunting discipline.

The probabilities of the $N-1$ group are expressed in the probabilities of the N group, which permits us to establish various expressions for the relation between the various probabilities. For our purpose we choose

$$sh[1 - {}_2X_{N \text{ in } N}(s,h)] = \sum_{i=1}^{N-1} i \cdot {}_2X_{i \text{ in } N}(s,h) + (s-N+1)h \cdot {}_2X_{N-1 \text{ in } N}(s,h).$$

The first term of this equation expresses the traffic offered to the $N-1$ group, as the observation period is reduced from one hour to $[1 - {}_2X_{N \text{ in } N}(s,h)]$. The second and third terms represent the traffic carried by the outlets and the overflow traffic.

After some simplification, the same equation becomes

$$(s-N+1)h \cdot {}_2X_{N-1 \text{ in } N}(s,h) = \sum_{i=0}^{N-1} (sh-i) \cdot {}_2X_{i \text{ in } N}(s,h).$$

By similar reasoning we can cover all limited groups of a size smaller than N and obtain as the final result the general equations

$$\left.\begin{aligned} &(s-c)h \cdot {}_2X_{c \text{ in } N}(s,h) = \sum_{i=0}^{c} (sh-i) \cdot {}_2X_{i \text{ in } N}(s,h) \\ &\sum_{i=0}^{N} {}_2X_{i \text{ in } N}(s,h) = 1, \\ &\qquad \text{where} \quad c = 1, 2, \cdots, N. \end{aligned}\right\} \quad (8)$$

These $N+1$ equations provide the answer to the subject problem. For their solution we refer to lemma (11).

The result is the Engset equation, which, if written with the symbols used throughout this paper, is

$${}_2X_{c \text{ in } N}(s,h) = {}_oT_{c \text{ in } N}(s,h) = B_c(s,h) / \sum_{i=0}^{N} B_i(s,h). \quad (9)$$

We use the last letter of Engset's name to distinguish these probabilities from others, reserving the E for the probabilities discovered by Erlang. The suffix o designates overflow in opposition to the suffix d, which is used to indicate delay.

By writing two consecutive equations (8), we obtain the following simple relation between two consecutive probabilities, which may be of some value in calculating tables or curves.

$$\frac{{}_oT_{c+1 \text{ in } N}(s,h)}{{}_oT_{c \text{ in } N}(s,h)} = \frac{(s-c)h}{(c+1)(1-h)},$$

where (10)

$$c = 0, 1, \cdots, N.$$

It will be recognized, on the basis of what has already been stated in connection with the admissible addition of partial probabilities, that the h appearing in equations (8), (9), and (10) represents a nonvariable and not an average. However, it need not be continuous in the sense that we used this word before.

Consequently, it can be stated that the Engset equations are applicable only to those cases in which the traffic offered to the group is equally distributed over the inlets. This imposes the restriction that the inlets, in their turn, should be hunted over in a random manner. Consequently, the Engset equations are not valid if the inlets are hunted over in a sequential manner, that is, if they are seized by the calls in their numerical order.

Another proof of the validity of the Engset equations for constant occupation time of the sources can also be obtained in the manner demonstrated in Sections 3.1 and 4.3, by the application of equation (1).

By increasing the number of sources to a point at which they become equal to the number of calls they carry, an equation resembling that of Engset is obtained, the only difference being that the s of the sources is replaced by the n of the number of calls. At the moment the number of sources reaches the limit n, the holding time of the connections reaches a constant continuous value. However, the practical value of such an equation is very minor, as traffic can in no way be separated from the equipment that carries the conversations.

4.2 Fry Equations

The equations dealt with in this section are commonly referred to as the Engset equations and are consequently attributed to him. They appear on closer examination to have been originated by Fry. This fact was well known among experts but was published for the first time by Syski.

The basic assumption of these equations is said to be that the holding time of the calls flowing over the limited group of outlets equals zero.

To clarify the following discussion of this equation it is reproduced as equation (11), but for greater clarity and uniformity within this paper we have introduced our own symbols.

$$F_{c \text{ in } N}(s, h) = \frac{C_c^s \left\{ \frac{sh}{s - [1 - {}_LF(s, h)]sh} \right\}^c}{\sum_{i=0}^{N} C_i^s \left\{ \frac{sh}{s - [1 - {}_LF(s, h)]sh} \right\}^i}. \tag{11}$$

The symbol ${}_LF(s, h)$ stands for the loss probability caused by the limited group of N outlets. The above equation can be found on page 341 of Fry's book, "Probability and its Engineering Uses." After some simple algebraic modifications the same equation may be written in the form

$$F_{c \text{ in } N}(s, h) = \frac{B_c\left[s, \frac{h}{1 + h \cdot {}_LF(s, h)}\right]}{\sum_{i=0}^{N} B_i\left[s, \frac{h}{1 + h \cdot {}_LF(s, h)}\right]}. \tag{12}$$

The resemblance between this equation and the Engset equation (9) will be noted. The difference is that in the former the occupation time h of the sources is reduced by the factor $[1 + h \cdot {}_LF(s, h)]^{-1}$.

Several objections may be raised against the validity of the equation under discussion. First of all we observe that, although the basic assumption is that the holding time of lost calls is zero, we still find some traffic overflowing the limited group of N outlets.

This traffic amounts to

$$\frac{(s - N)h}{1 + h \cdot {}_LF(s, h)} F_{N \text{ in } N}(s, h)$$

and always has a positive value. On the assumption that the lost calls are cleared, one might reasonably expect to find no lost traffic. As the equation stands, the holding time of the lost calls differs from zero, and the occupation time of the sources and consequently the holding time of all connections (whether lost or carried by the outlets) has been reduced by a factor smaller than one.

Secondly, one might observe that there is no logical reason why the traffic offered to the group should not be reduced a second time to eliminate the lost traffic we still found to be present. This would lead to the absurdity that after repeated application there might be no originating traffic left. The reduction of the traffic offered to the group of outlets leads therefore to a paradox.

Thirdly, in our opinion the assumption that lost calls have no holding time is in a broad sense not quite in agreement with telephone traffic experience. As the demand for telephone service is instigated by the needs of the subscribers, one cannot reasonably expect that the slightest obstruction (for example the 1-in-100 loss) will cause them to abandon their planned conversations with the wanted parties. Each subscriber will simply repeat his attempt until the person is reached. Consequently, the loss of calls will not reduce the traffic expressed in erlangs but simply increase the number of calls. The product of the two, number of calls and holding time, remains unaltered.

Fourthly, if reducing the originating traffic by the lost traffic is justifiable, one would also be entitled to reduce the originating traffic by some 15 percent, which is about the percentage of busy connections. The two types of obstruction (all outlets busy and called line busy) are different in cause but quite similar in effect.

In this connection it may be well to remember that a large part of the present automatic telephone equipment does not distinguish between lost calls caused by a shortage of connecting equipment and calls that find the wanted line engaged. In both instances the subscriber receives busy tone.

It is well recognized that under extreme conditions a subscriber may abandon all hope of obtaining the desired connection, but such strangling effect is not under discussion here.

In a sense one may state that under the actual conditions the calls are not lost but merely delayed, and as a consequence it would appear that equations expressing delay are more appropriate for the calculation of switching equipment, as their basis seems more in accordance with the actual conditions.

4.3 Erlang Overflow Equations

The Erlang overflow equations can be derived from the Engset equations of which they form a special case for $sh = y$ and $s \rightarrow \infty$ and $h \rightarrow 0$. This we intend to do in three steps.

(**A**) For the first of the sequence of Bernoulli probabilities we know that

$$\lim_{s=\infty} B_0(s, y/s) = \lim_{s=\infty} (1 - y/s)^s = e^{-y} = P_0(y).$$

(**B**) For the proportion between two consecutive probabilities we have the relation

$$\frac{B_{c+1}(s, h)}{B_c(s, h)} = \frac{h(s-c)}{(1-h)(c+1)}.$$

For the limit $h \rightarrow 0$ this proportion is equal to $y/(c+1)$, the same as for the proportion $P_{c+1}(y)/P_c(y)$. Consequently we obtain

$$\lim_{s=\infty} B_0(s, h) = P_0(y)$$

$$\lim_{s=\infty} B_1(s, h) = P_1(y)$$

et cetera.

(**C**) Further

$$\lim_{s=\infty} \left[\frac{B_c(s, h)}{\sum_{i=0}^{N} B_i(s, h)} \right] = \frac{\lim_{s=\infty} [B_c(s, h)]}{\sum_{i=0}^{N} \lim_{s-\infty} [B_i(s, h)]}.$$

Finally the Erlang overflow equations are obtained.

$$_{o}E_{c \text{ in } N}(y) = \frac{P_c(y)}{\sum_{i=0}^{N} P_i(y)}. \tag{13}$$

If we introduce this equation into equation (1) for both the time and space distribution functions (a method already applied to similar cases), the following equation should prove to be an identity.

$$P_c(y) \sum_{i=0}^{N} P_i(y)$$

$$= \sum_{k=0}^{c} \sum_{j=k} B_k(j, d) P_j(y)$$

$$\times \sum_{i=c-k} B_{c-k}(i, 1-d) P_i(y).$$

By application of lemma (2A), we obtain

$$= \sum_{k=0}^{c} P_k(dy) \times \sum_{j=k} P_{j-k}[(1-d)y]P_{c-k}[(1-d)y] \times \sum_{i=c-k} P_{i-c+k}(dy)$$

$$= \sum_{k=0}^{c} P_k(dy)P_{c-k}[(1-d)y] \times \sum_{j=k} P_{j-k}[(1-d)y] \sum_{i=c-k} P_{i-c+k}(dy)$$

$$= \sum_{k=0}^{c} P_k(dy)P_{c-k}[(1-d)y] \times \sum_{j=0}^{N} P_j[(1-d)y] \sum_{i=0}^{N-j} P_i(dy).$$

In the last equation we have introduced the limits for i and j. As the lower limit of j can be zero, its upper limit is N. The upper limit of i may not exceed $N-j$ so as to exclude the possibility of being left with probabilities for more than N simultaneous connections.

We now apply lemma (4B) to the first summation of this equation and lemma (14) to the latter two summations and thus obtain the identity looked for.

The Erlang overflow equations therefore satisfy equation (1) without any postulate for the distribution function of the holding time of the connections.

5. Groups with Fewer Outlets than Inlets (Delay)

The reasoning on the sequences of traffic patterns, explained in Section 4.1 and using Figure 2 as an example, applies to any problem of limited groups with hunting discipline and therefore, more specifically, also to the computation of delay probabilities. In this reasoning we have not made any reference to the origin of the calls, whether directly from the sources or from the sources via some intermediary waiting-call circuit equipment. Thus equation (8) also applies in this instance, but with one exception. In the equation describing the limiting case

$$sh = \sum_{i=1}^{N} i \cdot {}_2X_{i \text{ in } N}(s,h) + (s-N)h \cdot {}_2X_{N \text{ in } N}(s,h)$$

the last term, which describes the amount of overflow traffic, becomes equal to zero. It is well understood that with limited groups operated on a delay basis there are overflow calls during the periods all outlets are engaged, but the resulting delay traffic forms no part of the traffic offered to the group. Consequently the last term disappears, with the well-known result that

$$sh = \sum_{i=1}^{N} i \cdot {}_2X_{i \text{ in } N}(s,h).$$

The solution of these $N+1$ equations is facilitated by the aid of equation (10), which in this instance holds good for any value of $c < N$.

This solution is

$$\left.\begin{aligned} {}_dT_{c \text{ in } N}(s,h) &= \frac{B_c(s,h)}{\sum_{i=0}^{N-1} B_i(s,h) + \dfrac{N(1-h)}{N-sh} B_N(s,h)} \\ {}_dT_{N \text{ in } N}(s,h) &= \frac{\dfrac{N(1-h)}{N-sh} B_N(s,h)}{id}. \end{aligned}\right\} \quad (14)$$

With the aid of lemma (11) these equations may be written in the following form.

$$\left.\begin{aligned} {}_dT_{c \text{ in } N}(s,h) &= \frac{(N-sh)B_c(s,h)}{\sum_{i=0}^{N-1} (N-i)B_i(s,h)} \\ {}_dT_{N \text{ in } N}(s,h) &= \frac{(N-Nh)B_N(s,h)}{id}. \end{aligned}\right\} \quad (15)$$

Because of the correspondence between the above equations and the one of Engset, we have not introduced any new symbol for the probabilities shown. We have simply retained the Engset symbol, with the addition of the suffix d to indicate that the limited group operates on a delay basis.

The demonstration that this equation is valid only for constant but not necessarily continuous occupation time of the sources follows again from equation (1).

This demonstration is similar to that contained in Section 3.1, modified in accordance with the demonstration pertaining to the Erlang delay equations given below.

In the manner shown in Section 4.3, equations (14) are transformed for the limiting values $s \to \infty$ and $h \to 0$, while maintaining their product at the value $sh = y$, to

$$\left.\begin{aligned} {}_dE_{c \text{ in } N}(y) &= \frac{P_c(y)}{\sum_{i=0}^{N-1} P_i(y) + \frac{N}{N-y} P_N(y)} \\ {}_dE_{N \text{ in } N}(y) &= \frac{\frac{N}{N-y} P_N(y)}{id} \end{aligned}\right\} \quad (16)$$

which is the solution obtained by Erlang.

These equations may be simplified in the manner stated for equations (14). An alternative method runs as follows: In lemma (12) we lower the value of N by one. We obtain

$$\sum_{i=0}^{N-1} (y - i) P_i(y) = y P_{N-1}(y) = N P_N(y)$$

or

$$\sum_{i=0}^{N-1} \left[\frac{N-i}{N-y} - \frac{N-y}{N-y} \right] P_i(y) = \frac{N}{N-y} P_N(y)$$

or

$$\sum_{i=0}^{N-1} \frac{N-i}{N-y} P_i(y) = \sum_{i=0}^{N-1} P_i(y) + \frac{N}{N-y} P_N(y).$$

Introduction of this result into equations (16) leads to

$$\left.\begin{aligned} {}_dE_{c \text{ in } N}(y) &= \frac{(N-y) P_c(y)}{\sum_{i=0}^{N-1} (N-i) P_i(y)} \\ {}_dE_{N \text{ in } N}(y) &= \frac{N P_N(y)}{id}. \end{aligned}\right\} \quad (17)$$

The first of these two equations may be written in the form

$$\frac{{}_dE_{c \text{ in } N}(y)}{P_c(y)} = \frac{N-y}{\sum_{i=0}^{N-1} (N-i) P_i(y)}.$$

This equation states that the proportions between any Erlang delay probability for $c < N$ and the Poisson probabilities, if the traffics are identical, are equal to the proportion between the average number of idle outlets of the limited group and the average number of idle outlets below N of the main group.

The proof that the Erlang delay probabilities are valid for any holding-time distribution function follows from equation (1). We introduce, as in Section 4.3, equations (16) for both the time and space distribution functions X. We obtain in the manner demonstrated in Section 4.3

$$P_c(y) \left[\sum_{i=0}^{N-1} P_i(y) + \frac{N}{N-y} P_N(y) \right]$$

$$= \sum_{k=0}^{c} P_k(dy) P_{c-k}[(1-d)y] \sum_{j=0}^{N} P_j[(1-d)y]$$

$$\times \left[\sum_{i=0}^{N-j-1} P_i(dy) + \frac{N}{N-y} P_{N-j}(dy) \right].$$

This equation should again prove to be an identity. The limits we introduced are the same as those of Section 4.3.

We apply lemma (4B) to the first summation of the right-hand member and obtain $P_c(y)$. The remaining summations of this member are best written out, when lemma (14) can be applied. In this manner we obtain the identity looked for.

6. Conclusion

In conclusion it can be stated that the validity of the various equations of general application in the calculation of telephone switching equipment mostly depends on the distribution of the originating traffic over the sources. Equations comprising Bernoulli terms are valid only if this traffic is evenly distributed over these sources. Equations comprising Poisson terms are valid for any traffic distribution over the sources, either by random or by sequential hunting. However, the number of sources should be large compared with the number of outlets.

7. Lemmas

For the summation of functions comprising products of Bernoulli and Poisson functions possibly appearing in equations derived from equation (1), the lemmas listed below are of interest. The following symbols have been used.

$$B_a(n, t) = \binom{n}{a} t^a (1-t)^{n-a}$$

$$P_a(y) = y^a e^{-y}/a!$$

X stands for any type of distribution function.

$$a < N < n = 0, 1, 2, 3, \cdots$$
$$b \text{ and } t < 1$$
$$nt = y.$$

Most of these lemmas can be proved by simply writing them out. Where otherwise, the necessary indications are given.

7.1 First Group

$$\sum_{i=a}^{N} B_a(i, b) B_i(n, t)$$

$$= B_a(n, bt) \sum_{i=a}^{N} B_{i-a}\left[n - a, \frac{t(1-b)}{1-bt}\right]. \quad (1A)$$

And the following derived lemmas

$$a = 0, \quad \sum_{i=0}^{N} (1-b)^i B_i(n, t)$$

$$= B_0(n, bt) \sum_{i=0}^{N} B_i\left[n, \frac{t(1-b)}{1-bt}\right] \quad (1B)$$

$$N = n, \quad \sum_{i=a}^{n} B_a(i, b) B_i(n, t) = B_a(n, bt) \quad (1C)$$

$$N = n, \quad a = 0, \quad \sum_{i=0}^{n} (1-b)^i B_i(n, t)$$

$$= B_0(n, bt) = (1 - bt)^n \quad (1D)$$

$$N = n, \quad a = 0, \quad b = 0,$$

$$\sum_{i=0}^{n} B_i(n, t) = 1. \quad (1E)$$

For lemma (1A) we may also write

$$\sum_{i=a}^{N} B_a(i, b) B_i(n, t) = \frac{b^a}{(1-b)^a} (1 - bt)^n$$

$$\times \sum_{i=a}^{N} \binom{i}{a} B_i\left[n, \frac{t(1-b)}{1-bt}\right] \quad (1F)$$

or

$$\sum_{i=a}^{N} \binom{i}{a} B_i(n, t) = \binom{n}{a} t^a \sum_{i=a}^{N} B_{i-a}(n - a, t)$$

or

$$\sum_{i=a}^{N} \binom{i}{a} B_i(n, t) = \frac{(1 - b + bt)^n}{b^a (1-b)^{n-a}}$$

$$\times \sum_{i=a}^{N} B_a(i, b) B_i\left(n, \frac{t}{1 - b + bt}\right).$$

The corresponding group of lemmas comprises Poisson functions.

$$\sum_{i=a}^{N} B_a(i, b) P_i(y)$$

$$= P_a(by) \sum_{i=a}^{N} P_{i-a}[y(1-b)] \quad (2A)$$

$$a = 0, \quad \sum_{i=0}^{N} B_0(i, b) P_i(y)$$

$$= P_0(by) \sum_{i=0}^{N} P_i[y(1-b)] \quad (2B)$$

$$N = \infty, \quad \sum_{i=a}^{\infty} B_a(i, b) P_i(y) = P_a(by) \quad (2C)$$

$N = \infty, \quad a = 0,$

$$\sum_{i=0}^{\infty} (1-b)^i P_i(y) = P_0(by) = e^{-by} \quad (2D)$$

$N = \infty, \quad a = 0, \quad b = 0,$

$$\sum_{i=0}^{\infty} P_i(y) = 1 \quad (2E)$$

$$\sum_{i=a}^{N} \binom{i}{a} P_i(y) = \frac{y^a}{a!} \sum_{i=a}^{N} P_{i-a}(y). \quad (2F)$$

7.2 Second Group

$$\sum_{i=0}^{N} B_i(n_1, t) B_{a-i}(n_2, t)$$

$$= B_a(n_1 + n_2, t) \sum_{i=0}^{N} \frac{\binom{n_1}{i}\binom{n_2}{a-i}}{\binom{n_1+n_2}{a}} \quad (3A)$$

$N = a, \quad \sum_{i=0}^{a} B_i(n_1, t) B_{a-i}(n_2, t)$

$$= B_a(n_1 + n_2, t) \quad (3B)$$

as

$$\sum_{i=0}^{a} \binom{n_1}{i}\binom{n_2}{a-i} = \binom{n_1+n_2}{a}$$

or

$$\sum_{i=0}^{a} B_i(bn, t) B_{a-i}[(1-b)n, t] = B_a(n, t).$$

The corresponding group of lemmas comprises Poisson functions.

$$\sum_{i=0}^{N} P_i(by) P_{a-i}[(1-b)y]$$

$$= P_a(y) \sum_{i=0}^{N} B_i(a, b) \quad (4A)$$

or

$$\sum_{i=0}^{N} P_i(y_1) P_{a-i}(y_2)$$

$$= P_a(y_1 + y_2) \sum_{i=0}^{N} B_i\left(a, \frac{y_1}{y_1 + y_2}\right)$$

$N = a,$

$$\sum_{i=0}^{a} P_i(by) P_{a-i}[(1-b)y] = P_a(y) \quad (4B)$$

or

$$\sum_{i=0}^{a} P_i(y_i) P_{a-i}(y_2) = P_a(y_1 + y_2).$$

7.3 Third Group

$$\sum_{j=0}^{\infty} \sum_{i=0}^{j} j \cdot B_i(n_1, t) B_{j-i}(n_2, 1-t)$$

$$= tn_1 + (1-t)n_2. \quad (5)$$

After reversal of the sequence of summation

$$\sum_{i=0}^{n_1} B_i(n_1, t) \sum_{j=i}^{n_2} j \cdot B_{j-i}(n_2, 1-t)$$

$$= \sum_{i=0}^{n_1} B_i(n_1, t) [\sum_{j=i}^{n_2} (j-i) B_{j-i}(n_2, 1-t)$$

$$+ \sum_{j=i}^{n_2} i \cdot B_{j-i}(n_2, 1-t)]$$

$$= [n_2(1-t) + i] \sum_{i=0}^{n_1} B_i(n_1, t)$$

$$= n_2(1-t) \sum_{i=0}^{n_1} B_i(n_1, t)$$

$$+ \sum_{i=0}^{n_1} i \cdot B_i(n_1, t) = tn_1 + (1-t)n_2$$

$$\sum_{j=0}^{\infty} \sum_{i=0}^{j} j \cdot P_i(y_1) P_{j-i}(y_2) = y_1 + y_2 \quad (6)$$

$$\sum_{i=0}^{\infty} X_i(n, t) = 1 \quad (7)$$

$$B_i(n, t) = B_{n-i}(n, 1-t) \quad (8)$$

$$\sum_{i=0}^{n} i \cdot B_i(n, t) = nt \quad (9)$$

$$\sum_{i=0}^{\infty} i \cdot P_i(y) = y \quad (10)$$

$$\sum_{i=0}^{N} (st - i) B_i(s, t) = (s - N)\, t B_N(s, t). \quad (11)$$

The first part may be modified to

$$\sum_{i=0}^{N} (s - i)\, tB_i(s, t) - \sum_{i=1}^{N} i(1 - t)B_i(s, t).$$

We separate the last term from the first summation. In the latter summation we raise i by one and consequently lower the limits by one.

$$\sum_{i=0}^{N-1} (s - i)tB_i(s, t) + (s - N)tB_N(s, t)$$

$$- \sum_{i=0}^{N-1} (i + 1)(1 - t)B_{i+1}(s, t).$$

The above two summations are equal, so that only the second term remains.

The Poisson counterpart reads

$$\sum_{i=0}^{N} (y - i)P_i(y) = yP_N(y) \quad (12)$$

$$\sum_{j=0}^{N} B_j(bn, t) \sum_{i=0}^{N-j} B_i[(1 - b)n, t] = \sum_{i=0}^{N} B_i(n, t). \quad (13)$$

This lemma can be proved by writing it out and then applying lemma (3B).

$$\sum_{j=0}^{N} P_j(by) \sum_{i=0}^{N-j} P_i[(1 - b)y] = \sum_{i=0}^{N} P_i(y). \quad (14)$$

This lemma can be proved by writing it out and then applying lemma (4B).

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Planning of Telephone Systems Using Small-Diameter Coaxial Cable

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1. Introduction

Several statistical studies, which have been made separately, show that the development of telephone networks in different countries follows an exponential law [1–6]. Even in the highly developed networks of the United States of America and Sweden no sign of a saturation of demand is to be seen. It is forecast that within the next ten years the total telephone circuit length in many countries will be trebled. Although in principle it would be possible to reach this target by a proliferation of present-day techniques, it is easy to show that this is not the right method. When we consider that the demand for channels will continue to increase, a more-economical solution is to use carrier systems more widely.

The network planning engineer must, therefore, have at his disposal a complete range of carrier systems to be able to design an economical network which will allow for future development. The new small-diameter coaxial cable offers considerable advantages in solving this problem.

2. Reasons for Small-Diameter Coaxial Cable

Symmetric and coaxial pairs are both in general use for the transmission of frequency-division-multiplex signals. They have a loss characteristic that increases with frequency, making it necessary to place line repeaters closer to each other along a cable route to accommodate broader frequency bands. There is a remarkable difference, however, in the crosstalk characteristics of both types.

Between symmetric pairs, crosstalk performance is determined by the symmetry of the cables and by the different length of twist of the pairs. By additional balancing measures taken when the cable is being laid, the crosstalk attenuation can be improved further. The basic tendency still remains unaltered, however, that the crosstalk attenuation of symmetric cable pairs is worse the higher the frequency. This is the reason why systems with up to only 120 channels are in use on symmetric pair cables and even these require different cables for the two directions of transmission.

In the case of coaxial cables, crosstalk between pairs at high frequencies is very much less and no special balancing is required. This is because the crosstalk coupling is of a different type. A coaxial line has no electric field in the transverse direction outside the line and thus there is no crosstalk due to this cause. However, longitudinally there is an electric field, because a voltage exists between different points of the outer conductor of the coaxial pair due to the line losses. If several coaxial pairs are combined in the same cable the outer conductors are either in contact with each other or are coupled together by the high capacitance between the outer conductors. At higher frequencies the current flows mainly on the inner side of the outer conductor and therefore the crosstalk attenuation between coaxial pairs has its worst value at low frequencies, becoming better as the frequency increases.

Although the crosstalk attenuation at low frequencies can be improved by various constructional measures, they naturally add to the cost and it is more economical to limit the lowest frequency transmitted—usually to about 60 kilohertz.

Apart from some earlier types of coaxial pairs, the 2·6/9·5-millimetre cable has now been in worldwide use for a number of years. The diameter of the inner conductor of this pair is 2·6 millimetres while the inner diameter of the outer conductor is 9·5 millimetres. This coaxial pair is the subject of a recommendation of the International Telegraph and Telephone

Consultative Committee (CCITT) in which the mechanical and the electrical characteristics are specified [7, 19]. Figure 1 is based on the more-important carrier systems in operation in Europe before small-diameter coaxial-cable systems were designed. The abscissa shows the number of channels of the carrier system while the ordinate is a scale of the copper weight per telephone channel-kilometre in the cable. The weight of the copper in the cable per telephone channel-kilometre has been chosen as an objective measure for the comparison of different types of cables. This scale may also be used to some degree as a measure of the cable price. In this diagram there is an excessive separation between symmetric pair systems and coaxial pair systems. The gap is nearly one decade in number of channels.

From Figure 1, one can clearly see that the gap can be closed only by a coaxial pair system, because the symmetric pair systems are limited to around 120 channels.

If one attempted to close the gap by using a 2·6/9·5-millimetre coaxial pair, the copper per telephone channel-kilometre would be more than in the 120-channel system on symmetric pairs and would without doubt lead to an uneconomic system.

From these considerations, one can see why a new system had to be developed.

Figure 2 now shows how the gap is closed by the small-diameter coaxial-cable systems. The result is a range of systems with numbers of channels that are nearly uniformly distributed on a logarithmic scale. It is clear that future provision of 960-channel and 1260-channel systems will be on small-diameter coaxial cable.

3. Basic Considerations of System Characteristics

The small-diameter coaxial-cable technique has passed through several stages during its development up to its present point [8–10]. The cable itself has been considerably improved and it now appears that the developments of the different manufacturers have reached such a point that, apart from differences in the construction of the inner conductor insulation, the electrical characteristics are very similar [11]. The International Telegraph and Telephone Consultative Committee has, therefore, been able to describe these characteristics in a recommendation [12, 19].

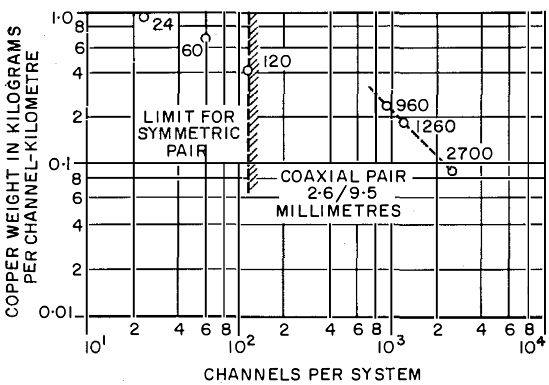


Figure 1—Weight of copper per kilometre length of telephone channel as a function of the indicated number of channels in the carrier system. This represents the more-important European broad-band carrier systems on cable before the introduction of small-diameter coaxial-cable systems.

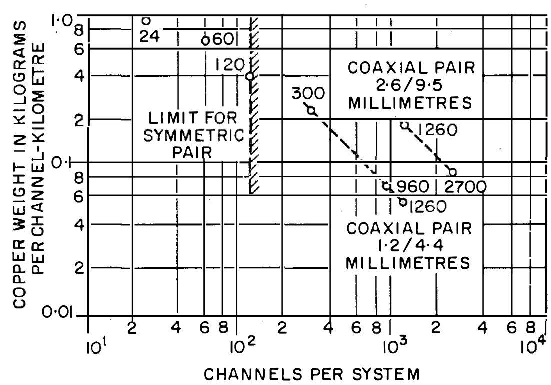


Figure 2—Relation between small-diameter coaxial-cable systems and other broad-band carrier systems.

3.1 Optimization of System Costs

After settling on the construction of the small-diameter coaxial pair, the main point now to be decided on is the diameter. Only the inner or the outer conductor can be chosen because the ratio between both should be 3·6.

In the equation for the line loss per kilometre

$$\alpha = 8{\cdot}686(\epsilon\rho\pi)^{1/2} f^{1/2} \times \frac{2}{d_a} \frac{\{(D/d)+1\}}{2\log_e (D/d)}, \text{ decibels/kilometre}$$

where

D = inner diameter of the outer conductor in millimetres
d = outer diameter of the inner conductor in millimetres
f = frequency in hertz
ρ = resistivity of conductor material in ohm-millimetre^2/kilometre
ϵ = dielectric constant of the insulation between conductors.

The expression $\{(D/d)+1\}/\{2\log_e (D/d)\}$ has a minimum at $D/d = 3{\cdot}6$.

The greater the diameter the smaller is the attenuation at a given frequency. This also means that fewer repeaters need to be inserted in the line. A large diameter of the coaxial pairs also means higher cable costs and lower equipment costs. By a suitable selection of the diameter, the total costs of cable and equipment can be brought to a minimum.

The first problem is to define which costs should be minimized. Should it be the initial costs, (that is, the price of the installed connection) or should it be related to the annual charges? Other important factors are maintenance costs, repair costs, power consumption, the paying of interests, et cetera. It is clear that in different countries these calculations would not give the same result.

Generally it can be said that the most economic cable diameter increases with the number of channels in the system. From the manufacturing point of view however it is not economic to have a large number of alternative sizes due to the high cost of cable manufacturing machines, and hence an effort is made to keep the number of different types of coaxial pairs as low as possible. At present only two types are recommended by the International Telegraph and Telephone Consultative Committee.

3.2 Power Feeding of Dependent Repeaters

In coaxial systems designed to achieve optimum economy, the repeater sections will in general be quite short. It is, therefore, vital that the power for these repeaters be transmitted over the cable itself from power feeding stations which, in the European area, should preferably be greater than 70 kilometres (45 miles) apart.

It is also clear that transistor systems having a low power drain would be essential. Transistor repeaters offer the further advantage of needing only one supply voltage. For this reason the repeaters would not require special supply circuits other than the usual power feed separation filter, if a direct-current power feeding system is used.

The power feeding current is generally transmitted on the inner conductors of the coaxial pairs because induction from other neighbouring circuits is less on the inner conductors and because the inner regions of the cable are better protected from accidental short-circuits.

3.3 Power Feeding Circuit

The power feeding circuit consists of the two inner conductors of both coaxial pairs comprising one carrier path. Each path has its own circuit. Thus the total reliability is increased, because in most cases a failure will interrupt only a part of the channels carried on one cable. Calculations show that with direct-current power feeding it is better to connect repeaters in series if more than 3 or 4 repeaters are to be supplied with power from one feeding point.

To gain an understanding of the power feeding problem, the different factors that influence the selection of the diameter of the coaxial pair will be presented.

The highest power feeding voltage that is allowed at the power feeding station depends on the diameter of the coaxial pair. In Figure 3 this relation is given. This is not a definite limit, because the necessary safety margins are a matter of agreement. The curve can be described by

$$E_{\mathrm{PF}} = E_1 d$$

where E_{PF} = highest power feeding voltage allowed

E_1 = highest voltage allowed for an inner conductor with a diameter of 1 millimetre

d = diameter of the inner conductor in millimetres.

When studying this curve it must be remembered that in the case of cable faults the whole power feeding voltage can be applied to one cable pair. The maximum voltage is limited to about 1500 to 2000 volts because at these voltages corona discharges may occur and produce a high noise level in the telephone channels of the system.

In these calculations a distance is assumed of 100 kilometres (60 miles) between the power feeding stations with a transformer at the halfway point to separate the power feeding circuits. Figure 4 shows the resistance of the inner conductor loop as a function of the diameter of the coaxial pair. The corresponding equation is

$$R_s = R_1/d^2$$

where R_s = resistance of inner conductor loop

R_1 = resistance of the loop with a conductor 1 millimetre in diameter

d = diameter of the inner conductor in millimetres.

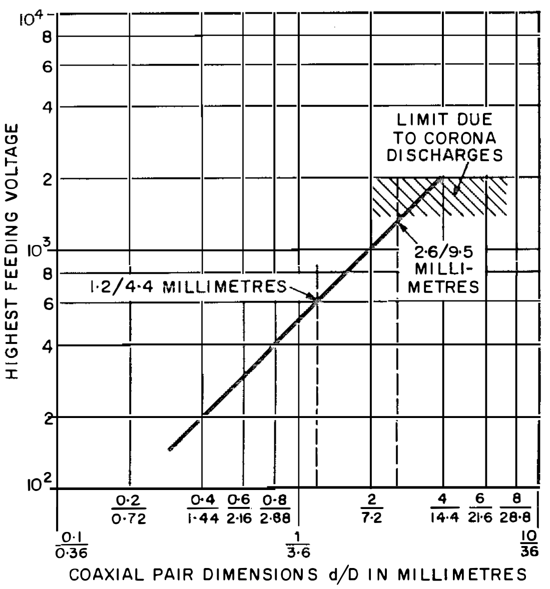


Figure 3—Highest allowable feeding voltage at the feeding point as a function of the dimensions of the coaxial pair.

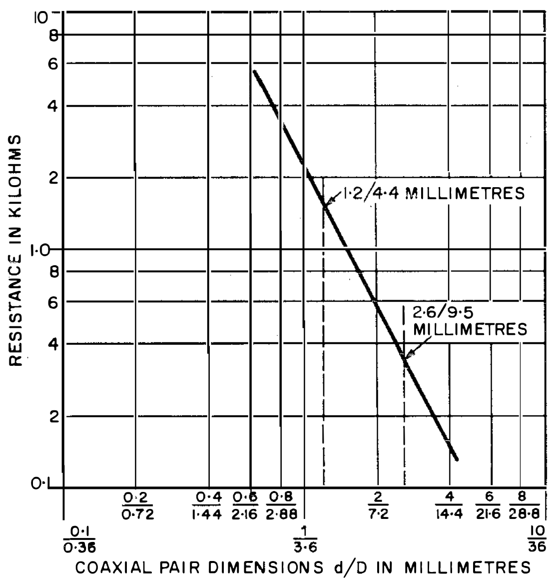


Figure 4—Loop resistance of the inner conductor of a coaxial pair for a 50-kilometre section as a function of the dimensions of the coaxial pair.

In this circuit in the case of power matching, the following power is available for all dependent repeaters.

$$N_{\max} = E_{PF}^2/(4R_s)$$
$$= d^4E_1^2/(4R_1).$$

In the following expression we define $N'_{\max}$ as $\frac{1}{2}N_{\max}$, which is the total power available to the line amplifiers for one direction of transmission on a 50-kilometre (30-mile) link. Thus

$$N'_{\max} = d^4E_1^2/(8R_1)$$

and is shown in Figure 5. The values for E_1 and R_1 have been taken from Figures 3 and 4.

In the case of power matching, half of the power feeding voltage can be used for the line amplifiers, the rest being lost in the resistance of the inner conductor loop. At the present state of the art of amplifier design, 10 volts is about the smallest supply voltage for an amplifier with suitable characteristics for carrier systems. Assuming these conditions a scale has been drawn in Figure 5 showing the greatest possible number of line amplifiers in a 50-kilometre (30-mile) link for power matching. If more than 10 volts is required for an amplifier, the number will be smaller. The figures in parentheses are the maximum numbers of line amplifiers and these figures are followed by the corresponding shortest repeater spacings in kilometres.

If the amplifiers do not need the greatest possible power, the current can be reduced and the voltage drop on the inner conductor loop can be smaller. Then more line amplifiers each with a 10-volt supply can be inserted in the loop without exceeding the allowable feeding voltage. As an example a curve is also drawn in Figure 5 giving the power for 25 line amplifiers if these amplifiers do not need the maximum power. For this curve the power feeding circuit is always fed with the highest allowable voltage shown in Figure 3. The curve is, therefore, also a limit, preventing the use of the left-hand area. If higher supply voltages for the amplifiers are necessary, the curve will move towards the top right-hand side of the diagram. The curve for 25 amplifiers in a 50-kilometre link has been chosen, because 2 kilometres is the repeater spacing of a proposed 2700-channel system on small-diameter coaxial cable (1·2/4·4 millimetres). It would appear that in the present state of the art, the use of a cable of smaller dimensions than this would have limited power feeding capability for long-distance systems.

3.4 Protection Against Lightning and Power Surges

Transistor repeaters on small-core cable, without special protection, are in general rather

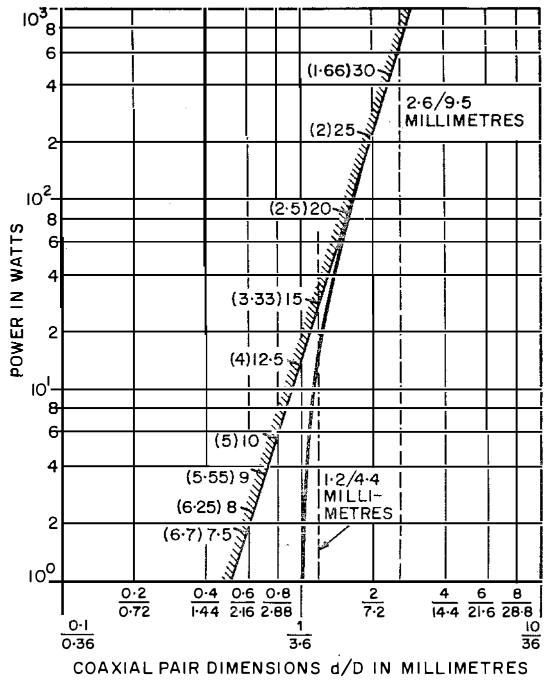


Figure 5—Remote feeding power available to the line amplifiers for a 50-kilometre (30-mile) section for only one direction of transmission. A minimum of 10 volts is provided for each line amplifier. The maximum number of line amplifiers is given with the corresponding spacing in kilometres (in parentheses). The solid curve is for 25 amplifiers spaced 2 kilometres (1·6 miles) apart.

vulnerable to induced electromotive forces from neighbouring power cables and lightning discharges [13, 14].

This is due partly to the greater sensitivity of semiconductors to transient overloads and partly because less metal is used in the cable sheath, which in turn has a higher resistance, allowing more of the induced current to flow in the coaxial tubes themselves.

With a direct-current series power feeding circuit, induced voltages accumulate over the whole power feeding circuit. But there are several procedures that can be adopted to reduce the dangers involved.

(A) Equipment can be installed at a distance from high-power circuits to considerably reduce the effects of induced voltages.

(B) The induced voltage on the coaxial pair can be reduced by a suitable sheathing and armouring of the cable.

(C) By preventing any connection to earth of the inner conductor as well as of the outer conductor, the current induced into the inner conductor can be considerably reduced.

(D) The length of the cable susceptible to induction can be shortened by establishing two independent power feeding stations. The two circuits are insulated from each other by line transformers capable of withstanding high voltages.

(E) By the installation of discharge devices at suitable points along the route.

In systems on the market, different combinations of these possibilities are in use. In all cases additional measures are necessary in the dependent repeaters to protect them against overvoltage.

3.5 Number of Channels and Repeater Section Length

When the dimensions of the coaxial pair are fixed, the next question is what should be the number of the channels carried by this pair. Many studies and investigations made on an international level have recommended that systems should have 300 and 960 or 1260 channels (Recommendations G.341, G.343, and G.344) of the International Telegraph and Telephone Consultative Committee [15].

In some situations it would be advantageous to be able to change from one system to the other when the demand arises for more circuits and hence the repeater spacings of such systems should be such that most of the repeater points can be used again. The situation is simplified if the 300-channel system has double the repeater spacing of the 960- or 1260-channel system.

Let us define a system (designated system 1), in which the repeater section loss at the frequency of the highest channel f_1 is a_1 and the line amplifiers fulfill definite requirements of thermal noise and harmonic distortion.

If this system is replaced by another system (system 2), the repeater spacing of which is only half that of the first system, then the number of the sources for thermal noise is doubled since each repeater contributes to the noise. To get the same signal-to-thermal-noise ratio at the end of the link, the repeater input level of the highest channel must be 3 decibels higher in the second system, if the repeaters of both systems have the same noise factor. The line attenuation is greatest in the highest channel and neglecting some irregularities at lower frequencies is proportional to the square root of the frequency. The repeater section attenuation a_2 of system 2 at the top frequency f_2 is, therefore:

$$a_2 = a_1 \cdot \tfrac{1}{2} (f_2/f_1)^{1/2}.$$

Figure 6 is derived from this calculation and shows the necessary changes of the transmit level of the second system compared with the first. The basis for this calculation is that of keeping the signal-to-thermal-noise ratio constant at the end of the link. The curves are given as a function of the repeater section attenuation a_1 of the first system, for example, a 300-channel system. Each curve belongs to a given frequency ratio f_2/f_1. It can be seen that by the alteration from a 300-channel system with 6-kilometre (3·7-mile) spacing to a 1260-channel system with a 3-kilometre (1·9-mile) spacing ($f_2/f_1 \approx 4$), the transmit level of the highest channel must be increased by 3 decibels. The transmit level of a 960-channel system with 4-kilometre (2·5-mile) spacing can be reduced by 2·6 decibels compared with a 300-channel system with 8-kilometre (5-mile) distance between the repeaters ($f_2/f_1 = 3 \cdot 1$).

By halving the repeater sections, the number of sources producing harmonic-distortion noise are doubled. To compensate for this, either the transmit level of system 2 must be 3 decibels lower or the linearity of the second system must be the same degree better. The 3-decibel change is enough both for second- and third-order harmonic distortion, because the second-order distortion adds mainly on a power basis and varies decibel by decibel with transmit level changes while the third-order distortion varies 2 decibels with a 1-decibel level change and adds mainly on a voltage basis.

If the number of channels is high enough (>240) the equivalent power is proportional to the number of channels and, therefore, is also nearly proportional to the frequency of the top channel. On the other hand the harmonic-distortion noise power arising in a telephone channel is independent of the number of channels if the equivalent power of the system is kept constant. Thereby it is assumed that the non-linearity is independent of frequency. Because the equivalent power increases in proportion to the number of channels, the repeaters of system 2 must meet the harmonic-distortion requirements of system 1 at a measuring level that is higher by the factor $\frac{1}{2} \log_e (f_2/f_1)$.

The result of these considerations is given in Figure 7. This figure explains at what level in system 2 the harmonic distortion (which is required for system 1) must be measured. (The following assumptions have been made: Half the repeater section of system 1 in system 2, the same thermal noise factor in both systems). Again the curves are given as a function of the repeater section loss a_1, for example, a 300-channel system, for several values of frequency ratio.

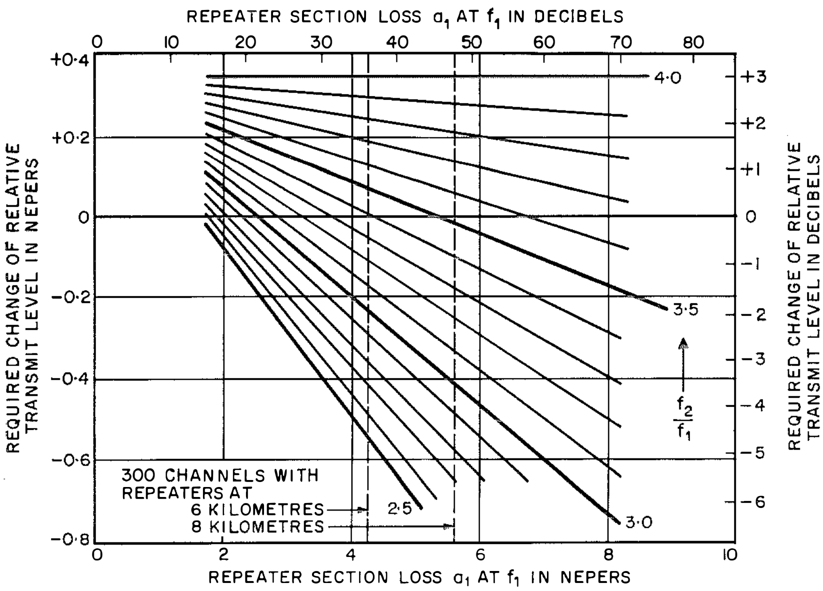


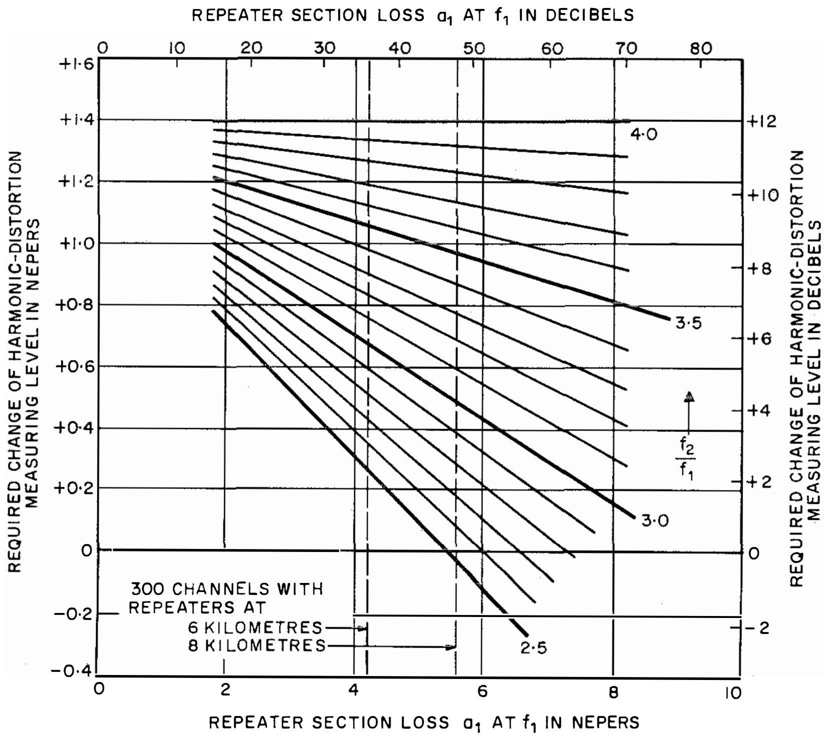
Figure 6 — Required change of the relative transmit level of the highest channel when replacing system 1 (subscript 1) by system 2 (subscript 2) having double the number of repeaters at half the original spacing on the same coaxial pair. Repeater noise factors are equal. Line loss varies as $f^{1/2}$.

In Figure 7 it can be seen that in a 1260-channel system with 3-kilometre (1·9-mile) repeater spacing the harmonic distortion must be 12 decibels better than in a 300-channel system with 6-kilometre (3·7-mile) spacing. The requirement of a 960-channel system with 4-kilometre (2·5-mile) spacing is only 5 decibels higher than that of a 300-channel system with 8-kilometre (5-mile) repeater spacing. To put it another way: if the harmonic distortion is always given for the relative transmit level of the highest telephone channel and comparing the latter two systems, the second-order-harmonic attenuation a_{k2} must be higher by 2·6 decibels + 5·2 decibels = 7·8 decibels (Figures 6 and 7). The third-order-harmonic attenuation a_{k3} must be higher by 2 × (2·6 decibels + 5·2 decibels) = 15·6 decibels. Effects of different pre-emphasis characteristics of the systems have not been taken into account in this calculation.

Figure 7 further shows that the "natural" frequency ratio f_2/f_1 for halving the repeater section is smaller than 3:1 in the range of interest of a_1. Besides this the number of channels in the systems have been chosen with 300 and 960 or 1260. This is mainly because existing systems had to be taken into account and because the systems must have frequency allocations combined from groups, supergroups, master groups, et cetera.

It can also be seen that in the case of the two systems with 300 channels and 6-kilometre (3·7-mile) spacing and with 1260 channels and 3-kilometre (1·9-mile) spacing, the 1260-channel system presents considerably greater problems to the designer than the 300-channel system. The two other systems with 300 and 960 channels and 8- and 4-kilometre (5- and 2·5-mile) repeater spacing respectively, offer about the same degree of difficulty. On the other hand, a comparison of the 1260-channel system with 3-kilometre (1·9-mile) spacing and the 960-channel system with 4-kilometre (2·5-mile) repeater spacing would show that these two

Figure 7 — Required change of harmonic-distortion measuring level when replacing system 1 (subscript 1) by system 2 (subscript 2) having double the number of repeaters at half the original spacing on the same coaxial pair. Repeater noise factors are equal. Line loss varies as $f^{1/2}$.



systems require about the same effort for their design.

Calculated from Figure 6, Figure 8 shows the changed requirements for the overload point. From this the power needed for the repeater can be estimated. These curves give only approximate values, because the exact values depend on the pre-emphasis and the margin necessary for level variations and equalization errors.

4. Other Special Features

After having given in the previous sections the ideas and considerations that have led to the small-diameter coaxial-cable technique, some special features from the planning of the new systems will now be described.

One characteristic is common to all the small-diameter coaxial-cable systems described in the following articles [16–18]. This is that the dependent repeaters are remote power fed and are designed to be used in underground repeater stations. This new method of locating the repeaters has been made possible only by the use of transistors. This means that no extra mains supply needs to be installed in the repeater stations, not even for measuring apparatus, which also use transistors and operate from a battery supply. The underground location of the repeaters brings savings in costs for ground and buildings and allows more freedom in routing the cable link.

The underground location of the repeaters would not be advantageous if it were not possible to make the repeaters sufficiently reliable that maintenance work at regular intervals is not required. With the exception of one adjustment during the installation of the repeater no further adjustment is required. All the additional facilities for equalization are provided in the power feeding stations, that is, the terminal and main repeater stations.

Due to the underground location of the dependent repeaters it is necessary to provide at the

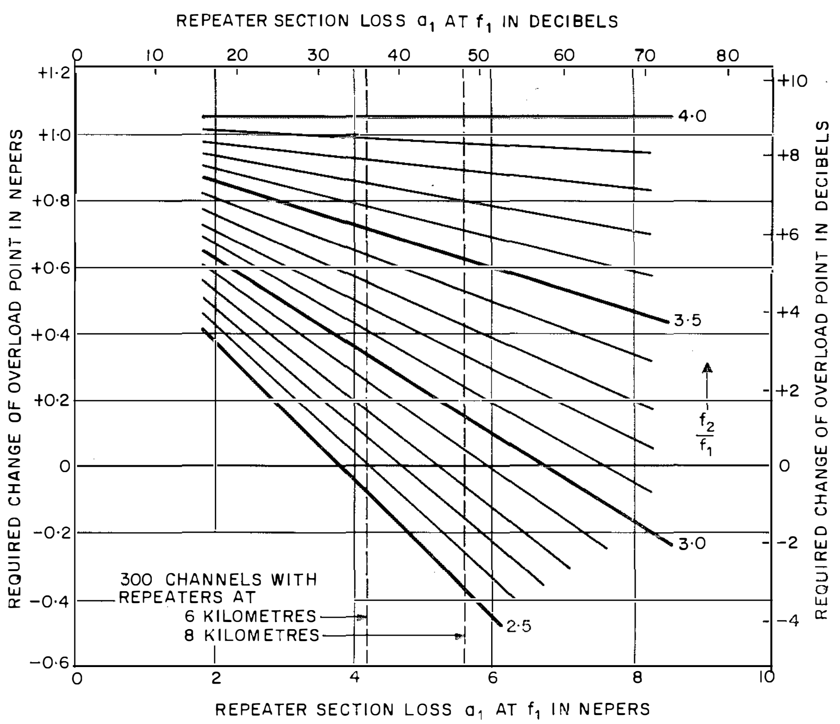


Figure 8 — Required change of the overload point requirement when replacing system 1 (subscript 1) by system 2 (subscript 2) having double the number of repeaters at half the original spacing on the same coaxial pair. Repeater noise factors are equal. Line loss varies as $f^{1/2}$.

power feeding station means of locating a fault in a repeater or in the cable whether this fault has interrupted the power feeding circuit or not.

Small-diameter coaxial-cable systems operate with pre-emphasis. This means that the relative transmit levels of the different channels are not equal, the channels of the upper frequencies being transmitted with a higher level than the channels at lower frequencies. The transmit levels of the highest-frequency channels are determined by the thermal noise of the repeater inputs, whereas the levels of the lower-frequency channels, being less attenuated by line loss, are less influenced by the thermal noise. The transmit level of these lower channels can therefore be reduced resulting in a lower requirement for the overload point and for harmonic distortion.

The small-diameter-cable systems are designed to meet fully all the recommendations of the International Telegraph and Telephone Consultative Committee concerning thermal noise, non-linear distortion noise, crosstalk, level tolerances, frequency allocations, interconnection levels, et cetera.

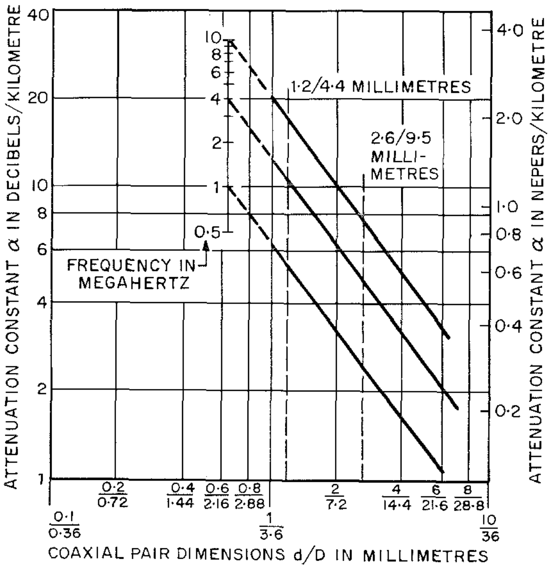


Figure 9—Attenuation constant of a coaxial pair with copper conductors as a function of size and frequency.

5. Use of the Equipment on Other Coaxial Pairs

It has been mentioned earlier in this article that the attenuation of a coaxial pair is proportional to the square root of the frequency. This is only correct if the frequency is high enough or the outer conductor is thick enough. For the present types of coaxial pairs the dielectric constant of the inner conductor insulation is nearly a constant value. Figure 9 has been drawn with these assumptions. It shows the relation between the attenuation constant, frequency, and dimensions of the coaxial pair. (Copper as the conducting material.) Below 1/3·6 millimetres the curves are only drawn as dotted lines, because with these dimensions the dielectric constant required could not be realized.

At lower frequencies the thickness of the outer conductor is a decisive factor and gives rise to different attenuation characteristics. The effect of screening the outer conductor with mild-steel tapes can also be seen on the low-frequency attenuation characteristic. Generally, equipment designed for a definite type of coaxial pair can also be used on other coaxial pairs if the error at the lower frequencies is corrected in the equalizers. With another coaxial cable the repeater section length is different, for example, repeaters of the 960-channel system with 4-kilometre (2·5-mile) repeater spacing on the small-diameter coaxial pair (1·2/4·4 millimetres) can also be used on the large coaxial pair (2·6/9·5 millimetres) with a spacing of about 9 kilometres (5·6 miles).

There is an advantage with this change to a thicker coaxial pair. With longer repeater sections the number of the sources of thermal and non-linear distortion noise is reduced, resulting in a better signal-to-noise ratio than the same link with small-diameter cable. However, before using equipment designed for the small-diameter cable on another coaxial pair, care must be taken to ensure that this cable is suitable for the power feeding system, fault location method, and earthing scheme.

6. Realization of the Equipment

The following three articles in this issue describe the equipment for the small-diameter coaxial-cable systems designed by Standard Elektrik Lorenz (Germany), Standard Telephones and Cables (Great Britain), and Standard Telephon und Radio (Switzerland).

These new products are all based on the same studies, which are outlined in this article. There are several reasons why it was necessary to design within the ITT System three different realizations of the same basic idea.

Each of the three manufacturing companies had to make the new systems in accordance with the requirements of its main customer, its corresponding telephone administration. These requirements include not only the special equipment practice standardized on a national basis, but also some differing level values and differences in the methods to ensure the safety of the maintenance personnel. The rules for the maintenance of the equipment also differ. There are further some differences in the general form of construction of cable networks, for example, in the field of cable pressurization, use of cable ducts, or in the use of combined cables. Finally there are some differences between the administrations as to the fields in which these systems can be used to best advantage. But the deciding reason for making three versions has been the requirement of each administration that the equipment must be able to work with the equipment of their other suppliers without any difficulties. In some cases this requirement is extended to such a degree that it must be possible to exchange the equipment subunits from different suppliers. The three new systems are described in separate articles. Although this results in some repetition, it means, on the other hand, that the systems are more-fully presented.

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He began his career as a student apprentice with Standard Telephones and Cables in 1944 and left in 1947 to do his National Service in Royal Electrical and Mechanical Engineers. After this service he worked in Canada on the development of microwave links for the Canadian General Electric Company.

In 1955 he joined Standard Telecommunication Laboratories, where he is now a project leader responsible for the Linear Systems Laboratory.

Mr. Barber was elected an Associate Member of the Institution of Electrical Engineers in 1957.

Multichannel Telephone Equipment of Standard Elektrik Lorenz for Small-Diameter Coaxial Cable

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1. General

The *V300* and *V960* carrier frequency line equipments for the small-diameter (1.2/4.4-millimeter) coaxial pairs were developed to meet not only the recommendations of the International Telegraph and Telephone Consultative Committee for high-quality speech circuits but also the requirements of the German Post Office for equipments supplied to it by various German manufacturers.

1.1 Frequency Allocations

1.1.1 *V960*

As shown in Figure 1, the *V960* system is capable of transmitting 960 channels in 16 supergroups in the frequency range from 60 to 4028 kilohertz or 3 master groups between 316 and 4188 kilohertz for 900 channels (*V900*). The frequencies of 60 and 4287 kilohertz serve as line pilot signals. Since the fault-location method uses pulses above these frequencies, an overall transmission bandwidth from 60 to approximately 4700 kilohertz is provided.

1.1.2 *V300*

The frequency allocation of the *V300* system for 300 channels is shown in Figure 2. The system is capable of transmitting 5 supergroups in the range from 60 to 1300 kilohertz or 1 master group between 64 and 1296 kilohertz. The higher-frequency line pilot is at 1364 kilohertz. Here again fault location enlarges the required bandwidth to a maximum frequency of approximately 1500 kilohertz.

1.2 Branching-off Facilities

At main repeater stations it is possible, with both types of line equipment, to drop any supergroups without reinjection, and to extract and reinsert supergroups 1 and 2 (60–552 kilohertz). Moreover, in the case of the *V960* main repeater stations, extraction and reinjection of supergroups 1 through 5 (60–1300 kilohertz) or supergroups 7 through 16 (1556–4028 kilohertz) may be carried out.

1.3 Equalization

The repeaters compensate for the attenuation in the small-diameter (1.2/4.4-millimeter) coaxial pair. Except for the lowest frequencies of the transmitted band, this attenuation is approximately proportional to the square root of the frequency. It has the following values at +10 degrees Celsius.

Frequency in Kilohertz	Nepers per Kilometer	Decibels per Mile
60	0.17	2.4
1300	0.7	9.7
4028	1.22	17.0

The attenuation is also temperature dependent and increases by about 0.2 percent per degree Celsius except at the lowest frequencies where it is up to 0.28 percent per degree Celsius. A temperature rise, therefore, appears as an increase in line length.

Attenuation as a function of cable length can be corrected by a fixed equalizer, but the temperature-dependent attenuation changes must be

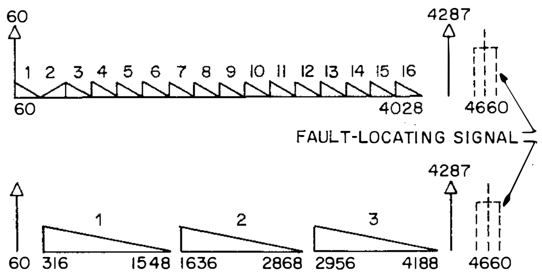

Figure 1—The *V960* system shown at top provides for 16 supergroups of 60 channels each between 60 and 4028 kilohertz. The lower diagram is for the *V900* system of 900 channels in 3 master groups. Fault-location signals are in the 4660-kilohertz region as indicated. Pilots are at 60 and 4287 kilohertz.

corrected by an equalizer which varies in sympathy with these changes. In Central Europe the mean temperature of a cable buried at a depth of 0.8 meter (32 inches) under the ground is about +10 degrees Celsius. At this depth the temperature changes by some ±8 degrees Celsius between summer and winter; daily variations being insignificant.

Transistors permit the dependent repeaters to be operated unattended and to be installed underground. By suitable heat insulation, the ambient temperature of the repeater and of the cable are approximately equal. This permits the use of a temperature-sensitive resistor to control gain. The small residual temperature difference between the internal temperature of the repeater housing and the ambient temperature of the cable can easily be accommodated in the design of the equalizer network. The cost of a temperature-controlled repeater is considerably less than of an automatic pilot-regulated repeater.

One cannot, however, completely give up the pilot-regulated repeater, since the temperature-sensitive device is only able to control the amplifier gain but not to regulate it. For this reason, a certain number of temperature-controlled repeaters must be followed by a pilot-regulated repeater. This avoids an accumulation of temperature-dependent equalization errors, resulting in the dependent repeaters being operated at the wrong output level. In any case, a pilot-regulated amplifier is equipped at the terminal or at main repeater stations above ground.

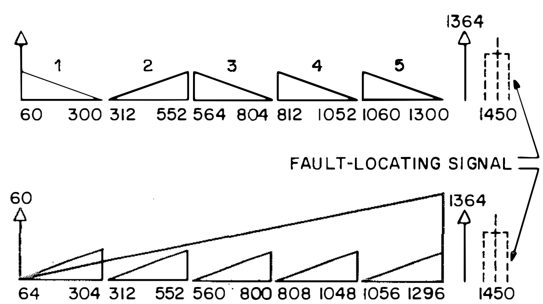

Figure 2—The *V300* system may accommodate 300 channels in either 5 supergroups or in a single master group with the frequency limits shown in kilohertz.

The pilot regulating devices operate automatically. They contain an integrating regulator, which responds slowly to relatively short-term variations of the pilot level, thereby eliminating any instability in the system. On the other hand, the time constant of the regulator is small when compared with attenuation changes due to cable temperature fluctuations, so that these changes are fully equalized.

Since the pilot-regulated amplifiers are able to equalize only at exactly the pilot frequency, it is necessary with both systems to include extra equalizers after a maximum of 24 dependent repeaters. These correct for the accumulated equalization errors at other frequencies and are fitted in only the terminal and main repeater stations.

1.4 Remote Power Feeding

Both systems are designed for remote series feeding of direct-current operating power. This minimizes the cost of power supply equipment inside the dependent repeaters. The constant current supplied to the dependent repeaters is 60 milliamperes for *V960* and 40 milliamperes for *V300*. It is fed over the inner conductors of the coaxial pairs. The power feeding equipment is accommodated in the terminal or main repeater stations above ground and can feed up to 12 buried dependent repeaters in each direction. This corresponds to the maximum number of dependent repeaters permissible in view of the need for an extra equalizer after 24 dependent repeaters. The surface repeater stations thus contain the power feeding equipment as well as the extra equalizers.

1.5 Overvoltage Protection

The German Post Administration requires that the voltage induced in a single interference section of such a cable be limited to 60 percent of the cable test voltage. This induced voltage from neighboring alternating-current power lines or electric railways is limited in the case of small-diameter coaxial cables to 1200 volts. To keep the interference sections as short as possible, the last dependent repeater of each power feeding section includes an isolating line transformer to isolate both the inner and the outer conductors of the two adjacent sections of the coaxial line. In addition, the outer conductor of the coaxial pair is insulated both from the cable sheath and earth (floating potential) to further reduce the effects of surges from alternating-current mains and atmospheric discharges.

As the metal frames of the buried repeaters are electrically connected to the outer conductor of the coaxial cable, these are also insulated from earth. This allows high interference voltages to appear between the metal parts of the buried repeater and its surroundings without causing damage to the equipment. The repeaters themselves are provided with insulated covers as protection for maintenance staff.

These precautions alone are not adequate to prevent damage to the transistor amplifiers, and overvoltage protection is required in the transmission path. A distinction is made between coarse and fine protection. Gas discharge arrestors are placed in the input and output circuits of all repeaters as coarse protectors. Two arrestors are used in parallel to increase reliability. The striking direct voltage of the discharge tubes is 600 volts, and this exceeds the highest possible power feeding voltage by a safe margin. This is also the case when under fault conditions low insulation resistance causes a connection between the inner and outer conductors of the coaxial pair in the opposite direction (such as through penetration of water into the cable). The discharge arrestor, having once fired, extinguishes immediately since the direct feed current together with the surge current from the line and equipment capacitances are too small to maintain it in a conducting state.

A surge voltage caused by lightning striking the ground close to the cable may rise to several thousand volts and since the arrestors take a finite time to ignite, the coarse protector does not respond immediately to the leading edge of the first voltage peak.

As there are other voltages impressed on the cable that are insufficient to cause the coarse arrestor to strike, the amplifiers are fitted with additional fine discharge protectors closer to the transistors.

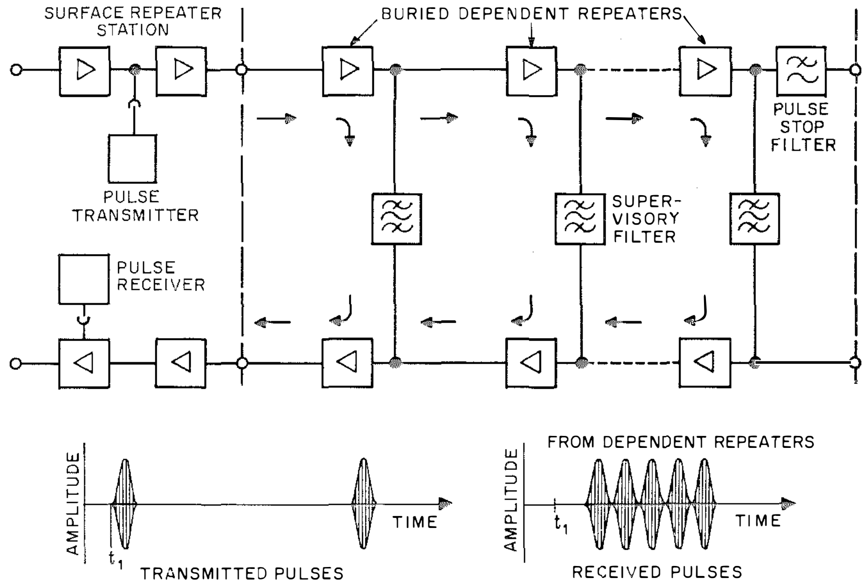

Figure 3—Fault location using pulses. A pulse sent from the pulse transmitter over the go path is selected by a filter at each dependent repeater and returned to the originating station over the return path. The transmitted pulses are repeated slowly enough to permit all the return pulses to be received without interfering with each other.

1.6 Supervisory Facilities

The pilots used at the repeater stations for automatic level regulation also indicate the condition of each transmission section. For this purpose the pilot level is indicated on a meter at the surface repeater station and a pilot alarm given if a decrease greater than 0.35 neper (3 decibels) from the nominal level is indicated.

1.7 Fault Location

Both pulses and direct-current measurements are available for locating faults.

1.7.1 *Pulse Method*

The pulse method presupposes an unbroken power feeding circuit. As shown in Figure 3 pulses of relatively low repetition frequency with a carrier frequency above the communication band are transmitted from the surface repeater stations. The pulses pass sequentially through all the dependent repeaters and are then suppressed in a stop filter at the end of the power feeding section. At each dependent repeater station, a supervisory filter at the output of its line amplifier transfers the pulse to the input of the line amplifier for the opposite direction. The returned pulses are displayed on an oscilloscope at the surface station that initiated the pulses. Each dependent repeater is a different distance from the surface station and is identified by the time taken for the round trip of its pulse. The absence of a pulse indicates which repeater has failed.

1.7.2 *Direct-Current Method*

When there is an interruption in the power feeding path, only the direct-current fault-location method can be used. This is shown in Figure 4. In each dependent repeater a high resistance in series with a diode is inserted between the inner conductors of the two coaxial pairs. Under normal working conditions the diode is reverse biased and nonconducting. The power feeding equipment is disconnected from an interrupted power feeding circuit and is replaced by the resistance bridge that provides an output voltage of the opposite polarity to make the diodes conductive and permit resistance measurements to be made. The resistance measurements permit the location of a fault to be determined.

1.8 Repeater Section Length

With the components now available, spacings of 4 and 8 kilometers (2.5 and 5 miles) for the *V960* and *V300*, respectively, have proved to

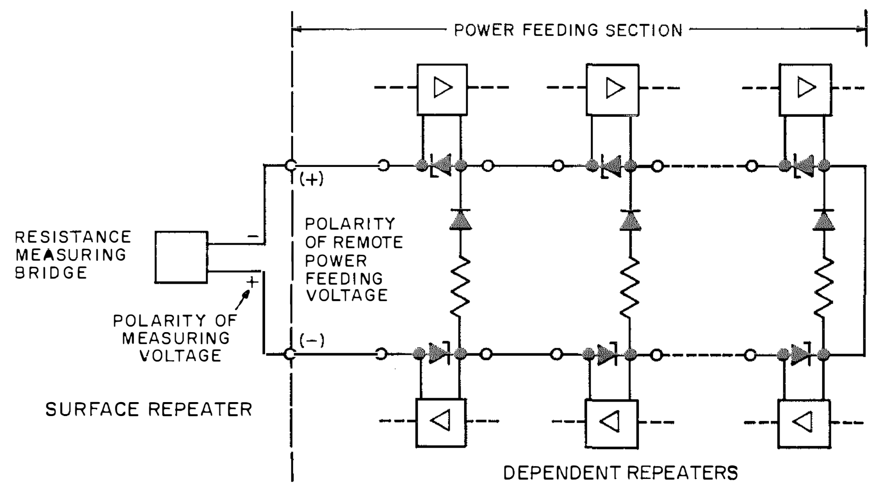

Figure 4 — Resistance measurement for fault location. The diodes are reverse biased and nonconducting with normal operating voltage polarity. For measuring, the operating voltage is removed and the bridge voltage polarity makes the diodes conducting.

be possible. The *V300* system can also be supplied for 6-kilometer (3.7-mile) repeater spacing. It is feasible to convert a cable initially equipped with *V300* repeaters to a *V960* system when the demand for telephone channels increases. This makes it necessary to provide for the siting of repeaters at 4-kilometer (2.5-mile) intervals when the cable is laid.

1.9 Noise

According to the recommendations of The International Telegraph and Telephone Consultative Committee, the average noise power at the end of a hypothetical reference circuit of 2500 kilometers (1550 miles) should not exceed 10 000 picowatts per channel at a point of zero relative level. Since 2500 picowatts are allowed for the modulation equipment in the reference circuit, the noise contribution of the line equipment must not exceed 3 picowatts per kilometer (4.8 picowatts per mile). Noise contributions are caused by the thermal noise of the line and of the transistors and as a result of nonlinear distortion produced in the amplifiers. As the linear crosstalk is intelligible, the requirements on crosstalk attenuation must be so high that the noise contributed by crosstalk is negligible.

1.10 Transmit Levels and Overload Points

The information given below relates to the supergroup method of the assembly of channels. Of course, the *V960* line equipment can transmit 3 master groups (*V900*) and the *V300* line equipment can transmit one master group without modification. Chosen on the basis of the requirements for thermal noise in a carrier channel and to provide adequate margins for signal level and equalization errors along the system, transmit levels of −1 and −0.5 neper (−8.7 and −4.3 decibels) relative a point of reference level are used for the top channels of the *V960* and of the *V300* systems, respectively. At a flat transmit level, the International Telegraph and Telephone Consultative Committee recommends a minimum overload point of +3.1 nepers (+27 decibels) for the *V960* system and of +2.7 nepers (+23 decibels) for the *V300* system referred to a point of zero relative level.

When a pre-emphasis network, which decreases the transmit level at low frequencies compared with high frequencies, is inserted in front of the transmit amplifier, requirements for the amplifier overload point are reduced and the distortion requirements are altered. This makes these requirements easier to satisfy, since they are frequency-dependent. By a suitable choice of pre-emphasis characteristic it is possible to achieve a situation where each channel has approximately the same total noise level (thermal noise and distortion). For the *V960* system and for the *V300* system a pre-emphasis of 1.15 nepers (10 decibels) was chosen. After allowing for erroneous signal levels due to inexact line lengths and accumulated equalization errors, a minimum value of +2.1 nepers (+18.5 decibels) referred to 1 milliwatt was obtained for both types of line equipment for the overload point, which corresponds to a power of 67 milliwatts.

2. Performance of Line Equipment

Several types of repeaters are provided. The surface terminal repeaters handle the signal band both to and from the modulation equipment. The buried dependent repeaters are of both the temperature-controlled and the pilot-regulated types. Finally surface main repeaters are required if the distance between two terminal repeaters exceeds 100 kilometers (63 miles) for the *V960* system or 200 kilometers (125 miles) for the *V300* system, that is, if more than 24 dependent repeaters are placed in tandem and provision must be made for extra equalizers and power feeding equipment. A further reason for the installation of main repeaters is the occasional necessity to drop or insert supergroups in the signal band.

The line equipments for the *V960* and *V300* systems form one family, using compatible repeater section lengths and identical mechanical design. This allows straightforward conversion from one type to the other if the line capacity is to be increased from 300 to 960 speech channels. In addition to the repeater housings and racks in the surface repeater stations, similar electrical design permits a number of units to be common to the two types of line equipment. Units that differ only in transmission frequency bands and levels use identical basic circuits wherever possible. This permits us to describe the *V960* and the *V300* repeater types together.

2.1 Temperature-Controlled Dependent Repeater

Figure 5 shows the layout of the temperature-controlled dependent repeater. Power separating filters are included at the inputs and outputs of each direction of transmission and these separate the direct-current power from the signal band. Overvoltage arrestors are placed across the signal jacks to protect the apparatus. The line amplifier is connected between the high-pass ends of these power separating filters. A zener diode, with a breakdown voltage exceeding the line-amplifier working voltage, connects the ends of the low-pass sections of the power separating filters together. The entire direct current normally flows through the line amplifier; when open-circuit occurs in the amplifier the direct-current power loop is maintained through the zener diode. Due to the limiting action of the zener diode when its breakdown voltage is exceeded, the diode constitutes, together with its paralleled capacitor, a further protection for the amplifier against interference. The capacitor bypasses the amplifier for relatively small alternating currents superimposed on the feed current and so prevents hum modulation. The block diagram also shows the supervisory filter and the series-connected diode and measuring resistor required for fault location.

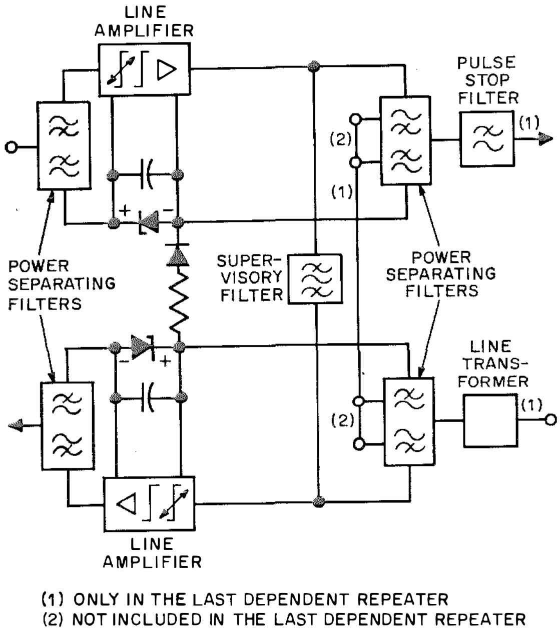

Figure 5—Temperature-controlled repeaters for *V960* and *V300* systems. In the last dependent repeaters in a power feeding section, the line transformers couple the signals to the next power section and a pulse stop filter is inserted in the go path to suppress the fault-locating pulses.

Each power feeding section is terminated by a last dependent repeater. As a pilot-regulated repeater is required only after a number of temperature-controlled repeaters, it is basically possible to employ a temperature-controlled repeater in this position. This is satisfactory from the size point of view, since it leaves the space normally occupied by the pilot regulating apparatus available for additional equipment. This last dependent repeater must also contain a pulse stop filter and a line isolating transformer for each direction of transmission as mentioned previously.

2.2 Pilot-Regulated Dependent Repeater

The pilot-regulated dependent repeater uses only one line pilot frequency of 4287 kilohertz for the *V960* or of 1364 kilohertz for the *V300* system. Thus it is not fully regulated as is the case with the main repeaters which, in addition, can use the 60-kilohertz pilot frequency. Figure 6 shows that the pilot-regulated dependent repeater contains, in addition to the units of the temperature-controlled repeater, a pilot regulating equipment consisting of band-pass filter, amplifier with rectifier, and regulator for each direction of transmission. A further zener diode and shunt capacitor are included in the power supply circuit for the pilot regulating equipment.

The outline circuit diagram of the line amplifier for *V960* is shown in Figure 7. The line amplifier consists of a pre-equalizer network, a preamplifier, and a final amplifier. The pre-equalizer, a bridged-T network, compensates for the slope of the gain of the preamplifier and final amplifier. As is seen in the section dealing with equalization, the required nominal overall gain at 60 kilohertz is 0.7 neper (6 decibels) and at 4028 kilohertz is 4.9 nepers (42.4 decibels) giving a slope in gain between high and low frequencies of 4.2 nepers (36.4 decibels). Using the pre-equalizer this slope is reduced to approximately 2 nepers (17.5 decibels).

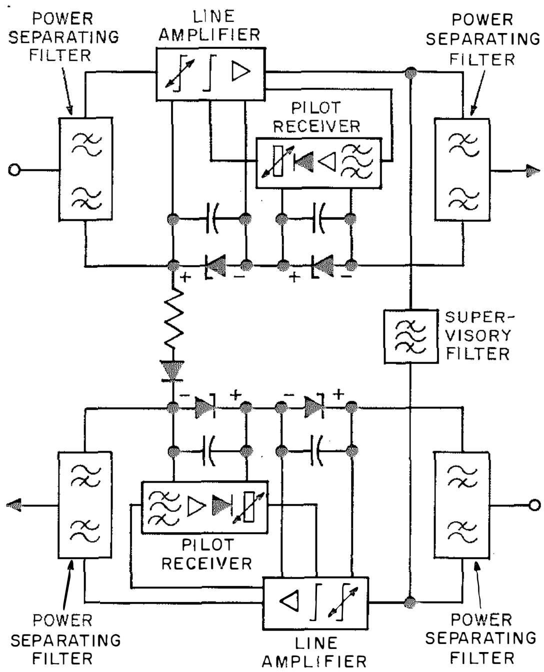

Figure 6—Pilot-regulated dependent repeaters for *V960* and *V300*.

This gives advantages regarding the singing margin of the amplifier and its noise figure at high frequencies. This worsening of the noise figure at low frequencies due to the attenuation of this pre-equalizer network is insignificant since the receive level at the input of the line amplifier is still considerably higher at low frequencies than at high frequencies even with pre-emphasis.

The preamplifier has two stages with high negative feedback and uses, as does the final amplifier, silicon planar transistors. The hybrid-coil bridges at the input and output are included in the negative feedback path and constitute an optimum circuit for low noise in the input stage and for maximum power output in the output stage.

Changes in attenuation due to changes in the temperature of the coaxial line are compensated for by an equalizing network inside the feedback loop of the preamplifier. Changes in the cable temperature at a depth of 0.8 meter (32 inches) of ± 10 degrees Celsius from a nominal temperature of $+10$ degrees Celsius cause changes in cable attenuation of ± 0.1 neper (± 0.9 decibel) at 4028 kilohertz per repeater section. In the case of the temperature-controlled repeater, the equalizing network of the preamplifier includes a resistor whose value depends on the ambient temperature. It is able to satisfactorily equalize attenuation changes of ± 0.15 neper (± 1.3 decibels) at 4028 kilohertz. In extreme cases a small residual equalization error of the order of 1 percent cannot however be avoided. As has already been mentioned, pilot-regulated repeaters are provided to compensate for this error. The only difference between the pilot-regulated and the temperature-controlled line amplifiers is that the ambient-temperature-controlled thermistor is replaced by a thermistor that is controlled from the pilot regulating equipment. The range of gain variation of the preamplifier is in this case ±0.3 neper (±2.6 decibels) at 4028 kilohertz, which is sufficient to equalize the attenuation changes due to variations in the temperature of the adjacent repeater section and also the accumulated temperature-dependent equalization errors of the preceding temperature-controlled repeaters. Finally the preamplifier is able to correct residual errors in the equalization of cable lengths, for which adjustments are in 0.1-neper (0.9-decibel) steps at 4028 kilohertz.

The final amplifier is a 3-stage unit also having high negative feedback. The second grounded-collector stage is necessary to drive the output stage. Since the noise figure of this amplifier is not critical, the hybrid-coil bridge at the input has been discarded. The network in the feedback loop allows correction of cable attenuation values deviating from the mean cable attenuation of 4.9 nepers (42.4 decibels) at 4028 kilohertz and 10 degrees Celsius, varying the amplifier output by ±0.3 neper (±2.6 decibels) in steps of 0.1 neper (0.9 decibel). Expressed as a cable length this is equivalent to ±250 meters (±820 feet). This length, however, can be fully used only if no allowance is required for tolerances in cable manufacturing.

As already mentioned a fine protection must be provided within the equipment for the protection of the transistors, and this is shown symbolically at the input and output of the line amplifier in Figure 7. The fine protection consists of 2 parallel discharge tubes having nominal striking voltages of 150 volts and a diode circuit. These are aided by the components of the repeater proper, the pre-equalizer network at the input, and the hybrid-bridge resistor and leakage inductance of the output transformer, which are utilized for current limitation. The diode circuit consists of 2 fast-acting radio-frequency diodes in parallel with opposite polarities and biased to avoid limitation of the output voltage and its attendant nonlinear distortion. These diode circuits together with the other protective devices provide protection against any kind of interference voltage.

The equalization error of the line amplifier is less than 0.04 neper (0.35 decibel) relative to the nominal loss of the coaxial line, and the systematic component of this is less than 0.02 neper (0.17 decibel). The second-order distortion margins obtained by the sum-and-difference-tones method and measured at the fundamental tone level of −1 neper (−8.7 decibels) referred to 1 milliwatt are 9.7 nepers (84 decibels) at 60 kilohertz and 8.6 nepers (75 decibels) at 4028 kilohertz below the fundamental tone level in the worst case. These margins vary almost linearly across the frequency range and

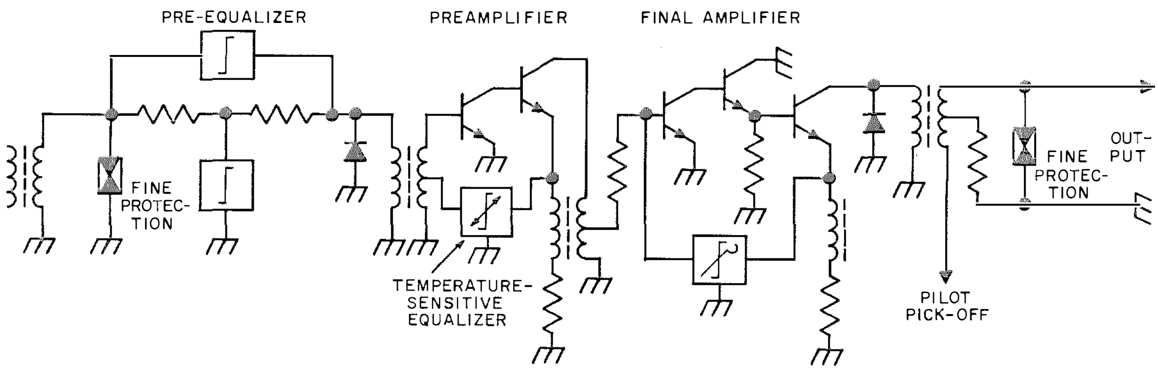

Figure 7—*V960* line amplifier.

exceed the required values by 0.4 to 1.1 nepers (3.5 to 9.5 decibels).

The third-order distortion margins at 60 and 4028 kilohertz of 14.3 nepers (124 decibels) and 12 nepers (104 decibels) are 2.4 to 0.8 nepers (21 to 7 decibels) better than the required values. At the same time, the noise figure measured at 4 megahertz is on the average 0.5 neper (4.5 decibels) against the planning value of 0.7 neper (6 decibels). Since the crosstalk meets its requirements satisfactorily and makes no noise contribution, the overall noise allowed by the recommendations of the International Telegraph and Telephone Consultative Committee is not fully used. The overload point at +2.25 nepers (+19.5 decibels) referred to 1 milliwatt is 0.15 neper (1.3 decibels) higher than required.

The mechanical construction of the *V960* line amplifier is shown in Figure 8. The two central boards accommodate the preamplifier and final amplifier including their power supplies while the outer boards mount the equalizer and fine-protection circuits together with the power separating filters.

The line amplifier for the *V300* system similarly consists of a preamplifier and final amplifier with pre-equalizing networks in the feedback loops. The equalizer and the grounded-collector stage in the power amplifier were omitted in this case, while an additional hybrid-coil bridge was fitted to the final amplifier input. The function is identical with the *V960* line amplifier.

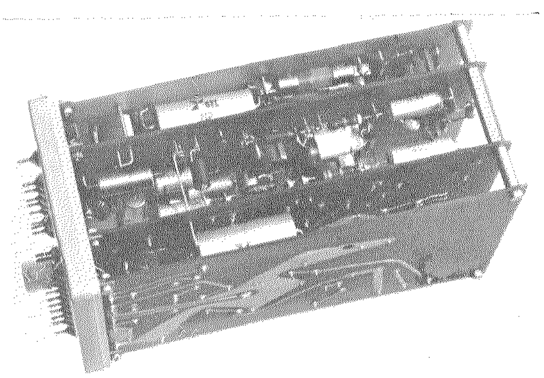

Figure 8—Line amplifier for *V960* system.

The network in the feedback loop of the preamplifier is utilized for temperature equalization. Variations in cable temperature of ±10 degrees Celsius result in attenuation changes at 1300 kilohertz not exceeding ±0.1 neper (±1 decibel) per repeater section. Again the temperature-controlled preamplifier is easily able to offset these and greater attenuation changes. The pilot-regulated preamplifier has a regulating range of ±0.3 neper (±2.6 decibels) at 1300 kilohertz, which is adequate for temperature equalization of the adjacent repeater section, correction of the accumulated residual errors of the previous temperature-controlled preamplifiers, and residual errors due to the equalization of cable lengths.

The network in the feedback loop of the final amplifier is used to correct deviations in the attenuation from the mean cable attenuation at 10 degrees Celsius, which amounts to 5.6 nepers (48.4 decibels) per section at 1300 kilohertz. The final amplifier is capable of being adjusted at 1300 kilohertz by ±0.3 neper (±2.6 decibels) in steps of 0.1 neper (0.9 decibel); this range is equivalent to ±430 meters (±1410 feet). This length can be fully used only if no allowance is required for tolerances in cable manufacturing. The fine protection against overvoltage at the input and output of the line amplifier can be slightly simplified compared with the *V960* line amplifier, as this *V300* amplifier can withstand a higher capacitance load, which allows zener diodes to be used.

The equalization error of the *V300* line amplifier is less than 0.04 neper (0.35 decibel) with a systematic component less than 0.02 neper (0.17 decibel) referred to the nominal loss of the coaxial line. Due to the favorable distortion, noise, and crosstalk values, again the overall noise allocation permitted by the recommendations of the International Telegraph and Telephone Consultative Committee is not fully used.

Figure 9 shows the line amplifier of the *V300* system. The two boards mount the preamplifier, final amplifier, power supply, fine protection, and power separating filters.

On both types of line equipment, the line pilot of 1364 or 4287 kilohertz is extracted at the output of the pilot-regulated line amplifier via a hybrid transformer. The pilot is selected by a band-pass filter and then passed to a rectifier circuit via an amplifier using silicon planar transistors. The rectified pilot is then compared with a reference voltage, and should it deviate by more than ±5 percent from the nominal value, an integrator regulator operates varying the current in the preamplifier thermistor until the rectified pilot voltage and the reference voltage coincide. To prevent the amplifier from being regulated to maximum gain in the event of pilot failure, the integrator regulator is stopped should the pilot level suddenly drop by more than 0.35 neper (3.0 decibels). The gain of the line amplifier is maintained at the value set at the instant of pilot failure.

The mechanical construction of the pilot-regulated dependent repeater is shown in Figure 10. Each of the two units at the bottom houses a line amplifier with two power separating filters and a power supply. On their front panels, the plugs for correcting amplifier gain for differences in the cable length can be seen. Above the line amplifiers is a unit containing the pulse selecting filter, the series-connected diode, and the associated measuring resistor for the direct-current fault locating circuits. The uppermost unit is left empty in temperature-controlled dependent repeaters. It houses the pilot regulating devices. At the last repeater in the power feed section, the amplifiers and fault locating units are moved upward and an additional unit containing the pulse stop filter and line isolating transformers is inserted at the bottom.

The plug-in repeater with its insulating cover is mounted in a cast metal casing. The German Post Office employs, in conjunction with the 12-pair coaxial cable it uses, a metal casing for 6 repeaters including cable seals.

2.3 Terminal Repeater

The terminal repeater station, which is always above ground, contains apparatus for the line pilots, residual equalization, and power feeding in addition to the transmit and receive amplifiers proper.

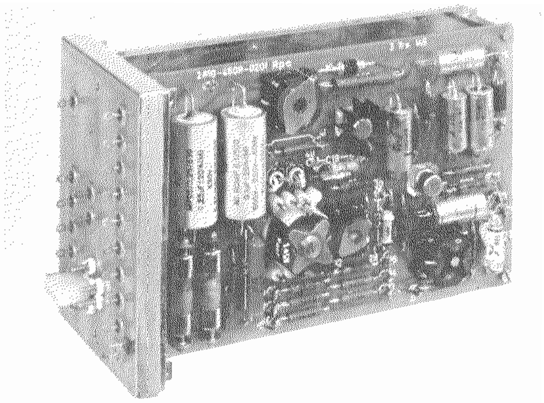

Figure 9—Line amplifier for *V300* system.

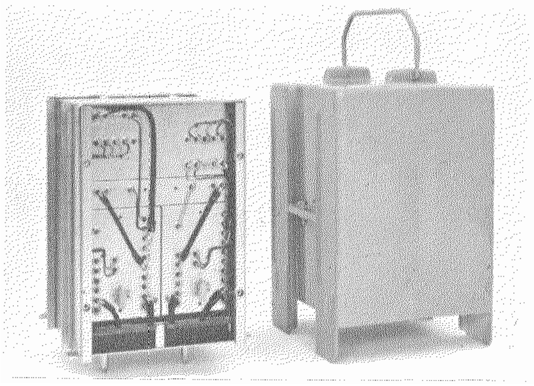

Figure 10—Pilot-regulated dependent repeater for *V960* and *V300* systems, with cover removed.

Figure 11 is a schematic diagram of a *V960* terminal repeater, which forms the point of interconnection with the modulation equipment. The terminal repeater is designed for the levels recommended by the International Telegraph and Telephone Consultative Committee at this point: In the transmit and receive directions −3.8 nepers (−33 decibels) referred to reference level or, optionally by strapping changes, −4.1 nepers (−36 decibels) in the transmit direction and −2.6 nepers (−23 decibels) in the receive direction both referred to reference level. Station-cabling equalizers are fitted in both directions of transmission to compensate for the frequency-dependent line loss occurring in wide frequency bands between the line distribution and terminal repeater racks. These equalizers are of the adjustable Bode type with constant input impedance [3] and have a maximum attenuation of 0.35 neper (3 decibels). They make up for the losses in the station cables, which are of various lengths, so that they are not frequency dependent. Adjustment is made by strapping into the circuit fixed resistors in steps of 0.05 neper (0.4 decibel) referred to the highest-frequency channel. Since the equalizer has to fan out on only one side, the cost of components could be greatly reduced. The calculations required for this equalizer and most of the other networks were made with the help of an electronic computer.

The transmit panel includes all high-frequency apparatus in the transmit direction. Pilot stop filters, employing crystals to meet the high

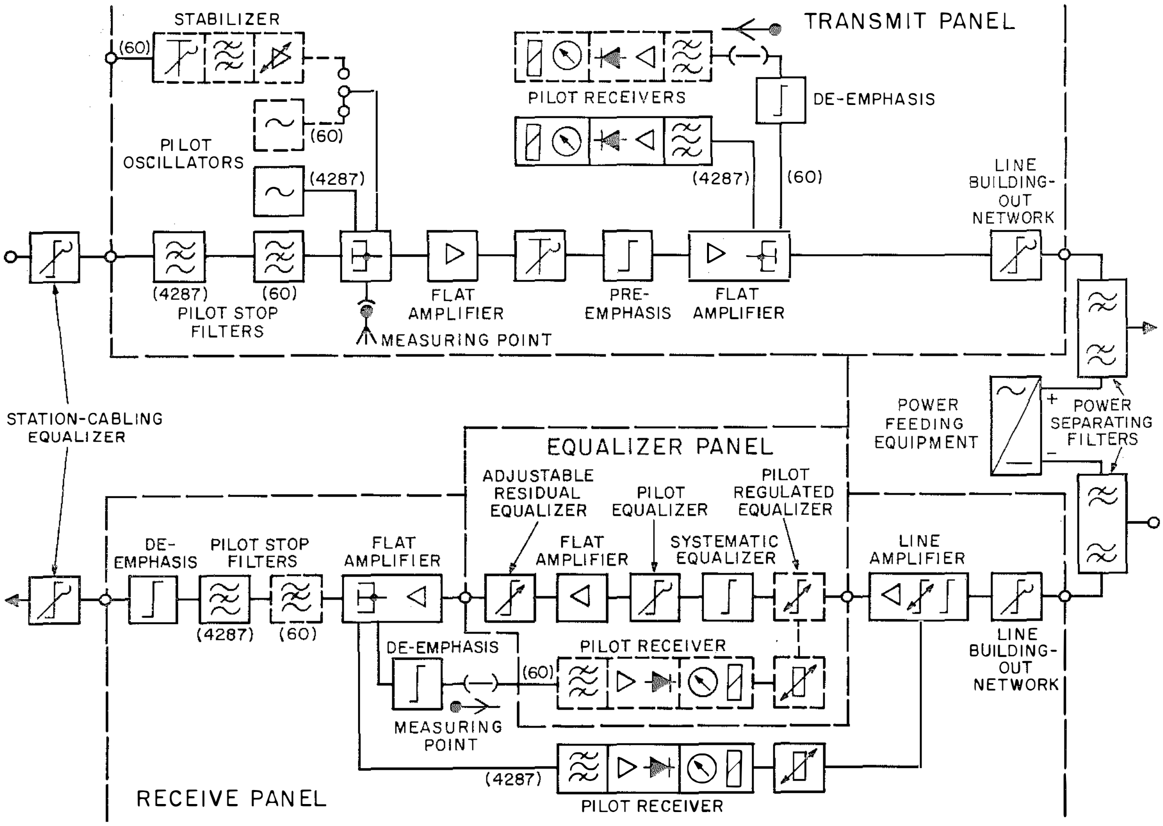

Figure 11—Schematic diagram of *V960* terminal repeater. Frequencies in kilohertz are in parentheses.

selectivity requirements, suppress unwanted tones from the modulation equipment at the line pilot frequencies of 60 and 4287 kilohertz, which are afterwards correctly injected. As the 60-kilohertz pilot regulating equipment may be omitted, it is shown by broken lines in Figure 11. After the pilot stop filters, the signal passes through the pilot combining unit made up of resistance and transformer hybrid bridges. This unit has three inputs each of which is decoupled with respect to the others; they permit the injection of the two line pilots and either test frequencies or the fault-locating pulses. The transmit panel also mounts the pilot oscillators required for each direction of transmission. To maintain the frequency stability of the 60-kilohertz pilot oscillators without using a temperature-controlled oven, the 60-kilohertz pilot is derived from 1200 kilohertz by frequency division. It is also possible to supply the 60-kilohertz pilot from an external source for frequency comparison purposes, in which case the oscillator is replaced by a level stabilizer.

The pilot injection unit is followed by two flat amplifiers to raise the signal to the required transmit level. A pre-emphasis network is inserted between the flat amplifiers. These flat amplifiers, which are also employed in other places in the surface repeater stations, are very similar to the final amplifier stage of the line amplifier. The equalizer network in the feedback loop is replaced by a resistor. Since the flat amplifier is frequently employed at points of low input level, it has, as has the preamplifier, a hybrid input transformer. When used as a transmit amplifier it is fitted with fine protection against overvoltage.

Each line pilot is extracted at the transmit amplifier outputs and fed to its pilot receiver consisting of a band-pass filter, amplifier with rectifier, and level indicator. These pilot receivers are electrically similar to the pilot regulating devices on the line. As an integrator regulator is not included, an indicator and relay unit provide the necessary visual and audible alarms. These are triggered when the pilot drops by more than 0.35 neper (3 decibels) from the nominal value.

It is also possible by detaching the 60-kilohertz pilot receiver to carry out flat measurements over the line via a de-emphasis network. If the length of the repeater section between the terminal repeater station and the first buried dependent repeater is shorter than can be covered by the adjustment provided in the line amplifier, the difference can be equalized by a line building-out network shown in Figure 12, in the transmit panel. For this purpose networks are provided with an attenuation of 0.3, 0.6, 1.2, and 2.4 nepers (2.6, 5.2, 10.4, and 20.8 decibels) at 4028 kilohertz. These networks can be used in tandem as required. The signal band is passed to the line via the high-pass section of the power separating filter, which is fitted with a coarse overvoltage arrestor.

The high-frequency equipment used in the receive direction is mounted in the receive and equalizer panels. The signal band received from the coaxial line is fed via the power separating filter and line building-out network to the line amplifier, which has a fine protector at its input. This line amplifier is identical in electrical design with the pilot-regulated line amplifier in the dependent repeaters and operates at the same output levels.

The equalizer panel mounts an equalizer regulated by the 60-kilohertz pilot. It corrects equalization errors in the lower part of the

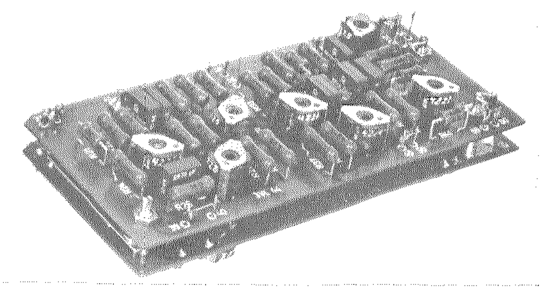

Figure 12—Line building-out network.

transmission band which are caused, for instance, by the temperature coefficient of the cable being somewhat higher at these frequencies. These equalization errors are only significant in the case of long transmission systems so that under certain circumstances the 60-kilohertz pilot may be omitted. From the 60-kilohertz pilot-regulated equalizer, the signal band is passed to a fixed equalizer which compensates for systematic errors. These are caused by the fact that the equalizing networks in the line amplifiers are of a simplified design and hence, even analytically, small differences result with respect to the frequency response of the coaxial line.

Finally, an adjustable residual equalizer compensates for the accumulated nonsystematic errors due to cable and equipment tolerances, while the pilot equalizers are used to set the pilots to their nominal values with great accuracy.

The residual equalizer consists of an echo equalizer and a resonance equalizer. Since the overall attenuation response of the echo equalizer builds up from the sum of a number of cosine curves, it is also called a cosine equalizer. The echo equalizer is shown in Figure 13. It consists of a chain of several equal delay networks with delay time t_0, extraction elements, and a summing amplifier. The whole arrangement is terminated in its characteristic impedance. At the input and output, as well as between delay elements, voltages can be extracted and summated. Figure 14 shows the locus and the amplitude of the first three delay

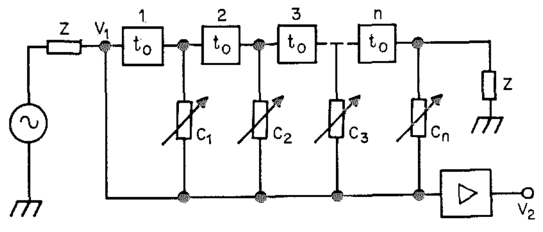

Figure 13—Echo equalizer.

elements of the echo equalizer. It can be seen that the echo equalizer is analogous to the mathematical Fourier analysis. Depending on the number of the delay elements used, it is possible to equalize a given attenuation curve with any desired accuracy. To be able to ensure close equalization even at low frequencies and at a low cost in delay elements, a resonance equalizer is also provided. This enables a resonance curve to be formed to raise or lower the levels. No reaction between resonance frequency, amplitude, or bandwidth takes place.

Between the pilot equalizer and the adjustable residual equalizer of Figure 11 is a flat amplifier to raise the signal band to its prescribed level. In the equalizer panel are also mounted the pilot band-pass filter, pilot amplifier with rectifier, pilot level indicator with a device for pilot alarm, and 60-kilohertz regulator. This pilot regulating equipment is similar to that of the dependent repeater. As in the transmit direction, a pilot level indicator and a device for visual and audible alarms are also incorporated.

The signal from the equalizer panel passes through another flat amplifier to two pilot stop filters. These are followed by a de-emphasis network, which compensates for the pre-emphasis imposed on the signal at the transmit terminal. Finally the signal passes through a station cabling equalizer to the line distributor, the pilot signals being picked off from the output of the last flat amplifier. A flat measuring point is provided at this point via a de-emphasis network and this is also used for connection of the pulse receiver for the fault location system. The pilot regulating equipment, which uses the 4287-kilohertz pilot, also includes a pilot level indicator and a device for initiating alarms.

The terminal repeaters for the *V300* system and for the *V960* are similar. The terminal repeater for the former is also designed to conform with the international recommendations with input levels of −4.1 nepers (−36 decibels) and output levels of −2.6 nepers (−23 decibles), both referred to reference level. The station cabling equalizer, also required in this equipment, compensates for the attenuation of the station cables of various lengths, producing a flat loss of 0.15 neper (1.3 decibels).

Power separating filters, power feeding equipments, and the 24-volt power supply panel for both terminal repeaters are identical. In most other panels, however, the *V300* system uses other units. The automatic pilot regulating equipment of 4287 kilohertz is replaced by one of 1364 kilohertz; pre- and de-emphasis characteristics differ, and the flat and line amplifiers must fulfill different requirements. In particular the equalizers are applicable only to the *V300* system. Three networks of the line building-out unit for the *V960* equipment, that is to say those with attenuations at 1300 kilohertz of 0.34, 0.68, and 1.36 nepers (2.95, 5.9, and 11.8 decibels), can be used for the *V300* equipment, in addition to a further network corresponding to a line length of 4 kilometers (2.5 miles). These panels are so constructed that they will also mount the printed boards for both *V960* and *V300* equipments.

In the *V300* transmit panel the cost was reduced a little when compared with the *V960* system. In addition to the 60-kilohertz pilot regulating equipment, the measuring and pilot combining units are identical for both terminal repeaters. For the *V300* system the pilot combining unit is followed by a pre-emphasis network and only one flat amplifier, this being sufficient to obtain the required transmit level. Here too, the flat amplifier was derived from the final amplifier of the dependent repeater. Thus it consists of two stages, the input and output of which employ hybrid coils, and all active units use silicon planar transistors. Similarly the cost of the equalizer panel was reduced. Since the signal band for the *V300* system is approximately a third of that of the *V960* system, the adjustable residual equalizer can be designed as a resonance equalizer, already described for the *V960* equipment for one resonance curve operating only at low frequencies. The *V300* system uses four such equalizers, each of them capable of covering the signal band and designed so that the resonance frequency, amplitude, and bandwidth of each equalizer section neither influences nor is disturbed by the adjacent equalizer.

The power supply units of both terminal repeaters deliver 24 volts required to drive the amplifiers, pilot oscillators, and pilot receivers and regulators. The input to the power supply can be 220 alternating volts +10 and −20 percent or 60, 48, or 24 direct volts with tolerances of +22 and −12 percent.

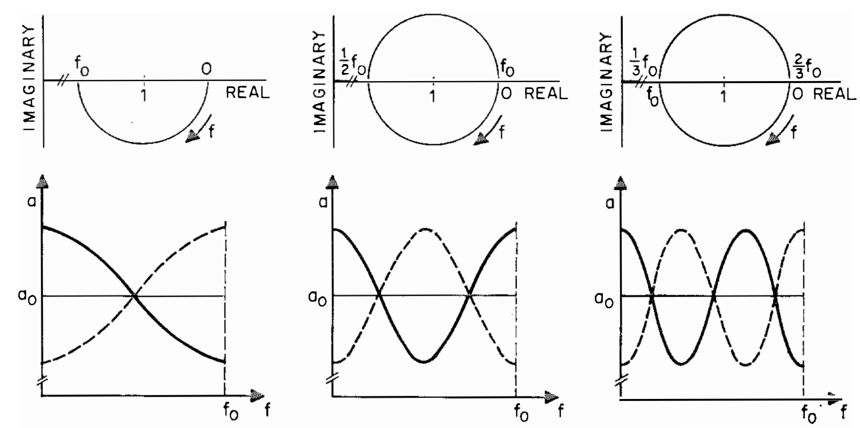

Figure 14—Locus and amplitude of the first three delay networks of the echo equalizer.

The remote power feeding equipment works from the same input voltages, delivering a constant feed current of 60 milliamperes for *V960* or 40 milliamperes for *V300* with an output voltage between 30 and 480 volts. The regulation characteristic is represented in Figure 15. Should the remote power feeding voltage exceed 520 volts through some fault such as a break in the power feeding circuit the supply automatically drops to 60 volts and at the same time an alarm is given. When the fault has been cleared, the voltage increases to its nominal value again.

The terminal and main repeater equipments (including branching-off facilities) are mounted in a rack 2.6 meters (8.6 feet) high. Figure 16 shows the cabinet complete with four repeaters. At the top of the rack the power separating filters and station cabling equalizers are placed and immediately beneath these are four sets of transmit, receive, and equalizer panels. These are followed by four power feeding equipments and finally by the 24-volt power supply at the bottom.

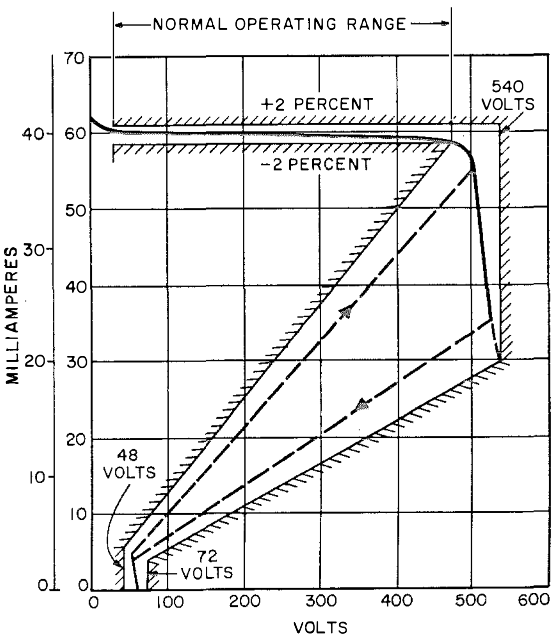

Figure 15—Regulation characteristic of the direct-current power feeding equipment. On open circuit, when the voltage exceeds 520 volts, the regulator reduces it to 60 volts automatically.

Since the individual panels use common items for both the *V960* and *V300* systems, a single standardized rack can be used. When changing the *V300* to *V960* equipment, the transmit, receive, and equalizer panels together with the station cabling equalizers at the top of the rack can be readily replaced by those of the *V960* equipment. The remote direct-current power feed can be altered from 40 to 60 milliamperes at will by resoldering straps. The power separation filters and the 24-volt supply remain unaltered.

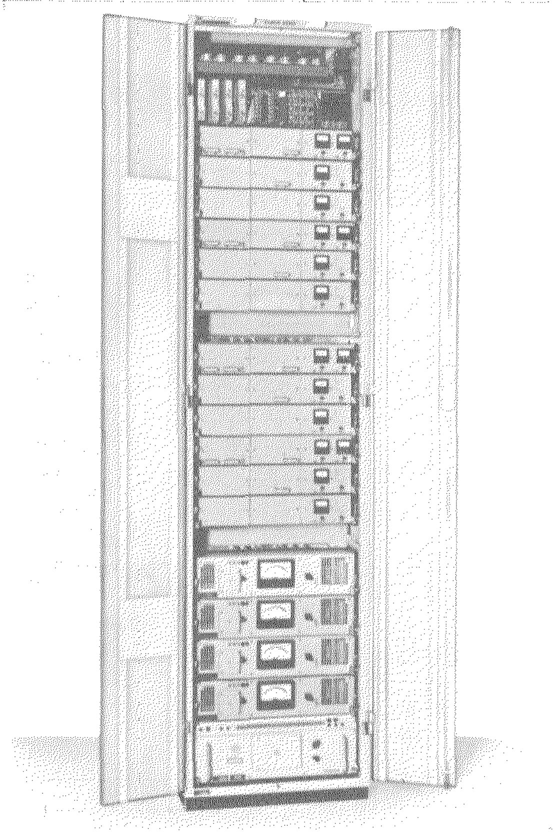

Figure 16—Four terminal repeaters mounted in a cabinet.

2.4 Main Repeater

The main repeater without branching-off facilities includes the essential receive and transmit sets and additionally the equipment required for extra equalization and remote power feeding. Figure 17 shows the layout of the *V960* main repeater. The block diagram of the *V300* main repeater is similarly derived from its terminal repeater. Both repeaters use the same units as the terminal repeaters, but a test frequency injection point with a pre-emphasis network is added.

Each transmission path is equipped with a receive panel, an equalizer panel, and an interconnection panel. The equalizer panel of the main repeater is the same as used in the terminal station. The receive panel is similar, only the pilot stop filters and the de-emphasis network are removed, because the main repeater station is inserted in the coaxial line.

The flat amplifier in the output is used as the transmit amplifier and is therefore provided with fine protection against overvoltage. The line building-out network, necessary if the station cannot be accurately sited, is mounted on the interconnection panel. This confers the advantage that the receive panel can be used without any modification as the receive panel for the main repeater with branching-off facilities. By plugging in the interconnection panel the terminal station rack can be easily turned into a main repeater rack without the necessity for any changeover.

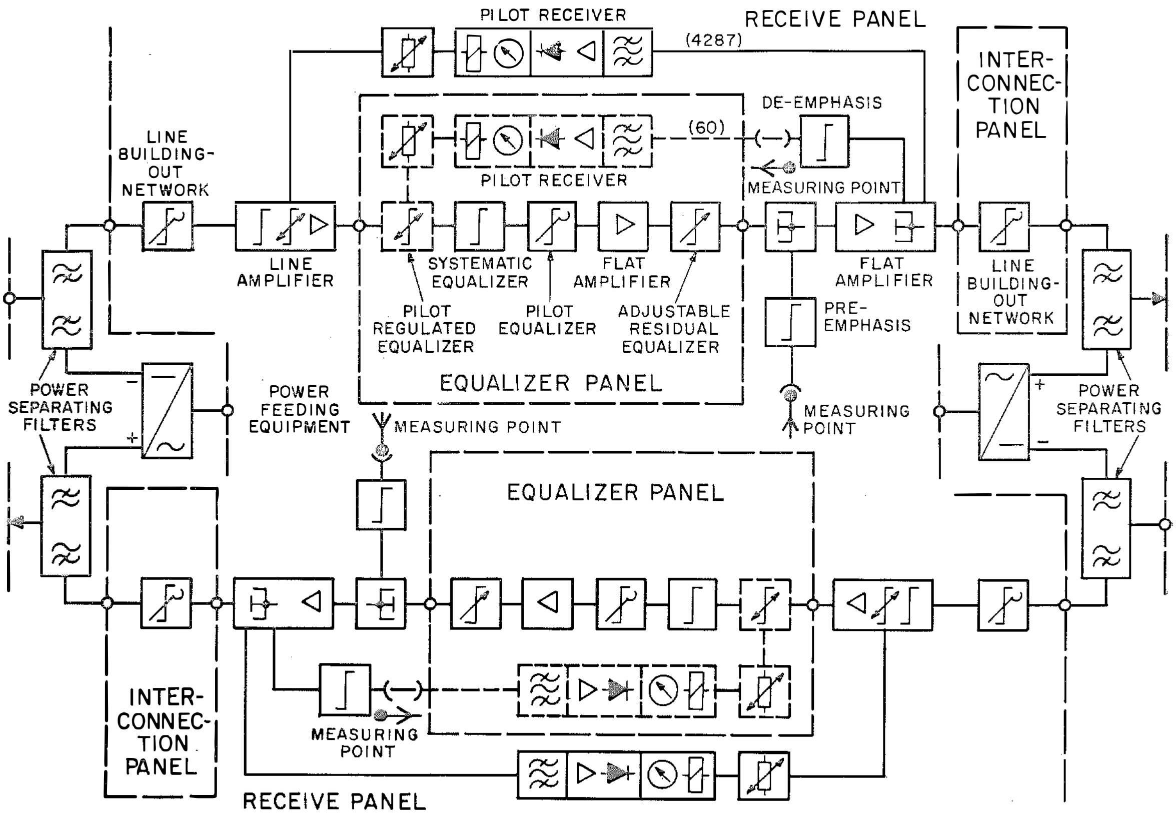

Figure 17—*V960* main repeater. Frequencies in kilohertz are in parentheses.

Thus by simply exchanging the transmit panels with the interconnection panels and exchanging the receive panels of the terminal repeater with the receive panels of the main repeater, the rack normally mounting four terminal repeaters becomes a main repeater rack mounting the equipment for two systems. To extract and reinject signals, additional branching-off and blocking filter panels are needed. These can also be mounted in the repeater rack.

In the case of the main repeater with branching-off facilities, the branching-off panels are mounted in place of the interconnection panels and the blocking filter panels are mounted in other unused spaces. As shown in Figures 18, 19, and 20, some units of the branching-off panel are already used elsewhere. Several arrangements are possible.

(A) The dropping of supergroups 1 and 2 (for *V300* the method is the same but with a pilot of 1364 kilohertz instead of 4287 kilohertz).

(B) The dropping of supergroups 1 through 5.

(C) The dropping of supergroups 7 through 16.

The amplifier and supervisory part of the repeater remains the same even though there are various arrangements for extraction and reinsertion units. Stop-band filters must be provided for the 60-kilohertz pilot when supergroups 1 and 2 and supergroups 1 through 5 are dropped and for the 4287-kilohertz pilot and the fault-location pulses when supergroups 7 through 16 are dropped. The station cabling equalizers, which are required when supergroups 1 through 5 or supergroups 7 through 16 are dropped, are mounted at the top of the rack. In the case of supergroups 1 and 2, a

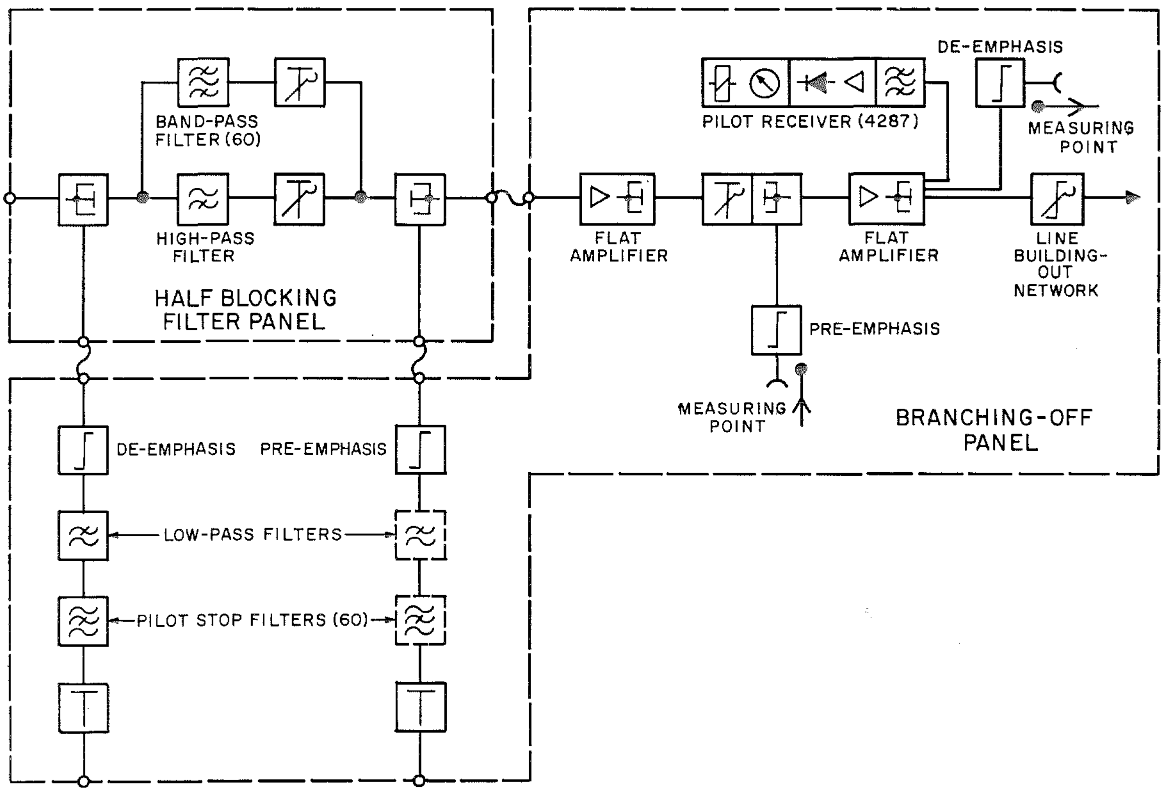

Figure 18—*V960* arrangement for dropping supergroups 1 and 2.

low-pass filter that admits only that band is also added.

Whereas a branching-off panel is required for each of the two directions of transmission in a system, only one blocking filter panel per system is needed, since the blocking filter units (as shown in Figure 18) are placed in a half panel. In the case of supergroups 1 and 2, a sharp-cut-off high-pass filter is required to stop the further transmission of those supergroups, but when dropping supergroups 1 through 5 and 7 through 16 it is possible to use separating filters. The blocking filters are bypassed to the line pilots by band-pass filters. For the *V300* system the blocking filter panel for the dropping of supergroups 1 and 2 differs from the *V960* arrangements in that the attenuators provided have different values.

Out of the existing arrangements for extraction with reinsertion can be obtained arrangements suitable for extraction without reinsertion. At the same time attenuators with similar losses are substituted for the blocking filters in the blocking-filter panel.

3. Data Summary

	V960 System	*V300* System
Line band for signals, pilots, and fault location (kilohertz)	≃60–4700	≃60–1500
Line band for signals and pilots (kilohertz)	60–4287	60–1364
Fault location signal, pulse carrier (kilohertz)	4660	1450
Mean repeater section lengths [kilometers (miles)]	4 (2.5)	8 (5)

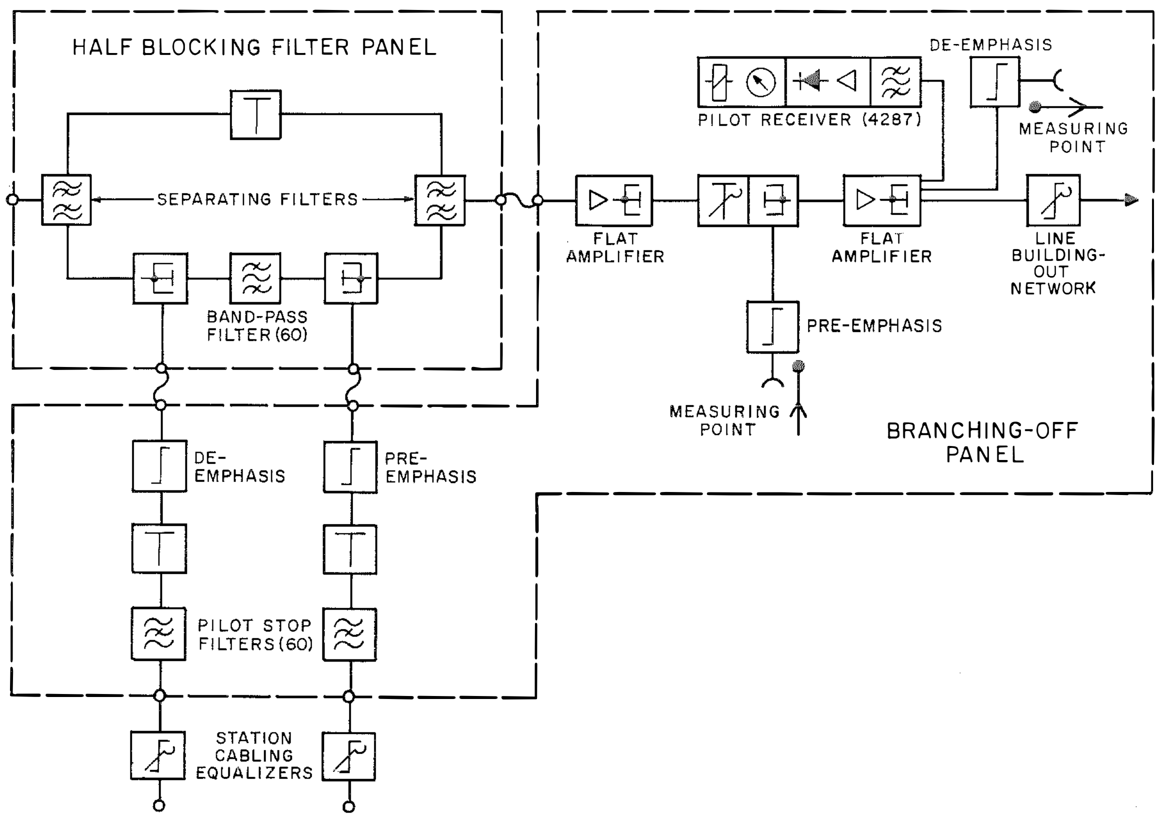

Figure 19—*V960* arrangement for dropping supergroups 1 through 5.

Mean amplifier gain [nepers (decibels)]	4.9 (42.4) at 4028 kilohertz	5.6 (48.4) at 1300 kilohertz
Range of gain variation [nepers (decibels)]	±0.6 (±5.2) at 4028 kilohertz	±0.6 (±5.2) at 1300 kilohertz
Transmit level [nepers (decibels)] referred to reference level	−1 (−8.7) at 4028 kilohertz	−0.5 (−4.3) at 1300 kilohertz
Pre-emphasis [nepers (decibels)]	1.15 (10)	1.15 (10)
Measured overload point at the output of the line amplifier [nepers (decibels) referred to 1 milliwatt]	≥ +2.25 (+19.5)	≥ +2.2 (+19)
Noise power* [picowatts per kilometer (picowatts per mile)]	<1 (1.6)	<1 (1.6)
Required direct-current power per temperature-controlled amplifier (watts, volts, milliamperes)	0.81, 13.5, 60	0.55, 13.7, 40
Required direct-current power per pilot-regulated amplifier (watts, volts, milliamperes)	1.53, 25.5, 60	1.07, 26.7, 40
Maximum distance between two surface repeater stations [kilometers (miles)]	100 (63)	200 (125)

* With loading in accordance with Recommendation G322 of the International Telegraph and Telephone Consultative Committee.

4. Acknowledgment

The author takes this opportunity of thanking his colleagues for all their help and assistance.

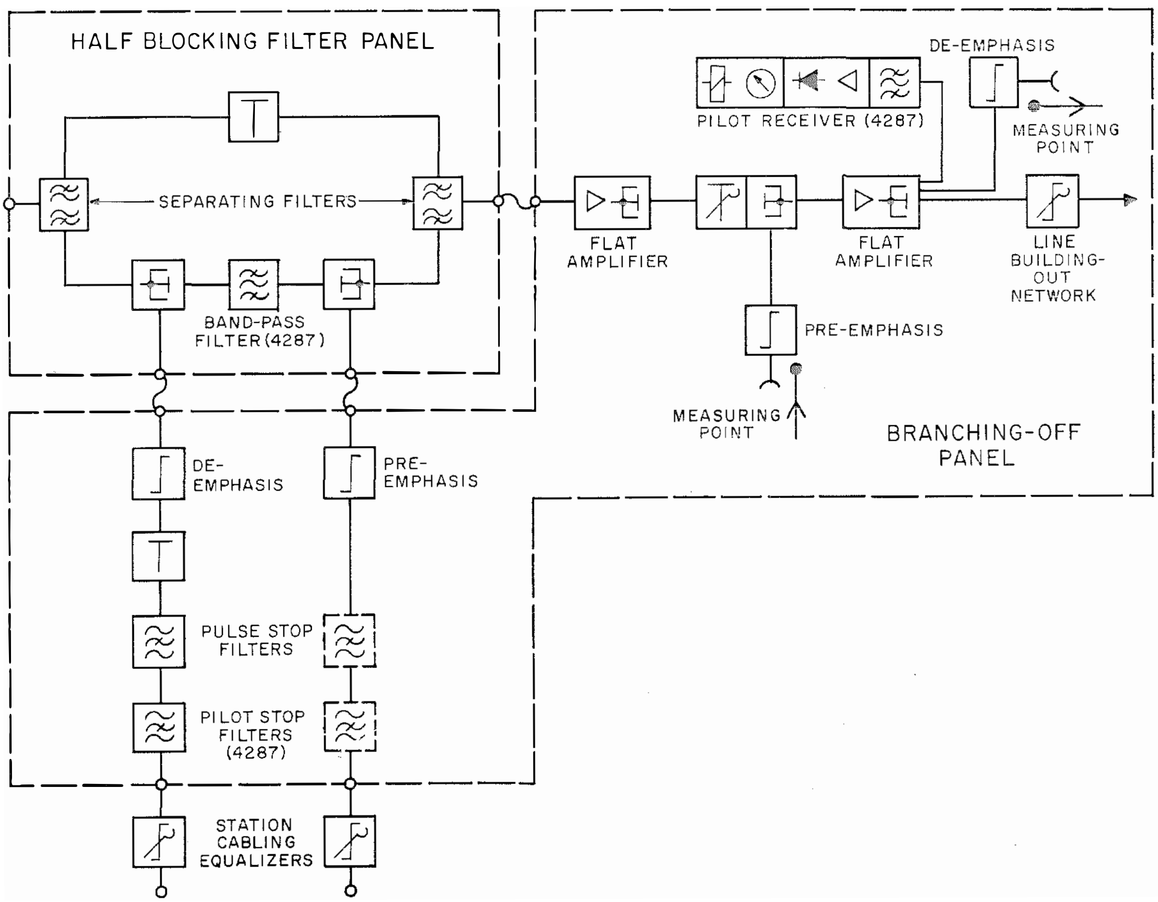

Figure 20—*V960* arrangement for dropping supergroups 7 through 16.

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Fritz Scheible was born in Kornwestheim, Germany, on 25 August 1923. In 1954 he received the Dipl.-Ing. degree from the Technischen Hochschule, Stuttgart.

He joined Standard Elektrik Lorenz in 1954 as a development engineer. Since 1962 he has been head of a laboratory for wide-band amplifiers for use in carrier and in radio link systems.

Pioneer Award to Otto Scheller

The Pioneer Award of the Group on Aerospace and Electronic Systems of the Institute of Electrical and Electronics Engineers was made posthumously to Otto Scheller for his conception of the principle of the radio range formed by overlapping directional radiation patterns and aurally identified equisignal courses.

Otto Scheller was born in 1876 and died in 1948. Most of his career was spent with C. Lorenz, now Standard Elektrik Lorenz, in Germany. His invention of the radio range is outlined in *Electrical Communication,* volume 40, number 3, pages 359–368; 1965.

Multichannel Telephone Equipment of Standard Telephones and Cables for Small-Diameter Coaxial Cable

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1. Introduction

In the United Kingdom and throughout Europe, a demand exists for 300- and 960-channel telephone systems for use on the small-diameter coaxial pairs of 0·174-inch (4·4-millimetre) diameter standardized by the International Telegraph and Telephone Consultative Committee (CCITT).

When originally introduced these systems were primarily intended for use in developed countries on spur routes of 20 to 60 miles (32 to 96 kilometres) to join provincial towns with main trunk routes and for use in underdeveloped countries as cheap main-line routes capable of forming a back-bone on which the telephone network could be extended.

Since their introduction, however, they have been increasingly used in developed countries as main-line systems of up to 100 miles (160 kilometres) and in less-developed areas routes of up to 500 miles (800 kilometres) have been proposed. In particular, railway administrations and highway departments have shown great interest in these private and secure telephone services.

To meet these various demands, great flexibility of equipment designs, high standards of reliability, and high performance must be coupled with low price. The two systems to be described have been produced by Standard Telephones and Cables to fulfill these needs and at the same time meet the specifications laid down by the British Post Office and the International Telegraph and Telephone Consultative Committee.

2. General System Considerations

The following frequency plans have been adopted.

2.1 Frequency Plan for 300 Channels

The 300-channel system is capable of transmitting 5 supergroups in the range from 60 to 1300 kilohertz or one mastergroup from 64 to 1296 kilohertz.

Line pilots are provided at 1364 kilohertz (the main regulating and control pilot) and 60 kilohertz. The 60-kilohertz tone may be derived from the station frequency-comparison pilot or be internally generated. It is used for the dual purpose of monitoring the line system at low frequencies and carrying the comparison pilot around the telephone network. It is level stabilized before application to the line system.

2.2 Frequency Plan for 960 Channels

The 960-channel system transmits 16 supergroups in the band from 60 to 4028 kilohertz with a regulating and control pilot at 4092 kilohertz. It conforms with all the existing 960-channel plant at present in service in the United Kingdom. The second line pilot, which is used solely for monitoring the system performance, is at 308 kilohertz.

A total bandwidth is available in the system to cover the range from 60 to 4500 kilohertz enabling the system, after modification of the line pilot frequency to 4287 kilohertz, to transmit the three master-group plans used elsewhere.

2.3 Supergroup Access Facilities

While the dividing line between passed and dropped supergroups may theoretically be set at any point, experience has shown that a limited number of combinations of filters will satisfy almost all requirements, and the standard arrangements follow.

Supergroups Passed	*Supergroups Dropped and Re-injected*
All supergroups but 2	Supergroup 2 only
Supergroups 2 upwards	Supergroup 1 only
Supergroups 3 upwards	Supergroups 1 and 2
Supergroups 5 upwards	Supergroups 1 through 4 (Group 1 of Supergroup 4 used for crossover)
Supergroups 7 upwards	Supergroups 1 through 5 and part of supergroup 6.*

* One or two groups are rendered unuseable in this case.

In addition bypass filters may be provided, if required, for 60 and 308 kilohertz pilots, when the corresponding supergroups are dropped to provide a through path for those pilots with frequencies within the stop band of the filter.

2.4 Noise Planning

The same basic arguments apply to noise on all types of coaxial equipment.

The hypothetical reference circuit of the International Telegraph and Telephone Consultative Committee of 1575 miles (2500 kilometres) for coaxial line equipments is made up of 9 homogeneous sections each of 175 miles (280 kilometres). This means that the line signal is demodulated and re-combined at intervals of 175 miles (280 kilometres) in such a manner that the re-assembled channels may be considered to be re-connected at these points in a random manner. Thus the total permitted line noise contribution of 7500 picowatts may be expressed in the form of 4·8 picowatts per mile (3 picowatts per kilometre) for purposes of practical comparison.

Since the crosstalk performance of coaxial cables is inherently very good at high frequencies, the overall crosstalk performance is within the province of equipment design, and so can be made very good. Hence no allocation of the system noise is made to this source and it is normally neglected. Radio-frequency breakthrough of broadcast channels is sometimes found on wide-band systems, but again these are generally at a very low level and occupy at most only a few channels so that they may also be neglected. Noise from power units and from other sources may also be overcome.

It is usual when planning the system to allocate proportions of the specified permitted noise to the thermal and intermodulation noise and this is determined in conjunction with the choice of the pre-emphasis characteristic to be used with the system.

To reach an optimum solution to this problem in the case of the 960-channel system, a computer was programmed to calculate the noise performance of the complete system as a function of amplifier gain, noise figure, non-linear distortion, phase characteristic, relative level, and pre-emphasis shape. The programme produced the proportion of noise in each channel due to thermal, second-order, and third-order distortion.

From these calculations and the chosen section lengths, the output level of the repeater at the highest channel was fixed.

2.5 Length of Repeater Section

The section length between repeaters is selected on the basis of the linearity, power handling capabilities, and bandwidth performance of the available transistors and is generally the subject of a recommendation of the International Telegraph and Telephone Consultative Committee.

Repeater section length of 3·6 miles (5·8 kilometres) for 300-channel systems and 2·5 miles (4 kilometres) for 960-channel equipments are employed in the systems.

2.6 Transmission and Overload Levels

Again the same basic arguments apply to both types of coaxial line equipment. The output levels of the 960-channel and 300-channel systems at 4028 and 1300 kilohertz are −9 and −13 decibels referred to reference level, respectively. The pre-emphasis is, as internationally recommended, to have a maximum loss of 5 decibels in the case of the 300-channel system and to follow a defined law with a maximum loss of 10 decibels in the case of the 960-channel system.

The minimum overload level for the 300-channel system is then calculated at +11 decibels referred to 1 milliwatt, and for the 960-channel system of +18·4 decibels referred to 1 milliwatt. These correspond to output powers of +12·6 milliwatts and 67 milliwatts, respectively. Both these output powers have been adequately exceeded, leaving substantial margins in the systems to allow for system level errors.

2.7 Equalization

A gain control range of approximately +4 decibels is provided within the line amplifiers of both systems to compensate principally for variations in the cable attenuation due to temperature. Other variations which are likely to occur are due to cable manufacturing tolerances, taken to be ±2 per cent of the nominal section attenuation but rarely reaching this; siting errors which are limited individually and cumulatively to ±220 yards (±0·2 kilometre), amplifier gain errors, and amplifier-regulator adjustment errors.

Taking the 300-channel system as an example, the variation in cable attenuation at the pilot frequency of 1364 kilohertz for 10-degree Celsius change on the mean temperature is ±0·72 decibel; the cable manufacturing tolerance is controlled to within ±0·72 decibel; the siting error gives ±1·24 decibel, and the amplifier gain errors taken over a large number of amplifiers is ±0·25 decibel from both causes.

From examination of these figures it will be seen that apart from the cable attenuation variations with temperature, the remaining errors are all of a fixed nature and can be assessed at the time of installation.

Providing that sufficient margin is available in the noise performance of the repeaters against level errors along the line and that some means of adjustment is available at the installation stage to cater for the fixed errors, it is possible to allocate all of the available range of control within the line amplifier to attenuation changes due to temperature and thus reduce the line plant to be installed. It is also feasible to allow these temperature changes to accumulate over a small number of repeater sections dependent on the range of cable temperature to be encountered.

The 960-channel and 300-channel systems both employ this method of equalization and hence each have two distinct types of repeaters. One repeater contains a pilot-controlled amplifier and a second amplifier the gain of which is adjustable as the square root of frequency whilst the other equips two gain-adjustable amplifiers. The pilot-controlled amplifier contains a Bode equalizer which is terminated in a thermistor driven proportionally by a pilot regulator which monitors the pilot level at the output of the amplifier. The adjustable-gain amplifier employs the same network, this time terminated in a series of fixed resistors which allow the gain to be offset in 0·5-decibel steps at the top channel frequency.

The pilot-controlled amplifier is sited so that it always works in the centre of its control range and is interspersed with the adjustable gain repeaters along the line. The 300-channel system in the United Kingdom is equipped with pilot-controlled repeaters at intervals of approximately one in four stations and the 960-channel system at intervals of one in three repeater stations.

Overall level errors due to minor discrepancies in the match between the gain of the amplifier and its adjacent line section, are compensated by a residual equalizer at the end of each line section.

On longer routes of more than 24 intermediate repeater stations, a switched equalizer is fitted at the terminal stations in the receive directions of transmission to compensate for minor cumulative level errors. These arise from the slight discrepancies which inevitably exist between the gain-control-network losses in each repeater and the attenuation changes in the cable due to temperature. These errors alter slowly with the temperature cycle over winter and summer.

Since the cable temperature co-efficient changes substantially at the bottom end of the frequency band (0·26 per cent per degree Celsius at 60 kilohertz and 0·19 per cent per degree Celsius at 300 kilohertz and above), care has been taken with the design of the gain-control network within each amplifier to accommodate this, avoiding large accumulating level variations at 60 kilohertz along the line system.

The points for compensation of minor equalization failures along the system coincide with the power feeding points.

2.8 Remote Power Feeding

The remote power supply for the repeaters is a constant direct current of 50 milliamperes as specified by the British Post Office, and is supplied over the centre conductors of the coaxial tubes. Stringent requirements are laid down so that under no circumstances can this current through a 2000-ohm resistor (intended to simulate a human being) be exceeded for longer than 250 milliseconds. The feeding voltage is limited under British Post Office regulations to 250-0-250 direct volts although an alternative of 300-0-300 volts is available within the power unit.

The power feeding arrangements for the two systems differ slightly due to the basic configurations of the repeaters and the operating conditions of the systems. The 960-channel repeater consists of a pre-amplifier and an output amplifier each of which requires 12 volts, a total of 24 volts per repeater. The power separating filter drops 0·25 volt and hence the total repeater (both directions of transmission) requires 48·5 volts at a direct current of 50 milliamperes (2·4 watts). The regulator unit is connected in shunt with the input amplifier and does not add to the total voltage.

The 300-channel system employs a 12-volt amplifier for each direction of transmission and one regulator fed in series with the amplifiers and driven from 15 volts. The power separating filter requires a voltage drop of 1·2 volts and the complete repeater therefore requires 40·2 volts (2 watts) for both directions of transmission.

Table 1 shows the characteristics of the two systems derived from these facts and is based on regulation being provided at alternate repeaters as in most exported systems; slightly longer

TABLE 1
CHARACTERISTICS OF TWO SYSTEMS

System (Channels)	Cable (Millimetres)	Supply (Volts)	Power Feed Distance		Repeater Sections
			(Kilometres)	(Miles)	
300	4·4	250-0-250	122	75·7	21
300	4·4	300-0-300	145	90	25
960	4·4	250-0-250	76	47·5	19
960	4·4	300-0-300	92	57	23
960	9·5	250-0-250	171	106	19
960	9·5	300-0-300	207	129	23

distances are possible if the regulators are spaced at wider intervals.

The power feeding stations are situated above ground at points where alternating-current mains or a central battery are available as a primary power supply. The equipment will operate from a central battery of 24 volts, which may vary between 28·15 and 21·8 volts, and from single-phase 45-to-66-hertz 90-to-130 volt or 180-to-260-volt alternating-current mains with the selected nominal voltage regulated to within ± 5 per cent.

2.9 Over-Voltage Protection

The surge protection provided may be divided into primary and secondary protection, the primary protection being employed to divert the bulk of the surge away from the repeater equipment and the secondary protection to reduce or clip the residual pulses to a level below which damage cannot occur in the equipment. In general the collector-emitter voltages of the transistors employed is limited by the manufacturers to 30 volts and thus the residual interfering pulses are reduced to less than 15 volts before they reach the transistors.

Primary protection takes the form of gas discharge tubes placed directly across the line terminals of the repeater at its input and output. Breakdown of these gas tubes occurs at 350 direct volts, which is higher than the maximum power feeding voltage to avoid continuous ignition from the power supply. The gas tube will also, should it inadvertently strike, carry the 50-milliampere feeding current permanently without damage and automatically extinguishes after the passage of a surge through the cable. Fast secondary protection is provided at the input and output of the repeater amplifiers by the connection of diodes and zener diodes, and the power supplies to the repeater are also individually safeguarded. Care was taken when adding the diode circuits to the amplifier to ensure that their non-linear performance did not interfere with the amplifier's harmonic performance.

In addition to these precautions a 50-hertz by-pass filter is provided within each repeater power supply. This avoids 50-hertz hum modulation interference due to induced alternating currents.

Each individual interstice wire is also protected by a 350-volt gas tube connected between the wire and the coaxial outer conductor.

2.10 Monitoring Facilities

The line regulating pilots are used to monitor the complete transmission section, and an alarm is given by chart recorders when the pilot level rises or falls by 3 decibels. Meters continuously display the pilot levels.

The regulating pilot level at the output of the regulated repeaters is monitored and an alarm passed to the terminal equipment when this changes by ± 4 decibels.

2.11 Fault Location

The method used for the location of faulty repeaters is imposed in the interests of standardization on all manufacturers of equipment tendering to the British Post Office. It employs a direct-current circuit using three of the interstice wires within the cable, selection of the appropriate two out of the three wires indicating the type of fault at the dependent repeater.

At the faulty repeater a large-value resistor is connected across the wires and the monitoring of a direct current flowing in this resistor at the terminal repeater indicates directly on a meter the position of the faulty repeater.

Power failure or pilot failure in either direction of transmission is shown. Location of faults on a second system on the same cable is possible with the original alarm scheme when the fault resistors of the second system are added in parallel with the first system. Diodes provide discrimination between the two systems.

2.12 Engineering Speech Circuit

Engineers visiting a repeater may communicate with one or both of the adjacent terminal repeaters. To do this audio amplifiers are employed at the surface stations and a portable telephone unit is taken to the repeaters and plugged into the line to complete the link. Power for the portable telephone, which uses direct-current loop calling and its loudspeaker as a microphone, is obtained from self-contained batteries. A 10-pound pair of interstice wires, loaded at 88 millihenries per mile (55 millihenries per kilometre) is used to make the circuit independent of the coaxial repeaters.

A single amplifier driven from and driving a loudspeaker at the terminal station is employed in conjunction with a send-receive switch for both directions of speech transmission.

3. Carrier Line Equipment

3.1 Manufacturing and Maintenance

A prime objective of the equipment design has been to obtain the lowest cost in both manufacture and maintenance consistent with reliability and high performance.

Silicon transistors have been used in general for the more-arduous circuit conditions where linearity, low saturation voltage, or temperature stability have been required. Circuits have been kept simple, and the number of different circuits kept to a minimum by using the same basic layout and configuration in the terminals and repeaters where appropriate.

Uniform cards, rack layouts, and assemblies have been used between the various shelves of the terminal repeaters, and in addition, great care has been taken to avoid the excessive use of lamps, relays, and other items which might waste power.

Reliability has been improved by the use of specially approved and vigorously tested components.

3.2 Dependent Repeater for 300 Channels

A regulated dependent repeater is shown in Figure 1 and its general schematic in Figure 2. An installed repeater in an early design of housing is shown in Figure 3. The repeater consists of 5 cards which slide into a position inside a shortened ITT Europe standard subframe. Each card has splashproof covers on its front and rear which also serve to electrically shield the individual transmission paths from each other. The subframe contains a power separating filter which splits the high-frequency bands from the power feeding direct current at the input to the repeater and re-combines them at the output. This power separating filter is made in two parts and one part is associated with each direction of transmission. The two

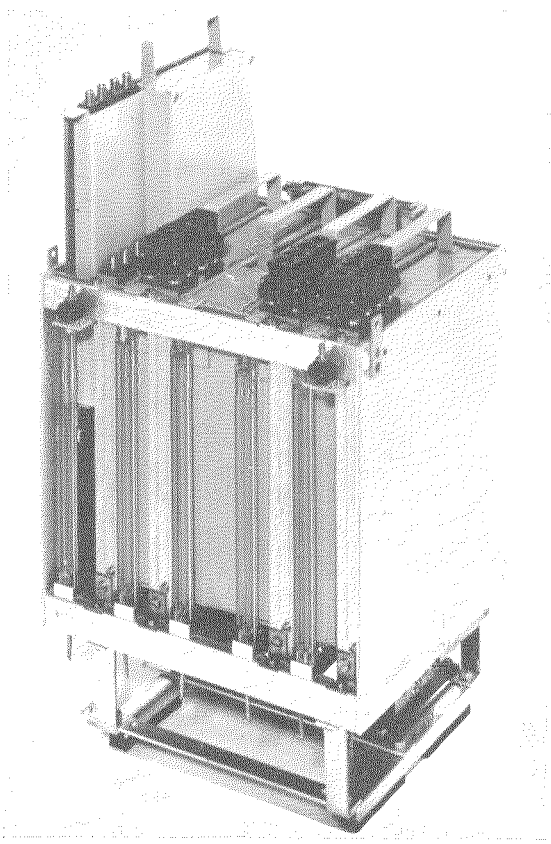

Figure 1—Dependent repeater for 300-channel system.

amplifiers required for transmission are placed at each end of the subframe adjacent to the power separating filters and the central position is allocated to the pilot control regulator which is permanently wired to one amplifier. This amplifier and regulator are employed at succeeding stations along the route in each direction of transmission as regulation is required. The other amplifier may be adjusted in discrete steps of gain and is employed to compensate for the fixed errors due to siting tolerance, et cetera, to be found within each regulated line section.

Power for the amplifiers is picked off from the centre of the power separating filters by means of a zener diode which is by-passed to 50 hertz by a large capacitor. Power separation is done by a simple prototype filter section, and direct-current blocking capacitors are fitted at the input and output of the repeater amplifiers.

The circuits for the alarm and fault location system are also equipped inside the separating filter. The amplifier, which has 4 stages of common-emitter transistors, has its gain control network fitted with a directly heated thermistor and is driven by the pilot regulator. This is placed in the centre of the amplifier between two input and two output stages. Overall and local feedback is employed and a beta network provides partial gain shaping. The gain shaping is completed by an input equalizer network.

The pilot regulator which is of the proportional type consists of high-pass coil-capacitor pick-off filter, tuned amplifier, comparator circuits, and direct-current amplifier. A transistor-operated switch provides the ±4-decibel level change alarm for the supervisory circuit.

Silicon epitaxial planar transistors are used throughout the amplifier and regulator high-frequency circuits.

The repeater is coupled by plug and sockets to the cable and is mounted in a housing which will take two such repeaters.

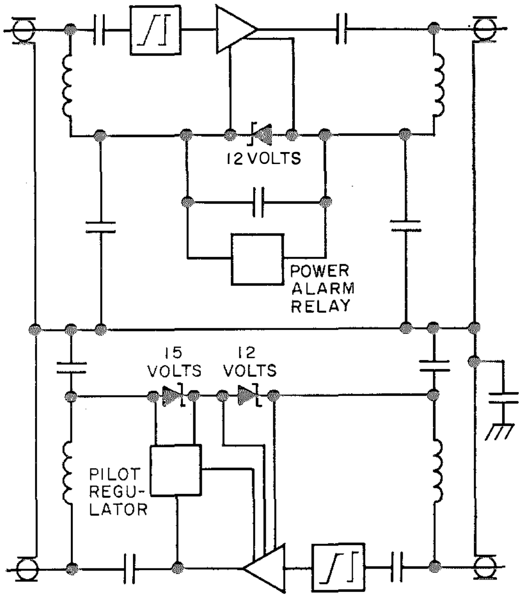

Figure 2—Diagram of 300-channel dependent repeater.

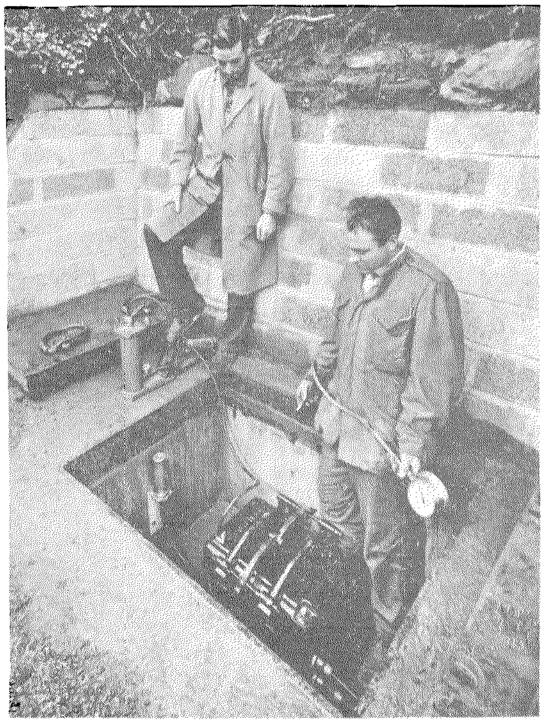

Figure 3—Repeater installed in an early type of housing.

The fixed-gain amplifier is exactly similar to the regulated amplifier except that the thermistor is replaced by a number of resistors which enable the control network to be offset at 1300 kilohertz by ± 4 decibels in $0 \cdot 5$-decibel steps. The nominal spacing for this repeater is $3 \cdot 6 \pm 0 \cdot 13$ miles ($5 \cdot 8 \pm 0 \cdot 2$ kilometres) at 10 degrees Celsius, which corresponds to a gain of 36 decibels at 1300 kilohertz.

The repeater meets the recommendations of the International Telegraph and Telephone Consultative Committee with regard to noise, crosstalk, et cetera, and matches the repeater section of cable to within $0 \cdot 35$ decibel. A cable-simulating network is available for occasional use, when it is geographically difficult to precisely site all the repeaters. It is installed inside the repeater housing but outside the repeater unit. The repeater performance is given in the data summary at the end of this paper.

3.3 Dependent Repeater for 960 Channels

The 960-channel dependent repeater shown in Figure 4 has exactly the same physical dimensions as the 300-channel repeater and contains four plug-in cards. The power separating filters for both directions of transmission are mounted together at one end of the repeater subframe. The small printed circuit required for the supervisory scheme is mounted beneath the power separating filters.

All the units are mounted on printed-circuit cards and are contained in front and rear cases which are connected directly to the outer conductor of the cable. Insulation is provided at the front of the repeater to avoid accidental hand contact with the insulated outer conductor. The units, as before, are two amplifiers, one for each direction of transmission, a regulator, and a cable-simulating network (this time inside the repeater unit).

A special feature of this system is the crosstalk performance which has been extended to produce programme performance throughout the whole band from 60 to 4028 kilohertz. Crosstalk margins of greater than 106 decibels at 4028 kilohertz have been obtained; these are some 8 decibels better than is required per repeater. To achieve this it has been necessary to eliminate earth loops and long earth connections; the power separating filters are grouped close together and each is totally enclosed.

The dependent repeater shown in Figure 5 consists of two amplifiers in each path; one has a sloping gain and provides some of the equalization required for the preceding cable section and the other has a flat gain characteristic and provides the output power. A Bode network between the two amplifiers controls the overall gain of the repeater amplifier, and the network loss is varied by the changing resistance of a directly heated thermistor or by fixed resistors. The thermistor, which is especially chosen and designed to impart good dynamic performance characteristics to the overall system, is used in pilot-controlled repeaters whilst resistors are employed in gain-adjustable repeaters again used to compensate for fixed errors in system

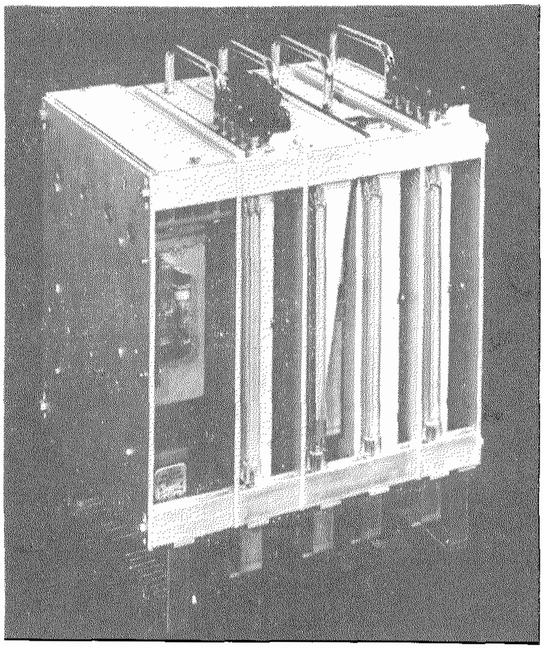

Figure 4—Dependent repeater for 960-channel system.

gains and losses. Silicon epitaxial planar transistors are used throughout the repeater amplifiers and again both local and overall feedback is applied individually to each amplifier. The input amplifier is gain shaped by a network in its feedback path and this is supplemented by an input equalizer network to complete the overall gain shaping.

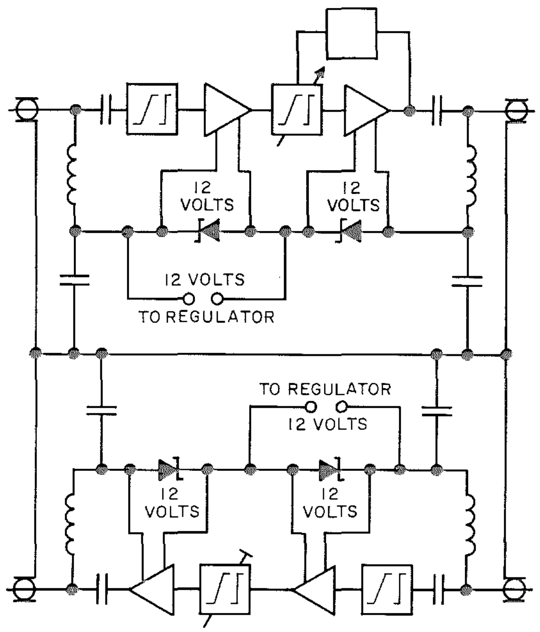

Figure 5—Diagram of dependent repeater for 960 channels.

The circuit arrangement and a photograph of the 960-channel amplifier are given in Figures 6 and 7. Typical performance figures appear in the data summary at the end of this paper and the total calculated system noise is shown in Figure 8.

The pilot regulator is similar in design to that used with the 300-channel equipment except that a crystal pilot pick-off filter is provided.

Since this transistor-type repeater is also capable of providing 960 channels as a direct replacement for the valve-type repeaters on the large-diameter 0·375-inch (9-millimetre) cable where repeater spacings vary considerably, the cable-simulating network has been built into the repeater subframe. It is not provided for 1·2/4·4-millimetre cable systems unless specifically required. It contains networks simulating 0·15, 0·3, 0·6, and 1·2 miles (0·25, 0·5, 1, and 2 kilometres) of small coaxial cable for each direction of transmission, each set of networks being in separate containers.

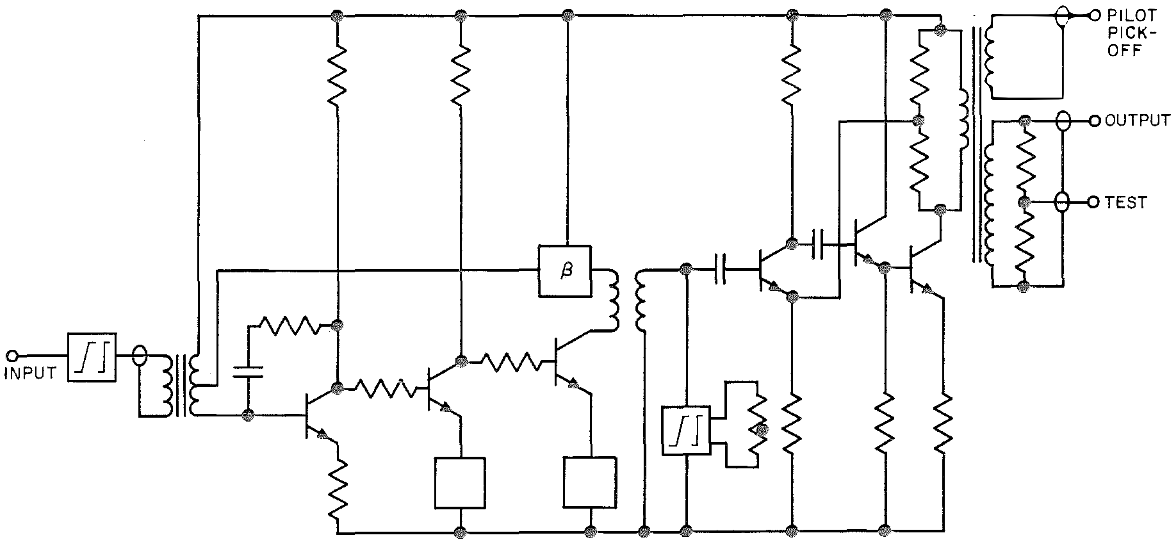

Figure 6—Explanatory circuit diagram for amplifiers in the dependent repeaters for 960-channel working.

3.4 Repeater Housing

Designed for use within an underground box or cable chamber (manhole), the repeater housing will take two subframes of either or both types side by side. In addition it provides a mounting for a flat plate which carries the interstice wire protectors and alarm diodes. A small, portable, battery-powered oscillator used for transmitting a monitoring pilot over short sections of the line for fault tracing purposes can also be slotted into this plate.

The box itself illustrated in Figures 9 and 10 is prefabricated from $\frac{3}{16}$-inch (4·7-millimetre) mild-steel plate, which is seam welded into the shape shown. Stiffening ribs are provided inside the slightly domed lid, and the ends of the box are drawn out to provide a mounting for the cable sealing ends and to ensure easy access

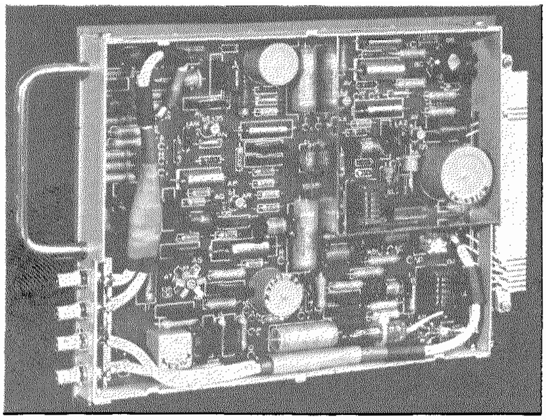

Figure 7—Line amplifier of 960-channel dependent repeaters.

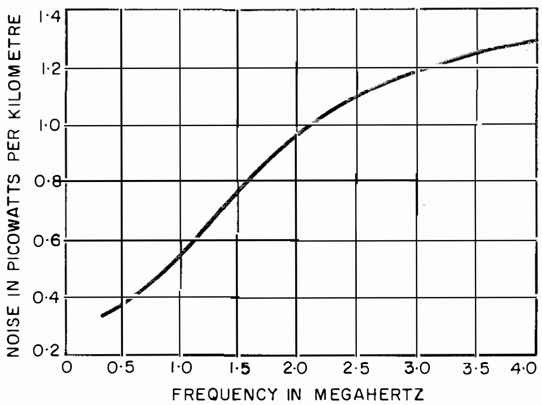

Figure 8—Calculated noise in picowatts referred to zero level psophometric for a 175-mile (280-kilometre) section of small-diameter coaxial cable for a 960-channel system. No allowance has been made for level misalignment or for terminal or power feeding stations required in this route.

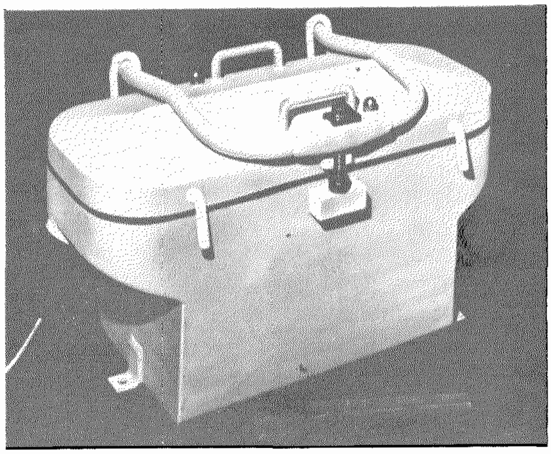

Figure 9—Repeater housing fully sealed.

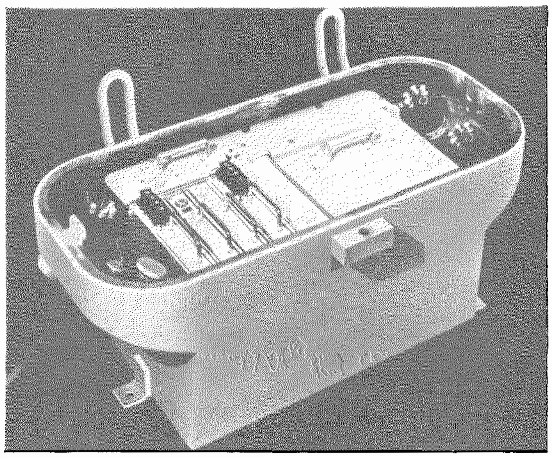

Figure 10—Repeater housing with cover removed showing one dependent repeater installed.

to them. The main cable may be brought in and out at one end of the box or at each end as desired and the third sealing end is used to provide a ground-level test point.

The use of a ground-level test point sited above the water line in the manhole enables the outputs of each repeater amplifier to be measured and easy access to the engineering speaker circuit without having to pump out the manhole or open the box.

The box may be equipped with a loading coil for the speech circuit and a gas-pressure contactor which will signal to the terminal repeater a fall in pressure from the normal internal box pressure of 8 pounds per square inch (0·6 kilogramme per square centimetre) to approximately 3 pounds per square inch (0·2 kilogramme per square centimetre). This allows remedial action to be taken to restore the seal before water can enter the box.

The box is equipped with an *0* ring to complete the lid seal and has feet on the lid to prevent the seal touching the ground when the lid is removed and placed to one side. Two Schrader valves in the lid allow the box to be flushed with dry air and the single-bolt fixing is safeguarded to avoid removal of the lid whilst still under pressure.

Earth lugs are provided inside and outside the housing, which may also be bolted to the floor of the manhole.

The cable sealing ends are epoxy resin mouldings equipped with neoprene *0* ring seals against the housing. Individual coaxial connectors are brought out for each coaxial tube and a small socket is equipped to terminate the 10 interstitial wires normally found within the cable.

3.5 Terminal Repeaters for 300 and 960 Channels

Manufactured in standard equipment practice, these equipments are mechanically and schematically similar, except that an extra amplifier having flat gain is required in the transmit shelf of the 960-channel version to compensate for higher network losses. The schematic for the 960-channel equipment is shown in Figure 11, and the terminal repeater is illustrated in Figure 12. A typical terminal repeater card for the 4092-kilohertz pilot oscillator is shown in Figure 13.

The high-frequency sections of the terminal repeater comprise transmit, receive, and common shelves, the latter providing functions common to both receive and transmit directions of transmission such as spare amplifiers and changeover apparatus. The supervisory shelf contains all the alarm and location apparatus for one terminal repeater including an inbuilt 20-volt to 120-volt direct-current to direct-current converter to drive the fault location apparatus. The power supply unit for the terminal and remote power supply unit for the repeaters each take up one shelf and the recorder and pilot display meters require two shelves.

The range of input levels possible on both equipments is designed to cater for all applications of line equipment including radio systems and recommendations of the International Telegraph and Telephone Consultative Committee and the British Post Office. Levels are −15 to −27 decibels referred to 1 milliwatt in 75-ohm unbalanced circuits at the output and −15 to −45 decibels referred to 1 milliwatt at the input of the systems.

Signals applied at the input to the high-frequency line are first passed through pilot stop filters of 308 and 4092 kilohertz for 960 channels or 60 and 1364 kilohertz for 300 channels to ensure that no extraneous signals at these frequencies cause interference with the line pilots. These pilots are than injected into the signal band after amplification, pre-emphasis, and, if required, pre-residual equalization. After further amplification the signals go to the cable via cable-simulating networks (which enable the first repeater section to be short if required) and power combining filter to the line. Duplicate pilot oscillators with fast changeover are provided for the regulating line pilot whilst the second pilot may be generated (in the 960-channel case of 308 kilohertz) or stabilized (in the 300-channel case of 60 kilohertz). Pilot alarms operating on variations of level of the outgoing pilot frequency are fitted to the output of the generators to operate at 0·5 decibel to give an alarm and if necessary to change oscillators. The changeover unit, which uses mercury-wetted relays, may be preferentially set to either oscillator.

Incoming signals from the coaxial line are received via a power separating filter and cable-simulating network similar to those in the transmit shelf. A line amplifier and regulator between them ensure that the correct levels are then passed on for de-emphasis and residual equalization. After amplification the line pilots are extracted and fed away to the monitoring and recording equipment, ensuring that all the amplifiers in the system are guarded by the pilots against failure. The main high-frequency signal band is then passed via pilot stop filters to the output from the line equipment for connection to the frequency-translating apparatus.

The receive line amplifier, except for the card on which it is mounted, is identical with that in the underground repeater. A switched simulation error equalizer may be fitted in the receive path on systems of greater than 24 repeater sections, should this prove to be necessary in service.

The monitor units consist of crystal pick-off filter, amplifier, and rectifier circuits, driving independently controlled chart recorders and decibel meters.

The 60-kilohertz pilot on the 300-channel equipment is, in addition, amplified and supplied to various distribution outlets for frequency comparison purposes.

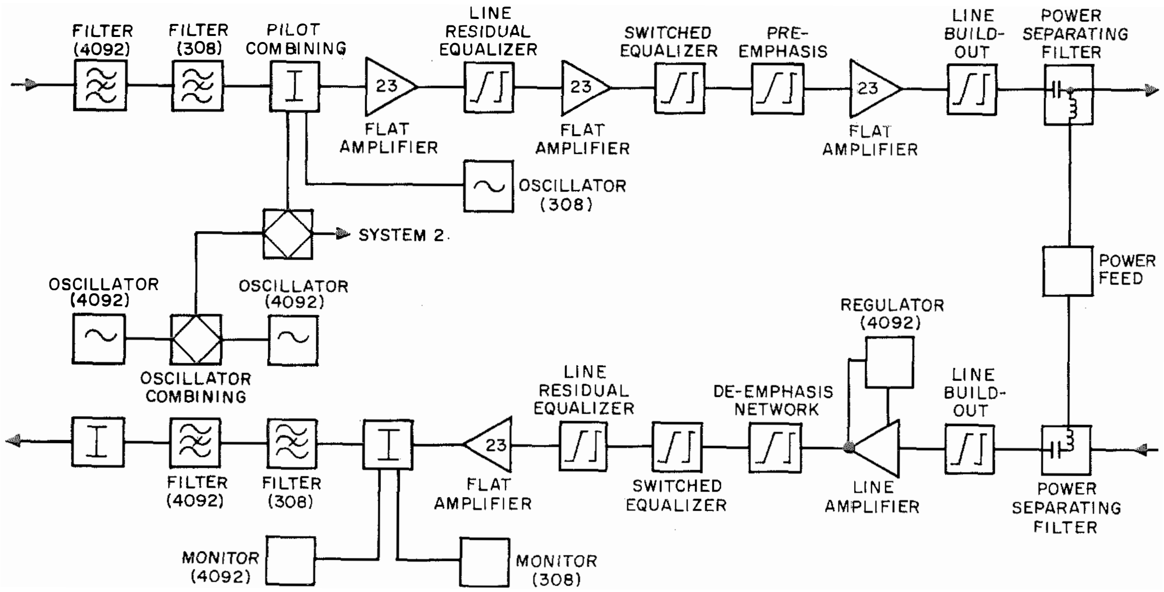

Figure 11—Terminal repeater for transmit and receive for 960-channel system. Frequencies are shown in kilohertz in parentheses. Amplifier gains are in decibels.

Protection is provided against current surges for the incoming and outgoing signal paths at the line terminals and for each interstice wire.

As a special requirement for the British Post Office and not generally supplied on export equipment, the common high-frequency shelf mounts a jack field for the selection of the input and output terminals of any working amplifier. This together with amplifier changeover circuits enable any amplifier to be replaced with less than 1-millisecond break in transmission.

The supervisory shelf provides a terminal alarm panel on which a variety of faults are shown such as power feed current failure, blown fuses, incoming or outgoing pilot failure, et cetera. In addition dependent alarm location and other subsidiary facilities are provided. The terminal apparatus for the speaker circuit is provided on this shelf, and this consists of a loudspeaking telephone and audio amplifiers. The circuit is normally loaded at 88 millihenries per mile (55 millihenries per kilometre) giving a useable frequency band from 300 hertz to 3·5 kilohertz up to 90 miles (144 kilometres) in length and may be of 2 or 4 wires.

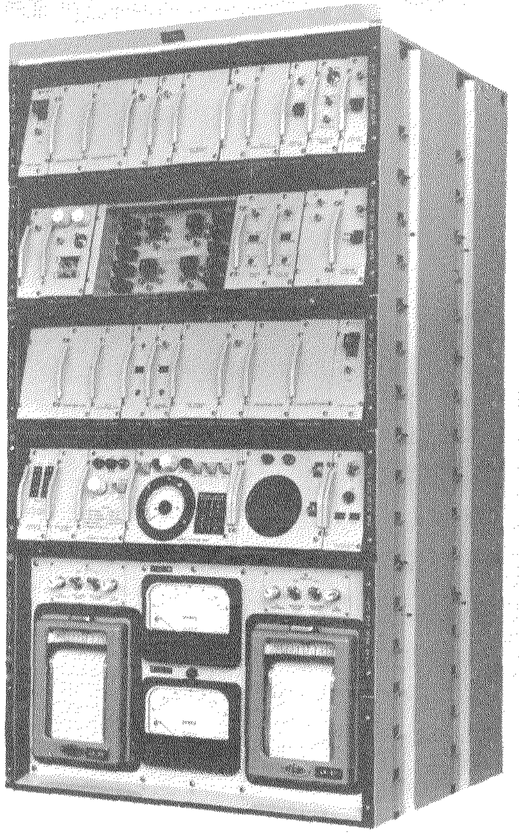

Figure 12—Terminal repeater for 960 channels.

3.6 Power Feeding Repeater

This is a rack mounted dependent repeater equipped with residual-equalization facilities and with power feeding units. This replaces the more conventional main station, which is not included in the range of equipment since it would have only a marginal noise advantage (one flat amplifier omitted) over a back-to-back terminal repeater and is rarely required. On occasions however, the route length exceeds the power feeding distance thus requiring extra power units. To avoid the relatively expensive solution of back-to-back terminal repeaters, the power feeding repeater is used.

This repeater is mounted in one shelf, the three power units (for power feeding each direction of transmission and for general power supplies) completing the repeater.

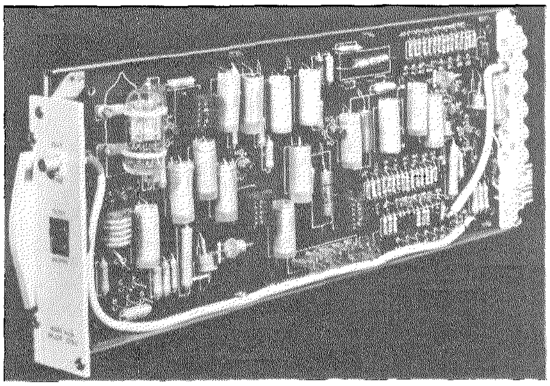

Figure 13—Typical card of a terminal repeater. It is for the 4092-kilohertz pilot oscillator in the 960-channel system.

4. Data Summary

4.1 Repeater Amplifier for 960 Channels

Nominal Gain: 6·3 decibels at 60 kilohertz and 43·28 decibels at 4092 kilohertz.

Gain Control Shape: ±0·75 decibel at 60 kilohertz and ±3·90 decibels at 4092 kilohertz.

Return Loss: Compared with 75 ohms, >23 decibels at 60 kilohertz and >20 decibels at 4092 kilohertz.

Overload Point: +20 decibels referred to 1 milliwatt with a series direct-current feed to the amplifier of 48 milliamperes and with a sine-wave input.

Noise Figure: 5·3 decibels average compared with −139 decibels referred to 1 milliwatt in a 3·1-kilohertz bandwidth as zero level.

Intermodulation: Measured as the sum and difference products of two tones, each fundamental tone giving an output of +5 decibels referred to 1 milliwatt with the amplifier adjusted to nominal gain.

Fundamental Frequencies in Megahertz A	Fundamental Frequencies in Megahertz B	Sum or Difference in Megahertz	Intermodulation in Decibels Referred to 1 Milliwatt
4·1	3·9	0·2	−75·6
2·2	1·9	4·1	−59·0
1·9(2A)	3·6	0·2	−84·5
3·6(2A)	3·1	4·1	−73·5

Transmit Levels: −19 decibels referred to reference level at 60 kilohertz and −9 decibels referred to reference level at 4028 kilohertz.

Ambient Temperature: Gain changes 0·02 decibel at 4 megahertz for a rise of 24 degrees Celsius in ambient.

Noise Power: Calculated to be 2·03 picowatts per mile (1·27 picowatts per kilometre) assuming ideal line levels throughout a homogeneous section of 175 miles (280 kilometres).

Power Consumption: Pre-amplifier is 25 milliamperes, power amplifier is 50 milliamperes, and regulation is 37 milliamperes, all at 12 volts.

Line Pilots: 308-kilohertz for monitoring and 4092 kilohertz for line regulating.

Repeater Spacing: 2·5 miles (4 kilometres).

4.2 Repeater Amplifier for 300 Channels

Nominal Gain: 8·6 decibels at 60 kilohertz and 35·8 decibels at 1364 kilohertz.

Gain Control Shape: ±1·3 decibels at 60 hertz and ±3·7 decibels at 1364 kilohertz.

Return Loss: Compared with 75 ohms, >25 decibels at the input and output of the amplifier.

Overload Point: +17 decibels referred to 1 milliwatt with a series direct-current feed of 28 milliamperes and with a sine-wave input.

Noise Figure: 12 decibels average compared with −139 decibels referred to 1 milliwatt in a 3·1-kilohertz bandwidth as zero level.

Harmonic Performance: Measured with the fundamental adjusted to +10 decibels referred to 1 milliwatt at output of amplifier.

Fundamental in Kilohertz	Decibels Referred to 1 Milliwatt: 2nd Harmonic	Decibels Referred to 1 Milliwatt: 3rd Harmonic
100	−60	−77
400	not measured	−65
650	−58	outside band

Transmit Levels: −18 decibels referred to reference level at 60 kilohertz and −13 decibels referred to reference level at 1364 kilohertz.

Ambient Temperature: Gain changes +0·1 decibel for a rise of 45 degrees Celsius in ambient.

Noise Power: Calculated to be 4·8 picowatts per mile (3 picowatts per kilometre) assuming ideal line levels throughout a homogeneous section of 175 miles (280 kilometres).

Power Consumption: Amplifier is 28 milliamperes at 12 volts and regulator is 37 milliamperes at 15 volts.

Line Pilots: 60 kilohertz for monitoring and frequency-comparison and 1364 kilohertz for line regulating.

Repeater Spacing: 3·75 miles (6 kilometres).

5. Acknowledgments

The author wishes to acknowledge the information and assistance given him within the Land Line Systems Division of Standard Telephones and Cables and at Standard Telecommunication Laboratories, Harlow.

6. References

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Multichannel Telephone Equipment of Standard Téléphone et Radio for Small-Diameter Coaxial Cable

P. GFELLER

Standard Téléphone et Radio S.A.; Zurich, Switzerland

1. Introduction

The first article in this series describes the primary planning for carrier telephone systems connected by small-diameter coaxial cable [1]. Parts 2 and 3 give the details of the equipment designs by Standard Elektrik Lorenz [2] and Standard Telephones and Cables [3] to meet the special requirements of their countries and customers. Part 4 now outlines the special features of the 300- and 1260-channel systems designed by Standard Téléphone et Radio (Switzerland).

The 300-channel system, the first of which was put into service recently, has a repeater spacing of 6 kilometers (3.75-miles). This system was specified in close collaboration with the Swiss Post, Telegraph, and Telephone Administration and in accordance with the recommendations of the International Telegraph and Telephone Consultative Committee. To extend the bandwidth of the small-diameter coaxial cables, the Swiss Post, Telegraph, and Telephone Administration has decided to divide the 6-kilometer (3.75-mile) repeater sections of the 300-channel system into two sections of half this length and transmit 1260 channels in accordance with the recommendations of the International Telegraph and Telephone Consultative Committee. The main reason for this decision is that the existing Swiss standard-diameter coaxial network, which is equipped with valve repeaters at a 9-kilometer (5.6-mile) spacing, has a capacity of 1260 channels. In addition, the 4000-megahertz radio relay network will in the near future be expanded from 960 to 1260 channels. Thus in a few years a uniform band of 1260 channels will be transmitted in the band from 60 to 5564 kilohertz on all long-haul connections of the Swiss coaxial-cable and radio relay network. The 1260-channel system is now being developed for a field trial in 1967.

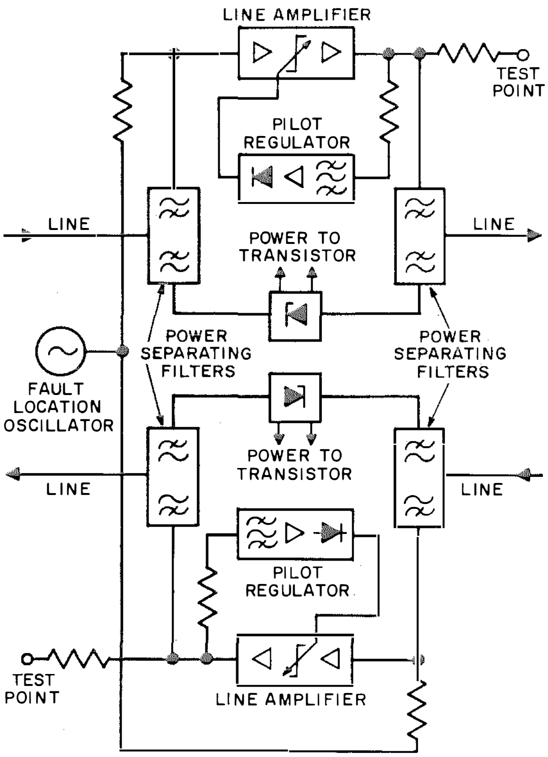

Figure 1—Block diagram of the 300-channel dependent repeater.

2. Characteristics for 300 Channels

The block diagrams of the dependent repeater (Figure 1) and of the terminal repeater (Figure 2), which are not explicitly dealt with in the text below, indicate the general layout of the line equipment.

2.1 Cable

The cable standardized by the Swiss Post, Telegraph, and Telephone Administration contains 10 coaxial pairs of the expanded-insulation type with 1.2/4.4-millimeter diameter coaxial conductors, the outer conductor being 0.15 millimeter thick. All outer conductors are connected together and earthed. Six intersticial pairs of 0.6-millimeter conductors are provided for operating and power feeding of the fault location oscillators, for the speaker circuit, and as spare pairs.

2.2 Repeater Sections

The nominal value of the repeater section length is 5.94 kilometers (3.7 miles) and the tolerance allowed is ±200 meters (±656 feet), for which no gain adjustment is provided. There is no cumulative inaccuracy of section length since the sum of n section lengths (neglecting the end sections) will be within n times 5.94 ±0.2 kilometers.

The end sections, which may be short, can be extended to 5.94 ±0.1 kilometers (3.7 ±0.06 miles) by a line building-out network. Up to 12 dependent repeaters may be employed between terminal repeaters.

2.3 Frequency Plan

The frequency band employed for telephone circuits extends from 60 to 1300 kilohertz, and the 14 repeater checking frequencies are located within the band 1320 to 1333 kilohertz. A 1364-kilohertz line regulating pilot is used, and a second line pilot is transmitted at 308 kilohertz. For frequency comparison purposes 60 kilohertz is also transmitted.

2.4 Line Levels and Noise

The transmit levels at the input of each repeater section at 1300 and 60 kilohertz are −13

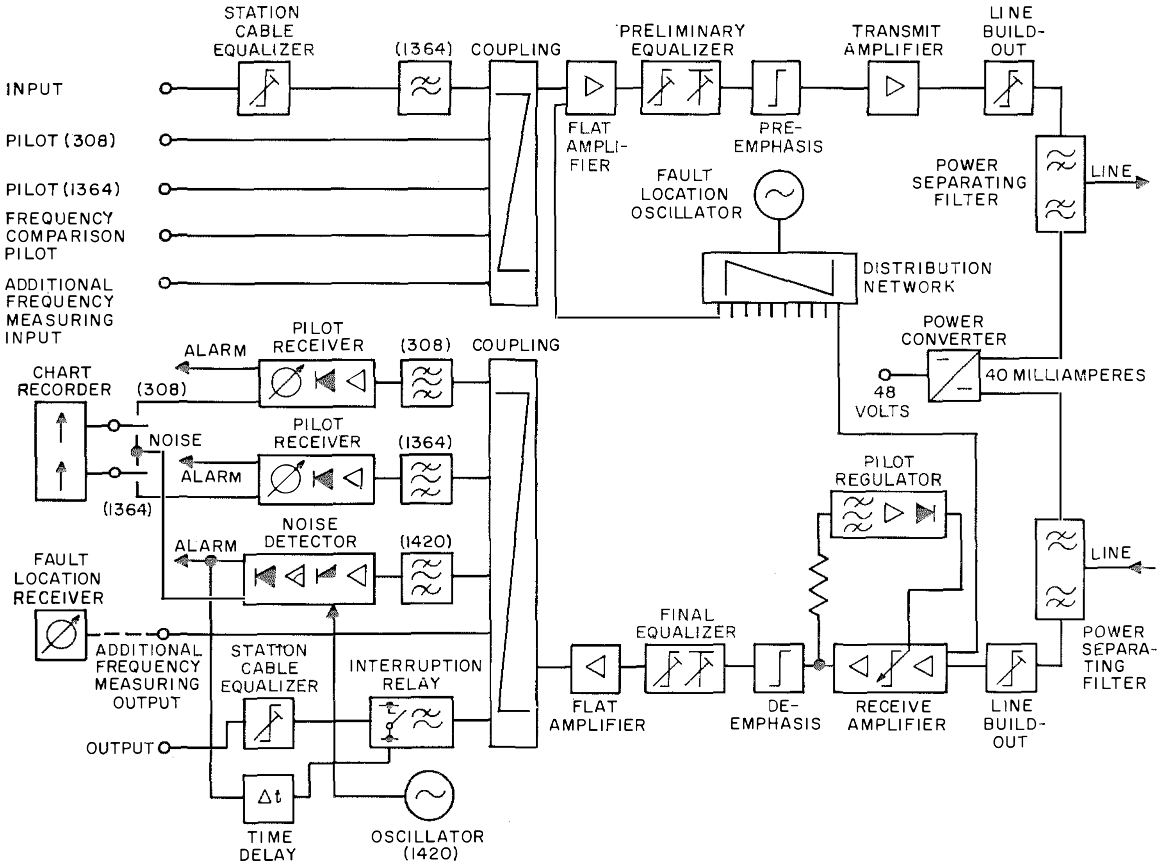

Figure 2—Block diagram of the 300-channel terminal repeater station. Frequencies in kilohertz are in parentheses.

and −22 decibels referred to reference level, respectively, and between these two frequencies the transmit level slopes approximately in accordance with 10 log $[1+6.4(f/f_{max})^4]$. The corresponding levels at the end of a repeater section are, at average cable temperature, −49 decibels referred to reference level at 1300 kilohertz and −31 decibels referred to reference level at 60 kilohertz.

With a repeater noise figure of less than 8.7 decibels at the highest frequency, the basic line noise per channel per kilometer will not exceed 0.7 picowatt psophometric referred to zero relative level.

The total noise measured with white-noise loading of +9.8 decibels relative 1 milliwatt referred to a point of zero relative level and with nominal levels is less than 2 picowatts psophometric referred to zero relative level per kilometer in each channel.

2.5 Level Regulation

Having regard to the large tolerance on repeater section length of ±200 meters (±656 feet) and due to some uncertainty, it has been decided to provide pilot regulation in each dependent repeater in the first systems. The control ratio of the regulators in the dependent repeaters is approximately 4 to obtain a good dynamic characteristic for the line. The regulators in the terminal repeaters are designed with a control ratio of 15. In the case of an interruption of the line regulating pilot, the gain of a pilot-regulated dependent repeater will be increased by about 4.3 decibels at 1300 kilohertz and by 1.7 decibels at 60 kilohertz.

2.6 Monitoring and Fault Location

In the terminal repeater stations the 308-kilohertz and 1364-kilohertz line pilots are monitored by pilot receivers that initiate an alarm if the pilot level decreases by more than 1.3 to 1.7 decibels. Individual measuring instruments show the level deviations of each of the pilots of the 5 systems, and an inbuilt centralized twin-channel chart recorder enables the two pilots of one system to be supervised simultaneously.

As the pilot regulators have no memory and since no alternative heating is provided for the thermistors, the line will increase its gain in the case of a failure of the line regulating pilot. Thus a noise-monitoring device is provided in the terminal repeater station. This device separates a 10-kilohertz band of noise at 1420 kilohertz by filtering, this noise band is amplified and then translated with the aid of a 1420-kilohertz auxiliary oscillator. The voice-frequency noise band received in this way is passed via a logarithmic network to a rectifier. Two alarm triggers, whose operating levels may be adjusted, evaluate the direct-current signal and operate urgent and nonurgent alarms. When the urgent alarm is given at a noise level of about 10^6 picowatts psophometric referred to a point of zero reference level, the system output after a short time (adjustable between 1 and 20 seconds) is short-circuited. The twin-channel chart recorder can also be used to monitor the system noise and one pilot simultaneously. Special paper allows the noise level to be read directly as picowatts psophometric referred to a point of zero reference level at 1420 kilohertz.

For fault location an oscillator is installed in each dependent repeater and terminal repeater station to generate a checking frequency that is specific to that repeater station, and injected at the input of the repeater. The 14 frequencies required have a nominal level of −17 decibels relative 1 milliwatt referred to a point of zero relative level and are spaced at 1 kilohertz between 1320 and 1333 kilohertz. In the terminal repeater station a fault location receiver is used to check the condition of the line. The fault location oscillators are switched on as necessary, this being done via an intersticial-pair loop from one of the two terminal stations.

2.7 Remote Power Feeding

As with the other systems described, the remote power feeding is by direct current with series-connected repeaters and the current is transmitted over the inner conductors of the system. The maximum feeding power allowed by the Swiss Post, Telegraph, and Telephone Administration is 50 milliamperes at 200-0-200 volts, and the maximum length of a power feeding section is fixed at 78 kilometers (49 miles) with 12 dependent repeaters. In this system up to 6 repeaters are power fed from one end with 40 milliamperes at 165-0-165 volts, and to feed 7 to 12 repeaters two feeding loops are required, one from each end. The coaxial outer conductors from this system are earthed over the whole cable length and are through connected.

2.8 Overvoltage Protection

The repeaters are protected by gas discharge tubes as primary protection and by diodes as secondary protection in such a way that they are able to withstand without damage transient surges of 4 kilovolts peak with a rise time of 1 microsecond and a fall time to half voltage of 50 microseconds as well as 1200 volts at 50 hertz for 1 second. In the case of permanent induction from electric railways, longitudinal voltages on the cable of 150 volts at 16.6 hertz will not cause any interference with transmission.

2.9 Mechanical Construction

The 300-channel system consists of the dependent repeater case, a terminal repeater rack, and an auxiliary rack. These are described below.

The two racks that house the terminal equipment are designed in the new equipment practice 62 (BW 62), which is the result of cooperation among the Swiss Post, Telegraph, and Telephone Administration and its suppliers [4]. The BW 62 racks have dimensions of 2736 by 540 by 225 millimeters (108 by 21 by 9 inches) and are subdivided into 19 shelves equipped with plug-in units of different widths.

The dependent repeater case, shown in Figures 3 and 4, is a watertight cast-iron box, standardized by the Swiss Post, Telegraph, and Telephone Administration. It contains 3 cable terminations. A frame for the plug-in units can be swung out of the case. Two of the cable terminations at the top provide for the 10 coaxial pairs and 6 intersticial pairs required for the two directions of transmission, and the third termination carries measuring points to the outside of the case. The frame for the plug-in units can be equipped with line amplifiers for 5 systems, which are also constructed in equipment practice 62. Turning the frame inward provides access to the cabling, while turning it outward exposes the units. The power separating filters are located to the left half of the top section adjacent to the preamplifier, output amplifier, and pilot regulators for the *A-B* direction of system 1. Equipment for the *B-A* direction of that system is placed on the right half of that section. The lower 4 sections

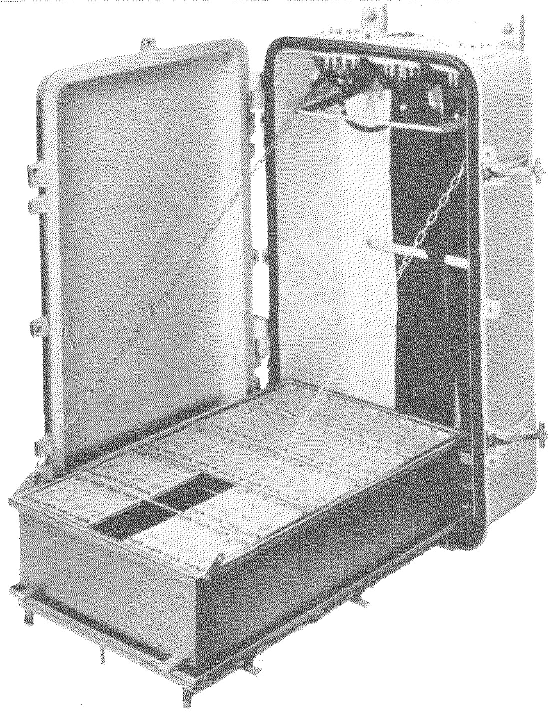

Figure 3—Dependent repeater in the turned-out position to provide access to the panels.

contain the repeater equipment for systems 2 through 5 in the same way.

The terminal repeater rack in Figure 5 houses the main part of the transmission and monitoring equipments for 5 systems. The plug-in units belonging to one system are arranged vertically, which is somewhat unusual in this equipment practice.

In the auxiliary rack to the left is the equipment directly associated with the cable. This includes cable terminations, line building-out networks, power separating filters, and the power feeding units. From these cable terminations each coaxial pair is led to *U* links that allow a through connection of two line regulating sections. This through connection is only possible when the two adjacent repeater sections are shorter than 2.97 kilometers (1.85 miles). The rack also contains the fault location receiver and the chart recorder. A jack field allows measurements to be made with additional measuring frequencies (auxiliary pilots) and also for the connection of the fault location receiver with any of the 5 systems.

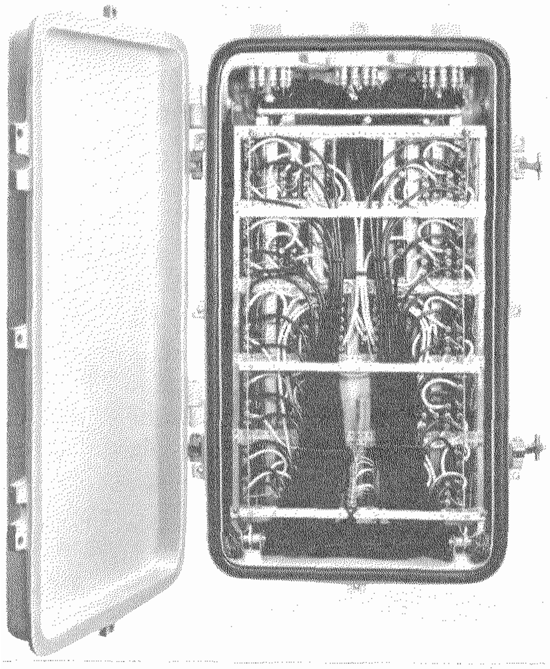

Figure 4—Dependent repeater in the turned-in position to provide access to the cabling.

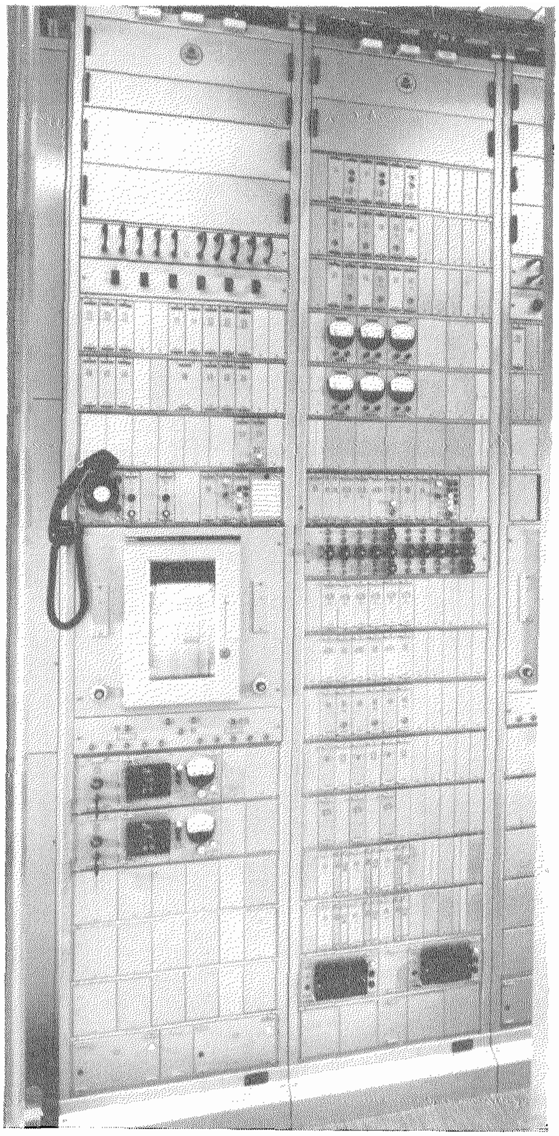

Figure 5—300-channel terminal repeater station. On the left-hand side is shown the auxiliary rack, equipped with two power feeding units, and on the right-hand side the terminal repeater rack, equipped with 3 systems each in a vertical array.

3. Characteristics for 1260 Channels

The 1260-channel system is an extension of the existing 300-channel system and not all facilities are yet fixed.

The repeater section length is 2.97 kilometers (1.85 miles) with a tolerance of ±200 meters (±656 feet) on existing routes and ±100 meters (±328 feet) on new routes. The maximum number of repeaters between two terminal stations is 26. The frequency band occupied by the telephone channels lies between 60 and 5564 kilohertz and the band for the checking frequencies is approximately 5600 to 5800 kilohertz. The frequency allocation is as follows.

Supergroups 1 through 16 are transmitted in the normal way between 60 and 4028 kilohertz, and supergroups 17 through 21 are translated into the band 4332 to 5564 kilohertz in an inverted position. This frequency band is the same as that of master group 4 in the 12-megahertz system.

The frequency of 4287 kilohertz is used as a main line regulating pilot and 308 kilohertz is provided as a second line pilot.

The transmit level and the pre-emphasis characteristic are in accordance with provisional Recommendation G344 of the International Telegraph and Telephone Consultative Committee, which calls for −13 decibels relative reference level at 5564 kilohertz, −17 decibels relative reference level at 4287 kilohertz, and −23.5 decibels relative reference level at 60 kilohertz. The pre-emphasis characteristic follows the law

$$A_{\mathrm{dB}} = 10 \log \left[1 + \frac{10}{1 + \dfrac{2.20}{(f/f_r - f_r/f)^2}} \right]$$

where f_r is the resonance frequency of 5.75 megahertz and the repeater input levels are −50 decibels relative reference level at 5564 kilohertz and −26.4 decibels relative reference level at 60 kilohertz.

The noise figure in the highest telephone channel is less than 6 decibels and thus the basic noise is approximately 1 picowatt psophometric relative a point of zero reference level per kilometer.

The overload point is +17.4 decibels referred to 1 milliwatt and this value is 7 decibels greater than the value recommended by the International Telegraph and Telephone Consultative Committee for a system with 1260 channels.

For comparison some typical data for the 300- and 1260-channel repeaters are given below.

	300 Channels	1260 Channels
Repeater spacing in kilometers	6	3
Repeater spacing in miles	3.75	1.88
Transmit level at maximum frequency in decibels relative reference level	−13	−13
Input level at maximum frequency in decibels relative reference level	−48.7	−50
Pre-emphasis in decibels	8.7	10.5
Required overload point with 7-decibel margin (power reduction due to pre-emphasis assumed to be 4.3 decibels) in decibels referred to 1 milliwatt	+12	+17.4
Noise figure of the repeater in decibels	8.7	6
Basic noise at maximum frequency in picowatts psophometric relative to a point of zero reference level per kilometer	0.7	1
Harmonic distortion requirement (the same effect of pre-emphasis assumed)	H2 H3	H2+9 decibels H3+18.3 decibels

This comparison shows that the transmit level at maximum frequency of both systems is the same. Therefore to get the same basic noise the noise figure of the repeater with the broader frequency band must be better by about 4.3 decibels. It can further be seen from this table that the 1260-channel repeater for 3 kilometers (1.9 miles) is somewhat easier to realize than the 960-channel repeater for 4-kilometer (2.5-mile) repeater spacing [2].

The 1260-channel system is planned with a level control that is a combination of pilot and thermometer regulation, and in the event of a failure of the line regulating pilot the increase in repeater gain will be limited either by an alternative method of heating the thermistor or by a memory-type regulator.

Monitoring of the noise will also be provided in this system at a frequency of about 4170 kilohertz.

The remote feeding power allowed is 60 milliamperes and 250-0-250 volts, with a maximum length of 81 kilometers (51 miles) for each power feeding section, and a maximum of 13 repeaters are power fed from one end of the main section.

The line equipment will be designed initially for telephone transmission only, with a possible requirement for television transmission at a later date. This will entail a specially designed repeater.

4. Conclusion

The two carrier line equipments described here will be used in a wide network of small-diameter coaxial cables and will help to satisfy the ever-increasing demand for long-distance telephone circuits.

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High-Power Varactor Frequency-Doubler Chains

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Economical design of high-power varactor frequency multipliers requires that the varactor diodes are operated under conditions of forward drive. Both lumped and distributed coupling networks may be used and experimental results are compared with design predictions for a 6-stage circuit producing an output of 1 watt at 4 gigahertz from an input of 10 watts at 67 megahertz. The techniques described have also been applied satisfactorily to the design of multistage circuits capable of an output of 5 watts at 4 gigahertz. Methods of testing complete multiplier chains are described.

1. Introduction

A considerable number of papers have been published in the past few years concerning the design and performance of varactor frequency multipliers, but the basis of design has invariably been the capacitance-voltage characteristic of the varactor in the reverse-bias condition $C \propto (\phi - V)^{-1/x}$, where x is 2 or 3 depending on the type of junction.

If, however, the voltage across the varactor is not restricted to the reverse-bias region, but is allowed to drive the varactor into the forward-bias region for a fraction of the cycle, charge storage effects are apparent, and the power handling capacity of the varactor is increased by a factor often in excess of 10 times that corresponding to reverse-bias operation. For operation in the forward-bias region, many of the considerations for reverse-bias operation no longer apply.

As there is no suitable theory to describe the operation of a varactor in the forward-bias region, the design of circuits must be based on the results of measurements made on typical varactors. The present work is restricted to a consideration of frequency doublers, however, for in this case the power handling capacity of the varactor is maximized.

The varactors used in the experimental work were all silicon types and they are representative of the products of a number of manufacturers.

2. Dynamic Impedance of a Varactor

To design the coupling networks between a source and a varactor and between the varactor and a load, it is necessary to know the input and output impedance of the varactor under dynamic conditions.

Measurements were made, therefore, on the simple doubler circuit shown in Figure 1, in which capacitor C_1 tunes the input circuit to resonance at f and C_2 tunes the output circuit to $2f$. The taps on the auto-transformers were adjusted to give optimum performance between 50-ohm terminations.

The input match was observed by means of a directional power meter, and the circuit was adjusted for maximum efficiency, using an input frequency of 67 megahertz, for a number of input power levels between 1 and 20 watts. At each power level, the varactor was disconnected from the circuit and replaced by a series resistance-capacitance circuit, which was then adjusted to give a good input match.

It was found that the input resistance of the varactor was essentially constant at 22 ohms for all power levels, but that the input

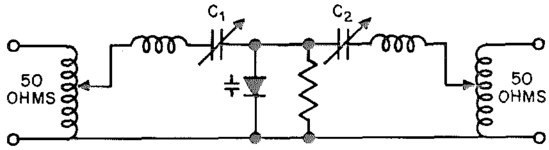


Figure 1—Circuit used to measure varactor input impedance.

capacitance varied as a function of power level as indicated in Figure 2. It is interesting to note that the input capacitance C is greater than the zero-bias capacitance C_0 of 28 picofarads, even at the 1-watt level. An equivalent input circuit of the varactor is, therefore, as shown in Figure 3, where r is the series resistance of the diode—a parameter frequently specified by the manufacturer—in series with a conversion resistance R, capacitance C, and inductance L. The circuit is shunted by the case capacitance of the diode C_c. The conversion resistance represents the ability of the varactor to convert power at frequency f to $2f$.

If one assumes that all silicon varactor types which exhibit charge storage effects are similar except for zero-bias capacitance, series resistance, and breakdown voltage, then it is possible to estimate the behaviour of the other types.

Thus, a varactor which has a nominal capacitance C behaves in a similar manner to two varactors of capacitance $C/2$ connected in parallel; that is, the conversion resistance is halved and the capacitance doubled.

The conversion resistance R was assumed to vary inversely as the input frequency. This is in accordance with the predictions of the theory of reverse-bias operation. Subsequent measurements at higher frequencies proved the validity of the assumption. The conversion resistance of a silicon varactor is plotted as a function of frequency in Figure 4, and the input capacitance is shown as a function of power level in Figure 2.

2.1 Choice of Optimum Load Resistor

To determine the optimum load to present to a varactor operating as a doubler, a circuit was set up in which the input circuit was the same as that used in Figure 1, but the output circuit was replaced by a double-tuned bandpass circuit employing a length of short-circuited 50-ohm line as the coupling inductor. It is well known that a circuit of this type has the property of matching resistors R_p and R_s connected to the primary and secondary

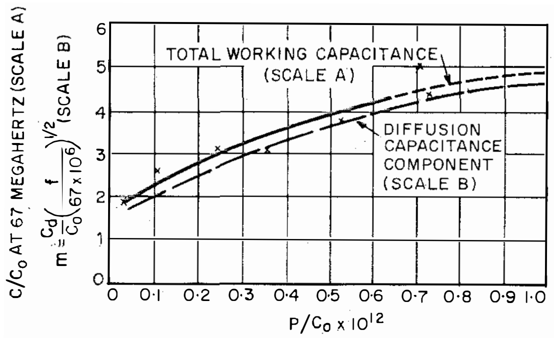


Figure 2—Working capacitance of silicon varactors.

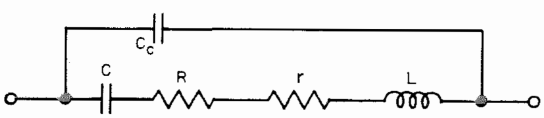


Figure 3—Equivalent circuit of varactor.

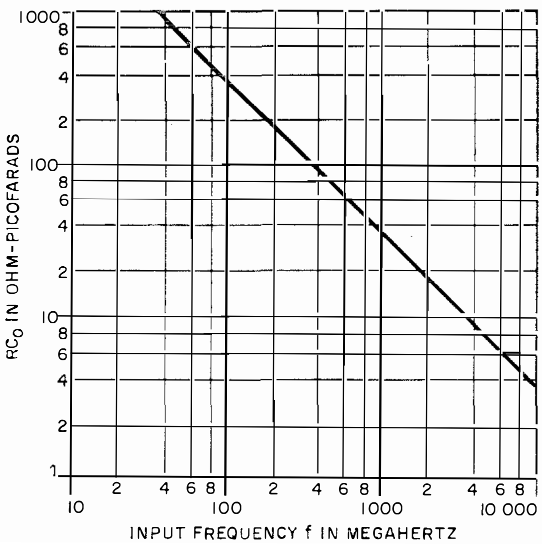


Figure 4—Conversion resistance of a doubler.

branches if $X^2 = R_p R_s$, where X is the coupling reactance. As $R_s = 50$ ohms and X can be accurately estimated from the length of line, this provides a method of calculating R_p under dynamic conditions. As might be expected, the change in efficiency with variation in R_p is quite slow, but a value of $R_p = R$, the conversion resistance, seemed to be the optimum value.

3. Varactor Efficiency

Once the conversion resistance and optimum load resistance are known, it is possible to calculate the efficiency of the varactor.

Thus, if the varactor receives an input power P at frequency f, the fraction $PR/(R+r)$ is converted to $2f$, the remainder being dissipated in the series resistance r of the varactor. As current at the output frequency $2f$ must also flow through the series resistance r, the useful output power is further reduced by the ratio $R/(R+r)$. It follows, therefore, that the ratio of output power at $2f$ to input power at f is

$$\eta = R^2/(R+r)^2 \qquad (1)$$

which is the varactor efficiency.

The varactor loss P_d is obviously

$$P_d = P(1-\eta) = \frac{Pr(2R+r)}{(R+r)^2}. \qquad (2)$$

In addition to the varactor loss, circuit losses will occur in both the input and output circuits. The circuit loss depends on the circuit configuration, and this will be discussed in Section 4.

3.1 Efficiency as a Function of Cut-Off Frequency

The cut-off frequency f_c of the varactor, defined as the frequency at which the reactance of the varactor at break-down voltage is equal to the series resistance, is often used as a figure of merit. Thus

$$1/2\pi f_c C_b = r$$

where C_b is the capacitance at break-down voltage.

Now from Figure 4, the conversion resistance is proportional to the reactance of the varactor at input frequency f. Hence

$$R = \frac{K}{2\pi f C_0} = \frac{Krf_c}{f}\frac{C_b}{C_0} \qquad (3)$$

where K is a constant.

Substituting (3) into (1) gives

$$\eta = \frac{K^2 f_c^2 C_b^2/f^2 C_0^2}{(1 + K f_c C_b / f C_0)^2}.$$

Now from the data relating to the varactor used in the experiment we have $R + r = 22$ ohms, $r = 1{\cdot}65$ ohms, $C_b = 3{\cdot}9$ picofarads, and $C_0 = 28$ picofarads. Hence $f_c = 24{\cdot}7$ gigahertz and

$$K = \frac{C_0 f R}{C_b f_c r} = 0{\cdot}24$$

which gives

$$\eta = \frac{5{\cdot}76 \times 10^{-2} f_c^2 C_b^2 / f^2 C_0^2}{(1 + 0{\cdot}24 C_b f_c / f C_0)^2}.$$

The varactor loss is plotted in Figure 5 as a function of the parameter $C_b f_c / C_0 f$.

3.2 Power Handling Capacity

The power handling capacity of a varactor may be limited by either the varactor dissipa-

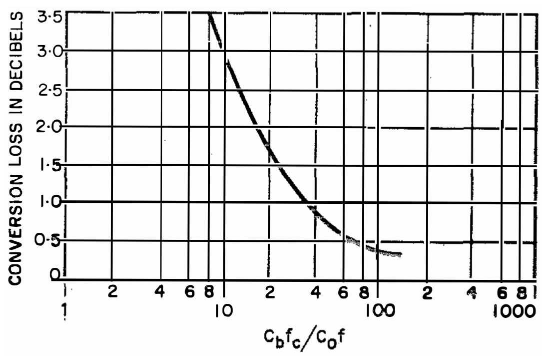


Figure 5—Conversion loss of a varactor diode in a doubler circuit.

tion P_d given by (2) reaching the maximum permissible level or by the peak voltage developed across the varactor reaching the break-down value. The voltage developed across the varactor has two main components, the voltages at f and $2f$, respectively. The peak current I_f at frequency f is given by

$$I_f = \left(\frac{2P}{R+r}\right)^{1/2}$$

where P is the input power.

From Figure 3, the input impedance Z_f of the junction at frequency f is

$$Z_f = \left(R^2 + \frac{1}{\omega^2 C^2}\right)^{1/2}$$

and consequently the peak voltage V_f at frequency f developed across the varactor is

$$V_f = I_f Z_f = \left\{\left(\frac{2P}{R+r}\right)\left(R^2 + \frac{1}{\omega^2 C^2}\right)\right\}^{1/2}. \quad (4)$$

Similarly, the peak output current at frequency $2f$ is

$$I_{2f} = \left(\frac{2\eta P}{R}\right)^{1/2}.$$

As the maximum power transfer occurs when the load circuit is the conjugate of the varactor impedance, it follows that the varactor impedance at $2f$ is

$$Z_{2f} = \left(R^2 + \frac{1}{4\omega^2 C^2}\right)^{1/2}.$$

The peak voltage developed across the varactor at $2f$ is therefore

$$V_{2f} = \left\{\frac{2\eta P}{R}\left(R^2 + \frac{1}{4\omega^2 C^2}\right)\right\}^{1/2}. \quad (5)$$

Now it may be shown that the maximum peak-to-peak value of the function $\sin(\omega t + \theta) + A \sin 2\omega t$ is approximately $3^{1/2}(1 + A)$ for the range $1 > A > 0{\cdot}25$.

It follows, therefore, that the peak-to-peak voltage developed across the varactor junction is

$$\hat{V} = 3^{1/2}(V_f + V_{2f}) \quad (6)$$

and if it is assumed that the voltage drive into the forward region is small, then the value given by (6) must be less than the break-down voltage of the varactor.

3.3 Working Capacitance as a Function of Frequency

The capacitance C in the above equations is the working capacitance of the varactor at a power level P and frequency f. It is necessary therefore to consider the variation of this capacitance as a function of power and frequency.

The capacitance C can be considered as the sum of the mean depletion-layer capacitance $C_{\bar{v}}$ and the mean diffusion capacitance C_d of the varactor, the former being independent of frequency whilst the latter varies [1] as $(f)^{-1/2}$. Thus

$$C/C_0 = (C_{\bar{v}} + C_d)/C_0.$$

The mean depletion-layer capacitance $C_{\bar{v}}$ is the capacitance corresponding to the negative bias voltage across the varactor, and hence $C_{\bar{v}}$ is generally small compared with the zero-bias capacitance C_0. It is shown later that $C_{\bar{v}} \simeq 0{\cdot}2\ C_0$ for the conditions commonly encountered in practice, and so the diffusion-capacitance component of the total working capacitance may be estimated by subtracting 0·2 from the ordinate of the solid curve in Figure 2. It follows therefore that the diffusion capacitance can be obtained at a frequency f from the relation $C_d = mC_0(67 \times 10^6/f)^{1/2}$, where m is the ordinate of the dashed curve in Figure 2. Clearly at infinite frequency the diffusion-capacitance contribution to the working capacitance is zero, and the latter is then equal to the depletion-layer capacitance. The diffusion capacitance is, of course, controlled by the magnitude of the forward drive, which in turn depends on the value of the self-bias resistor. The value of the self-bias resistor is not very critical, and experience has shown that a value in the range from $3 \times 10^{-7}/C_0$ to $8 \times 10^{-7}/C_0$ is suitable. If we consider a varactor handling a power P watts at an input frequency f, then from (3) and (4) we have

$$V_f = \left\{ \frac{2\eta P \omega C_0}{K} \left(\frac{K^2}{\omega^2 C_0^2} + \frac{1}{\omega^2 C^2} \right) \right\}^{1/2}$$

where $C = C_{\bar{v}} + C_d$. Let

$$C_{\bar{v}} = A C_0$$

and

$$C_d = m C_0 \left(\frac{67 \times 10^6}{f} \right)^{1/2} = B C_0.$$

Then

$$V_f = \left\{ \frac{2\eta P K}{\omega C_0} \left(1 + \frac{1}{K^2 (A + B)^2} \right) \right\}^{1/2}.$$

Similarly

$$V_{2f} = \left\{ \frac{2\eta P K}{\omega C_0} \left(1 + \frac{1}{4K^2 (A + B)^2} \right) \right\}^{1/2}.$$

Hence

$$\frac{\hat{V}^2}{\eta} = \frac{6PK}{\omega C_0} \left\{ \left(1 + \frac{1}{K^2(A + B)^2} \right)^{1/2} + \left(1 + \frac{1}{4K^2(A + B)^2} \right)^{1/2} \right\}^2. \quad (7)$$

Equation (7) is plotted in Figure 6 for a range of the parameter P/C_0. It has been assumed in evaluating (7) that the diffusion capacitance is given by the dashed curve in Figure 2 and that the depletion-layer capacitance $C_{\bar{v}} = 0{\cdot}2\, C_0$. Now for the abrupt-junction diode, the capacitance C_v at a voltage $\hat{V}$ is

$$C_v / C_0 = \{\phi / (\phi - \hat{V})\}^{1/2}$$

where ϕ is the contact potential $\simeq 0{\cdot}6$ volt for silicon.

The mean depletion-layer capacitance $C_{\bar{v}}$ however [2] is approximately $2C_v$. Thus the constant A is given by

$$2\{\phi / (\phi - \hat{V})\}^{1/2}.$$

From Figures 4 and 6, it is seen that maximum efficiency is obtained when the largest practicable value of the parameter P/C_0 is chosen. If one assumes reasonable efficiency of operation, then $\hat{V} > 30$ is usually possible provided that diode dissipation is not a problem.

Hence $A \leq 0{\cdot}28$. A lower limit to A is set by the maximum available varactor break-down voltage, which is usually 150 volts. Thus A is restricted to the range $0{\cdot}125$ to $0{\cdot}28$ in practice. The result of a variation of the parameter A is indicated in the case of the $P/C_0 = 4 \times 10^{10}$ curve in Figure 6. The effect of a variation in A will, of course, be correspondingly smaller for larger values of the parameter P/C_0.

The choice of a suitable varactor to operate at a given input power level and frequency in a doubler circuit, with a specified conversion loss, may be made with the aid of Figures 5 and 6. An example of the use of these curves appears in Section 9.1.

4. Coupling Networks

The networks used to couple the varactor to the power source and to the load must have the following properties:—

(**A**) Provide the required band-pass characteristic.

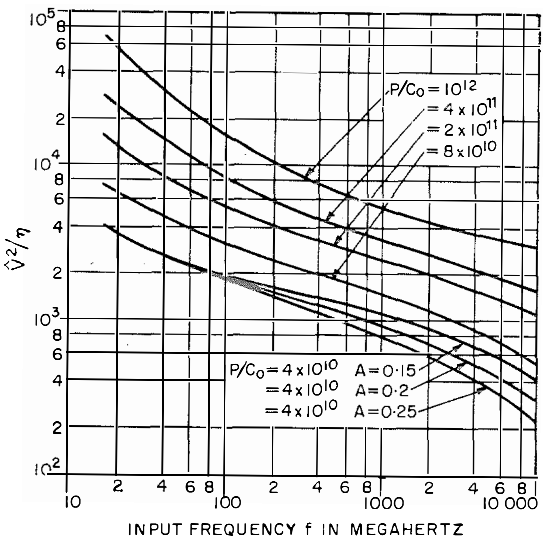


Figure 6—Variation of $\hat{V}^2/\eta$ with frequency.

(**B**) Minimize the flow of unwanted harmonic currents through the varactor.

(**C**) Provide the required degree of suppression of unwanted harmonic power at the load terminals.

(**D**) Provide impedance transformation to match source and load impedances to the varactor impedance.

(**E**) Not be excessively lossy in the pass band.

Requirement (**A**) generally requires a moderately low loaded Q factor whilst (**B**) and (**C**) both require high values of loaded Q. This is a common problem, and the usual solution is to employ a pair of circuits that give both an improved amplitude and phase characteristic within the pass band and twice the rate of cut off outside the pass band. To provide impedance transformation it would, in any case, be necessary to add an extra reactance to the single tuned circuit, making a total of 3 reactances compared with 5 for the coupled circuit.

The circuit loss of a series-tuned or double-tuned coupling circuit is indicated in Figure 7. As the unloaded Q value of a lumped, microstrip, small-diameter coaxial, or waveguide coupling circuit is usually greater than 300, it is not difficult to limit the loss per coupling circuit to about 0·25 decibel.

4.1 Bandwidth and Methods of Cascading

Multipliers operating in the very-high- and ultra-high-frequency range generally have quite high efficiencies per stage of doubling, and it appears that the overall bandwidth of a chain of cascaded stages is only slightly smaller than that of any one stage. That is, the chain behaves substantially as if it were composed of pure reactances. The bandwidth of a coupling network depends on the loaded Q of the tuned circuits, but the range of loaded Q values which may be used is limited to a minimum value of 1 to 3, depending on power level, by the Q of the varactor itself. A more-practical limit is set by the difficulties of making the resulting extremely small values of inductance required in the coupling networks.

Very low Q values also present problems of preventing unwanted harmonics from reaching the output. There is no upper limit to the loaded Q factors of the coupling networks, provided that the circuit loss can be tolerated (or very high unloaded Q values can be obtained) and the pass band is stable with respect to temperature fluctuations, ageing, et cetera. Many workers in the varactor frequency-multiplier field make individual multiplier stages with 50-ohm input and output impedances, and merely cascade a number of such stages to make a complete chain. The basic idea appears to be that testing a single stage is very much easier than testing a complete chain.

This method of design appears to be unnecessarily complicated in that extra circuit components are required to provide the necessary matching arrangements at the input and output of each unit, and consequently the chains described in this paper were designed as integral units with no provision for testing individual stages. This method of design offers the possibility of making very compact circuits as the number of components is reduced. Methods of testing complete multiplier chains are described in Section 10.

As the design procedures differ for multipliers employing lumped and distributed circuits, it

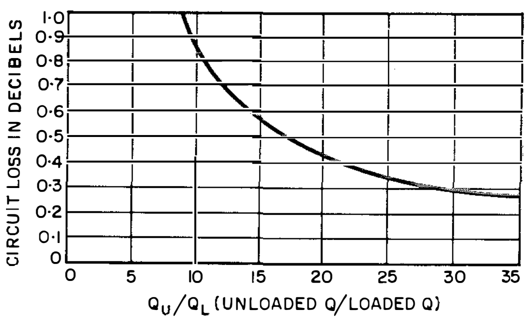


Figure 7—Circuit loss.

is proposed to deal with them separately. Thus lumped-circuit design is described in Sections 4.2 through 6, and distributed coupling circuits are discussed in Sections 7 through 9.

4.2 Choice of Coupling Network for Lumped Circuits

Having decided to employ double-tuned transformers as coupling networks, one then has a choice of coupling methods:—

(**A**) The type of coupling impedance: mutual inductance, self inductance, or capacitance.

(**B**) The choice of T or π networks.

The choice of network is, to some extent, a matter of convenience, but capacitive coupling has a major advantage since a wide range of low-loss capacitors is commercially available. Inductive coupling involves the adjustment of tapped coils, the relative positions of components, or measurements in situ of the coupling impedances. Capacitance coupling is therefore to be preferred.

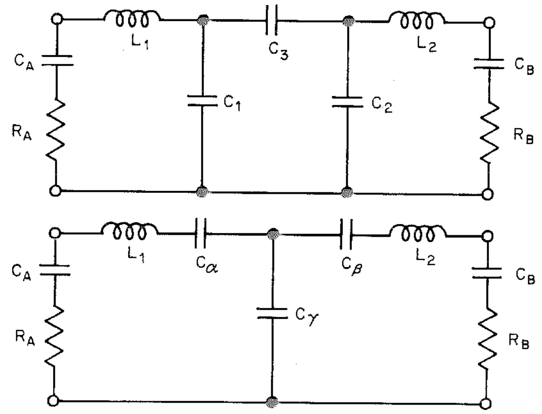
The choice between T and π configurations may be made on purely practical grounds; in the example to be described it was decided to use the π type of circuit where possible, as this resulted in the rotors of both tuning capacitors being at earth potential. In some cases, however, the coupling capacitance required in the π circuit becomes inconveniently small. The T circuit is then employed.

5. Network Design

The general coupling network used as an interstage between two varactors is shown at the top of Figure 8, and it is required to calculate the component values necessary to couple varactor A (represented by C_A and R_A) to B (represented by C_B and R_B).

The equivalent circuit at the bottom of Figure 8 is rather easier to solve, and hence its components are first calculated and then the T network of capacitors C_α, C_β, and C_γ replaced by the π arrangement C_1, C_2, and C_3 by means of the transformation from T to π. The steps in the design are as follows:—

Figure 8—At top is shown a π coupling network, while at bottom is shown the equivalent T network.



(**A**) Using selectivity curves for a double-tuned circuit [3] and the specified bandwidth of the chain, calculate the loaded Q of all coupling networks.* If the multiplier chain is

*It should be noted that the curve given in [3] is for the case of infinite Q, and that the frequency response of a coupled circuit becomes asymmetric for low Q values. This results in a slope in the frequency response corresponding to 0·5 decibel for a 10-per-cent bandwidth with primary and secondary Q values of 10, the sign of the slope depending on the type of coupling employed. The infinite Q curves give the mean response for either side of the centre frequency. If greater accuracy is required the ratio of the power transmitted P_ω at an angular frequency ω to that at the centre of the band ω_0 may be calculated from

$$\frac{P_\omega}{P_{\omega 0}} = \frac{4U}{(1+U)^2 + Q_p^2 x^2 (1-U)^2}$$

where

$$U = \frac{\omega_0^2}{\omega^2(1+Q_s^2x^2)}$$

for a shunt coupling capacitor

$$= \frac{\omega^2}{\omega_0^2(1+Q_s^2x^2)}$$

for a shunt coupling inductor, and

$$x = \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}.$$

required to operate over a very-small-percentage bandwidth, leading to high values of Q, it may be advantageous to choose an arbitrary Q value which is quite low to avoid circuit difficulties due to dynamic detuning, temperature effects, et cetera.

(**B**) Calculate the inductances L_1 and L_2 using the relationships $L_1 = QR_A/\omega_0$ and $L_2 = QR_B/\omega_0$ where ω_0 is the angular frequency at the centre of the pass band.

(**C**) Calculate the coupling capacitance C_r from the relation $C_r = 1/\omega_0(R_A R_B)^{1/2}$.

(**D**) It now remains to calculate the tuning capacitances C_α and C_β.

Both primary and secondary circuits are resonant when completed through the coupling capacitor C_γ. That is

$$\omega_0 L_1 = \frac{1}{\omega_0}\left(\frac{1}{C_A} + \frac{1}{C_\alpha} + \frac{1}{C_\gamma}\right)$$

and similarly

$$\omega_0 L_2 = \frac{1}{\omega_0}\left(\frac{1}{C_B} + \frac{1}{C_\beta} + \frac{1}{C_\gamma}\right).$$

Hence

$$C_\alpha = \frac{C_A}{\omega_0 C_A\{QR_A - (R_A R_B)^{1/2}\} - 1}$$

and

$$C_\beta = \frac{C_B}{\omega_0 C_B\{QR_B - (R_A R_B)^{1/2}\} - 1}.$$

Finally C_α, C_β, and C_γ are transformed to the π network values using the relationship

$$C_1 = \frac{C_\alpha C_\gamma}{C}$$

$$C_2 = \frac{C_\beta C_\gamma}{C}$$

$$C_3 = \frac{C_\alpha C_\beta}{C}$$

where $C = C_\alpha + C_\beta + C_\gamma$.

5.1 Self Capacitance of Inductors

If the inductance values calculated by the methods outlined above are so large that they are resonant with less than a 1-picofarad tuning capacitance, then the circuit must be re-calculated using a lower value of loaded Q, for the minimum capacitance of a trimming capacitor will be of this order.

If the inductors are measured at a frequency lower than that at which they are required to operate, it may be necessary to allow for the effects of self capacitance.

Thus, if an inductor having several turns of wire is required to tune with a capacitance of 2 or 3 picofarads, the effects of self capacitance will not be negligible, for the self capacitance C_L in picofarads is of the order of 0·75 times the radius of the coil in centimetres. The equivalent circuit of the inductor is then as shown in Figure 9 and the impedance is

$$Z = \frac{r + j\omega\{L(1 - \omega^2 L C_L) - C_L r^2\}}{(1 - \omega^2 L C_L)^2 + \omega^2 C_L^2 r^2}.$$

If the resonant capacitance is nC_L, then the impedance Z is given by

$$\frac{r + j\omega L\left\{\left(1 - \frac{1}{n}\right) - \frac{1}{nQ^2}\right\}}{\left(1 - \frac{1}{n}\right)^2 + \frac{1}{n^2 Q^2}}.$$

Neglecting the terms $1/nQ^2$ and $1/n^2Q^2$ gives

$$Z = \frac{rn^2}{(n-1)^2} + j\omega L\left(\frac{n}{n-1}\right).$$

Thus if $C_L = 0{\cdot}4$ picofarad (say 1-centimetre-diameter coil) and $n = 3{\cdot}5$ corresponding to a 1·4-picofarad tuning capacitance, then the

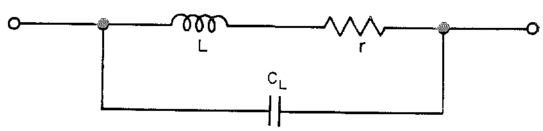


Figure 9—Self capacitance of inductor.

effective inductance is 40 per cent too large and the effective series resistance is doubled.

5.2 Small Values of Inductance

Values of inductance smaller than about 0·05 microhenry cannot be adjusted satisfactorily using an inductance bridge because the exact shape of the circuit then becomes important. This difficulty can be overcome by calibrating the trimming capacitor and adjusting the inductance value in situ, using a grid-dip meter as an indicator of resonance.

5.3 Screening and Layout

To ensure that the coupled circuits behave correctly, it is necessary to ensure that no stray couplings exist between the primary and secondary circuits. A convenient method of providing adequate screening is to build the individual halves of each coupled circuit on opposite sides of the chassis as shown in Figure 10. By careful attention to the layout and by choice of component types, it is possible to arrange that all the trimming capacitors are accessible from one side of the chassis.

6. Example of Design

An example of the design methods outlined above is given in Figure 11, which corresponds to the physical arrangement of Figure 10. The unit comprises 4 cascaded doubler stages operating at an input frequency of 67 megahertz and an output frequency of 1072 megahertz. The circuit is designed to operate from a 50-ohm source, using the transforming properties of the input network, and to deliver power via a quarter-wave transformer to a 50-ohm load. It is necessary to use the quarter-wave transformer in the output circuit because the output circuit matches 10 ohms with the values indicated in the diagram. If the secondary circuit was required to match to 50 ohms directly, then the tuning capacitance would be one-fifth of the indicated value, which is impracticably small if the designed Q factor of 10 is maintained. It is also seen that a T coupling network is used instead of a π network at the output frequency. This is because the coupling capacitor required for the π network would be about 0·17 picofarad, which is too small for practical use. All the coupling capacitors in the chain are fixed values—in some cases it was necessary to use 2 capacitors in parallel to approximate the theoretical value, and in some cases it was necessary to select values from the nearest commercial classification.

The coupling capacitor in the output circuit comprises the capacitance to ground from the small box which contains the tuning capacitors associated with the T network. The screening box was employed to reduce the stray capacitance to ground from the tuning capacitors indicated by the dashed lines in Figure 12. If the stray capacitance is comparable to the required value of tuning capacitance, the remaining components in the T network become very small. By placing a screen around the tuning capacitors, however, as in Figure 13, the stray capacitances fall in parallel with their respective tuning capacitors and consequently they have no deleterious effects.

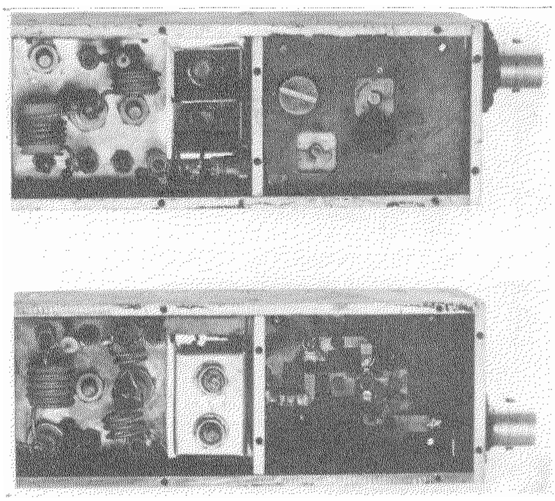


Figure 10—Opposite sides of ×64 multiplier chain.

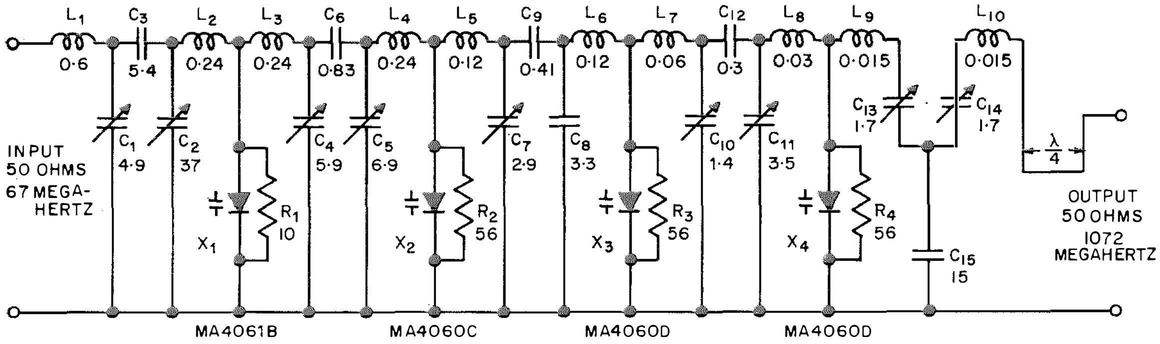


Figure 11—Circuit of ×16 multiplier. Capacitance values are in picofarads, inductance values in microhenries, and resistance values in kilohms.

6.1 Varactor Bias

It will be seen from Figure 11 that all the varactors have shunt bias resistors connected, the resistor being 10 kilohms for the *MA4061B* and 56 kilohms for the *MA4060C* and *MA4060D* types.

In addition to establishing the required charge storage conditions, the bias resistors stabilize the input-level–output-level characteristic [4] of the varactor and minimize the hysteresis effects* which otherwise occur.

The bias resistors shown have little effect on the varactor efficiencies, circuit loss, et cetera.

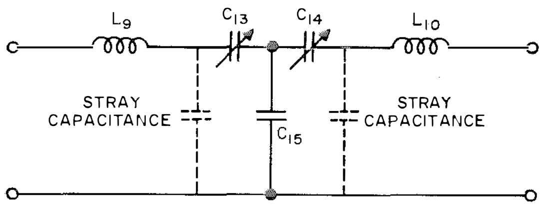


Figure 12—Stray capacitance in output circuit of ×16 multiplier.

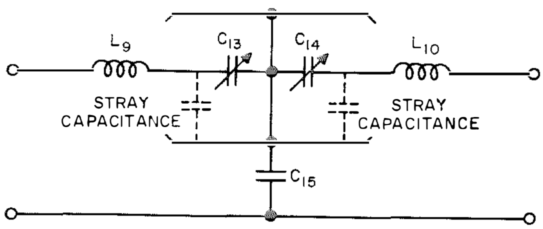


Figure 13—Effect of screening in output circuit of ×16 multiplier.

* Suitable choice of self-bias resistor results in a smaller variation of varactor impedance with change in power level than can be obtained with fixed-bias operation. Thus, the input capacitance of a self-biased varactor has a range from C_0 to mC_0 corresponding to a power-level range from zero to a level P watts, where m is given by curves of the type shown in Figure 2. The corresponding bias voltage will also vary from zero to some value E. A varactor operating with a fixed negative bias E_b, however, will have a working capacitance C which is considerably smaller than C_0 for very small input powers and will remain smaller than the self-bias value at all power levels for which $E < E_b$. For $E_b < E$, however, the input capacitance for fixed-bias operation will be greater than that for the case of self-bias working. The conversion resistance of the varactor would also be expected to show larger variations with change of power input for the case of operation with fixed bias, particularly for small values of input; for example, for small input powers the conversion resistance would be in accordance with the predictions of reverse-bias theory and thus would exceed the value corresponding to operation in the forward region by a factor of several times.

6.2 Calculated and Measured Loss

The following data are relevant:—

Input frequency in megahertz	67
Input power level in watts	10
Output frequency in megahertz	1072
Output power level in watts	3·2
1-decibel-down bandwidth	±4 per cent
Loaded Q values of coupling networks:	
Input circuits to first varactor	$Q_L = 5$
All other circuits	$Q_L = 10$

From the above data, the estimated loss of the varactors and coupling circuits is 4 decibels, whilst the measured loss is 5 decibels.

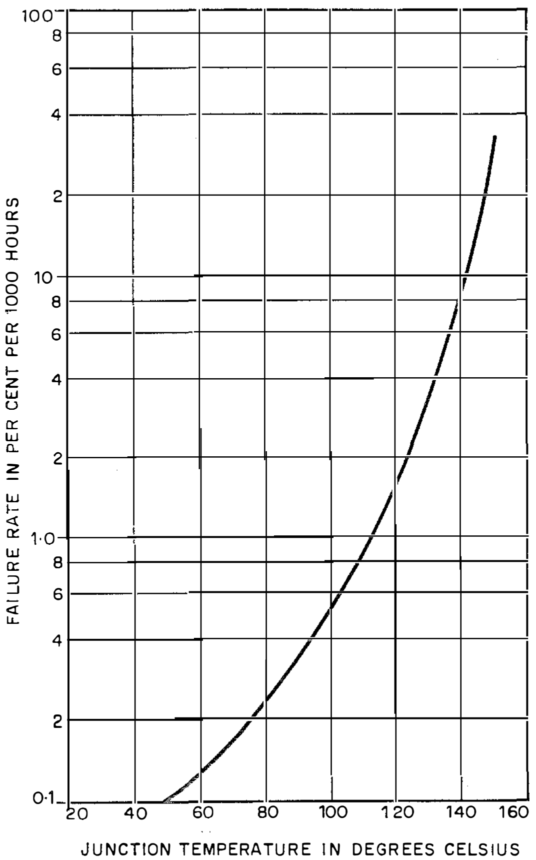


Figure 14—Life data for semiconductor junctions.

This discrepancy may be due in part to errors in estimating the unloaded Q of the high-frequency circuits.

7. Choice of Varactors for Distributed Circuits

In the foregoing sections, power handling capacity and cut-off frequency have been the major considerations in the selection of varactors for use at very-high frequency. At these frequencies the inductance L and shunt capacitance C_c of the diode equivalent circuit shown in Figure 3 can be ignored, but at microwave frequencies the effect of these parameters is considerable and has a major effect on circuit design. This is discussed in greater detail in Section 7.1. Problems also arise in the microwave region due to difficulties of maintaining a high ratio of cut-off frequency to input frequency (f_c/f) leading to increased conversion loss. This is particularly so if it is required to handle high input powers. As the thermal resistance of a microwave diode is higher than one suitable for use at very-high frequency, the increased conversion loss produces additional difficulties due to the increase in temperature of the junction. It is desirable to minimize junction temperature as this leads to improved reliability. A plot of failure rate against junction temperature is shown in Figure 14, which was produced from data supplied by a semiconductor manufacturer. These data are now over 2 years old and may be pessimistic.*

To select a diode for a given application, it is of primary importance that it should be

* Since the paper was written, additional life data have become available from a number of sources. There is broad agreement between the recent sources of data and it now appears that the failure rate of a silicon junction is of the order of 0·005 per cent per 1000 hours at 80 degrees Celsius, and the failure rate doubles each 12-degree Celsius increase in junction temperature above 80 degrees Celsius.

capable of handling the input power without going into reverse break-down. The diode may be selected by the following method. If the required varactor efficiency is η then the varactor dissipation P_d is given by

$$P_d = P(1 - \eta). \tag{8}$$

If the permissible rise in junction temperature consistent with minimum life is T, it follows that the maximum thermal resistance θ is

$$\theta = T/P_d. \tag{9}$$

The value of θ required to satisfy (9) may be obtained with a number of varactor types having different values of C_0 and V_b.

The value of $\hat{V}$ is calculated with the aid of Figure 6 for the varactor having the smallest value of C_0, and if this is less than the rated break-down voltage, it will be satisfactory provided that the series resistance is low enough to give the required efficiency. In some cases it may be preferable to employ pairs of varactors instead of a single varactor of higher capacitance, as higher values of cut-off frequency are usually associated with low-capacitance diodes.

Should there be diodes of more than one type of encapsulation that satisfy all the foregoing conditions, then it may be desirable to tolerate the higher thermal resistance of, say, the "pill-with-prongs" type of encapsulation in exchange for the lower frequencies of parallel and series resonance of, say, the cartridge encapsulation as this permits greater bandwidths and larger tolerance of diode parameters.

7.1 Diode Impedance

To design a suitable circuit to have the characteristics outlined in Section 4, it is essential to know the diode impedance at the working power level and frequency.

Figure 4 shows a plot of conversion resistance multiplied by zero-bias capacitance versus frequency. This curve was obtained from (3), the constant K being determined from measurements made at 67 megahertz. However, from work carried out at microwave frequencies using silicon diodes in a pill-with-prongs encapsulation, it was evident that this curve also applied at 2 gigahertz.

It should be appreciated that the shunt capacitance C_c and series inductance L of the encapsulation seriously modify the value of the conversion resistance R as seen at the terminals of the diode. Reference to Figure 15, which is a plot of the impedance at the diode terminals, illustrates this point. Component values are shown on the diagram and are those quoted in a manufacturer's catalogue for a cartridge diode. Further complication is introduced by the parasitics of the varactor mount.

To find the working impedance of a diode in its mount, a simple microstrip circuit containing the varactor is adjusted to give maximum efficiency at the required power level and frequency. If all the controls are within their range of adjustment and the efficiency is approximately that expected, it may be assumed that the condition for maximum efficiency is the condition of correct tuning and matching. The diode is then removed and the conjugate impedance of the diode measured.

8. Circuit Design

The circuits that couple adjacent stages at microwave frequencies should have all the properties of lumped circuits outlined in Section 4, but with the additional requirement of providing adequate heat sinks.

Generating high powers at microwave frequencies will, in general, mean that large amounts of power are dissipated in the diodes and hence good heat-sinking of the diode is essential to keep the junction temperature low. Voltage-driven diodes as shown in Figure 16A usually have both ends of the diode isolated from earth and so in general cannot be connected to an infinite heat sink. In the special case of waveguide circuits it is possible for a voltage-driven diode to be connected to one wall of the waveguide, which in practice can be considered an infinite heat sink. When using microstrip or coaxial circuits as a diode coupling circuit it is essential to use the current-driven mode shown in Figure 16B to get effective heat-sinking. This is achieved by connecting one end of the diode to a massive earth conductor.

8.1 Power Wastage

Power wastage can come about for several reasons and a well-designed circuit should keep this power loss to a minimum. The unloaded Q of the circuit should be made a maximum by using lengths of high-quality transmission line as inductors and tuning these with lumped capacitors having low-loss dielectric. It should not be difficult to achieve Q's greater than 500 by this means.

In a circuit containing distributed elements, the main band-pass characteristic does not necessarily indicate the performance at frequencies harmonically related to the main response. This is due to the impedance transformation effect of the lengths of transmission line. A simple example is illustrated by the well-known quarter-wave transformer. A length of line of characteristic impedance $Z_0 = 1$ and $\lambda/4$ at f is terminated at one end by a resistance $R_1 = 0{\cdot}05$. This is transformed to a resistance $R_2 = 20$ at the remote end. At $2f$, however, the length of line becomes $\lambda/2$ and $R_2 = 0{\cdot}05$.

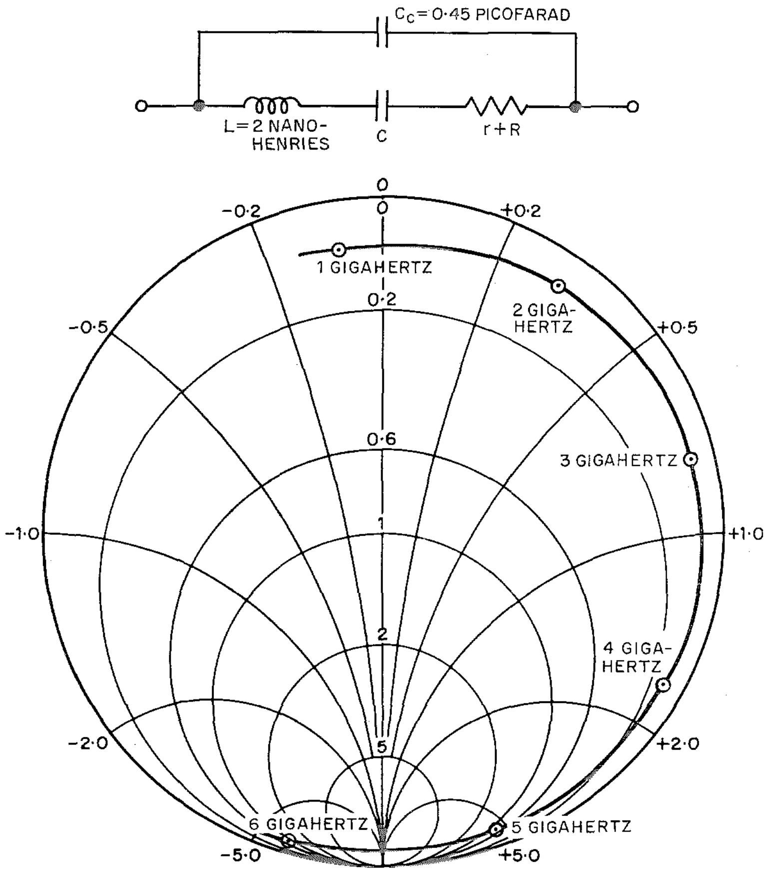


Figure 15—Impedance plot of cartridge diode normalized to 50 ohms. At top is equivalent circuit of a varactor where $P = 5$ watts, $C_0 = 10$ picofarads, $C = C_0\{0{\cdot}2 + m(67 \times 10^6/f)^{1/2}\}$, $r = 1$ ohm, and R is given by Figure 4.

Because of this type of impedance transformation, the circuit performance has to be considered separately at all the frequencies which are likely to exist. The condition for low power wastage due to unwanted circulating currents in the circuit and the output load is satisfied by presenting a high impedance to all the unwanted frequencies at the varactor terminals.

When the loss in a transmission line is small, the loss per half wavelength is given by [5]

$$\frac{\text{Power loss on mismatched line}}{\text{Power loss on same line matched}} \simeq \tfrac{1}{2}(S+1/S)$$

where S is the voltage standing-wave ratio. The necessity of keeping the standing-wave ratio close to unity is thus obvious, and hence the transformation along the line should be kept to a minimum.

8.2 Coupling Circuit Design

The calculations involved in the mathematical design of pairs of coupling circuits between diodes would prove to be very laborious when all the circuit parameters were taken into account, that is, tuning, impedance matching, suppression of unwanted harmonics, desired frequency response, and minimum loss. An empirical method of design which has consistently produced good results will be described. The method involves plotting the circuit components on a Smith chart.

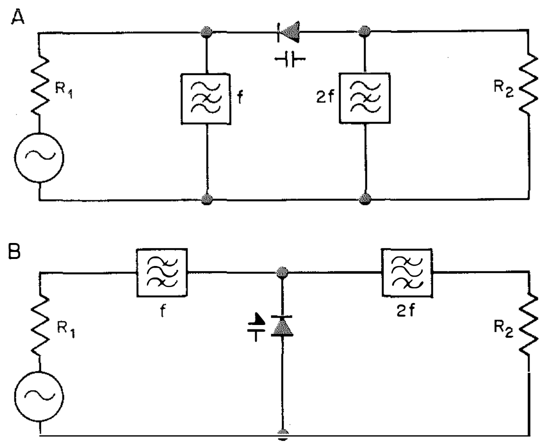


Figure 16—At top is a voltage-driven diode and below is a current-driven diode.

In general the circuit of Figure 17 will have inferior harmonic rejection to the lumped-circuit equivalent unless precautions are taken. As previously indicated, it is possible for the length of tuning line to present a lower impedance at the harmonic than at the fundamental frequency. Also the value of the shunt reactance required for critical coupling is numerically equal to the real impedance at the coupling point, and if this impedance happens to be higher than the input or output impedance, then the rejection of the higher frequencies will in general be reduced.

If C_1 and C_2 in the circuit shown in Figure 18 are equal, no impedance transformation is effected, but for C_1 and C_2 unequal the higher impedance occurs at the end with the smaller capacitance. With one capacitance infinite, the impedance transformation is a maximum and depends primarily on the length of line; it is almost independent of the characteristic impedance Z_0 unless this impedance is less than the loading resistance at the higher-impedance end.

The Q depends primarily on the line length and increases with increase in length. When using a single tuning capacitance, its reactance is approximately proportional to Z_0 and hence may vary considerably as Z_0 is changed without significantly changing the Q. By inserting a second capacitor at the opposite end of the line (thereby reducing the capacitance of the first) and by increasing the line length, the

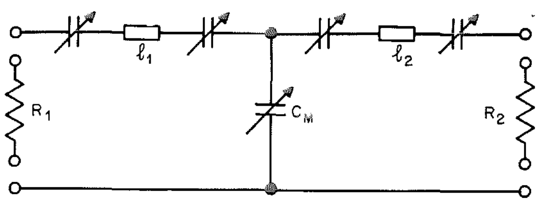


Figure 17—Distributed coupled circuits.

Q can be increased without changing the transformation ratio. Some of these properties are illustrated in the Smith chart of Figure 18.

It should also be realized that whilst a single tuned circuit consisting of a length of line divided by a series capacitor behaves similarly to the circuit of Figure 18 at the fundamental resonance, the impedance presented at the harmonic frequencies may differ considerably. Hence it is possible to choose the circuit arrangement which gives the best suppression of the unwanted harmonic currents whilst maintaining the desired band-pass characteristic.

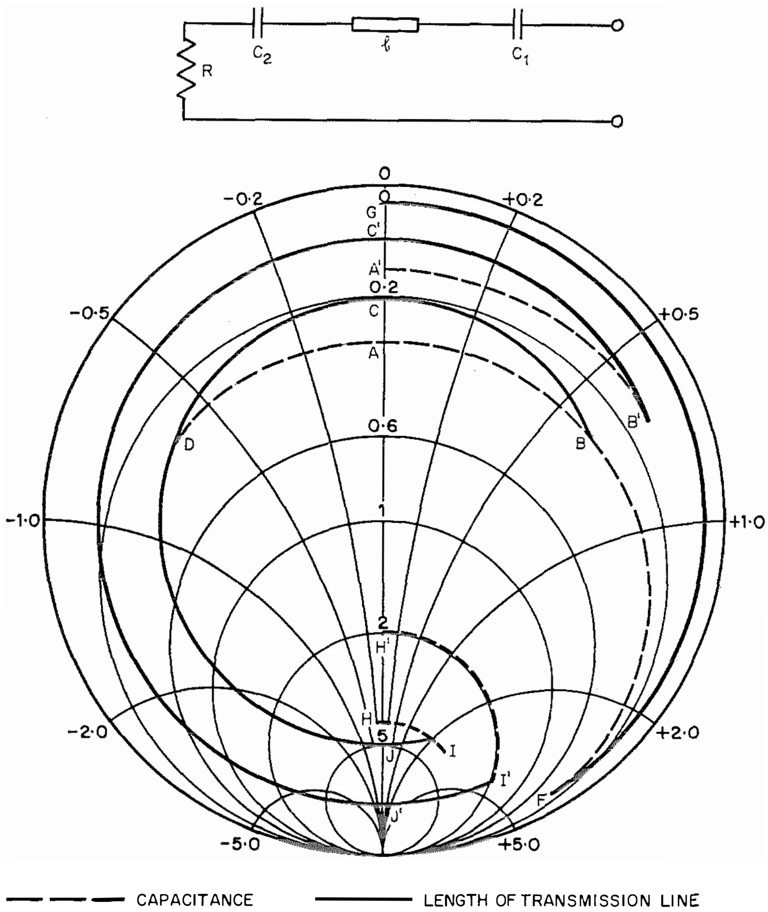


Figure 18—Some transformation effects of the distributed circuit shown.

(a) ABCDA—Circuit with zero impedance transformation.

(b) ABFG—Circuit with maximum impedance transformation (C_2 infinite).

(c) ABC—Case (b) with C_2 infinite and l reduced, illustrating reduced impedance transformation.

(d) A′B′C′—Case (c) with Z_0 doubled. Reactance of C_1 is approximately doubled but Q and transformation ratio are approximately as in (c).

(e) HIJC—Largest terminating resistance larger than Z_0.

(f) H′I′J′C′—Case (e) with Z_0 doubled. Impedance transformation arranged to be as in (e) by adjustment of l. Reactance of C_1 increased by approximately a factor of 4.

(g) ABKLMC—Same impedance transformation as (c) but higher Q.

Since a low coupling reactance contributes significantly to the high-frequency rejection characteristics, it is normally preferable to use the impedance transformation along the length of line to effect the transformation rather than use a higher value of shunt reactance. In Figure 17, for example, if $R_1 > R_2$, the components would be chosen so that the real component of the impedance was R_2 at the common point on both l_1, l_2, and the shunt capacitor C_M, and hence the magnitude of the coupling reactance would be R_2. Consequently, when transforming from the diode conversion resistance to, say, a 50-ohm load impedance, it is preferable to effect the impedance transformation with the line length of the final circuit of the coupled pair.

Rejection of the input frequency by the circuit resonant at second harmonic mainly depends on the reactance of the series tuning capacitors.

Minimum loss is achieved by using high-quality lines and capacitors and having minimum reflection on the line, consistent with maintaining the required harmonic rejection. Minimum reflection is achieved by having minimum impedance transformation along the line.

Using the Smith chart it is possible to select the appropriate component values to achieve tuning and matching. It is then necessary to check the circuit performance relating to bandwidth and harmonic rejection. Adjustment of the component values may then be necessary to achieve the required results. An estimate of bandwidth can be made by determining component reactances at small frequency differences from the centre of the band and plotting on the Smith chart (renormalized to the diode conversion resistance) the locus of the impedance seen by the diode. The reflection loss at any frequency deviation can then be calculated from the position on the Smith chart of the impedance locus at that frequency from

$$\text{Reflection loss} = 10 \log \frac{1}{1-\rho^2} \text{ decibels}$$

where $\rho = (1-S)/(1+S)$ = voltage reflection coefficient, and S = voltage standing-wave ratio.

Calling the full radius of the Smith chart unity, ρ is the radius of the reflection-loss circle.

To simplify the design it is assumed that the output impedance of the preceding stage is constant within the pass band of the coupling network.

9. Design Example

9.1 Selection of Diode

The method of diode selection and circuit design will now be illustrated by considering a particular example.

It was desired to produce 1·0 watt at 4·288 gigahertz from an input of 1·8 watts at 2·144 gigahertz, and the output was required to vary by less than 1 decibel over a frequency band of 2·5 per cent.

The maximum dissipated power is 0·8 watt and so, if the circuit loss is assumed to be 0·25 decibel, $P_d = 0{\cdot}74$ watt and $\eta = 0{\cdot}59$. Assuming a maximum permissible failure rate better than 1 per cent per 1000 hours, Figure 14 indicates a maximum permissible junction temperature of 110 degrees Celsius. Using (9) and the fact that a maximum heat-sink temperature of 70 degrees Celsius is specified, the maximum value of thermal resistance is found to be 54 degrees Celsius per watt.

From the discussion in Section 7.1 it is apparent that a pill-with-prongs encapsulation is preferable at microwave frequencies, and a thermal resistance of the order of 50 degrees Celsius per watt is available in this encapsulation for values of C_0 around 5 picofarads. Thus $P/C_0 \simeq 3{\cdot}6 \times 10^{11}$, which from Figure 6 corresponds to a value of $\hat{V}^2/\eta = 2500$, that is, $\hat{V} = 38{\cdot}4$ volts.

An *MA4055C* having the following characteristics meets the above requirements.

$C_0 =$ 5·75 picofarads
$C_{vb} =$ 1·42 picofarads
$V_b =$ 53 volts (minimum value = 48 volts)
$\theta =$ 50 degrees Celsius per watt.

The power delivered by the diode for 1 watt at the output terminal must be 1·06 watts, that is, a conversion loss of 2·3 decibels. From Figure 5 it follows that

$$\frac{C_b f_c}{C_0 f} \geq 13{\cdot}8$$

and this requires $f_c \geq 120$ gigahertz.

This value of f_c is better than the standard classification and hence diodes must be specially selected. The actual diode used had the characteristics listed above together with a cut-off frequency of 190 gigahertz. This should result in a conversion loss of 1·55 decibels, which together with a circuit loss of 0·25 decibel leads to an overall loss of 1·8 decibels.

9.2 Design of Circuit

The steps of the circuit design can now be outlined, and an illustrative example is given below.

(**A**) Select the diode as outlined in Section 9.1.

(**B**) Ascertain the diode conversion resistance from Figure 4, and the working reactance from

$$C = m\left\{\left(\frac{67 \times 10^6}{f}\right)^{1/2} + 0{\cdot}2\right\} C_0$$

where m is given by Figure 2. If working close to the parallel resonance of the diode, its impedance is best found by measurement at the appropriate frequency and power level as described in Section 7.1.

(**C**) Using the Smith chart construction, design the primary side of the coupled circuit. Choose the load impedance to be purely resistive and approximately the same magnitude as the output impedance of the diode. Check that the circuit has the desired Q_L and gives good suppression of unwanted harmonics. If the desired results are not realized, adjust the component values using the principles outlined in Section 8.

(**D**) Design the secondary of the coupled circuit using the Smith chart construction again (the input impedance being equal to the load resistance of the primary circuit, and the output impedance being either equal to the complex conjugate of the input impedance of the next diode or equal to the terminating load). If the output load is in the form of a further stage of harmonic generation, ensure that a high impedance is presented to the varactor at the frequencies generated by the following stages. The value of loaded Q should also be checked as before and the necessary adjustments made.

(**E**) Assuming a coupling circuit of the type shown at the top of Figure 19, an impedance plot of the primary and secondary circuits similar to that shown at the bottom of Figure 19 results, where the loci denoted by the symbols A through J represent the impedances at the corresponding points in the coupling circuit above.

The chart shown in Figure 19 is normalized to the characteristic impedance of the line elements, which in this case is 100 ohms. The suffixes indicate the percentage frequency deviations from the band centre. It now remains to choose the reactance X of the shunt coupling capacitor, which is numerically equal to the load resistance R of the primary circuit. The load resistance is given by the point where the locus of the primary impedance plot, produced in step (**C**), intersects the real axis of the chart.

The inclusion of the shunt coupling capacitor necessitates either an increase in the lengths of line in both the primary and secondary circuits or an increase in the series tuning capacitors equivalent to a change in series reactance of jR ohms.

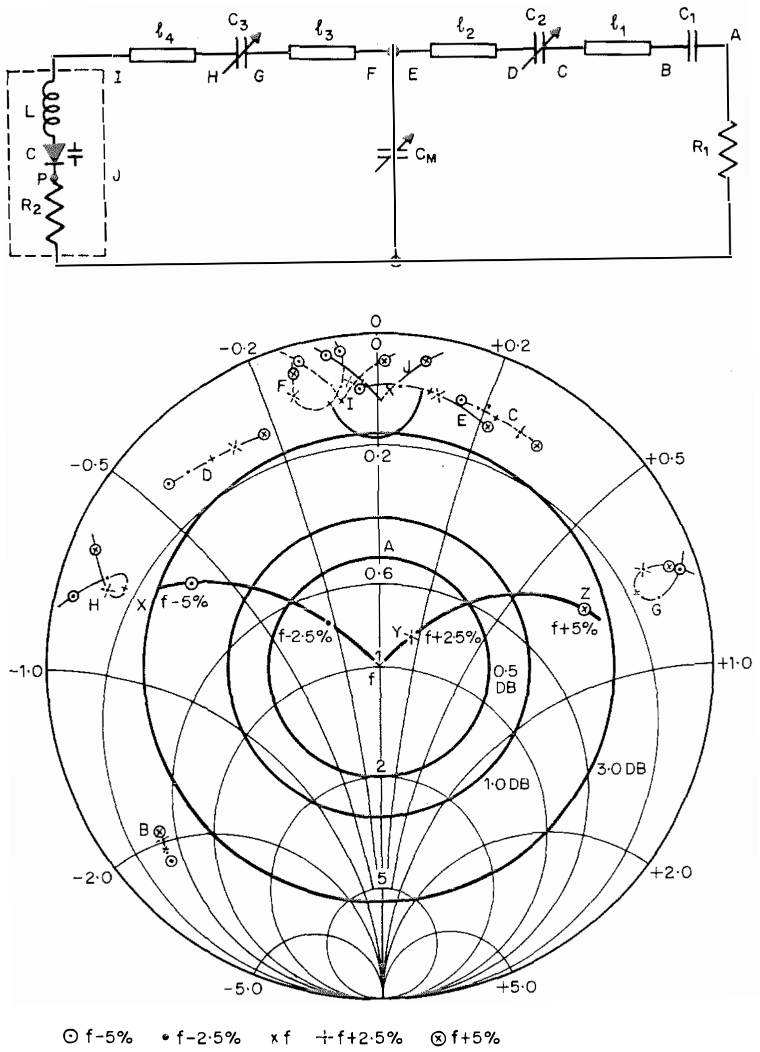


Figure 19—The normalized values of the components of the 4-gigahertz output coupled circuit shown above are given on the following table.

The Smith chart gives the impedance values of the circuit. The designations A through J refer to the impedance of R_1 when it is transformed to the corresponding points in the chart.

	Deviation in Per Cent from Frequency f				
	−5	−2·5	•	+2·5	+5
R_1 (ohms)	0·5	0·5	0·5	0·5	0·5
C_1 (ohms)	2·10	2·05	2·00	1·95	1·90
l_1 (λ)	0·206	0·211	0·216	0·222	0·227
C_2 (ohms)	0·588	0·575	0·560	0·546	0·533
l_2 (λ)	0·063	0·064	0·066	0·068	0·069
C_M (mhos)	8·50	8·75	9·00	9·25	9·50
l_3 (λ)	0·109	0·113	1·116	0·119	0·122
C_3 (ohms)	1·51	1·48	1·44	1·41	1·37
l_4 (λ)	0·086	0·088	0·090	0·092	0·095
L (•hms)	0·190	0·195	0·200	0·205	0·210
C (ohms)	0·147	0·144	0·140	0·137	0·133
R_2 (•hms)			0·11		

These modifications are in general small, and so the effect on the performance of the circuit can normally be ignored. In the example, l_2 is the minimum practical length in which the necessary components can be accommodated so it is necessary to adjust l_1.

(**F**) The final step is to ensure that the desired frequency response of the stage is obtained. The locus J in Figure 19 represents the impedance of the load resistance R_1 when transformed through the entire coupling network to the terminal P in the coupling circuit at top. Renormalizing the chart to the diode conversion resistance transforms the locus J such that at the centre frequency the impedance remains real. The locus XYZ in Figure 19 results, and hence the frequency response of the stage can be estimated by comparing the Smith chart reflection-loss circles with the frequency steps marked on the locus.

9.3 Results

This section gives typical results where the measured efficiency and bandwidth are compared with the predicted values together with the measured values of spurious outputs when driven by a varactor frequency-multiplier chain comprising 3 lumped-circuit stages and a distributed circuit of similar design to the one described. These stages can be identified easily in Figure 10.

Combining the predicted frequency response of the 4-gigahertz circuit under discussion with the known variation of output power with frequency of the preceding stages, a 1-decibel bandwidth of 6 per cent was expected from the complete chain. The measured output from the chain indicated that the loss of the last stage was 2·2 decibels and the overall frequency response was 1 decibel down for a bandwidth of 4 per cent. The measured loss is in reasonable agreement with the 1·8 decibels predicted, whilst the discrepancy between the measured value of 4 per cent and the predicted value of 6-per-cent bandwidth for 1-decibel loss was probably due to dynamic detuning and the variation of both source impedance and conversion resistance with frequency (both of these parameters being assumed constant in estimating the bandwidth).

The input-output power relationship was approximately linear and the frequency response remained continuous with variations in power level down to zero input, showing that the varactor had a good resistive match over a wide range of power levels.

Measured power levels of harmonics of 1 gigahertz, relative to peak power output, are given below.

Harmonic Frequency	Power Level in Decibels
Fundamental (1 gigahertz)	−49
2nd	−18
3rd	−19
5th	−30
6th	−37
7th	−51
8th	−47
9th	−80
10th	−28

In the particular example described, an improvement in the suppression of the 2nd and 3rd harmonics of 1 gigahertz would be desirable. Some adjustment of C_1 and C_2 of Figure 19 would probably improve the 2-gigahertz and 3-gigahertz suppression without appreciably degrading the remainder of the harmonic performance.

It is of interest to note that where close agreement between predicted and measured efficiencies have differed appreciably it has always been found that the diodes had either poor charge storage capability, an actual cut-off frequency lower than that quoted by the manufacturer, or both.

10. Methods of Testing Multiplier Chains

To adequately test a frequency-multiplier chain the following equipment is essential:—

(**A**) A power source, having the required output level, which may be swept automatically over a range of output frequencies somewhat in excess of the required operating range.

(**B**) A directional power meter, which may be used for checking the input match of the multiplier chain.

(**C**) A power meter, preferably of the calorimetric type, for measuring the output power.

(**D**) A band-pass filter centred on the output frequency, or a suitable wavemeter.

(**E**) A range of suitable attenuators.

10.1 Adjustment of the Chain

A test circuit comprising power source, directional power meter, multiplier chain under test, and wavemeter should be set up. The power source should be set to the normal input level for the chain, and to the nominal centre frequency. The input stages of the chain should then be tuned for the best value of input match, as indicated by the directional power meter. When a reasonable input match has been achieved, it will be found that the third or fourth tuned circuit has some effect on the reflected power, and consequently approximate tuning conditions can be judged for these circuits. Examination of the output with a wavemeter, or a power meter connected through a suitable filter, will almost certainly indicate some output at the required frequency. Adjustment of the output circuits to increase this output, adding attenuation when necessary to the wavemeter circuit, will in general affect the input match. Slight adjustment should then be made to the whole circuit, watching both the input match and the output power. The output power level should be measured and, if it is of the right order, the input frequency should be swept and the amplitude response observed on an oscilloscope. Small adjustments should be made to eliminate steps and bursts of oscillation which may occur in the amplitude characteristic within the pass band.

If the chain comprises a large number of multiplier stages in cascade, it may be advantageous to divide the chains into two sections for the purpose of testing. The low-frequency section should be adjusted first using a quarter-wave transformer to match the power at the point of separation to a 50-ohm load. Subsequently the second section should be directly coupled to the first and then adjusted. The condition of correct tuning should result from this procedure.

11. Conclusions

Information has been presented enabling the design of high-power frequency-multiplier chains employing doublers. The design data are applicable over a wide range of frequencies from high frequency to super-high frequency, and examples of design over an appreciable part of this spectrum have been given. Comparison between predicted and measured efficiency of multiplier chains is good, whilst the discrepancy between the predicted and measured bandwidths is probably due mainly to the effects of dynamic detuning. It is emphasized that the data given relate to operation in the charge-storage mode, without which the construction of economical high-power multiplier chains is impossible.

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Röhren (Electron Tubes)

The 13th edition of this manual by Friedrich Fritz of Standard Elektrik Lorenz gives technical data and base diagrams of standard receiving and picture tubes, of special tubes, and of small transmitting tubes available at the end of 1965.

Over 100 new types were added in this edition. The most important United States types, including compactrons, are given as well as a list of existing equivalent German types.

The manual, 20.5 by 14.5 centimeters (8.1 by 5.7 inches), is published by Codex-Verlag, Böblingen, and is available at DM 4 from Standard Elektrik Lorenz, Components Division, 85 Nürnberg, Germany.

Trends in Radio and Television Receiver Components in Germany

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The trend in the development of components for broadcast and television receivers is controlled by the manifold specifications related to receiver appearance, to electrical functions and their controls, and to quality and price.

This trend is particularly sensitive to the introduction of new services such as radio-frequency stereophonic reception, ultra-high-frequency and color television, and to modern production methods such as the introduction of printed-wiring boards and laminated plastic cards that replace the conventional metal chassis and permit automation of component mounting. Another stimulus is the appearance of novel kinds of transistors, tunnel and zener diodes, and voltage-adjustable capacitors. This constant flow of suggestions to the equipment designer results in new solutions to existing problems and these in turn flow back to the component designer as suggestions for further modifications and improvements.

In this constant exchange of information between equipment and component designers, the former must often assume the lead. In the end, each component within an equipment affects its operation, and the equipment designer necessarily assumes the responsibility for component quality. People will readily say "This equipment of the *X* Company was a failure" and rarely "The component of the *Y* Company caused the failure."

The purpose of the common work devoted to a new component is therefore to make sure that it will operate under specified conditions (temperature range, humidity, overload, vibration, et cetera), which may be simulated by severe life tests.

Since the weakest link controls the strength of a chain, one of the objectives in this work is to equalize the durability and life expectancy of all the components used in an equipment or possibly in a product line. If the life is known to be limited, provisions are made for convenient replacement by means of plug-in connections.

In choosing a component, the equipment designer is inclined to favor one that can be obtained from several sources; he is bound to seek independence from a single supplier in the first place because both he and his competitors might one day demand quantities far exceeding the supplier's capacity. This is particularly true of components used in many equipment types. The desirable interchangeability of components requires that the component producers provide accurate, complete, and comparable technical information. In actual practice, however, this requirement often conflicts with the component producer's attempt to coordinate with as many equipment producers as possible. These statements essentially apply to manufacturers of passive components; in the case of active components the production of which is difficult and involves considerable investment in tooling, the few producers will tend to standardize and coordinate their products.

Thus such German producers of active components as Siemens, Standard Elektrik Lorenz, Telefunken, and Valvo follow a pattern of cooperation that is widely different from the methods employed by the countless manufacturers of passive components. A characteristic feature is the application-engineering service provided by such companies for their customers. This involves not only working out examples of applications and models, but also advising, assisting, and testing in the early stage of an equipment development. Of course all this work is undertaken in strict confidence.

1. Trends in Development of Active Components

For years interest has been focused on the competition between the vacuum tube and the semiconductor. It is wise to prepare for surprising turns and inventions that will result from the attempts to decide this case.

About 15 years ago, the first target of semiconductor engineering was the power rectifier tube, known as a sensitive component that caused frequent failures. The substitution of the selenium rectifier for this tube, however, led to many early difficulties that had to be overcome before the claim of "infinite life" became actual fact. Today the flat-type selenium rectifier, absolutely reliable in operation, prevails in power supplies of broadcast receivers; in television receivers it has been overtaken by highly efficient silicon rectifiers that feature reliability and smaller size as is evident in Figure 1.

The rectifier tube is still found in the high-voltage section of a television receiver where it is indispensable for handling up to 20 kilovolts. In portable television receivers in which these voltages are much lower, however, semiconductor rectifier stacks are beginning to replace the tube.

The next success of semiconductor devices occurred when matched pairs of germanium diodes began to be used in demodulators of frequency-modulation receivers. It was the Schaub Company in close cooperation with Süddeutsch Apparate Fabrik who assumed the risk of introducing this new component on a large scale. Today germanium diodes are primarily employed in ratio detectors of frequency-modulation equipment and demodulators of television sets.

The use of semiconductor diodes in receiver circuits paved the way for the transistor. There

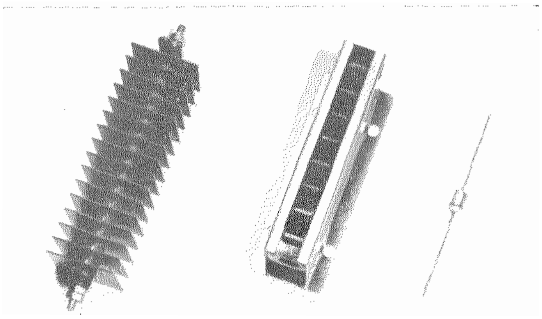

Figure 1—Power rectifiers for television sets. From left to right are shown early selenium, later selenium, and silicon types.

is no decline in the unmatched success of the transistor from its first application in 1956, just 8 years after its invention had been disclosed to the public. The tentative application experiments with transistors in audio circuits of the first battery-operated broadcast receivers were rapidly followed by all-transistor amplitude-modulated receivers, and the first all-transistor portable receivers that featured a very-high-frequency section appeared on the market as early as 1959.

The Graetz Company, now a division of Standard Elektrik Lorenz but then still independent, presented the first experimental all-transistor very-high-frequency receiver at the Radio Exhibition in Duesseldorf in 1957, while the Schaub Company was the first to show, at the Hannover Fair in 1959, a 4-band (including very-high-frequency) all-transistor universal portable, home, and automobile receiver, and to manufacture this set immediately thereafter. This receiver, named *TOURING*, has since been improved and is being produced in quantities unpredicted by its forerunner, the first tube-type portable adapted to automobile operation in 1954. Transistors, wiring boards, and a simpler design make modern receivers much-less likely to need repairs.

When transistors had been proved to be superior to tubes for all receiver applications, all portables almost simultaneously were equipped with transistors in 1960. The purely mechanical advantages alone were stimulating: simpler power supply, no *B* battery, no heaters, substantially reduced cost of operation, and reduced space requirements. On top of that, there were electrical advantages. With transistors it became possible to provide a radio-frequency input stage for the very-high-frequency tuner shown in Figure 2. This resulted in a marked increase of sensitivity while the spurious oscillator radiation was simultaneously reduced. Reduction of spurious radiation had been a particular problem in portable sets and was aggravated by the changing of their locations with respect to other receivers with which they could interfere. Tubes in battery-operated receivers had noise figures of 20 to 30 kT_0; with mesa transistors in the very-high-frequency input stage, noise figures have been reduced to 3 to 4 kT_0 and are now comparable to the optimum values that can be achieved with tubes operated from power mains.

From the very beginning the power output of portable sets was comparable to the 200 milliwatts offered with conventional tubes in class-A operation. The equivalent transistor receiver used a push-pull class-B output stage. Its power output has been increased in several development steps so that 6 watts are available from first-class receivers in motorcar operation, while in portable operation their output is limited to 2 or 2.5 watts to reduce the battery current drain. High-power output circuits require careful heat-dissipation control by mounting on ample cooling surfaces, compensating circuits employing thermistors, and negative temperature coefficients in the base-circuit resistances, in addition to careful adjustment of the idle current and satisfactory dimensioning of the driver stage. The transistor pairs are sorted for identical characteristics and current gain, and stringent requirements apply to current gain with high peak currents. The development of germanium transistors suitable for portable receivers has left nothing of importance to be desired; the function, price, and quality of these transistors are fully satisfactory. Development work is therefore going on mainly in power-stage circuit design where transformers can be eliminated if pairs of *p-n-p* and *n-p-n* transistors are used. However, the power output of this type of stage drops below the rated value when the operating voltage is low.

Battery-operated receivers must function satisfactorily down to half the rated battery voltage. This has required the development of stabilizing elements. Originally, nickel-cadmium cells with short charging periods and a stabilized voltage of 1.35 volts proved suitable; now, however, these are being replaced increasingly by circuits with silicon diodes using the steep-slope knee at 0.7 volt of the diode characteristic.

In the beginning, zener diodes were preferred to stabilize the voltage in motorcar operation; they prevented the undesirable variation in the frequency of the very-high-frequency oscillator as a result of battery voltage fluctuations caused by the necessary changes in the speed of the car engine. Now the use of zener diodes is virtually limited to voltage dividers in which the 12 volts of the car battery are reduced to the receiver rating—in most cases 7.5 volts—to equal the voltage of the battery in the receiver.

Varicaps, which are diodes having a capacitance that can be voltage controlled when biased in the nonconducting region, are generally used for automatic frequency control of very-high-frequency oscillators in universal portable radios, because an automatic-frequency-control voltage is available from the ratio detector. Development trends indicate the possibility of using varicaps for tuning through the very-high-frequency band. This feature would be of particular interest because it could be combined with push-button tuning and remote control at moderate cost. The operating voltage needed to tune through the very-high-frequency band is about 20 volts; it can be provided from a direct-current-to-direct-current system.

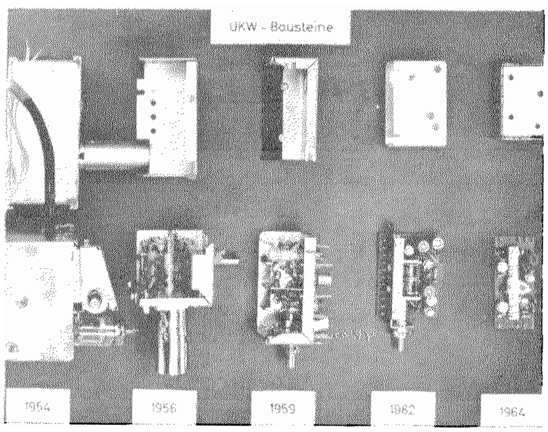

Figure 2—Reduction in size of the very-high-frequency tuner between 1954 and 1964.

In future years the possibility of electric tuning by varicaps will surely be closely scrutinized, and the component designers will have to meet tight specifications from equipment designers.

The introduction of transistors into mains-operated broadcast home receivers and consoles is at a much slower pace. Here vacuum tubes have reached a state of relative perfection after many years of development, and special combination tubes permit the design of optimum circuits. Moreover, the tubes have excellent regulation properties and are capable of processing the high input signal voltages obtained from outdoor and community antennas. Although new up-controlled [1] transistors permit the problem of automatic gain control to be solved, and although progress has been made in their handling of high input signal voltages, price considerations have so far favored full transistorization only of simple mains-operated receivers with single-channel audio amplifiers having power outputs not exceeding 3 watts in class-*A* operation. High-quality receivers with two-channel audio sections for stereophonic reproduction are predominantly equipped with tubes. An exception is the built-in radio-frequency stereo decoder, which makes use of the transistor advantages of low space requirements and low current drain; usually this circuit is of the plug-in type for convenient addition to the set. Occasionally the very-high-frequency section also uses transistors because the achievable noise figure is comparable to conventional values obtained with vacuum tubes.

Transistors in the power stages would require a costly power supply, as high currents are required at low operating voltages. An additional transistor stage in the base circuit, controlled by a zener diode, is a standard requirement to stabilize the power supply.

[1] Siemens "Technische Mitteilungen Halbleiter," number 1-6300-091, pages 4–16; October 1964.

Commercial tuners for high-fidelity systems with excellent electrical properties have proved the feasibility of all-transistor high-quality receivers; here, however, price considerations are not as important as in the conventional home receiver.

In television receivers, the introduction of transistors started, not unexpectedly, in the limiter stage of the intermediate-frequency amplifier. When the *AF 139* mesa transistor became available, however, transistors suddenly appeared in the ultra-high-frequency unit, for it markedly increased the receiver sensitivity. Equipping this unit with two *AF 139* transistors suddenly became compulsory; the rate of introduction of this transistor was determined by the production capacity at Siemens.

While transistors were originally used to replace tubes, the present state of the art in ultra-high-frequency units exploits all inherent possibilities of the transistor. Characteristic features are quarter-wave lines with 4-gang tuning capacitors in miniature compact units. The very-high-frequency tuner, a drum-switch unit derived from American models, had finally reached a state of development characterized by a low-noise cascode input stage with the special double-triode *PCC 88* tube, a mixer, and a separate oscillator using the *PCF 86* tube and having low spurious radiation. The designs used today feature transistors and station preselection by keys or multiposition switches as had been used earlier, a simplified switch-type tuner with neutrode input stage, and variometer tuning.

Two solutions are particularly interesting. One of them is the so-called integrated tuner, equipped with 2 to 5 transistors and only ultra-high-frequency-type tuning capacitors, tunable through all television bands. In switching from band to band, additional coils become series-connected to the ultra-high-frequency tuning lines while some of the transistors are involved in the switching. It is good practice to assemble this unit with a high-precision mechanical preselecting unit that permits rapid access to preset channels. The inherent features in this solution will influence the development trend in future years.

The other solution is the all-transistor very-high-frequency tuner with diode tuning. Three varicaps are tuned by a special voltage control, the operation of which includes a cam that switches the coils between bands 1 and 3. The frequency response is uniform with a flat characteristic of the control resistor. The main advantages of diode tuning are reduced space requirements, better operating reliability due to the small number of changeover contacts, and low torque. The last-mentioned advantage is being utilized by Grundig, the originator of this solution, for smooth drive of a single-knob control for 6 stations that can be selected as desired from the very-high- or ultra-high-frequency bands.

Vacuum-tube stages are still being used in the video intermediate-frequency amplifiers. Examples of such tubes are the *EF 183* and *EF 184* wide-band pentodes offering high gain. Also employed are amplifiers with an automatic-gain-control tube in the input. Generally 3 or 4 transistor stages are used with up-control types in the input stages. The existing trends point to all-transistor versions, but the pace is slow because economic considerations are at least as important as technical factors.

An accelerated development appears feasible in connection with the new silicon transistors *BF 167* and *BF 168,* which have very-low feedback capacitances as a result of a novel combined evaporation-and-diffusion method that permits the generation of a screening electrode between base and collector.

Standard Elektrik Lorenz produced the first silicon transistor actually used in a receiver. It is the type *BFY 41,* forming part of the circuit of the *HANSEAT* receiver manufactured by Mende. This transistor has been further developed and now has the voltage-breakdown stability between collector and emitter that is needed for the full control range of the television picture tube. Combined with a common-emitter input stage *BFY 39,* this *n-p-n* planar transistor, now coded *BFY 43,* enjoys an ever-increasing field of application.

Disregarding obvious transistor applications in the intermediate-frequency amplifier and the clipper, considerable time will elapse before transistors are exclusively used in mains-operated receivers. This is so because the power stages will require an additional costly power supply and because the problem of a reliable low-cost transistor for the horizontal output stage remains to be solved.

On the other hand, most portable receivers are fully equipped with transistors because low-power output stages produce enough power to deflect the electron beam in picture tubes up to 30 centimeters (12 inches) in diameter. Such receivers can be operated from the power mains as well as from a storage battery. However, the latter gives only a few hours of operation from one charge.

In Germany there is a trend toward 28- and 30-centimeter (11- and 12-inch) picture tubes in portable television sets, while smaller tube sizes are less favored. In the case of nonportable receivers, the 65-centimeter (25-inch) tube will compete with the 59-centimeter (23-inch) tube that has been the standard for several years. In 1964 the public seemed to be impressed by the idea of a picture aspect ratio of 3:4. However, manufacturers seem to be moving away from this by their reluctance to produce 3:4 picture tubes on the ground of difficult glass processing, high weight, and high price.

The design of the implosion-proof picture tube has covered virtually the whole field. Very few picture tubes are still being coated with a transparent plastic sheet for protection against the consequences of implosion.

Silicon transistors seem to be invading the field of television receivers, but their greatest prospects should be in motorcar receivers, where their excellent performance under varying ambient temperatures will be particularly appreciated.

In conclusion, it may be stated that all tubes could be replaced with equivalent-function transistors, excepting the picture tube, high-voltage rectifiers, and indicator tubes. Technical and price developments will set the pace and determine the percentage of silicon transistors. Totally unanticipated changes may result from new semiconductor devices.

2. Trends in Development of Passive Components

Traditional receiver components are resistors, capacitors, inductors, electromechanical transducers, relays, and loudspeakers.

Fixed resistors are available in a wide range of resistance values with tolerances of ±20, ±10, and ±5 percent. Their power ratings range down to 1/20 watt. In the future, physical size will be of increased importance and the range of sizes will be reduced in favor of optimum properties. Since the power rating depends on the ambient temperature, it will be stated in the form of a characteristic.

Resistors are available in batches of 1000, color coded to identify resistance values, and belted on paper tape for use in automatic assembly lines. In the future, physical designs for mounting normal to the circuit board are to be expected, and this may frustrate the automation of the assembly line. Connecting leads cut to size and reinforced or bent to fit board mounting holes will increase reliability and shorten assembly time.

Almost fully automatic production methods have resulted in fixed resistors of very-uniform quality and very-low price.

Adjustable resistors for conventional volume, tone, contrast, and brightness controls, open or enclosed, have now been supplemented by double adjustable resistors for ganged control of stereo amplifiers.

Controls for electronic tuning are under development. High accuracy of repetitive adjustments and close tolerance of resistance versus angular position are required for indication of the tuning position on a calibrated scale. Possibly the slide resistor will be revived for this application, the slide being used as a scale pointer. Vernier controls with preset station selection, combined with switches or keys, can be expected to appear on the market.

Capacitors are being offered in bewildering varieties. Among them are fixed ceramic types with excellent radio-frequency properties, types having closely staggered temperature-coefficient ranges for compensating circuits in filters and oscillators, and numerous types resembling stalks, discs, caps, et cetera. The standard series are no longer satisfactory, as radio-frequency circuits often call for close capacitance tolerances to meet a particular design. Apart from ceramic types, wound styroflex capacitors are often used in filters because of their relatively low price; their weakness is a heat sensitivity that presents difficulties in dip soldering or in physical contact with a hot soldering iron.

A clearly noticeable trend in large wound capacitors is in size reduction where paper is replaced by plastic as the dielectric. (See Figure 3.)

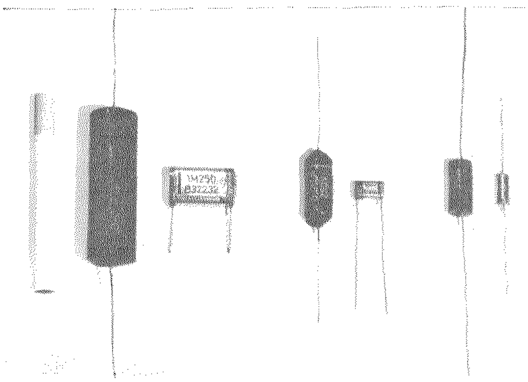

Figure 3—Pairs of equivalent wound capacitors using paper (left) and plastic (right) dielectrics.

Automatic assembly of capacitors is not being employed, as thin connecting leads of various diameters are necessary to avoid contact difficulties.

Electrolytic capacitors for transistor circuits show the same trend toward miniature size for convenient mounting on boards. The solid-electrolyte tantalum capacitors that may be favored in tight locations of future designs are particularly small. Generally the use of boards and cards controls the standardizing of capacitor and connecting-lead shapes and sizes in compliance with standard grids. Special capacitors are available for booster circuits in television sets, for protecting circuits in accordance with various national standards, and for lead-through applications for radio-frequency blocking.

Adjustable capacitors are still the most-suitable tuning means for broadcast receivers. Two-gang capacitors for superheterodyne receivers have frequency-corrected stator-plate designs and built-in transmission gears stressed to prevent lost motion. In the very-high-frequency case tuning variometers and varicaps are also employed. Equipment designers have demanded varicaps suitable for the medium-wave range, too. Early realization of this requirement by component producers is not probable because the relatively large bandwidth presents serious difficulties. The use of tuning varicaps in the ultra-high-frequency section of television sets seems feasible if the losses can be kept small for frequencies up to 1 gigahertz.

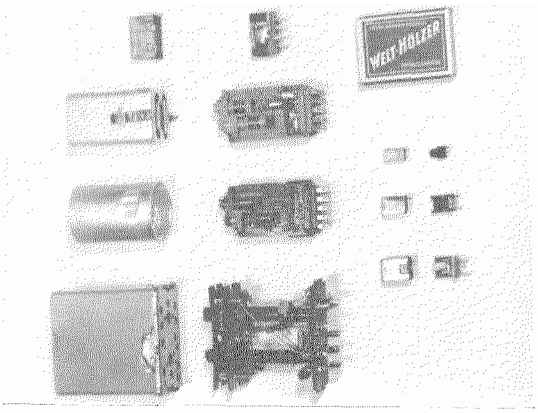

Figure 4—Reduction in size of intermediate-frequency filters and coils.

Coils are now also being used with cards, especially in radio-frequency circuits in transistor sets. They are tapped for matching or neutralizing. Various methods are used in the coil-winding process to connect the proper wire end to the proper terminal. A possible solution of this problem is double dip soldering, the first time after the coil-winding process and the second after insertion into the card. Cam-controlled moving cross-coils are often being employed in input circuits where ferrite antennas or tuning variometers are provided.

Intermediate-frequency filters are designed for alignment from one side and simple mounting (see Figure 4) ; in their production, automatic coil winding is applied. The era of printed coils seems to have begun with the wide-band circuits of the video intermediate-frequency amplifiers. Forerunners of printed coils may be found in very-high-frequency channel selectors, in printed-wiring matching circuits for very-high-frequency antennas, and in printed Guanella transformers (2 pairs of coupled coils giving $Z/2$ to $2Z$ impedance conversion). Radio-frequency coils present primarily a problem of printing technology because the spacing of the coil turns must be very close to produce useful inductance values.

Only electromechanical filters can be expected to supersede the conventional amplitude-modulation intermediate-frequency coils. The first experiments with ceramic resonators called for additional wound coils for matching and for suppression of undesired radiation. These problems should be solved in the near future.

Transformer design is also receiving attention, particularly for a suitable production method for the manufacture of low-leakage tape-wound cores. Pocket receivers require either small transformers or output stages that need no transformer. Thus the automatic winding of coreless transformer coils and their automated assembly with core laminations is in competition with new circuit designs. The output stages without transformer need higher voltages, however, than can be obtained from a car battery.

Horizontal output transformers of television sets (see Figure 5) have become much more reliable because of reduced leakage in the coupling of the windings, better impregnation of high-voltage coils, more-effective arrangement of parts, and improved core material. Plug-in types are still being favored for more-convenient servicing. The trend toward higher and higher voltages entails occasional failures, although designers have learned to keep temperatures of core and windings within safe limits even under adverse conditions.

Further development of line transformers is necessitated by more-stringent requirements for television receivers and by the miniature size required for portable sets. The same is true of deflection coils, the development trends of which are controlled by transistor application, miniature parts, color, and picture-tube dimensions.

The loudspeaker is the most-important transducer. Its development is also determined by a number of factors. The range of application extends from the miniature type in the pocket radio to the bass loudspeaker in a high-fidelity system. The former calls for small dimensions and high efficiency; the latter should have a flat response-frequency characteristic and no distortion. Ferrite magnets help to reduce the depth of components in portable sets; however, their leakage field may affect the radio-frequency and intermediate-frequency circuit tuning. Therefore, either problems in the alignment of these tuned circuits or the use of a leakage-compensated and more-expensive loudspeaker must be faced.

In high-fidelity reproduction, a new loudspeaker arrangement has been introduced. It has low efficiency but excellent tone reproduction. The housing volume is relatively small. The desired result is achieved by use of a very-low resonant frequency of the speaker and by filling the housing with damping mineral fibers.

The loudspeakers used in home receivers are mostly oval, with axis ratios between 1.5:1 and 4:1. High ratios such as the last-mentioned meet the requirements of industrial designers better than those of acoustical engineers.

The electrodynamic loudspeaker is clearly predominant while the dielectric speaker, used for the upper audio range for many years, is now of very-limited significance. The demand by the equipment designer for a loudspeaker with much-higher efficiency has not been met in the past 40 years.

It is hardly necessary to dwell for long on relays. Their range of application is limited and their use is costly; even today there is no inexpensive manufactured product that will retain the required reliability over a long life. Relays are used only in special equipment such as multistandard receivers for television, remote-control equipment, et cetera.

Keyboards for band switching, preferred in broadcast receivers, matured to separate units during the wired-chassis era. Apart from the switching contacts, they comprised also tone controls with indicators, tone-control keys, scale holder, et cetera. Today they are much less expensive because of automated production; they are light, made primarily of plastic, and can be directly inserted into wiring boards. These keyboards have inherent disadvantages: The number of keys cannot exceed a certain limit

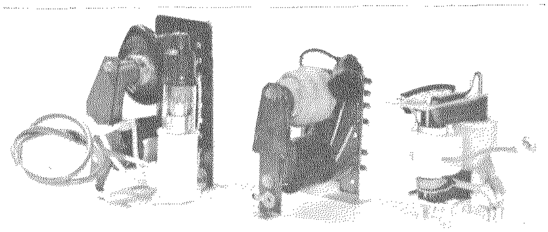

Figure 5—Horizontal output transformers. The newest type at the right is a plug-in unit.

if mechanical stability is to be maintained. Mechanical switching of solid duplex drives presents difficulties. It is impossible to use secondary switches controlled by levers. However, the process of adaptation to the new conditions is in full swing, so improved solutions may still come forth.

Other keyboards are used in television sets to preselect desired stations. A precise mechanism is indispensable, as the accuracy of adjustment must be within 10 to 20 micrometers. Available mechanisms operate with an accuracy that eliminates the need for automatic frequency control in some cases. These arrangements, complemented by multiposition switches of comparable accuracy, have superseded costlier solutions such as automatic station hunting and motor-operated channel selectors.

Extreme precision is also necessary in keyboards of car receivers, in which the job is complicated by the difficult coil-winding process for the variometer required for tuning. In Germany these devices are manufactured by producers of car radios for their own use and are not commercially available.

3. Summary

The trend in the development of components for broadcast and television receivers is essentially following the specifications resulting from the use of transistors, miniature parts, and printed-wiring boards.

In the future, most of the progress will be in the area of miniature parts. Thin-film devices are going to be used in the very near future. Attention is being paid to thick-film circuits and integrated semiconductor circuits; the prices for these devices soon may drop very rapidly to attractive levels.

Jan Harmans was born on 17 February 1912 in Dresden, Germany. He graduated from the Technische Hochschule in Dresden as a Diplom-Ingenieur in 1936 and received his Dr.-Ing. degree in 1940. Thereupon he worked on the development of aircraft guidance systems and radar techniques.

In 1949 Dr. Harmans joined Standard Elektrik Lorenz, where he is now in charge of the development of radio receivers.

Millionth Line Cut Over in New Carnot Exchange—The cutover of a new 10 000-line Pentaconta telephone exchange in Paris also completed the installation of the millionth line of automatic telephone subscriber equipment by Le Matériel Téléphonique.

The new exchange was officially inaugurated by Jacques Marette, Minister of Post and Telecommunications, in the presence of about 300 persons including R. Croze, General Director of Telecommunications, and M. Jambenoire, Director of Telecommunications of the Paris area. See Figure 1. The rotary *7A1* Carnot exchange being replaced by the "all-number" *227* was the first automatic central office in Paris. It has operated continuously since 1928 including the war years. It is now planned to replace all rotary in Paris with Pentaconta switching systems.

The new center will handle part of the incoming traffic for the MacMahon rotary *7B1* office. Its building now houses 50 000 lines, of which 14 000 are Pentaconta equipment and the remainder are rotary. It includes the transit center for the northwest Paris suburbs.

Of the million subscriber lines, 480 000 are rotary *7A1,* 430 000 are rotary *7B1,* and 90 000 are in Pentaconta system. In addition, several automatic toll exchanges and 2000-line concentrators have been produced and installed in France and abroad.

Le Matériel Téléphonique
France

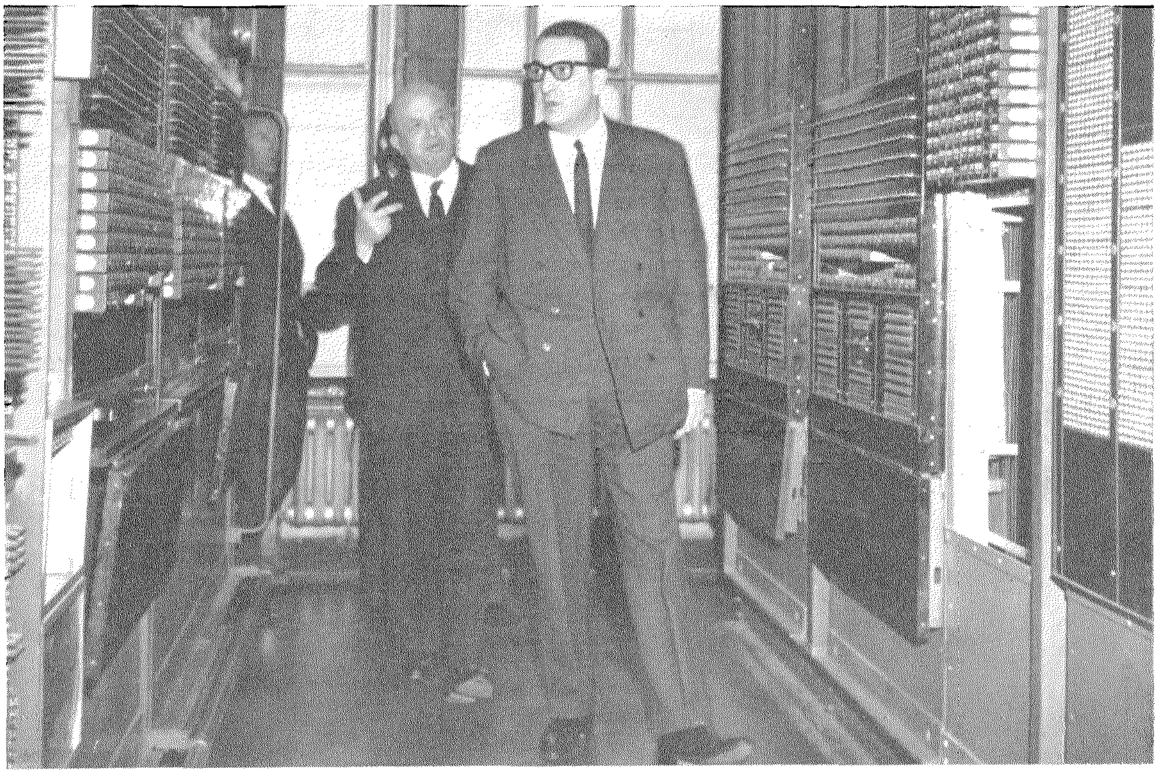

Figure 1—Jacques Marette, French Minister of Post and Telecommunications, and M. Jambenoire, Director of Telecommunications of the Paris area, visiting the old Carnot exchange before its replacement by the new *227* Pentaconta switching office.

Automation of Brussels Postal Check Office— The Belgian Minister of Post and Telecommunications inaugurated the operation of the first equipment installed in Brussels for machine sorting and accounting of postal checks. See Figure 2. The installation of 8 more equipments, each consisting of 2 jacketers, 1 processing complex, 3 classifiers, and 2 unjacketers, will provide for processing 1 250 000 transactions per day.

A jacketing machine first inserts documents (which vary in size and paper quality) individually into transparent mylar jackets. Each jacket is provided with a strip of magnetic tape on which the pertinent information from the document is recorded.

The jacketed documents then go to the processing complex, which consists of a computer, tape units, high-speed printer, and jacket reader-sorter. These are used for coding, sorting, and bookkeeping operations.

The classifying machine sorts a batch of 1200 jackets according to their account numbers. In the final operation, the unjacketing machine removes the documents from the jackets and dates the documents.

Bell Telephone Manufacturing Company
Belgium

Quasi-Electronic Telephone Exchange in Vienna —The first quasi-electronic telephone central office in Austria was cut over in March 1966 by Otto Probst, Austrian Minister of Telecommunications and Transport. Several officials of the Austrian telecommunications administration led by General Director Benno Schaginger also attended.

In the business district of Vienna, the exchange serves 500 subscriber sets of dial and push-button types. Abbreviated-number calling is available.

The exchange is a development of the *HE-60L* system initially installed in Stuttgart in 1963. It uses transistors, diodes, magnetic cores, and integrated circuits for storage and selection of routes, and Herkon reed relays as switching

Figure 2—Postal check processing office in Brussels.

crosspoints. The control center is shown in Figure 3.

Standard Telephone & Telegraphen
Austria
Standard Elektrik Lorenz
Germany

Pulse-Code-Modulation Multiplex Set—The *WC-101B* pulse-code-modulation multiplex unit shown in Figure 4 and a *CXL-5* microwave set combine to provide a complete 12-channel radio terminal approximately 1 cubic foot (0.03 cubic meter) in volume and about 55 pounds (25 kilograms) in weight. Less than 100 watts is drawn from a self-contained power supply.

The time-division-multiplex unit provides for bit synchronization of the 6-bit code and for bipolar modulation. Two multiplexers can be used with a combiner to provide 24-channel capability.

Built-in facilities for alignment, fault detection, and isolation make external test equipment unnecessary. The low-power microelectronic circuits have extremely high stability and reliability with minimum maintenance.

ITT Federal Laboratories
United States of America

Pulse-Code Modulation for Mobile Radiotelephony—A coder-decoder to permit pulse-code modulation to be used on single-channel mobile radiotelephone sets can be accommodated in a volume of 0.1 liter (6 cubic inches) with a weight smaller than 200 grams (7 ounces).

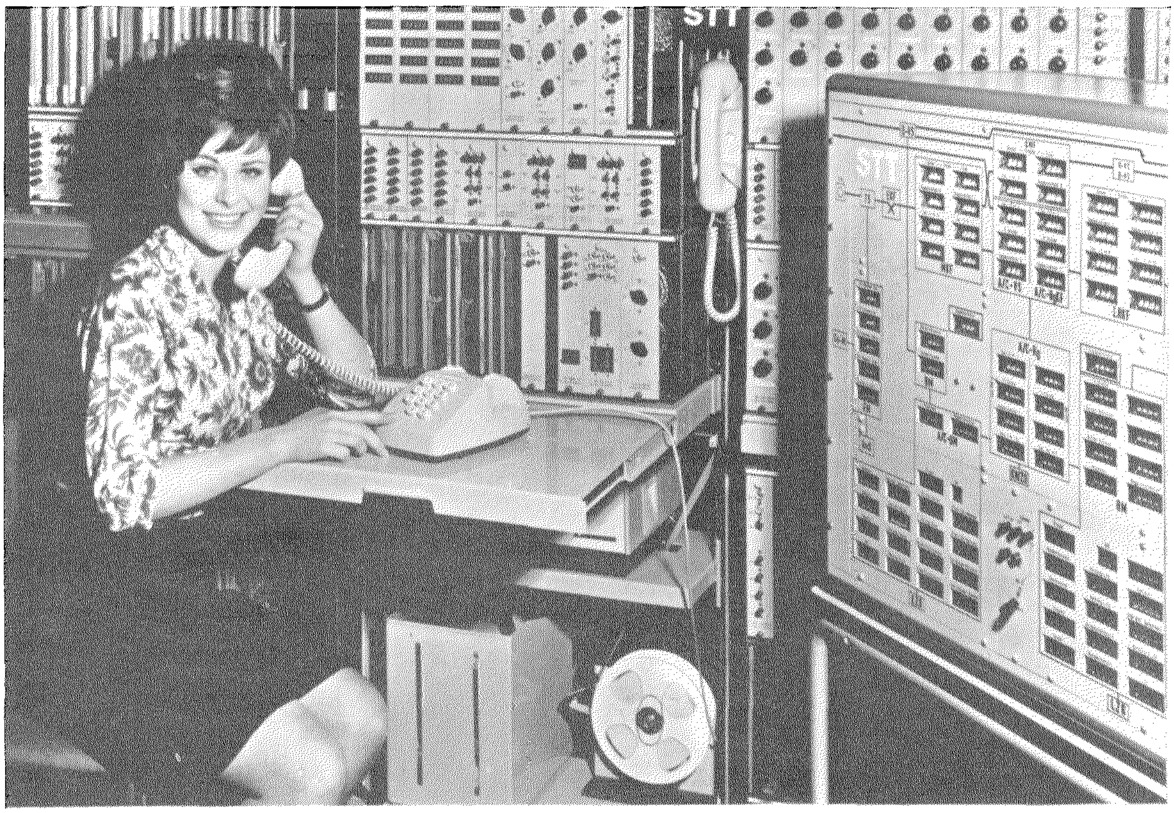

Figure 3—Testing and control center of quasi-electronic telephone exchange in Vienna.

Using integrated circuits, 6-bit codes are produced at an 8-kilohertz sampling frequency. A 4-slope multilinear companding law gives a companding improvement of 26 decibels. It is compatible with a military integrated telephone network.

A digital counting coder uses a varying rate of counting with amplitude to introduce amplitude compression.

Laboratoire Central de Télécommunications
France

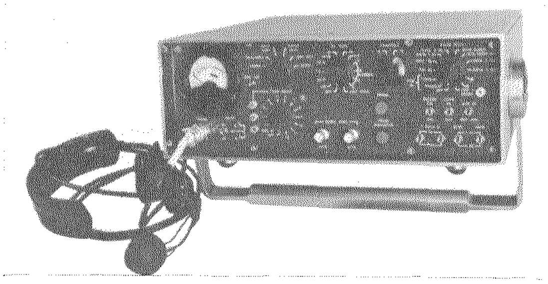

Figure 4—Microelectronic multiplexer for 12-channel pulse-code modulation.

Computer-Based Communications System 6300-ADX—The *6300ADX* is an on-line stored-program digital system designed to operate on a real-time basis in telecommunication networks having up to 60 duplex lines.

A message received at a typical station in a telegraph network may be retransmitted to several other stations and then in turn rerouted and processed in a complex pattern. The *6300ADX* controls each relay center according to a stored program and automatically routes messages and data from manual keyboards, automatic machines, or from other computers.

Figure 5 shows the method of operation. Information from incoming lines is stored magnetically to drive the outgoing lines. Routing and queuing decisions are performed by the central processor at electronic speed and a message need not be fully received before it is processed and placed on an outgoing queue.

The central processor stores 12-bit words in ferrite cores in modules of 4096 words, up to a maximum of 32 768 words, for programing and for buffering of input and output information. The cycle time is 1.5 microseconds. The main store is a magnetic drum with a maximum capacity of 262 144 words. Magnetic tape can be used for auxiliary storage.

Multiplexors enable up to 60 duplex land lines or radio circuits to be served in real time. Transmission speeds between 50 and 2400 bauds can be accommodated. Speed, code, and format conversion can be provided for intercommunication irrespective of such differences.

Typical applications of the system include: telegraph message and data switching, message collection and distribution, man-to-computer information interchange, integration of multiple computer networks, message accounting, traffic statistics, stock control, and reservation systems.

Standard Telephones and Cables
United Kingdom

Advances in Components—The ant-size tantalum capacitor shown in Figure 6 is available in ratings up to 35 volts and up to 150 microcoulombs. Its size, excellent frequency performance, working temperature from −55 to +85 degrees Celsius, and stable low leakage current

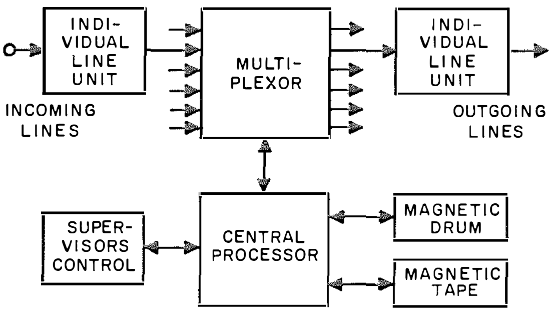

Figure 5—Block diagram of *6300ADX* system. Additional multiplexors with their individual lines and additional storage can be connected to the central processor to increase the line capacity of the system.

make it superior to the hitherto-used wet electrolytic aluminum capacitor for portable radio and television sets and car radios.

A new metallized paper capacitor will withstand a case temperature of 100 degrees Celsius against 85 degrees for conventional units.

Taking advantage of the development of selenium high-voltage rectifier stacks by Standard Telephones and Cables, the rated reverse voltage has been increased to meet the requirements of the German market. A 20-kilovolt stack has a length of 130 millimeters (5.1 inches) and is 4.5 millimeters (0.18 inch) in diameter.

Developments in passive and hybrid thin-film circuits have resulted in production of thin-film circuits for electronic desk-size calculators and car radios to obtain the advantages of small size and reliability.

Standard Elektrik Lorenz
Germany

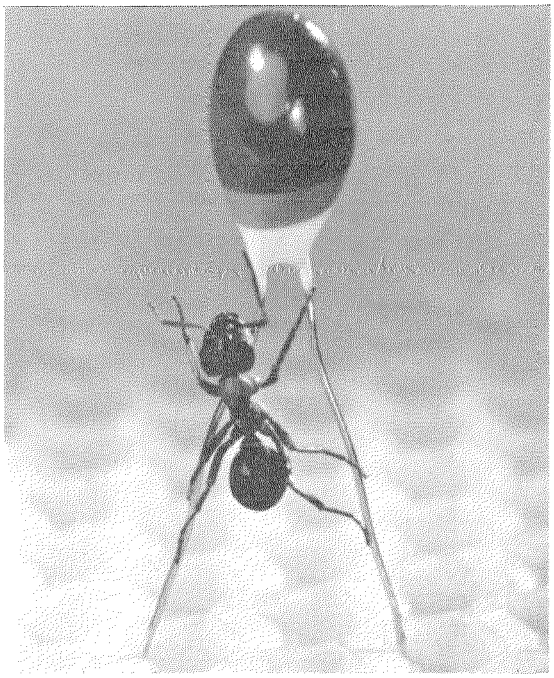

Figure 6—Bead-type tantalum capacitor compared with an ant.

Very-High-Frequency Radiotelephone *TransITT-8*—Smaller than most car radios is the *TransITT-8* transmitter-receiver shown in Figure 7. Intended for dashboard mounting, it is powered directly by the 12-volt car battery. It is designed for operation in the 68–88 and 146–174 megahertz bands with phase modulation and channel separation of 20, 25, or 50 kilohertz.

This set is part of a new line of very-high-frequency equipments for portable, mobile, fixed, and marine stations. Ministac construction allows very small modules even when using reliable components of normal size. Most modules are identical for the whole line of equipments, which simplifies maintenance and production.

Standard Electric
Denmark

From Learning to Computing—A task presented to the STeLLA learning machine is in the form of a changing binary pattern, a number of actions, and a binary channel for "reward signals." STeLLA is designed to obtain a reward signal as often as possible by performing the actions, given no other information than the binary pattern. She learns the consequences of

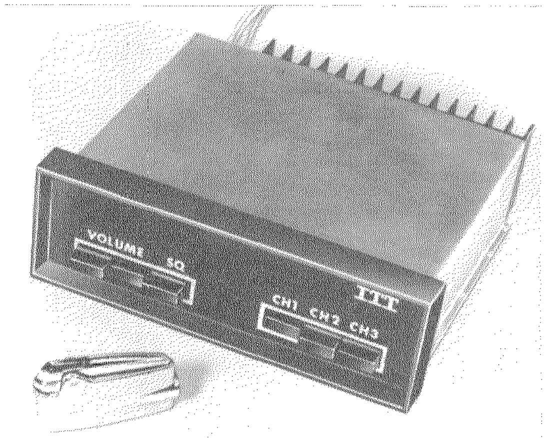

Figure 7—Radiotelephone set *TransITT-8* for mobile use.

the actions, the relevance of the input, and the nature of the task for which she is being rewarded. Experience of her tasks is stored in the form of paths of pattern clusters on a adaptive hill, rising to reward. General experience of her environment is stored in an adaptive model which enables her to construct forward trajectories by prediction and to use such "hypotheses" to optimize her strategies.

Various tasks have been presented to STeLLA with encouraging results. Figure 8 shows a "hill of paths" developed in her memory for an automobile-steering task. She is given the position across the road of the vehicle and its angular position relative to the direction of the road, coded as a 10-bit word. STeLLA neither starts with any information about the significance of this coded word nor of the three actions which cause the vehicle to be steered either ahead, to the left, or to the right. Shown in the figure above a position-angle plot, shaded areas represent the positions and angles of the vehicle for which STeLLA has found particular actions appropriate. Given quite different problems, she would develop a similar hill of paths. Efficient strategies have been developed for STeLLA by observing her behavior with difficult tasks when her memory store is severely limited.

So far, STeLLA and her environment have had to be simulated on a digital computer, but with the progress of microcircuitry the possibility of building a model is becoming more feasible. The ADDIE adaptive digital circuit was developed to this end. Now it has become apparent that the ADDIE is but one element appropriate to a new form of computing which employs stochastic sequences to represent variables. The stochastic computer promises the realization of complex systems for process control, aircraft and missile guidance, and pattern recognition, with all the benefits of microminiaturization and digital compatibility.

Standard Telecommunication Laboratories
United Kingdom

Miniature High-Voltage Power Supplies—Outputs of 12 and 16 kilovolts at 1 microampere with a maximum ripple of 1 percent are obtainable from inputs of 1.5 volts at 18 milliamperes or 1.34 volts at 15 milliamperes with a standard line of power supplies shown in Figure 9. The units are 2.7 inches (69 millimeters) long by 1 inch (25.4 millimeters) in diameter and weigh 4 ounces (113 grams).

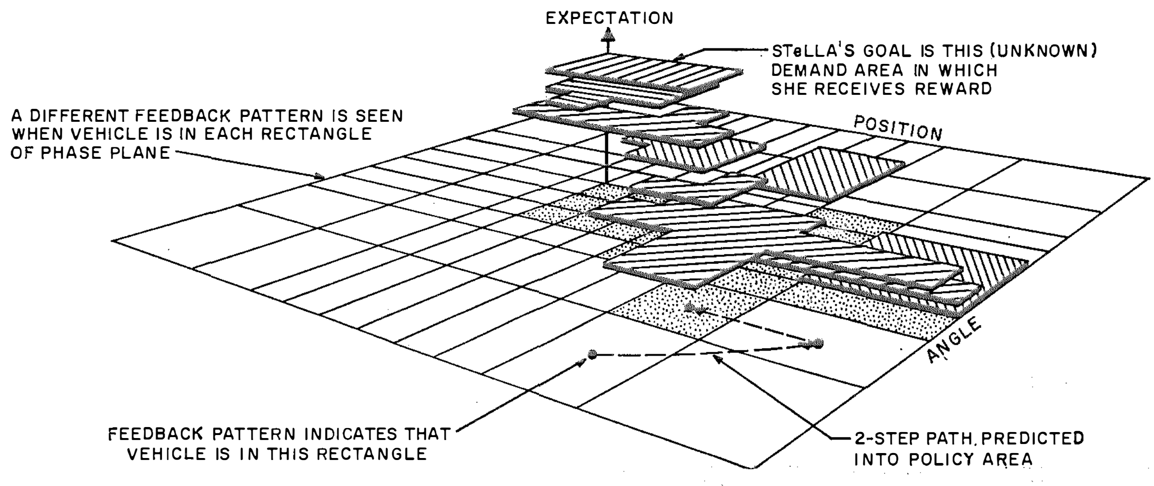

Figure 8—Experience of learning machine in the form of steps leading up to award.

Other models in rectangular cases 1 inch (25 millimeters) square by 4 inches (100 millimeters) long feature inputs of 12, 20, or 28 volts with outputs in 300-volt steps between 1200 and 3000 volts. Ripple is 1 volt peak-to-peak.

These supplies are useful for operating photoemission tubes, ultraviolet and infrared detectors, photometers, spectrometers, and electro-optical imaging systems.

ITT Industrial Laboratories Division
United States of America

Jewel Box Subscriber Set—Soon to be distributed internationally is the Jewel Box subscriber set shown in Figure 10. Starting with the finest quality for transmission and reception, it offers a new shape of harmonious proportions coupled with ease of handling, 5 different colors, an electronic bell of pleasing tone and adjustable intensity, automatic rewind of handset cord into the base, and sturdy construction. Weighing only 950 grams (33.5 ounces) it measures 216 by 120 by 92 millimeters (8.5 by 4.7 by 3.6 inches).

Fabbrica Apparecchiature per Comunicazioni
Elettriche Standard
Italy

Relay for Printed Circuits—The *PZ* relay shown in Figure 11 is designed for direct soldering on printed-wiring boards. This is aided by keeping the terminals at least two modules (a module is 2.5 or 2.54 millimeters) apart. The height of the relay has been reduced as this is often a problem. Four changeovers are provided with double contacts. Relays are available for direct-current operation at 12, 24, 36, and 48 volts. A dust-proof protection cover is supplied. Dimensions are 15 by 23 by 28 millimeters (0.6 by 0.9 by 1.1 inches). The weight is 17 grams (0.55 ounce).

Contact pressure ≥ 10 grams, operate time is 5 milliseconds, and release time ≈ 2.5 milliseconds.

Standard Téléphone et Radio
Switzerland

Test Jack Strip—A new test jack strip for use on main distributing frames in telephone offices has the same size but twice the capacity of the present units (40 pairs instead of 20 pairs) and also replaces the existing protector mounts. A variant used on the horizontal side of the distributing frame has 1 or 2 extra pairs used

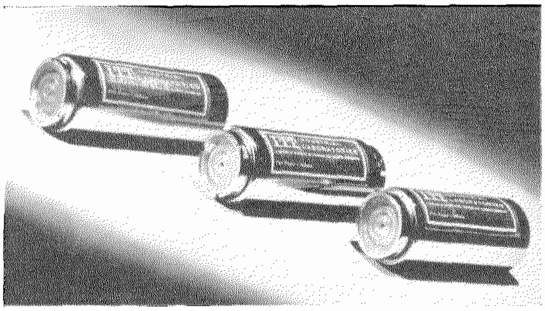

Figure 9—High-voltage power supplies for phototubes and similar devices.

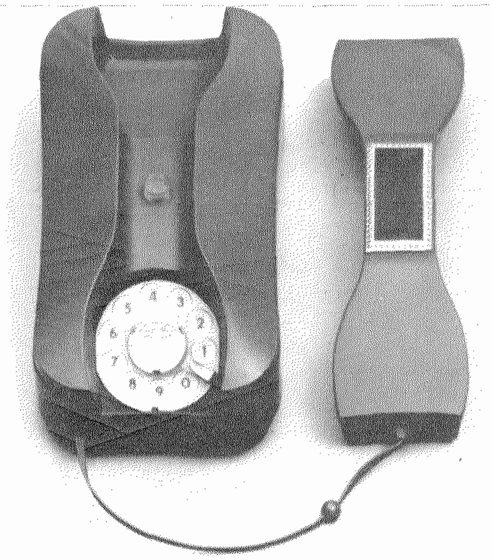

Figure 10—Jewel Box subscriber set.

as pilot lines. The required floor space and amount of jumper wire is reduced to about 40 percent.

Independent wafers are stacked with a change in color every 5 wafers for identification as shown in Figure 12. Individual wafers may be replaced without disturbing the others.

Color-coded plugs are inserted to perform the following functions: connect tone or voice to a subscriber line, test outer and inner lines, test insulation of 1 or 5 lines simultaneously, discriminate the type of subscriber line and current supply, and protect against abnormal voltages and currents. These latter plugs carry 2 delay-action fuses and replace existing protectors.

Standard Eléctrica
Spain

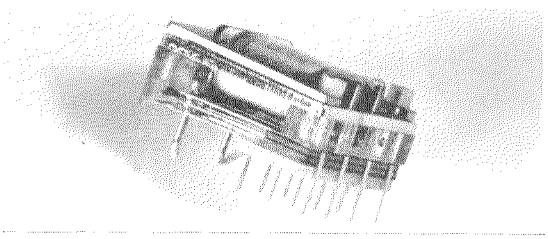

Figure 11—Relay for mounting on printed-circuit boards.

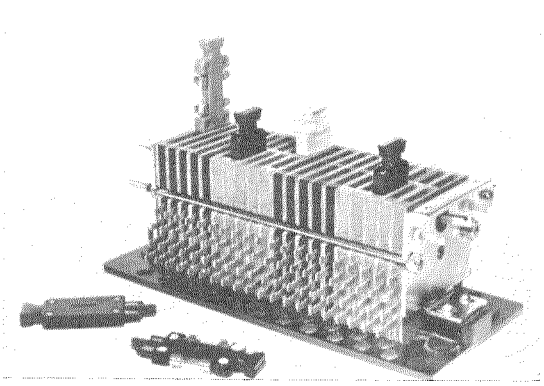

Figure 12—New test jack strip for telephone main distributing frames.

Optical Document Sorter ODS 2—The *ODS 2* shown in Figure 13 reads and sorts documents varying in size between 105 by 65 millimeters (4.1 by 2.5 inches) and 210 by 105 millimeters (8.3 by 4.1 inches). Paper weight may be from 80 to 170 grams per square meter (approximately 20 to 40 pounds per 500 sheets of 17 by 22 inches). The constant transport speed at the reader of approximately 2 meters (79 inches) per second permits sorting 45 000 of the smallest documents per hour.

Special characters, *OCR-A,* provided for in German standard *DIN 66 008,* are shown in Figure 14. There are 4 characters per centimeter (10 per inch). Only numbers and four special signs are used. Printer's type, high-speed printers, typewriters, and adding machines are available for producing these special characters.

Sorting may be to 10 destination pockets with an additional pocket for nonidentified items and another pocket for special items. The sorter

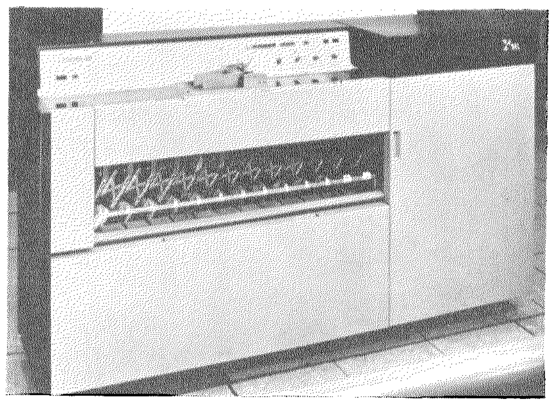

Figure 13—Optical sorter *ODS 2.*

Figure 14—The optical sorter uses these special characters.

Figure 15—Multimeter with clip-on current-transformer probe.

Figure 16—Bergamo N transit exchange connected with the local exchanges A and B.

can be adapted for on-line operation with a computer.

Standard Elektrik Lorenz
Germany

Multimeter With Clip-On Current Probe—The *404B* alternating-current instrument shown in Figure 15 has a 12-position rotary switch to select function and range. A clip-on current transformer permits current measurements without breaking the circuit to be measured. A transistor amplifier is used for some ranges.

With the clip-on transformer, alternating-current ranges are from 0.1 to 1000 amperes, and without it from 0.1 milliampere to 1 ampere. Voltage ranges are 150 and 600 volts with 5000 ohms per volt. The single resistance range is of 100 kilohms with the center-scale value being 2 kilohms.

Everything is contained in a leather case 180 by 120 by 260 millimeters (7.1 by 4.7 by 10.2 inches) and weighs 3.6 kilograms (8 pounds).

Compagnie Générale de Métrologie
France

Expansion of Bergamo Pentaconta Installation—Two new Pentaconta switching offices were cut over in Bergamo, Italy. One was a transit exchange with 327 incoming and 273 outgoing junctions, Bergamo N, and the other was a 6000-line extension, Bergamo B, of the existing local exchange, Bergamo A.

Bergamo is now the largest Pentaconta installation in the world, consisting of a local exchange in Pentaconta 500 switching having 20 000 subscriber numbers, the new extension in Pentaconta 1000 switching with 6000 subscriber lines and capacity for 30 000 lines, and the trunk transit exchange for subscriber direct toll dialing and semiautomatic traffic originating and terminating in the local exchanges. Thus the local exchanges have 26 000 subscriber numbers in use and there are 1565 junctions to and from Bergamo. Figure 16 shows Bergamo N connected with A and B.

Fabbrica Apparecchiature per Comunicazioni Elettriche Standard
Italy

Transistor Test Set—The *74163B* test set shown in Figure 17 will measure dynamic current gain of both silicon and germanium *p-n-p* and *n-p-n* transistors. Gains up to 200 are indicated on a direct-calibrated dual-scale meter. The emitter current is approximately 1 milliampere and collector voltage is approximately 5 volts.

The appropriate circuit is obtained by push-button switching. The transistor can be connected to either a 3-pin socket or 3 terminals. The voltage of the operating battery can be checked on the meter.

Overall dimensions including the handle are $7\frac{3}{4}$ by $5\frac{1}{4}$ by $3\frac{5}{8}$ inches (197 by 133 by 92 millimeters). The weight is $3\frac{1}{4}$ pounds (1.5 kilograms).

Standard Telephones and Cables
United Kingdom

Secondary-Radar Transmission System *FAS-3*—The additional data available from secondary radar beyond that obtained with primary radar makes it an important element in air-traffic control. Further improvement results from placing some secondary radars at a distance from the air-traffic-control center.

The *FAS-3* system extracts only the essential information from a secondary radar and encodes it for transmission over conventional telephone lines to the air-traffic-control center. The 4096 possible codes are decoded at the receive end for actuating the display equipment. Either the position of an identified aircraft (passive decoding) or the identity of a visible target (active decoding) can be obtained.

Standard Elektrik Lorenz
Germany

Alternating-Current Millivoltmeter — Useful over the frequency range from 20 hertz to 1 megahertz is the model *777A* voltmeter shown in Figure 18. Its 10 ranges with full-scale values from 1 millivolt to 30 volts may be

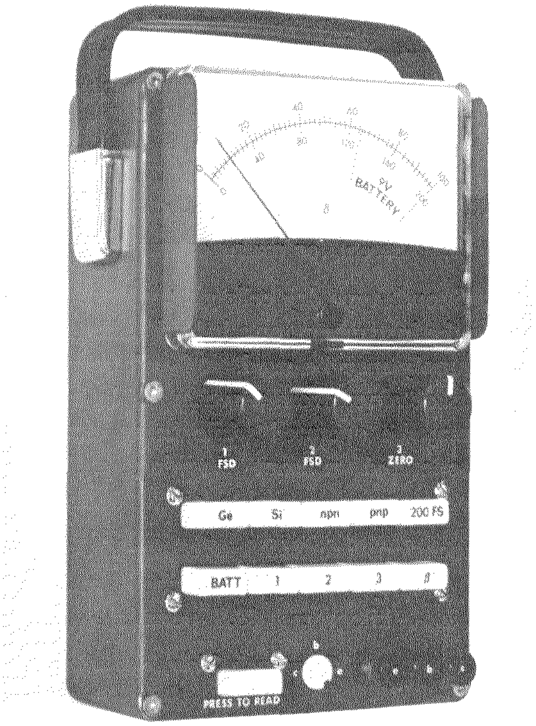

Figure 17—The *74163B* transistor test set features portability, robustness, and extreme simplicity of operation.

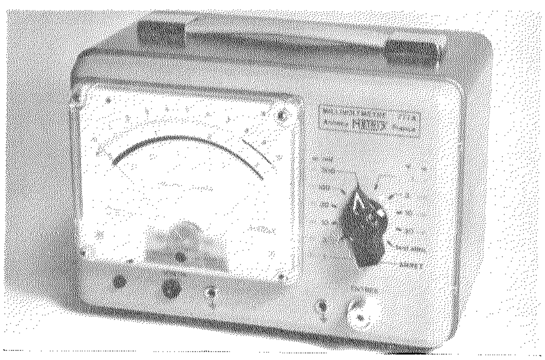

Figure 18—Alternating-current millivoltmeter.

multiplied by 10 with a probe having a built-in voltage divider. The probe reduces the input capacitance at high frequencies. The input impedance without probe consists of 1 megohm in parallel with 30 picofarads.

A transistor amplifier on printed-circuit boards is flat within ±0.3 decibel over the frequency range and operates a galvanometer. A negative-feedback loop gives a linear scale. An amplifier outlet on the panel permits adjustment of the gain or attenuation of the amplifier. Its maximum output of 1 volt direct current corresponds to full deflection of the galvanometer. Error does not exceed 1.5 percent for battery operation or 2.5 percent for mains operation.

Either 6 torch-type dry cells giving 9 volts will operate the instrument, or a hybrid battery-and-mains supply may be used. A 9-volt regulated supply is obtained from a rechargeable battery which may be operated while being charged. When operated without charging, a 1-volt reference voltage is available for calibration.

The dimensions are 235 by 180 by 200 millimeters (9.3 by 7.1 by 7.9 inches) and the weight is 4 kilograms (8.8 pounds).

Compagnie Générale de Métrologie
France

Lillo Subscriber Set for Hotels—Derived from the well-known Lillo subscriber set, the hotel model shown in Figure 19 offers high electro-acoustic quality at a low price. It is available in several colors for wall mounting or for table use.

Fabbrica Apparecchiature per Comunicazioni
Elettriche Standard
Italy

Improved Storage-Type Cathode-Ray Tube—The *AS17-21* dark-trace cathode-ray tube for storing oscillographic images has been replaced by the *AS17-21A,* which is capable of higher erasing frequency. It is shown in Figure 20. With intermittent operation, life is approximately 40 000 write/erase cycles and with continuous erasing it is more than 3500 hours. It is useful in measuring, recording, and control instruments where the graphic presentation of a process has a duration greater than 20 to 30 seconds and where a complete series of a large number of successive transients at long intervals must be recorded.

Standard Telephones and Cables
United Kingdom

Telephone Terminal Equipment for Mobile Radio—In many mobile radiotelephone systems, the fixed station must interconnect with existing telephone networks. As both radio and telephone systems vary greatly, the interface equipment has been designed to be readily modified to fit each particular set of conditions.

Among the numerous variations are simplex or duplex operation, 2-wire or 4-wire telephone connections, selective radio calling, direct dialing in both directions, and the distances between the fixed radio station, radio operating room, and telephone exchange.

The interface equipment shown in Figure 21 is constructed in International Standard Equip-

Figure 19—Hotel-type Lillo subscriber set.

ment Practice using printed-circuit boards and modules. Semiconductors and Herkon relays are used in the voice-frequency circuits, and miniature plug-in relays perform auxiliary functions.

Without changing the overall wiring, printed-circuit boards can be selected to fit the needs of a particular installation. For instance, a linear amplifier or a voice-operated gain-adjusting device (*VOGAD*) may be used, either a 2-wire or a 4-wire relay may be plugged in, or a transistor chopper may provide for 25-hertz ringing voltage.

If a hybrid is needed having a high safety factor against radiation of the echo of the received signal by the transmitter regardless of variations in impedance of the 2-wire system, a self-regulating unit can be inserted without disturbing the rest of the equipment.

Standard Elektrik Lorenz
Germany

Precision Signal Generator 932ATR—The *932-ATR* generator covers from 50 kilohertz to 50 megahertz in 9 bands with direct indicating scales accurate to within ± 1 percent.

The use of symmetrical circuits, output level control, low-frequency negative feedback, and stabilized power supply gives a frequency stability better than $\pm 5 \times 10^{-5}$ over 10 minutes. Frequency modulation due to hum is lower than ± 5 hertz over the entire frequency range and that resulting from the intended 30-percent amplitude modulation is less than $\pm 10^{-6}$ of the indicated carrier frequency. Step and continuously adjustable attenuators provide for all amplitudes between 2.23 volts and 0.2 microvolt into a 50-ohm load. The attenuators are calibrated in volts and in decibels referred to 1 milliwatt.

Modulation at 1000 hertz ± 5 percent is produced by an internal resistance-capacitance oscillator. Modulation depth is indicated on a galvanometer to ± 5 percent and is adjustable to 100 percent. External modulation between 20 and 20 000 hertz may be applied. The modulating frequency must not exceed 6 percent of the lowest value of the frequency range being used for 30-percent depth of modulation and 2 percent for 80-percent modulation. A root-mean-square voltage of 0.6 volt is required for 100-percent modulation.

A quartz-controlled harmonic generator operable at 100 kilohertz and at 2 megahertz with a precision of 5×10^{-5} provides a check of the frequency of the main generator. The maximum error is 0.1 percent for any frequency between two check points in a band.

The *932ATR* will operate from a 115- or 250-volt source between 48 and 400 hertz, requiring

Figure 20—Dark-trace cathode-ray tube *AS17-21A* for use in storage-type oscillograph instruments.

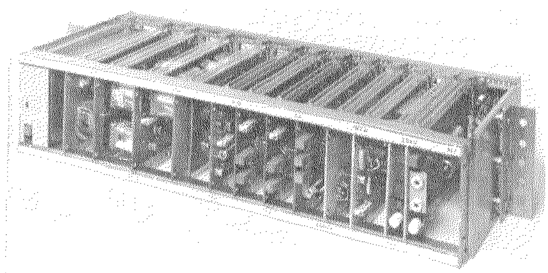

Figure 21—Modular equipment for interconnecting mobile radio systems and telephone networks.

150 voltamperes. Its weight is 44 kilograms (97 pounds) and its dimensions are 610 by 420 by 370 millimeters (24.1 by 16.6 by 14.6 inches).

Compagnie Générale de Métrologie
France

Italian Coaxial Cable System to Be Expanded—Following extensive favorable field trials, the Italian telecommunications administration has placed orders with us for terminal and line equipment for 12-megahertz coaxial cable systems handling 2700 channels over a pair of tubes. It will be installed along the main telephone trunk lines between Milan and Rome and between Ferrara and Pescara, about 1000 kilometers (625 miles).

Fabbrica Apparecchiature per Comunicazioni
Elettriche Standard
Italy

Milliwatt Test Set Restyled—The redesigned *74166A* milliwatt test set shown in Figure 22 has the same facilities in only half the volume. Push buttons replace the rotary switch and the meter is behind the panel to give more protection. Two models differ only in input connectors, *H* being fitted with British Post Office Number 1 coaxial plugs and *J* with *BNC* connectors. The sets provide for accurate terminated level measurements of +1 to −1 decibel referred to 1 milliwatt on 75-ohm unbalanced circuits at frequencies up to 30 megahertz. They also measure heater voltages between 5.6 and 7.0 volts.

A wired-in thermocouple is used with its heater impedance shunted down to 75 ohms and therefore the instrument cannot be used for measuring through levels. High measuring accuracy is possible from a built-in miniature Weston standard cell, against which the circuit is standardized by a series of simple switching and adjusting operations. A further feature is that the standardizing circuit can be used to send a direct-current power of 1 milliwatt into an external 75-ohm circuit for calibrating other apparatus.

Both models operate from three 1.5-volt dry cells. Their dimensions are 12¼ by 8 by 7⅛ inches (311 by 203 by 181 millimeters) with the lid in position. Weight is 12 pounds (5.5 kilograms).

Standard Telephones and Cables
United Kingdom

German Administration Adopts *GH2011* Data Modem—The *GH2011* data modem has been adopted by the German Federal Republic for handling data over the public telephone network. This modem permits use of any code required by the input and output devices, such as tape and card punches and readers, keyboards, computers, and magnetic-tape stores. Signaling rates of 1200 and 600 bits per second are provided.

The interface performance of these modems complies with the recommendations of the International Telegraph and Telephone Consultative Committee. They are available to telephone subscribers on a rental basis.

Standard Elektrik Lorenz
Germany

Signal Generator *919A*—The frequency range from 50 kilohertz to 50 megahertz is covered

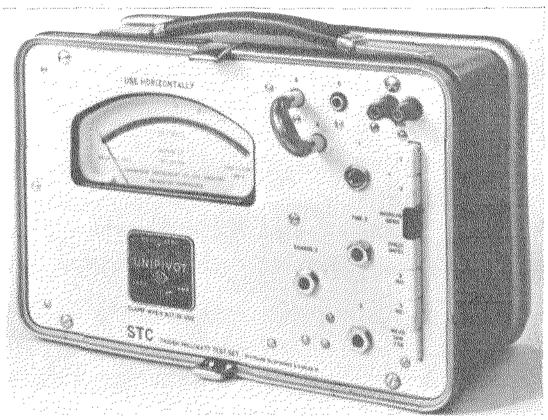

Figure 22—Milliwatt test set.

in 7 bands by the *919A* signal generator shown in Figure 23. An expanded band from 400 to 500 kilohertz is a convenience for adjusting intermediate-frequency amplifiers.

The stabilized oscillator requires no amplitude adjustment as frequency is varied. Step and continuously adjustable attenuators permit any output level between 100 millivolts and 1 microvolt into a 75-ohm load. An internal oscillator at 1000 hertz ±5 percent will modulate the radio-frequency output to 30 percent ±5 percent. Envelope distortion is less than 5 percent over the entire frequency range. An external source between 50 and 10 000 hertz at 5 volts root-mean-square for an input impedance of 10 kilohms will produce 30-percent modulation.

Power consumption is 60 voltamperes from the 50-hertz 115–250-volt mains. Cabinet dimensions are 495 by 250 by 310 millimeters (19.5 by 9.9 by 12.2 inches) and weighs 17.5 kilograms (38.5 pounds). For rack mounting the dimensions and weight are slightly smaller.

Compagnie Générale de Métrologie
France

London Borough Mobile Radiotelephone System—A London Borough has installed a radiotelephone base station and 34 vehicle stations to improve communication in its public works department.

The frequency-modulation base station delivers 25 watts to a 75-ohm load measured at the extremes of the band with a spurious output less than 2.5 microwatts. The frequency stability is ±1.5 kilohertz for a temperature variation between −10 and +40 degrees Celsius and a ±10-percent change in power supply voltage. The solid-state double-superheterodyne receiver incorporates muting during the absence of a signal.

Mobile station transmitters provide 10 watts into a 50-ohm load at the upper end of the band. For the conditions noted above, frequency stability is ±2 kilohertz. Solid-state muted double-superheterodyne receivers are used in the vehicles also. Operation is from the 12-volt car battery with a drain of 6 amperes during transmit. Mobile sets normally intercommunicate through the base station but talk-through facilities between vehicles are available for emergencies. All receivers and transmitters are crystal controlled at the chosen operating frequency in the band from 71 to 175 megahertz.

Standard Telephones and Cables
United Kingdom

Static Inverter—A static inverter has been developed for demonstration to the Italian and German telecommunication administrations. Two units may be operated in parallel with control and supervisory elements that ensure proper operation under all load conditions as well as shutdown of a unit in trouble. Hall generators with amplifiers assess the wattage and reactive power outputs. The electronic control circuits employ silicon semiconductor elements.

Standard Elektrik Lorenz
Germany

Control for Parallel-Operated Power Supplies—A power supply employing three parallelled rotary converters has been installed in a junction center of the German public telephone network. In addition, three parallelled rectifier equipments, each rated at 212 volts and 400 amperes, have been put in operation. Standard

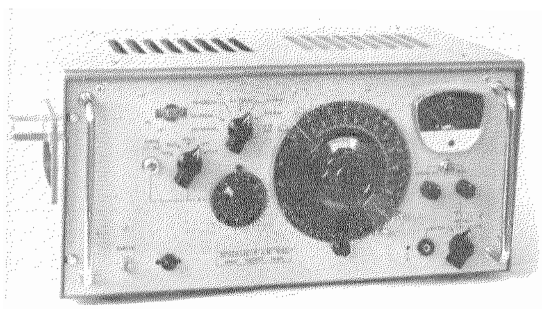

Figure 23—Signal generator covering 50 kilohertz to 50 megahertz.

Elektrik Lorenz supplied the thyristor sets for the rectifiers as well as the electronic controls and regulators required for parallel operation of the three converters and three rectifiers.

Standard Elektrik Lorenz
Germany

Television Test Pattern Generator *266A*—Test signals for both 625- and 819-line television receivers are generated in the very- and ultra-high-frequency television bands used in Europe by the *266A* test set shown in Figure 24.

A carrier frequency accuracy of 2×10^{-4} is obtained in the very-high-frequency band with crystal-controlled oscillators. The output level is 10 millivolts into a 75-ohm load and can be continuously adjusted from 0 to −60 decibels. The sound level is 10 decibels below the picture level.

The rotary switch at the lower left permits any one of 11 interchangeable tuning straps to be selected for the very-high-frequency channels. In the 12th position it places the ultra-high-frequency system in operation, which is controlled by the continuous tuning dial at the lower right. Here the sound and picture carriers are obtained by adding crystal-controlled frequencies to a continuously adjustable frequency. The picture carrier frequency is indicated directly on the control dial.

For 819 lines a 20 475-hertz oscillator and for 625 lines a 15 625-hertz oscillator, both adjustable over ±1 percent, and the 50-hertz mains frequency produce line and picture synchronizing signals together with a controllable number of pulses and square waves to form horizontal and vertical bars to test scanning linearity.

A definition oscillator, the frequency of which can be controlled from 3.5 to 10 megahertz by a calibrated dial, permits the bandwidth of the receiver to be measured. This video signal modulates the picture carrier either positively or negatively as required and is adjustable for up to 70-percent modulation. This signal is available at a video outlet.

The sound carrier may be internally modulated at 1000 hertz or by an external source of 1.5 volts between 50 and 10 000 hertz. For frequency modulation, the audio-frequency signal modulates the definition oscillator, which is set at 5.5 megahertz as a subcarrier, over an excursion of ±15 kilohertz.

The test set is operated from the 50-hertz power mains at any of several voltages between 115 and 240. It takes 55 voltamperes. Dimensions are 407 by 225 by 217 millimeters (16.1 by 8.9 by 8.6 inches) and the weight is 11 kilograms (24.2 pounds).

Compagnie Générale de Métrologie
France

Air Navigation Equipment for Bulgaria—Ground navigation aids, ground-to-air communication equipment, airborne calibration sets, omnidirectional radio range beacons, and direction finders are being built for installation in Bulgaria.

Very-high-frequency direction finders of the Doppler type *DDF1* will be installed on the Black Sea coast at Burgas and Varna. The very-high-frequency omnidirectional radio range beacons (VOR) will be the 200-watt *BQ3A* type, which provides a limaçon rotating pattern at 30 hertz from 12 dipole slots with mechanical commutation. The operating frequency can be in the range from 112 to 118 megahertz.

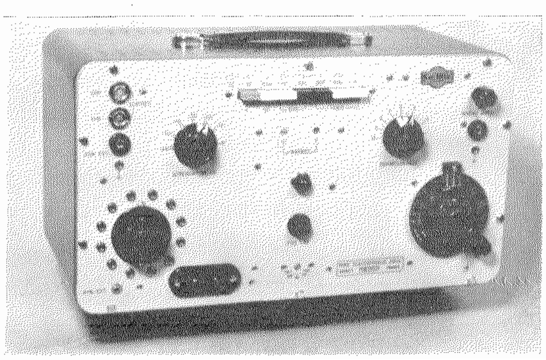

Figure 24—Test set for 625- and 819-line European television receivers.

The solid-state calibration equipment is completely demountable and is for use in aircraft to check the ground beacons for the very-high-frequency omnidirectional range and for an instrument low-approach system previously installed in Sofia. The calibration equipment can operate a pen recorder which will simultaneously record 5 channels of information.

The very-high-frequency communication equipment consists of a *DV8B* 60-watt transmitter and a *RX2SA* receiver.

The Bulgarian maintenance personnel will be trained in the United Kingdom.

Standard Telephones and Cables
United Kingdom

Fire Alarm System for Large Ferries—The Bremerhaven Seebeck dockyard has launched two large ferries capable of carrying 235 passengers and 200 cars for service between Europe and England. An alarm system that responds quickly to an outbreak of fire and indicates the location of the fire protects both ships. It was developed and installed in cooperation with the Selbsttätige Feuerlöschgesellschaft (Automatic Fire Extinguishing Company) of Hamburg.

Standard Elektrik Lorenz
Germany

Electronic Thermostat for Electric Appliances—To improve the performance of thermostats used in washing machines and similar appliances, an electronic thermostat uses a thermistor in a resistance bridge as the temperature sensing element and a transistor amplifier to provide operation within ±1 degree Celsius instead of ±5 degrees for former designs. Several temperature steps can be provided for.

Standard Elektrik Lorenz
Germany

Television Receiver Alignment Wobulator *241A*—The *241A* wobulator produces any frequency between 470 and 870 megahertz with frequency modulation continuously adjustable to ±15 megahertz around the center frequency.

The operating center frequency, indicated on the large dial of Figure 25, is produced by a Lecher wire oscillator. It is frequency modulated by a capacitor vibrating at 50 hertz. An automatic control of the oscillator supply voltage reduces parasitic amplitude modulation and maintains a constant output level independent of frequency. For a total frequency excursion of 20 megahertz, the parasitic amplitude modulation does not exceed 0.5 decibel, the parasitic frequency modulation is less than 0.002 of the frequency indicated on the dial, and the error is less than 5 percent.

For models with an attenuator, the output level to a matched load is adjustable from 1 to 250 millivolts root-mean-square, calibrated from 0 to −40 decibels with a zero level of 100 millivolts. For models without this attenuator, the output is 1 volt root-mean-square and a 6-decibel pad can be inserted by pushing a button.

Output impedance may be 50 or 75 ohms. An outlet is provided for a marker signal. For the horizontal deflection of the oscilloscope, a 10-volt peak-to-peak 50-hertz output is available. Phase adjustment over about 120 degrees is provided.

The equipment operates from the 115–220-volt 50-hertz mains. It is available in a cabinet 518

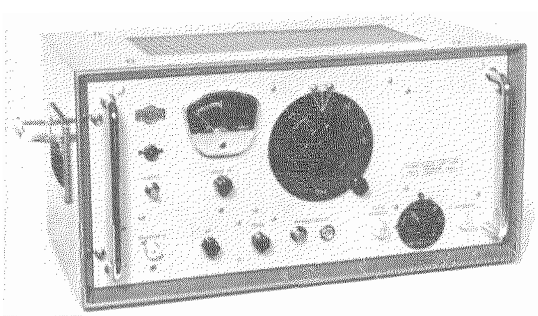

Figure 25—Television receiver alignment wobulator *241A*.

by 261 by 320 millimeters (20.5 by 10.3 by 12.6 inches) or in a rack 483 by 221.5 (5 units) by 275 millimeters (19.1 by 8.7 by 10.9 inches) and weighs 17.5 kilograms (38.5 pounds).

Compagnie Générale de Métrologie
France

Greek Telephone Network Expansion—The Hellenic Telecommunication Organization has ordered crossbar switching equipment to serve 68 000 lines. This expansion of service will be made in the automatic local exchanges for Keramikos, Patissia, Daphne, Nikea and Salonica.

This order permits us to continue our leading role in expanding and modernizing the Greek toll dialing service and long-haul communications network.

Standard Elektrik Lorenz
Germany

United States Patents Issued to International Telephone and Telegraph System; May 1965–July 1965

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R. T. Adams and K. J. Staller, ITT Federal Laboratories, Diversity Communication System, 3 195 048.

H. H. Adelaar, Bell Telephone Manufacturing Company (Antwerp), Resonant Transfer Time Division Multiplex System Utilizing Negative Impedance Amplification Means, 3 187 100.

H. H. Adelaar, F. Clemens, and J. L. Masure, Bell Telephone Manufacturing Company (Antwerp), Master-Slave Memory Controlled Switching Among a Plurality of TDM Highways, 3 187 099.

B. Alexander and G. A. Deschamps, ITT Federal Laboratories, Velocity and Position Computer, 3 194 948.

F. J. Altman and A. T. Brown, 3rd, ITT Federal Laboratories, Radio Diversity Receiving System With Automatic Phase Control, 3 195 049.

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R. C. Bassett, ITT Federal Laboratories, Perspective Vectorcardioscope, 3 186 403.

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W. J. Frankton, Standard Telephones and Cables (London), Carrier Current Communication Systems Incorporating Repeaters, 3 189 694.

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G. G. Gassmann, C. Lorenz (Stuttgart), Circuit for Producing a Directive Voltage as a Function of the Phase Difference Between Two A.C. Voltages, 3 194 971.

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Master-Slave Memory Controlled Switching Among a Plurality of TDM Highways

3 187 099

H. H. Adelaar, F. Clemens, and J. L. Masure

The patent discloses a time-division telephone multiplex system in which a plurality of lines may be selectively connected with any one of a plurality of highways, the lines being sequentially connected to the highways selectively during recurring time intervals. The system is provided with a slave control equipment associated with each highway to memorize the time positions when certain line gates are unblocked. A common equipment supplies a memory associated with a plurality of the slave control equipments to control the routing and release of calls over the highways.

Pulse-Count Control Circuit Wherein the Input is Sampled and Inhibited Upon Input Exceeding Predetermined Frequency

3 187 202

T. Casc

In certain systems of pulse communication, such as air-to-air communication systems, it is necessary to limit the times that pulses may be received in order to prevent overload of transponder modulators. This invention provides a pulse-count control circuit to control receipt of pulses by sampling the received pulses and setting a reference pulse-count rate. If this rate is exceeded, an inhibit gate is operated for a predetermined period to limit the pulses received to the reference count rate.

Contact Positioning Structure for a Resilient Connector Insulator

3 184 701

R. H. Ellis

In connectors, particularly for small-size connectors in which the connector insulators are of resilient material which facilitates the construction of multiple-contact connectors, there is provided a rigid insulating body molded into the insulator to form a stop shoulder limiting the depth of insertion of the contact pins into the receiving bore.

Carrier Current Communication Systems Incorporating Repeaters

3 189 694

W. J. Frankton

A system is described for testing the response of tandem connected repeaters of a submarine cable, for example, in which the successive repeaters each respond to two different test frequencies. At the repeater being tested, one of the test frequencies is multiplied and used to modulate the other test frequency. One modulation sideband is selected and returned to the attended station from which the test signals are sent.

Radio Transmitter Overload Protection System

3 182 260

L. J. Heaton-Armstrong

This is an overload protection system for a radio transmitter designed to provide a rapid cutoff to avoid difficulty found in anode current and anode heat type cutoff systems. It provides a cutoff relay responsive to the difference between a current proportional to the power output of the transmitter and a current proportional to the anode-cathode current in a tube of the transmitter.

Telephone Subscriber Sets Using Amplifiers

3 197 570

E. H. Lambourn

A transistorized telephone subset is described in which a loudspeaker is energized by an incoming ringing signal which serves to bias the transistor to operate as a class-A amplifier, but on removal of the handset the switch hook connects the transistor to amplify outgoing speech.

Gyromagnetic Resonance Waveguide Isolator With Ferrite Strips and Overlapping Ferrite Bar

3 197 718

V. Sigrist and A. Muller

This patent describes a gyromatic resonance waveguide isolator using ferrite, particularly adapted for use at high frequencies in which the conventional ferrite gyrators are difficult to manufacture and to position sufficiently close to each other. The ferrite gyrators are made in longitudinal strips in the walls intermediate the side walls of the rectangular waveguide, and a third ferrite strip is positioned within the waveguide overlapping these strips.

Principal ITT System Products

Communication Equipment and Systems

automatic telephone and telegraph central office switching systems...private telephone and telegraph exchanges—PABX and PAX, electromechanical and electronic...carrier systems: telephone, telegraph, power-line, radio multiplex...long-distance dialing and signaling equipment...automatic message accounting and ticketing equipment...switchboards: manual (local, toll), dial-assistance...test boards and desk...telephones: desk, wall, pay-station, special-environment, field sets...automatic answering and recording equipment...microwave radio systems: line-of-sight, over-the-horizon ...teleprinters and facsimile equipment... broadcast transmitters: AM, FM, TV...studio equipment...point-to-point radio communication...mobile communication: air, ground, marine, portable...closed-circuit television: industrial, aircraft, nuclear radiation...slow-scan television...intercommunication, paging, and public-address systems...submarine cable systems...coaxial cable systems

Data Handling and Transmission

data storage, transmission, display...data-link systems...railway and power control and signaling systems...information-processing and document-handling systems...analog-digital converters...alarm and signaling systems... telemetering

Navigation and Radar

electronic navigation...radar: ground and airborne...simulators: aircraft, radar...antisubmarine warfare systems...distance-measuring and bearing systems: Tacan, DMET, Vortac, Loran...Instrument Landing Systems (ILS) ...air-traffic-control systems...direction finders: aircraft, marine...altimeters..flight systems

Space Equipment and Systems

simulators: missile...missile fuzing, launching, guidance, tracking, recording, and control systems...missile-range control and instrumentation...electronic countermeasures...power systems: ground-support, aircraft, spacecraft, missile...ground and environmental test equipment...programmers, automatic...infrared detection and guidance equipment...global and space communication, control, and data systems ...system management: worldwide, local... ground transportable satellite tracking stations

Commercial/Industrial Equipment and Systems

inverters: static, high-power...power-supply systems...mail-handling systems...pneumatic tube systems...instruments: test, measuring ...oscilloscopes: large-screen, bar-graph...vibration test equipment...pumps: centrifugal, circulating (for domestic and industrial heating)...industrial heating and cooling equipment...automatic controls, valves, instruments, and accessories...nuclear instrumentation

Components and Materials

power rectifiers: selenium, silicon...transistors...diodes: signal, zener, parametric, tunnel ...semiconductor materials: germanium, silicon, gallium arsenide...picture tubes...tubes: receiving, transmitting, rectifier, thyratron, image, storage, microwave, klystron, magnetron, traveling-wave...capacitors: paper, metalized paper, electrolytic, mica, plastic film, tantalum...ferrites...magnetic cores...relays: telephone, industrial, vacuum...switches: telephone (including crossbar), industrial...magnetic counters...magnetic amplifiers and systems...resistors...varistors, thermistors, Silistor devices...quartz crystals...filters: mechanical, quartz, optical...circuits: printed, thin-film, integrated...hermetic seals...photocells, photomultipliers, infrared detectors... antennas...motors: subfractional, fractional, integral...connectors: standard, miniature, microminiature...speakers and turntables

Cable and Wire Products

multiconductor telephone cable...telephone wire: bridle, distribution, drop...switchboard and terminating cable...telephone cords... submarine cable and repeaters...coaxial cable: air and solid dielectric...waveguides...aircraft cable...power cable...domestic cord sets ...fuses and wiring devices...wire, general-purpose

Consumer Products

television and radio receivers...high-fidelity phonographs and equipment...tape recorders ...microphones and loudspeakers...refrigerators and freezers...air conditioners...hearing aids...home intercommunication equipment... electrical housewares

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