tions inevitably will appear as well and they are a very important factor in filter-parameter instability.

In the studies available, the effect of detuning is taken into account for the special case of an unloaded balanced twin-$T$ bridge. Such an analysis is far from adequate in every case for the purposes of optimum engineering design: initially, as has been shown elsewhere, it is precisely the unbalanced bridge that provides the maximum filter $Q$ in the absence of regeneration; second, in actual circuits (especially those using transistors) it is very frequently impermissible to idealize the operating conditions for a twin-$T$ bridge.

The present study considered the effect of detunings on filter properties in the general case of a loaded unbalanced twin-$T$ bridge. A statistical approach to random deviations makes it possible to replace semiqualitative considerations with rigorous mathematical methods from the theory of probability in the analysis of instabilities and to furnish a basis in particular for recommendations as to the magnitude of tolerances in the assembly of a twin-$T$ bridge.

**Differentiation of Random Processes,** E. Ya. Ryskin, pp. 73–76.

**Number 7, July 1962**

**Analysis of a New Circuit for a Square Wave Pulse Generator,** A. S. Vladimirov, pp. 17–28.

Generators shaping symmetrical oscillations of accurate square waves are frequently required in the design and development of various radio devices. A number of methods for shaping such oscillations are known, although, as a rule, the circuits of such generators are relatively complex and cumbersome. In particular, a frequently used method for forming square wave pulses from a sinusoidal voltage requires the use of shaping and amplifying stages. The difficulties incurred in using such a circuit are greatly increased when low frequency (several cps) pulses of high quality are required. When used by itself, a multivibrator cannot form accurate square wave pulses without spikes. As a result, such generators must also contain shaping and amplifying stages.

Ideally shaped square wave pulses, obtainable under easily fulfilled conditions, may be produced by triggered circuits having two stable equilibrium states in which constant plate voltage for tubes is insured. However, the triggered circuit is not a generator and consequently requires external control pulses. A new circuit for a one-tube square wave pulse generator in which the forementioned triggered circuit is used is discussed.

**Application of Laguerre Functions to Parametric Coding of Speech Signals,** V. I. Kulya, pp. 34–42.

An investigation is made of the possibility of constructing a system for compressing the speech spectrum by using the principle of transmitting a limited number of parametized signals proportional to the coefficients of a Laguerre function series expansion of the instantaneous autocorrelation function of the speech signal. The use of Laguerre functions makes it possible to construct an analyzer and a synthesizer for a synthetic-telephony channel using RC elements. A special feature of the method is that the accuracy of approximation to the envelope of the instantaneous spectrum varies along the frequency scale proportionally with the properties of auditory reception of speech.

## TELECOMMUNICATIONS AND RADIO ENGINEERING. PART 2. RADIO ENGINEERING (*Radiotekhnika*). Published by American Institute of Electrical Engineers in conjunction with Scripta Technica, Inc., New York

**Number 1, January 1962**

**Reaction of an Inertial System to Zero Beats,** M. I. Dorman, pp. 11–20.

The peculiarities of the passage of zero beats through linear inertial systems are analyzed. Time and spectral relations are presented which characterize the output process under periodic frequency modulation.

Sequential conversion of the spectrum by the use of heterodyning is often found in different apparatus in which frequency modulation is used. The process of heterodyning an FM signal when its center frequency is close to the heterodyne frequency is accompanied by the formation of so-called "zero beats," which are a complex oscillation containing difference and sum components. If the mixer load is sufficiently inert and its passband is in the comparatively low frequency range, then only the difference frequency component need be considered. At those instants when the difference frequency passes through the zero value, the shape of the oscillations at the mixer output differs radically from a sinusoid and a more or less wide domain distorted by zero beats is formed. These distortions also determine the unique nature of the output process.

In the sequel we shall analyze the two simplest and, moreover, the most characteristic selective systems; a single-section, low-frequency RC filter and a single resonant circuit, whose impulse responses equal, respectively:

$$g(t) = \alpha \cdot e^{-\alpha t} \qquad \alpha = 1/\mathrm{RC} \tag{1}$$

and

$$g(t) = \Omega_p e^{-\alpha t} \sin \Omega_p t \tag{2}$$

where $\Omega_p$ is the resonant frequency; $\alpha = \Omega_p/2Q;$ and $Q$ is the quality factor.

**Delay Lines with Distributed Constants for Nanosecond Pulse Circuits,** V. A. Solov'yev, pp. 21–31.

Delay lines are widely used in pulse technique and particularly in electronic automation and computer technique. The tendency to raise the efficiency of electronic automation and computer technique substantially leads at present to systems operating with extremely narrow pulses at the greatest repetition frequency. In this connection, the following requirements are imposed on delay lines: 1) high resolving power; 2) low wave resistance; 3) miniature construction; and 4) simplicity of construction facilitating the organization of mass production.

Only delay lines with distributed constants satisfy the requirement of high resolving power or low rise time of the transient characteristic $\tau_l$ to the required degree. The great advantage of these lines, as compared with delay lines with lumped constants, is the absence of separate structural elements of inductance cores and capacitors. This is particularly important for small-scale constructions, which are based on monolithic elements in order to simplify the technology and give the greatest possibility for diminution of the volume.

Existing constructions of delay lines with distributed constants do not satisfy the requirements of prospective high-frequency circuits [pulse duration $\tau_p = (25\text{–}100) \times 10^{-9}$ sec]. They do not permit wave resistances $\rho < 300$ ohm to be obtained, have large dimensions (a volume $v > 10$ cm$^3$ for a delay of $T_d \geqslant 6\tau_l$), and are structurally complex; the simplest among them have poor resolving power (rise time of $\tau > 50 \times 10^{-9}$ sec).

In this connection, the delay lines with distributed constants considered here were developed and they permit structurally simple lines to be obtained with $\tau_l \geqslant 10 \times 10^{-9}$ sec, $\rho \geqslant 30$ ohm, magnitude of the delay $T_d \geqslant 6\tau$, in a volume $v = 0.7\text{–}1.4$ cm$^3$. For the computation it is proposed to use a method recommended in practice, which is based on time relations. On the basis of a computation of the phase-frequency distortions which are characteristic for helical lines, expressions are derived for the computation of the rise time of the transient characteristic. On the basis of an analysis of the line construction, a formula is derived for the computation of the distributed capacitance.

*Conclusion:* Use of the method of computation explained and the formulas derived in order to estimate the magnitude of the distributed capacitance and the rise time is highly recommended in practice. They guarantee the efficiency of constructing small-scale delay lines for high frequency circuits. Use of the new lines, with a multiconductor helix and twin reverse conductor, would permit obtaining low resistance ($\rho = 30\text{–}300$ ohm) lines which would surpass other lines in resolving power at a considerably reduced size.

**Design of Chebyshev Directional Couplers with Weak Coupling,** A. L. Fel'dshteyn and Ye. S. Zhavoronkova, pp. 39–50.

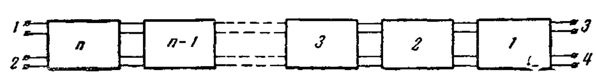



**Fig. 1**

A method is proposed for the synthesis of multielement directional couplers which possess optimum properties. Typical problems are tabulated for two to eleven elements. The technique of the computations is illustrated by examples.

The cascade connection of eight-terminal networks (Fig. 1) with a logical choice of its elements is a directional coupler; it delivers a prescribed part of the power from the main 1-3 line to arm 4. The undesired passage of power into arm 2 is usually limited to a certain admissible small quantity. In the general case the appropriate transmission coefficients $S_{14}$ and $S_{12}$ are frequency dependent; the optimum form of these dependences (in a certain approximate sense) should be achieved in computing directional couplers.

It is known that if the mentioned functions of the frequency are Chebyshev functions (isoextremal), then the required coupler parameters are realized for the least number of elements, i.e., in the most economical manner. Regrettably, information on this question in the literature is fragmentary; the work of foreign authors suffers from an unsystematic treatment. In the present work a method of computing multielement Chebyshev directional couplers with weak coupling is systematized on the basis of certain premises developed in previous work. Tables of the solutions of typical problems including a considerable part of practical inquiries are the essential result of this paper.

*Conclusion:* When using the tables of the appendix, the method explained guarantees a sufficiently simple and rapid engineering computation of Chebyshev directional couplers. We use the method for forward attenuation values of $10 \log (1/|S_{14}|^2) \geqslant$ 10 db.

The effective parameters of the directional coupler can be rather worse than the computed (intrinsic) parameters because of the imperfection of the wave loads, and also because of the interactions not taken into account between the coupling elements and errors in production.

The influence of mismatching the final loads is illustrated by the graph in Fig. 6. Evidently, realization of a minimum directivity greater than 40 db is very difficult for loads achievable in practice.

The second source of error is the frequency response of the coupling elements. Although it is usually small in comparison with factor $\exp i2(n - \alpha)\theta$, the induced error can become perceptible in a broad frequency band. Taking numerical account of this factor of error requires a separate investigation.

**Statistical Estimation of the Reliability of "Aging" Devices,** V. I. Siforov and G. B. Linkovskiy, pp. 62–67.

The statistical estimation of the reliability of "aging" devices on the basis of an approximation of the "danger" of failure by a polynomial of high degree is analyzed. A system of algebraic equations is obtained in the parameters of the "danger" of failure being estimated, which is solved by numerical methods. A construction is indicated for a confidence interval for the estimated "danger" of failure and the reliability of the device.

*Conclusion:* Let us note that the necessity for using numerical methods in the individual steps of the statistical estimate of the reliability is caused by the specificity of the "aging" process of the devices. The statistical estimate mentioned should be made separately for each class of devices. Let us describe briefly the course of the computations on the basis of the analyses carried out:

1) The primary quantities $t_1, t_2, \ldots, t_n$ are taken from experimental data for each kind of device separately, where $n \geqslant 30$–$50$ is recommended.

2) We select $m$ (usually $m \approx 12$–$18$).

3) Solving the system numerically, we find the parameters $\lambda_k{}^*$ ($k = 0, 1, \ldots, m$), and on this basis we find the empirical function $a^*(t)$ of the "danger" of failure.

4) The confidence interval for the actual "danger" of failure $a(t)$ is found.

5) We find the concrete expression for the empirical reliability of the device $P^*(t)$.

6) The confidence interval for the actual reliability of the device $P(t)$ is found.

We note that such a method for statistically estimating the reliability of devices has great generality and is applicable for any device or machine and also in problems of the "aging" of biological systems.

## Number 2, February 1962

**Noise Immunity of a Receiver Generating a Set of Measurements of the Autocorrelation Function,** N. F. Vollerner, N. G. Gatkin, and M. I. Karnovskiy, pp. 1–7.

The noise immunity of a correlation receiver of pulse signals is analyzed under the condition that the signal at the output is generated by a set of measurements of the running autocorrelation function taken with definite weighting factors. Values of the weighting factors which would guarantee the greatest noise immunity are determined.

**Comparison of Methods of Receiving Discrete Signals,** B. A. Varshaver, pp. 8–12.

On the basis of the theory of potential noise immunity and the theory of information, the information transmission rate in a communication channel is compared in cases when the reception of the code combinations is accomplished "as a unit" and by elements in the appropriate ideal Kotel'nikov receiver. The comparison is made on the example of a 4-digit correcting code with the base 3.

Block reception of code combinations permits a higher transmission rate to be obtained if other conditions are equal and the restrictions on the noise immunity are identical, or, correspondingly, greater noise immunity can be achieved if the rates are equal. However, a considerably more complex apparatus is required for the reception of block code combinations than is needed for reception by elements. Consequently, the question of the usefulness of using block reception should be resolved while taking into account whether the gain in the transmission rate or in the noise immunity during block reception compensates for the complication of the apparatus involved in this kind of reception.

The problem is posed herein of comparing the information transmission rate in a communication channel on the basis of the theories of potential noise immunity and information for two cases: a) the code combination is received by an ideal Kotel'nikov receiver "as a unit"; b) the code combination is received by an ideal Kotel'nikov receiver in elements.

The general investigation of the range of questions referring to the problem under consideration is very complicated. No study is known to the author in which an attempt has been made to compare numerically both kinds of receptions in the customary formulation of the problem.

An example is given of a comparison according to the plan mentioned by using a 4-digit code ($n = 4$) with the base 3 ($m = 3$) correcting a single error ($\Delta = 1$) for the message transmission. All the signals represented by the code combinations are assumed to be equally likely. Results obtained by the author in other papers are used in the comparison.

**Distribution of Overshoots and Maxima of Fluctuations,** V. I. Tikhonov and Ye. I. Kulikov, pp. 12–21.

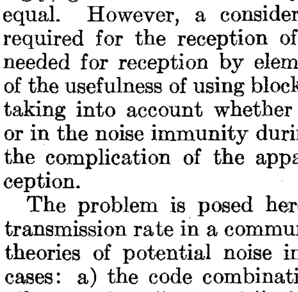



Fig. 1

A method of producing oscillograms of the intrinsic fluctuations at the output of a radio receiver is described briefly. A statistical treatment of the oscillograms is mentioned which permits a sample function distribution to be obtained in terms of the number of overshoots at various levels as well as the distribution of the absolute maxima for sample functions of various durations. An approximate method is mentioned for obtaining the last distribution by theoretical means.

An attempt is made herein to analyze two problems: 1) to find the distribution of the sample function of noise of fixed duration as a function of the number of overshoots; 2) to establish the distribution of the absolute maximum in the sample function of noise of fixed duration.

Let us explain the content of both problems and mention their practical value. The substance of the first problem is the following. We assume that there is sufficiently large number $N$ of sample functions (photographs) of stationary noise of fixed duration $T$ which have been obtained under invariant external conditions (Fig. 1). Let us take some constant level $H$ and compute the number of overshoots (the number of crossings of the level $H$ from below upward) in the first, second, etc., sample function. Let $m$ denote the number of sample functions having $n$ crossings of the level $H$. Then the quantity $p = m/N$ will characterize the distribution of sample functions of fixed duration by means of the number of overshoots.

Such a problem is encountered in the theory of signal detection in a noise background. As is known, the threshold of detection is determined by a given false alarm probability when detecting signals in a noise background by means of the Neyman-Pearson criterion. Knowledge of the distribution of the sample functions in terms of number of overshoots permits the determination of this threshold and yields a graphic clarification of the false alarm probability itself.

Let us note that particular cases of the problem mentioned were analyzed by a number of authors. From these works it follows that the theoretical solution of the problem is difficult. Consequently, in the first stage an experimental investigation was undertaken of the distribution of sample functions of fixed duration by means of the number of overshoots. The method of the experimental investigation is mentioned briefly in Sec. 2 and the results obtained are presented.

The meaning of the second problem is the following. Let there be the sample function of a stationary fluctuating process $\xi(t)$ with known characteristics within the finite time interval $[0,T]$. This sample function has a greatest value which we shall denote by $H_m$ (Fig. 1). We are required to find the one-dimensional probability density for $H_m$ which will characterize the distribution of the greatest values in different sample functions of the fluctuations of given duration being considered.

In addition to its purely physical value, this problem is also of practical interest. The method of the maximum of the equal-likelihood function is often used in estimating the unknown parameters of a signal received jointly with fluctuating noise. That value of the parameters is taken as the true value for which the equal-likelihood function has an absolute maximum in the prescribed time interval. However, for small signal-to-noise ratios, this absolute maximum can be due to noise overshoots and can be removed considerably from the true value of the parameter. For such signal-to-noise ratios, the method of the maximum of the equal-likelihood function yields an inadmissible error and loses its value.

**Predistortion and Filtering in a Channel with Variable Parameters,** V. I. Kulya, pp. 21–27.

A formula is derived for the mean-square error caused by the transmission of a signal with noise through a channel with random changes of the parameters. It is shown that in the general case the expressions for the transfer coefficients of optimum linear filters which minimize the power of the error satisfy a nonlinear integral equation with a nonsymmetric kernel. Certain special cases are discussed.

The purpose of the present paper is to discuss questions associated with the formulation and solution of problems of finding optimum linear operations on a signal during transmission over a channel whose system function $H(-i\omega, t)$ depends on time.

*Conclusion:* The conclusions obtained in Sec. 3 are substantially a generalization of the theory of optimum linear filtering of a stationary, random signal in the presence of noise when the signal passes through a linear channel with variable parameters. Here, for simple filtering (without predistortion), formulas are obtained in closed form which outwardly recall the analogous known results for channels with constant parameters. The details of the computations consist only of operations with the two-dimensional spectral function of the channel $G(-i\lambda, -i\omega)$. This latter must be known a priori for each specific problem.

The results of Sec. 4 have been obtained as a consequence of the material of Sec. 3 under the assumption that the channel parameters are slowly varying functions of time. Formulas given yield the complete solution of the problem of predistortion and filtering and are easily applicable to the specific case of transmission over a two-path channel with random delay. In this case $H(-i\omega, \zeta) = (1 + e^{-i\omega\zeta})/2$. Formula (19) yields $H_0(-i\omega) = [1 + \varphi(-i\omega)]/2$, where $\varphi(-i\omega)$ is the characteristic function of the random variable $\zeta$. We find from (21):

$$G^2(i\omega) = 1 + \int_0^\infty \cos\omega\zeta \, dF(\zeta) = 1 + |\varphi(-i\omega)|$$

**Design of a Multistage Frequency Multiplier,** M. Ye. Zhabotinskiy and Yu. L. Sverdlov, pp. 28–39.

A method and results are given for the design of an experimental multistage frequency multiplier consisting of multiplier stages of a new type which guarantee a very small side frequency content in the output signal spectrum.

One of the fundamental restrictions imposed on multistage frequency multipliers is that the output of any spectral component, other than the required harmonic, must be absent or sufficiently small. However, conventional multistage multiplier circuits do not meet this requirement. The presence of secondary components in the spectrum of each multiplier stage seriously hampers the design of multistage amplifiers, since the computation of each stage cannot be carried out in isolation from the preceding one.

In this paper, by starting from the results of other research, a new multiplier circuit is analyzed which will guarantee a high degree of filtering of the secondary harmonics. A multistage amplifier composed of these units permits very slight side components to be obtained in the spectrum of the output frequency and it is easily amenable to engineering computation.

*Conclusion:* The multistage amplifier analyzed herein is intended to discriminate high harmonics of a fixed frequency $f_0$. The frequency deviation of a real generator from $f_0$, which is generated particularly by fluctuating processes, leads to an increase in the effective value of the side components in the output oscillation spectrum. However, as computations and experiment show, a relative instability of the master oscillator, on the order of $10^{-6}$–$10^{-7}$, does not noticeably affect the results. Conventional quartz oscillators, as were used in the experiment described, have such instability.

Let us consider the question of stability of the given circuit with respect to parasitic generation. Since all the multiplier units in the multistage multiplier are tuned to substantially different frequencies, the stability of the whole circuit is determined mainly by the stability of each unit. Computations carried out according to another paper show that the last unit, most favorable in this respect and tuned to a very high frequency, is stable up to a frequency of 100–150 Mc (for a tube with $C_{ga} \sim 2.10^{-2}$ pf and $S_0 \sim 5$ ma/v). Experiment confirms the correctness of this deduction.

**Investigation of the Locking Operation of an Automatic Tracking Circuit,** S. V. Pervachev, pp. 48–53.

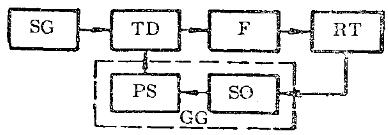



Fig. 1

The locking region of an automatic tracker is analyzed. On the basis of the generality of the mathematical apparatus describing the behavior of an automatic tracker and an automatic CW phase control system, there is determined the dependence of the locking band of an automatic tracker with an integrating filter on the system parameters. Tracking automatic tracker can be used to discriminate a periodic pulse signal in an interference background in pulse radio-communication lines. The block diagram of such an automatic selector is presented in Fig. 1.

A comparison of the time position of pulses presented from the pulse signal generator (SG) and the gate generator (GG) occurs in the time discriminator (TD). The gate generator consists of a sine wave oscillator (SO) and a pulse shaping circuit (PS). If the signals and the gating pulses overlap in time but their centers do not coincide, then a voltage appears at the output of the time discriminator and filter (F), which affects the frequency and phase of the oscillator through the reactance tube (RT), so that the initial pulse mismatch is diminished. A synchronized operation is possible in the automatic selector, in which the initially different repetition rates of the signal and gating pulses become equal because of the aligning effect of the reactance tube. However, synchronized operation is established in the selector when the initial difference in the repetition rates and the inertia of the tracking system, determined by the filter, are sufficiently small. Otherwise, synchronization, i.e., locking, does not occur.

Let the locking band be understood to be the maximum initial difference in the repetition rates of the signal and gating pulses for which buildup of a synchronized operation is possible. Let us investigate the dependence of the magnitude of the locking band on the selector parameters. An automatic tracking circuit is described by a differential equation which is analogous to the equation for an automatic frequency control system of CW signals. This permits the use of a method of analysis developed by M. V. Kapranov to determine the locking band of an AFC system in the analysis of the automatic tracker. The locking band of the tracker is determined herein under the assumption that the filter F is an inertial loop with the transfer function $1/(1 + pT)$.

## Number 3, March 1962

**One Method of Reducing a Multipole Network to an Equivalent Four-Terminal Network,** A. A. Tyutin, pp. 7–17.

A method is examined of reducing a multipole network to an equivalent four-terminal network, based on the separation of a group of coordinates with the condition that the components of the given vector, corresponding to continuous loops (circuits) of the network, are equal to zero. Formulas are introduced for calculating the elements of the reduced matrix for the separation of two, three, and four coordinates with arbitrary indices that correspond to the reduction of the original circuit to a short-circuited four-terminal network and to a four-terminal transfer network for various theorems as the basis relative to the separate loops (circuits). Formulas are introduced for the calculation of certain physical parameters of a short-circuited four-terminal network and a way is shown to calculate these parameters for transfer four-terminal networks.

**Input Impedance of Stepped Transitions,** L. P. Yavich, pp. 20–24.

The problem of determining the input impedance of stepped transitions is solved. The advantage of a transition with a maximally flat frequency characteristic, as compared with Chebyshev characteristics, is demonstrated in relation to the constancy of the resistive and slightly reactive components. At the present time, two types of stepped transitions find greatest practical application: Chebyshev transitions and those with a maximally flat frequency characteristic. A comparative evaluation of these transitions is ordinarily based on bandwidth and indicates superiority of the Chebyshev transition. Nevertheless, in numerous instances there are imposed very severe requirements on the input impedance of the stepped transitions (stability of resistive components and a small value of the reactive). Under these conditions, the selection of the type of transition becomes more difficult. On these grounds, the aim of this paper is to compare the input impedance of transitions of the indicated types.

1) In stepped transitions where a resistive component near a constant value and a reactive component approaching zero are required, it is preferable to select transitions with a maximally flat frequency characteristic and the number of steps $n = 5$ or 6.

2) Similar requirements can be satisfied by Chebyshev step transitions under very strict mismatch tolerances. In this case, the broadband superiority of the Chebyshev transitions over the maximally flat frequency characteristic transitions are lost.

**Analysis of Two-Circuit Parametric Frequency Converters,** M. Ye. Movshovich, pp. 24–34.

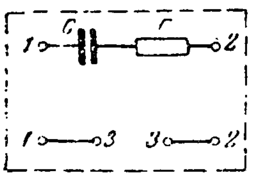


Fig. 12

The gain coefficient and the band and noise characteristics of a two-circuit parametric converter are examined from the viewpoint of general converter theory and the general theory of linear circuits.

The discussion shows that, for investigation of the described characteristics of two-circuit parametric converters, there is no necessity for any new criteria (conditions) in comparison with those employed in tube and transistor practice. The description indicates the possibility of creating a single engineering method for designing tube, diode, transistor, and parametric converters. It is understood that the analysis method and calculations discussed can be successfully applied to a higher frequency band, in which the diode properties can be determined from the circuit of Fig. 12.

**Threshold Sensitivity of Phase Radio Direction Finders,** V. V. Tsvetnov, pp. 44–57.

Using by way of example a number of widespread types of phase radio direction finders with a large base, a method of computing their threshold sensitivites in a background of noise of natural origin is given.

**Method of Analyzing Harmonic Frequency Dividers,** I. Kh. Rizkin, pp. 73–75.

It has been shown in another paper that the analysis of complex frequency divider circuits of harmonic type can reduce, in a number of cases, to the analysis of a certain equivalent system described by an equation of just the second order. The result formulated in the work mentioned referred to the stationary region in a system described by a differential equation of the form

$$K(D)\,Y = \mu\psi(Y, Y', \ldots, Y^{(m-1)}) + B_0 \cos\omega_{1n} t \qquad [1]$$

where $\psi$ is an analytic function (the notation used here and later agrees with that used in the first reference. Questions of the stability of the stationary division region were not analyzed.

The purpose of the present work is: 1) to ascertain in what cases stability of the stationary division region in the original system is a consequence of stability of an analogous region of the equivalent second-order system; 2) to establish the possibility of making a transition to the equivalent system in the case of a circuit with discontinuous (but not analytic, as in the paper noted), nonlinear characteristics.

## Number 4, April 1962

**Analog Division by the Linear Charging of a Capacitance,** L. Ya. Il'nitskiy, pp. 11–16.

As a rule the division operation is performed in analog computers either by multiplier circuits connected in the feedback loop of the resolver or by multiplier circuits with appropriate functional converters of the analog voltages. Hence, circuits of division units are mainly derivatives of circuits of multipliers. However, there are possibilities of producing division circuits by other means. Open-type division circuits, i.e., without feedback, are of great interest for practical work. Such circuits contain the minimum quantity of elements, are sufficiently reliable in operation, and are simple to repair.

The present paper is devoted to an open-loop pulse divider based on the linear charging of a capacitance. The analog voltages controlling the charging of the capacitor in the unit described form a sawtooth voltage whose amplitude is proportional to the frequency. In those cases where high accuracy in computing the quotient is not required, the circuit of such a unit can be very simple; the unit can be made sufficiently exact by slightly complicating the circuit.

**Spectral Analysis of a Pulse Train,** M. I. Finkel'shteyn, pp. 16–21.

The spectrum of a pulse train with symmetric envelope is determined and a method of estimating the degree to which the spectrum is concentrated around the "principal lobes" is given. The concept of the concentration coefficient is introduced which has the order of magnitude of the cube of the number of pulses for a cosine-squared envelope.

**Theory of a "Comb with a Tapered Tooth,"** Ye. G. Solov'yev, pp. 33–37.

Solution of the problem of electromagnetic wave propagation in waveguides of arbitrary cross section is usually associated with the selection of a coordinate system in which the coordinate surfaces coincide with the waveguide walls. In the presence of anisotropic conductivity of the walls it is moreover desirable that the corresponding coordinate lines should coincide with the lines of waveguide wall conductivity. In certain cases, such formulation of the problem leads to the necessity of using a non-orthogonal coordinate system in order to simplify satisfaction of the boundary conditions because of complication of the form of the wave equation. As will be shown below, the wave equation is successfully simplified for the considered problem so that the introduction of a non-orthogonal coordinate system yields a shorter and more illustrative means of solving the problem.

The equations and formulas are written in the rational, practical system of units. Attenuation in the walls is not taken into account. The time dependence is taken in the form $e^{-i\omega t}$

**Tuning of a Selective RC Amplifier with Amplified Selective Negative Feedback,** A. I. Belichenko, pp. 43–50.

Criteria for tuning a high-Q RC amplifier with amplified selective negative feedback and also phase relations in the amplifier circuits are discussed, and it is shown that it is necessary to eliminate positive feedback near the resonance frequency by correcting the phase shift caused by the input impedance of a parallel-T RC bridge.

**Fundamentals of Statistical Calculation of Radio Circuits,** I. M. Aynbinder, pp. 51–59.

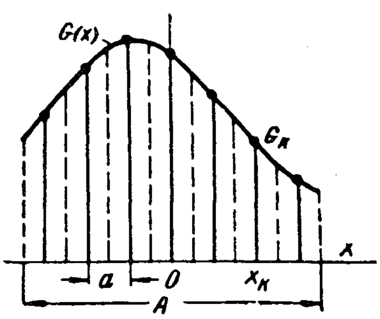
Engineering calculation of radio engineering circuits can completely satisfy modern design requirements for serial production only when, as a result of this calculation, it is possible at the design stage to determine with prescribed accuracy and reliability the principal statistical characteristics of those variations in circuit parameters which may be expected after the circuits have been placed in production.

The methods of calculation described in the literature and used at the present time do not satisfy the mentioned requirements; questions of statistical calculation are touched upon only partially in another paper. The present paper discusses the elements of calculation based on methods of mathematical statistics and probability theory methods with consideration of specific peculiarities of the circuits of radio engineering apparatus. The practical importance of using these methods at the stage of selecting and developing the circuit is due to the fact that determination of parameter variations and refinement of tolerances by experimental means, as is customary at the present time, is associated with considerable labor and expense in the preparation of large test batches and with carrying out a great volume of work in measurements of parameters of this circuit. Moreover, in the presence of a preliminary analysis of the expected spread in parameters of different versions of a circuit in the development stage, the most suitable can be selected without special expenditures.

## Number 5, May 1962

**Directivity Pattern of Linear Radiators,** N. A. Semenov, pp. 25–32.

Fig. 1




Most complex antenna systems with high directivity can be represented as groups of linear systems of radiators, each of which is characterized by a certain spread along its axis and a certain distribution of excitation (current, voltage) along the same line. Since the directivity pattern of the group in this case is the product of the patterns of the component linear systems, the attention given in antenna theory to the analysis of the directivity characteristics of radiators distributed arbitrarily along a line cannot seem unnatural.

In the present paper, the discussion revolves around linear systems excited by two methods:

a) A system consisting of a series of equidistant, identical, discrete radiators, each with arbitrary excitation amplitude and phase. The total number of emitters is $N$: the spacing between them $a = A/N$. The excitation of each of the antennas of the system is characterized by the complex coefficient $G_k$, which is the ratio between current, voltage or directivity of the field exciting the given antenna and the basic exciting force (e.g., the maximum excitation, or the excitation of the central antenna). In the following, we shall designate this sytem as discrete.

b) A continuous radiation system of length A, along which the distribution of the excitation is described by the normalized function $G(x)$.

Let us consider the discrete and continuous systems to be similar if the ratio $A/\lambda$ in both systems are equal; and let the modulus and phase of the function $G(x)$ in the intervals between the measurement points vary according to some given rule (Fig. 1).

The purpose of the present work is to determine the conditions under which the directivity patterns of similar [i.e., with an excitation satisfying condition (1)] discrete and continuous linear systems are identical in practice, and also to estimate the magnitude of a possible difference between their patterns.

The directivity pattern of a continuous system is determined by using a Fourier integral transform, and that of a discrete system is determined using a Fourier series. If the law of the variation of the excitation along the system is represented in the form of some simple function, then an integral transformation permits the general expression for the directivity pattern to be obtained rapidly and, possibly, the directivity coefficients as well. In this case, it is expedient to reduce the analysis of a discrete system to the analysis of a continuous system.

If the law of the excitation distribution is given numerically or in the form of a function not Fourier-integrable, then it is convenient to discretize the continuous radiator by replacing the continuous function $G(x)$ by a finite number of measurements $G_k$.

In both cases it is necessary to know how to replace one system by the other without introducing significant errors in the final result.

**Cross Modulation in Resonance Transistor Amplifiers,** M. Ye. Movshovich, pp. 71–78.

A rigorous theoretical analysis of the manner in which cross modulation arises when transistors are operated requires the consideration of a great number of factors and is highly complex. Consequently, in the present paper, only the fundamental relationships between the cross-modulation coefficient $k_{\text{nep}}$ and interference are determined, by an analysis of the simplest theoretical models of transistors; subsequently, the correctness of these relationships and their values for real transistors are evaluated experimentally. All the investigations are carried out for the case when there is no impedance in the common electrode circuit of the transistor which may cause feedback at the signal frequencies and no products arise due to nonlinearity of the input loop.

## Number 6, June 1962

**Diffraction by a Plate Located at the Earth in the "Line-of-Sight" Zone,** V. P. Peresada, pp. 23–27.

On the basis of research of V. A. Fok on diffraction at a spherical earth, an expression is derived (in the Kirchhoff approximation) for the diffraction pattern of the secondary radiation field of a plate located in the "line-of-sight" zone. Auxiliary graphs are plotted for computation of the attenuation function with consideration of curvature of the earth, and comparative calculations are performed for several patterns.

**Raising the Quality of a Transistorized Pulse Amplifier Stage,** T. M. Agakhanyan, pp. 31–36.

One of the fundamental characteristics of a pulse amplifier stage, whether of vacuum tubes or of transistors, is the stage quality. This is defined as the ratio of gain to the rise time of the output pulse. This ratio permits a determination of the magnitude of the expected pulse gain for a given rise time or of the pulse rise time for a given gain.

Raising the quality of a stage permits an increase in the gain or a decrease in the pulse rise time. Consequently, a great deal of attention has been paid to questions of raising the stage quality in amplifier engineering. Correcting two-terminal or four-terminal networks containing inductance coils are widely used in vacuum tube amplifiers in order to raise the quality. The stage quality can also be raised by using feedback.

The forementioned methods of raising the stage quality are also used in transistor engineering. However, the use of inductance coils and feedback does not exhaust the possibilities of raising the quality of a transistor stage. On the basis of an elementary amplifier stage, peculiarities of a transistor amplifier are examined from the viewpoint of obtaining the greatest possible quality.

*Conclusion:* The analysis of a transistor stage and its experimental investigation, carried out for transistors of P-15, P-401, and P-402 type, permit the following conclusions to be made:

1) The transistors possess sufficiently high quality. In considering their quality, modern drift transistors are as good as the best high-frequency vacuum tubes.

2) The transistor quality is defined by the product of the collector junction capacity and the base volume resistance and is independent of the magnitude of the time constant of the transfer constant of the base current $\tau_\beta$ or the emitter current $\tau_\alpha$. But the smaller the $\tau_\beta$, the larger the quality of the stage for given $R_c$ and $E_c$.

3) The quality of a transistor stage increases as the resistance in the collector circuit $R_c$ increases. This circumstance obliges the use of comparatively high-voltage sources to supply pulse amplifiers.

4) The quality of a transistor stage is usually considerably less than the quality of the transistor itself because of the influence of the internal resistance of the input signal source. One of the effective methods of raising the quality of the stage is the use of a matching transformer at the stage input.

**Suppressed-Carrier Double-Sideband Radio Communication,** Z. I. Model', V. N. Arzumanov, and I. A. Tsikin, pp. 41–53.

The disadvantages in principle of conventional amplitude modulation (AM) and the great advantages of suppressed-carrier radio communication became obvious about forty years ago. It was considered indisputable that suppressed-carrier double-sideband communication is much poorer than single sideband communication, since it occupies twice the bandwidth and requires recovery of the carrier in the receiver, not only in frequency, but also in phase. After a great deal of experience in single-sideband communication had been accumulated and all the fundamental problems of constructing single-sideband transmitters and receivers had been solved, many countries began converting all long-distance communication to the single sideband. In this connection a lively discussion arose in radio engineering literature (primarily American) as to which form of suppressed-carrier communication has greater prospects, double- (DSB) or single-sideband (SSB).

Keeping in mind the statements of double-sideband communication supporters (J. Costas et al.), let us examine the fundamental questions of constructing double-sideband transmitters and receivers, and let us also compare the indexes of double- and single-sideband communications.

*Conclusion:* Let us dwell briefly on certain questions of comparing SSB and DSB systems which were not touched on in the foregoing.

a) Width of the occupied frequency band. It is generally assumed that SSB permits doubling of the number of independent communication channels as compared with AM and DSB. Actually, the advantage of SSB is not this great, since the bandwidth carried in the communication channel is determined not only by the useful signal but also by the protective zone $\Delta f$. Thus, for example, if it is assumed, starting from frequency instability, that $\Delta f = 2 \cdot 10^{-5} f$, then the single-sideband gain is $(2F + \Delta f)/(F + \Delta f) = 1.6$ for $f = 1000$ Mc and an $F = 3000$ kc modulating frequency band. It is remarked in the literature that in a number of cases in practice inadequate suppression of the sideband in a SSB transmitter eliminates the possibility of its use by another single-sideband transmitter.

b) Multichannel operation. This kind of operation imposes very rigid demands on the linearity of the transmitting and receiving channels. Hence, the forementioned advantage of DSB transmitters and receivers is not realized successfully.

c) Use of AM equipment in the transition period. Supporters of the DSB system state that the existing AM equipment is much more easily adapted to DSB communication than to SSB. It is impossible to make a general conclusion on this question since conversion to SSB is simpler in a number of cases in practice.

It follows from the foregoing that the DSB system is not competitive with SSB in the majority of communication areas. Apparently DSB is more promising: in peripheral service characterized by low transmitter power, single-channel operation, a low number of modes, not very high frequency stability, etc.; in television. The DSB system is also competitive with SSB in those cases when single-sideband communication without a pilot signal is impossible (reduced frequency stability, influence of the Doppler effect, necessity of conserving the shape of the modulating voltage, etc.).

**Use of Semiconductor Diodes as the Element of a Timing Loop,** A. M. Mishin, pp. 53–58.

Results of certain investigations on the determination of new possibilities of connecting semiconductor diodes in pulse and time-setting units are presented.

## Number 7, July 1962

**Method of Signal Extrapolation by Power Series,** A. A. Gorbachev, pp. 9–12.

A method of extrapolating signals interrupted by pulse interference is examined. The method is based on the use of loops of series-connected RC sections.

**Nonlinear Distortions of an FM Signal in a Channel with Limited Passband,** L. M. Kononovich, pp. 32–37.

On the basis of spectrum analysis starting from the general formulas for the shape of an FM output signal, computational formulas are given to determine the nonlinear distortion of an FM signal in a channel with a rectangular amplitude-frequency and a linear phase-frequency characteristic.

**Computation of the Operation Condition of Transistor Circuits on the Basis of the Theory of an Autonomous Four-Terminal Network,** V. A. Tsar'kov, pp. 47–54.

A computation is given of currents and voltages in a transistor circuit in terms of the parameters of the associated four-terminal network connecting all the circuit elements with the exception of the transistor itself, which is also represented as a four-terminal network. The possibility is proved of determining the stabilizing properties of circuits by using new circuit characteristics independent of the transistor parameters. A general formula for all circuits, permitting an estimate of the influence of the transistor input impedance on the circuit operation is derived.

**Correlation Function and Spectral Density of a Quantized Process,** A. I. Velichkin, pp. 70–77.

Several investigations of the correlation function and spectral density of a quantized process are known. The work of Bennet may be mentioned first. B. V. Abramov made a similar investigation. The authors in these investigations were limited to an examination of the case when the level-quantization unit had a uniform characteristic. Moreover, the level quantization and the time quantization were analyzed jointly and this led to such complex expressions for the correlation function that the spectral density could be determined only approximately.

Another solution of the problem is proposed here which is valid for any characteristic of the unit carrying out the level quantization. Moreover, the level and time quantizations are examined separately. This permits more simple and, moreover, exact mathematical expressions to be obtained for the correlation function and the spectral density.