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7 August 1980

USSR Report

ELECTRONICS AND ELECTRICAL ENGINEERING

(FOUO 13/80)

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ANTENNAS

UDC 621.396.676

METHOD FOR CALIBRATING LOAD RESISTORS OF TWO-DIMENSIONAL PERIODIC STRUCTURES

Kiev, IZVESTIYA VUZ SSSR-RADIOELEKTRONIKA in Russian No 3, 1980 pp 90-93

[Article by O. N. Tereshin, V. A. Konskiy, and V. I. Kornyukhin, submitted 26 Feb 1979]

[Text] At the present time, two-dimensional periodic loaded structures [1] are beginning to be used widely in the synthesis of low Q-factor antennas working in the leaky wave mode [2]. Such antennas continuously transform the feed wave along the antenna into the radiated electromagnetic wave. They can be constructed, for example, on the basis of a stripline one of whose plates goes into a two-dimentional periodic loaded structure or on the basis of a coaxial line whose external braiding also goes into a two-dimensional periodic cylindrical loaded structure. The reactive load resistors of two-dimensional periodic structures played a role of couplers between the radiating slots and the feed wave. The main parameters of the antenna, such as the constancy of the input resistance and the directional diagram in the operating frequency band, depend considerably on the degree of couping between the radiating elements and the feed wave. Therefore, the accuracy of the realization of the calculated values of load resistors must be sufficiently high.

Unfortunately, the existing measuring equipment makes it impossible to measure the reactive loads (capacitance and inductance realized, for example, on the basis of segments of a two-wire line) in the microwave band with a high degree of accuracy.

This article describes a method for measuring reactive loads in the microwave band which makes it possible to achieve a sufficiently high degree of accuracy.

A plane homogeneous electromagnetic wave propagating along the axix z falls onto a two-dimensional periodic loaded structure situated in the plane (x,y) which is characterized by the periods T, T_1, and the value of the load resistance Z_H (Figure 1). It is assumed that there is no variations of the field and load values along x. For this case, with the structure of TEM-waves in half-spaces I ($z > 0$) and II ($z < 0$), there exist two components of the electromagnetic field [3]:

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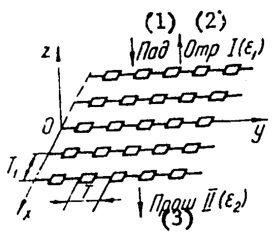
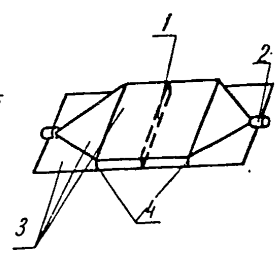
$$H_x;\quad E_y = -\frac{i}{\omega\varepsilon}\frac{\partial H_x}{\partial z}.$$





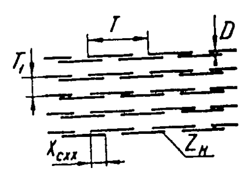


Figure 1. Two-Dimensional Periodic Loaded Structure
Key: 1. Incident
2. Reflected
3. Transmitted

Figure 2. Stripline Used in Calibration: 1 -- two-dimensional periodic structure; 2 -- high-frequency connector; 3 -- metal plates; 4 -- dielectric bushings

Figure 3. Two-Dimensional Periodic Structure with Capacitive Loads Realized on the Basis of Open Segments of a Two-Wire Line

With consideration of the above assumptions, the field of the incident electromagnetic wave can be written in the following form:

$$H_{x\text{пад}} = A_0 e^{ikz};\qquad E_{y\text{пад}} = \frac{k}{\omega\varepsilon} A_0 e^{ikz},$$

where A_0 -- amplitude coefficient of the incident wave; $k = 2\pi/\lambda$ -- wave number.

The plane homogeneous electromagnetic wave falling on the two-dimensional periodic loaded structure partially reflects from the structure and partially passes through it.

Thus, the resulting electromagnetic field in the I half-space can be written in the following form:

$$H_{x1} = H^{(1)}_{x\text{пад}} + H^{(2)}_{x\text{отр}} = A_0 e^{ikz} + A_1 e^{-ikz};$$
$$E_{y1} = \frac{k}{\omega\varepsilon_1}\left(A_0 e^{ikz} - A_1 e^{-ikz}\right), \tag{1}$$

Key: 1. Incident
2. Reflected

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where A_1 -- amplitude coefficient of the reflected waves; ε_1 -- dielectric permeability of the medium in I half-space, and in II half-space:

(2)

Key: 1. Transmitted

where B -- amplitude coefficient of the transmitted wave; ε_2 -- dielectric permeability of the medium in II half-space.

According to [1], on the surface of the location of a two-dimensional periodic loaded structure the following boundary conditions must be fulfilled for z = 0:

$$H_{x2} = \overset{(1)}{H_{x\text{пром.}}} = Be^{ikz}; \quad E_{y2} = \frac{k}{\omega\varepsilon_2} Be^{ikz}, \tag{3}$$

$$\underset{z=0}{E_{y1}} = \underset{z=0}{E_{y2}},$$

$$Z_н = \frac{T}{T_1} \frac{E_{y,z=0}}{\underset{z=0}{H_{x1}} - \underset{z=0}{H_{x2}}}. \tag{4}$$

With consideration of (1) and (2) the conditions (3) and (4) will, respectively, assume the following form:

$$A_1 = A_0 - \frac{\varepsilon_1}{\varepsilon_2} B, \tag{5}$$

$$Z_н = \frac{T}{T_1} W_2 \frac{B}{2A_0 - B\left(1 + \frac{\varepsilon_1}{\varepsilon_2}\right)}, \tag{6}$$

where $W_2 = k/\omega\varepsilon_2$ -- wave resistance of the medium of II half-space.

Assuming that the loads are purely reactive -- $Z_H = iX_H$ and $\varepsilon_1 = \varepsilon_2$ and using (5) and (6), we obtain an expression for the reflection coefficient in the form of

$$\Gamma = \frac{A_1}{A_0} = \frac{1}{1 + 2iX'_н \frac{T_1}{T}}, \tag{7}$$

where $X'_н = X_н/W$; $W = k/\omega\varepsilon$ -- wave resistance of the medium.

The modulus and the phase of the reflection coefficient can be determined from (7) by the formulas:

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$$|\Gamma| = \frac{1}{\sqrt{1+\left(2X'_{\text{н}}\frac{T_1}{T}\right)^2}}; \tag{8}$$

$$\varphi = \operatorname{arctg}\left(-2X'_{\text{н}}\frac{T_1}{T}\right). \tag{9}$$

The obtained calculated relations (8) and (9) make it possible to determine the value of the reactive resistance of the loads $Z_{H\ measured}$ of the two-dimensional periodic structure by the measured values of the modulus and phase of the reflection coefficient.

For measuring the modulus and the phase of the reflection coefficient, it is possible to use a stripline matched with a feeding coaxial cable. The two-dimensional periodic loaded structure whose value of the reactive resistances of the loads has to be measured is placed between the plates of the stripline perpendicular to the direction of the propagation of the electromagnetic wave of the stripline (Figure 2). This makes it possible to measure the modulus and the phase of the reflection coefficient with a high degree of accuracy by the known methods [4, 5].

The method described above was used in developing models of antennas discussed in [2], which made it possible to improve substantially the electrical parameters of the antennas, bringing them close to the calculated values. Reactive loads of the antenna models were designed on the basis of segments of a two-wire line.

A two-dimensional periodic structure with capacitive loads achieved on the basis of open segments of a two-wire line is shown in Figure 3. The length of the wire segments, the distance between them, and the diameter of the wire were selected on the basis of the following formula [6]:

$$X_{cxx} = \frac{\lambda}{2\pi}\operatorname{arcctg}\left(\frac{Z_{\text{н}}^{(1)}}{W_{\text{дл}}^{(2)}}\right).$$

Key: 1. Z_{load}
2. W_{length}

where X_{cxx} -- length of the open segment of the two-wire line equivalent to the prescribed capacitive load $Z_{\text{н}}=-i/\omega C$; λ -- calculated wave length:

$W_{\text{дл}}=276\sqrt{\mu/\varepsilon}\lg D/r$ -- wave resistance of the two-wire line; D -- distance between wires: r -- wire radius.

In the process of the calibration of X_{cxx}, D, and r were changing until the value of the load resistance $Z_{H\ measured}$ determined by the proposed method reached the required calculated value Z_H.

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COMMUNICATIONS, COMMUNICATION EQUIPMENT, RECEIVERS AND TRANSMITTERS, NETWORKS, RADIO PHYSICS, DATA TRANSMISSION AND PROCESSING, INFORMATION THEORY

UDC 621.391.7

DATA TRANSMISSION VIA A TELEPHONE EXCHANGE NETWORK

Moscow ELEKTROSVYAZ' in Russian No 1, 1980 pp 25-27 manuscript received 29 Jun 79

[Article by V.O. Shvartsman and V.G. Osipov]

[Text] The ever increasing demand for data transmission (PD) is leading to the necessity of using the most branched communications network for this purpose: the general service telephone exchange network (TF-OP). The fast rates of growth of automation in the long distance portion of the network are making data transmission available between practically any points in the nation.

Despite the development of specialized data transmission networks, usually no less than 50 to 55 percent of data transmission subscribers make use of the switched TF-OP network. However, in the organization of data transmission via the TF-OP network, it is necessary to take into account the fact that its characteristics do not meet all of the requirements of data transmission subscribers, while data transmission, in turn, can theoretically prove to have an undesirable impact on the operational quality of the telephone network.

The characteristics of the TF-OP network. Since in the design of this network, it was not intended for data transmission, some characteristics of the TF-OP network (the nonuniformity of the amplitude-frequency response and the group delay time short-term interruption, pulse interference, etc.), which do not exert a marked influence on the quality of speech transmission, considerably degrade the fidelity of data transmission. The reduction in accuracy is primarily due to interference arising in the electromechanical equipment of municipal ATS's [automatic telephone exchanges], and its level depends on the type of ATS [2, 3]: in crossbar exchanges, the error factor runs on the order of $2 \cdot 10^{-5}$; in computer exchanges it is $5 \cdot 10^{-5}$; and in 10-step exchanges, it runs up to $2 \cdot 10^{-2}$. The load on an ATS likewise has an impact on transmission accuracy. Thus, even 10-step exchanges in a light load period (for example, at night) provide a fidelity 2 to 3 orders of magnitude better than in the peak load. Long distance calls usually have enormously better

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indicators with respect to fidelity (by one to two orders of magnitude). The introduction of crossbar and quasi-electronic, and thereafter even electronic ATS's and MTS's [long distance telephone exchanges] will promote a significant increase in the quality of data transmission via a TF-OP network.

When data is transmitted via a TF-OP network, the discrete data signals are converted as a result of the modulation into voice frequency signals, while back conversion is realized for the reception. Because of the nearness or equality of the frequencies for telephone signaling and the characteristic frequencies of the most widespread modems (modulators-demodulators), it is possible for the data signals to imitate the telephone signaling, as a result of which, a connection can be broken. Thus, two of the four characteristic frequencies (980, 1,180, 1,650 and 1,850 Hz) of a modem, operating at a rate of 200 bit/sec, are close to the frequencies of the dual frequency telephone signaling system (1,200 and 1,600 Hz), while one of the characteristic modem frequencies at 1,200 bit/sec (2,100 Hz) matches the frequency of the single frequency signaling system.

All of the difficulties mentioned here in the utilization of a TF-OP network for data transmission can be completely overcome [3]. One of the ways of boosting transmission fidelity up to the requisite level (an error factor of $K_{err} \leq 10^{-6}$) is the use of error protection devices (UZO), based on the method of resolving feedback in conjunction with an error detecting code.

The danger of data signals simulating the telephone signaling currents can be eliminated by means of limiting the transmission time for signals of one polarity (and correspondingly, one frequency) to such short values that the telephone signaling equipment does not have time to actuate. This limiting is usually achieved either by the use of a specially chosen correcting code which does not have code combinations incorporated in it which consist of continuous zeros or ones, or by using a scrambler. The latter converts the signal being transmitted to a quasi-random sequence, a characteristic distinction of which is the low probability of the appearance of long sequences of one polarity, something practically precludes the possibility of telephone signaling equipment actuating from data signals.

Until recently, the data transmission rate via a TF-OP network did not exceed 1,200 bit/sec. Success in the sophistication of modems and adaptive correction devices have recently made it possible to go over to higher rates: 2,400 and 4,800 bit/sec. The characteristics of such modems have already been standardized by the CCITT. Tests of 2,400 and 4,800 bit/sec modems on a TF-OP network, which were performed by the TsNIIS [Central Scientific Research Institute for Communications] and the KONIIS [Kiev Branch of the Central Scientific Research Institute for

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[redacted]

Communications] in 1974, showed that the value of the error factor is practically the same when operating at speeds of both 1,200 bit/sec and 2,400 bit.sec. When operating at a speed of 4,800 bit/sec, despite the increase in the error factor (by 4 to 6 times), the utilization of a TF-OP network is justified at the present time since this increase can easily be compensated by using error protection devices in the data transmission equipment (APD). Modern APD makes it possible to assure a transmission accuracy of no less than 10^{-6} on a TF-OP network.

Thus, one can argue that the specific features and poor characteristics of the TF-OP network for data transmission, when the appropriate terminal equipment is present, are not an obstacle to data transmission at rather high speeds via the PF-OP network.

The impact of data transmission on the characteristics of an TF-OP network. This problem involves two aspects. One of them consists in the influence of the additional load (the data transmission) on the quality of service to telephone subscribers of a TF-OP network, while the other consists in the possibility of the degradation of the operational quality of the system due to the increased output power of the terminal data transmission equipment as compared to a telephone set. It is not difficult to convince oneself that these dangers are without any serious bases.

The number of data transmission subscribers is many times less than the number of telephone subscribers. Thus, for example, based on data from the U.S. (the nation which makes the greatest use of data transmission), the ratio of the number of APD's to the capacity of the TF-OP network is so small that even in the future it will amount to only units of percent [1, 4]. The relative number of data transmission sets is even less in Western Europe and Japan, where it is not planned that they will number more than one percent. Such a low relative number of data transmission subscribers attest to the theoretical impossibility of a negative impact of data transmission on the operation of a TF-OP network, even in those cases where the parameters of the data load do not meet the norms adopted in this network.

However, the data transmission load parameters do not basically differ from the corresponding characteristics of the telephone load. Thus, the data signal level at the input to data transmission equipment inserted in a TF-OP network is set at 50 or 32 µW 0 for various types of equipment. A power of 50 µW 0 somewhat exceeds that level at which the channels can be used without limitations. At the same time, overloading the transmission systems of the TF-OP network with data signals is practically impossible because of the miniscule probability of signals from two or more data transmission sets getting into one channel generating system, because of the small number of them. Also to be taken into account is the fact that a power of 50 µW 0 is characteristic of data transmission equipment produced in earlier years. The new terminal equipment developed

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after 1 January, 1974 in accordance with OST 4. GO. 208. 004, should operate with a level of 32 μW 0, something which generally eliminates the question of limitations in the use of transmission system channels for data transmission.

As far as the question of the load on subscriber lines is concerned, it is well known that low data volumes are characteristic of the overwhelming portion of data transmission subscribers: on the order of 10,000 characters per 24 hours and less. Such a volume even for a transmission rate of 1,200 bit/sec (≈ 100 char/sec) can be transmitted in 100 sec. Considering the fact that data exchange is usually accomplished two to three times a day, the duration of one data transmission session amounts to tens of seconds, and taking into account the service telephone conversations, the length of a session obviously should not exceed units of minutes (hundredths of an Erlang), i.e., the duration of one telephone conversation. However, the utilization of lower transmission rates (200 bit/sec and below) is not desirable, since the prerequisites for a marked increase in the load can be created. Transmission rates of 600 bit/sec and above are recommended. The possibility of utilizing rather high data transmission rates on a TF-OP network makes it possible to reduce the load on the network. For data transmission at low rates, it is expedient to use the PD-200 network which has been created.

The conclusion concerning the low data transmission load which has been drawn here applies to systems with so called batch transmission, where the information is generated in the form of one data file ("packet"), following the transmission of which, the communications session is terminated. Packet transmission is characteristic of planning, statistical, reporting and other kinds of data transmitted in a relative time scale. Along with such systems, interactive data transmission systems are finding application, which provide for interaction in real time between man and computer or computer and computer.

Included among such systems are some queuing systems (reference services, ticket ordering systems, systems for the solution of various problems on remote computers, etc.). The interactive mode includes pauses, during which the results are estimated, the situation is thought out, etc. The pauses lead to an increase in the overall length of a data transmission session, because of which, it is possible in some cases for interactive data transmission systems to have a negative impact on the operation of a TF-OP network. For this reason, the interactive transmission mode basically requires the construction of specialized data transmission networks.

The specific features of the data load presented here raise the question of monitoring the power of data signals and the length of time a subscriber line is busy. However, because of the small number of data transmission subscribers and the absolutely insignificant differences between the major parameters of data and telephone loads, it is not

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expedient to organize systematic monitoring of the parameters. Exceeding the established duration of subscriber line occupancy in the peak load hour (an average of 0.1 Erl, the maximum permissible mode is 0.15 Erl) is not a real possibility because of the low data volumes characteristic of data transmission subscribers, while operation at an elevated level is improbable in practice because of the existence of the practice of placing a lead seal on the output signal level regulator. Nonetheless, it is useful to organize selective monitoring of the transmit level. Such monitoring, which it is obviously expedient to accomplish manually, will provide an adequate guarantee of protection of the telephone exchange network against any negative influences.

The advantages of the use of a TF-OP network for data transmission. The TF-OP network makes it possible to design not only data transmission systems accessible to a wide group of users, but also maximally simplified systems. Such systems, usually designed for the transmission of low data volumes, can be built on the basis of multifrequency keyboard telephone sets, which, besides their main function, play the part of an extremely simple data transmission set. Data transmission is accomplished by punching the buttons on the telephone set keyboard following the establishing of the call connection, and data reception is accomplished at the computer center using a modem which received multifrequency signals from push-button telephone sets.

In some cases, data transmission can be accomplished in practice only by using the accessible and widely branched TF-OP network: for example, in the public service - for the remote monitoring of the status of non-transportable seriously ill patients by means of a computer or for the transmission of cardiograms from the apartment of the sick person to the cardiological center for analysis; in the sphere of everyday life - for the cashless payment for electrical power, gas, etc.

The utilization of a TF-OP network for data transmission is being stimulated by its universal nature: the transmission of the most diverse information is possible via this network (speech, data, facsimile signals). Considering the fact that managing the national economy, as a rule, requires the organization of all kinds of communications (or at least, telephone and data transmission), the orientation towards the telephone network proves a natural one.

Data transmission via the TF-OP network in our country is regulated by the "Regulations on the Conditions for the Utilization of the Switched Telephone Network for Data Transmission". The basic stipulations of this document, which opens up broad prospects for data transmission users. consist in the following: data can be transmitted without limitations at any time of the day; the average load on a subscriber line should not exceed 0.1 Erl in the peak load hours, and the maximal load should not exceed 0.15 Erl; the average data signal power at the input to the

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closest transmission system should not exceed 32 μW 0 (for terminal data transmission equipment developed prior to 1 January, 1974, a level of 50 μW 0 is permitted); the data transmission rate, as a rule, should be no lower than 600 bit/sec: the maximum permissible rate is established at 4,800 bit/sec; the type of terminal set connected to the TF-OP network should be included in the products list approved by the USSR Ministry of Communications; complaints concerning data transmission quality during the normal course of a telephone conversation are not accepted by the communications organs.

The area of TF-OP network utilization is an extremely wide one, however, it is far from unlimited. It is apparent that this network cannot be used in cases where: the load generated by a data transmission subscriber exceeds the intrinsic level for telephone subscribers; data transmission subscribers need specific services not provided by the TF-OP network (for example, multi-address communications); the network does not provide a subscriber with the requisite operational mode in the needed directions (for example, automatic dialing of a long distance call); the time making a call connection or the network reliability do not satisfy the subscriber requirements.

There is no automatic identification of the called subscriber in the TF-OP network, since it is accomplished by the subscribers themselves in telephone communications. For this reason, when exchanging data via the TF-OP network, it is necessary at the start of a session to make sure that the connection has been correctly established. This precludes the possibility of sending data to the wrong address. It should also be kept in mind that the use of radio relay and satellite communications links in the TF-OP network places well known limitations on the possibility of using the TF-OP network for data transmission.

Conclusion. The capacity of the TF-OP network has substantially increased over recent years and the level of long distance automation is growing at a fast pace. The quality indicators of the network are being improved through the introduction of increasingly refined equipment. In the future is the saturation of the network with equipment, the design principle of which is based on digital transmission and switching methods. Such equipment provides for high quality characteristics, something which is especially important for data transmission. However, even a digital switched telephone network does not eliminate the need of designing specialized data transmission networks, since the requirements of many data transmission subscribers as regards a number of network parameters (signaling, services, switching methods, etc.) can be specific, and moreover, change rather rapidly, while the requirements of telephone subscribers have primarily remained the same.

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Considering the comparatively small number of data transmission subscribers it is not expedient to extend these requirements to the entire telephone network. The most efficient approach to the development of data transmission technology is the design of specialized data transmission networks in conjunction with the use of the TF-OP network for data transmission, taking into account the limitations which follow from the specific features of this network.

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SOME PROBLEMS OF TELEPHONE NETWORK CONTROL

Moscow ELEKTROSVYAZ' in Russian No 1, Jan 80 pp 21-24 manuscript received 5 Apr 78

[Article by M.A. Shneps]

[Text] Increased demands are placed on the theoretical basis for the carrying capacity of a communications network and on the substantiation of the standards set for communications quality because of the need for the large capital investments in the Unified Automated Communications Network [1]. This confronts telephone network designers with problems, for the solution of which there is unfortunately still no adequate theoretical basis.

Many problems in telephone traffic theory are solved without sufficiently taking into account the actual operating conditions of a telephone network. Specifically: the capacity of the trunk group of channels is computed from Erlang's first formula, which does not take into account repeat attempts made by the calling subscriber with the blocking of the next attempt to make a connection with the called subscriber; reliability indicators are not reflected in the calculations of the carrying capacity of the network, or in the standards set for losses; when planning bypass routes, the fact that the utilization of them increases the carrying capacity of a network only in the nonoverloaded state of the network or with overloads (failures) of individual trunk groups is not taken into account.

However, the elimination of the difficulties enumerated above on an individual basis does not answer all of the questions which arise in the design of a network, especially those related to network control in overload modes. A general approach to the solution of telephone network control problems as a whole is presented in this paper, including the calculation of the carrying capacity of the switching equipment, the control devices, and considerations for the setting of loss standards and reliability indicators are given, where these considerations are based on economic calculations.

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An approximate network design method. We shall begin with a simple example which illustrates the complexity of the problems which come up in network design. We shall consider a fragment of a channel switching network (Figure 1). Communications are established between the calling subscriber A and the called subscriber B through four switching stations, each of which has its own switching circuit, KS, and controller, UU. We shall assume that the subscriber is α-persistent, i.e., following an unsuccessful attempt to make a call, the subscriber repeats the call with a probability α after a random time, and gives up on transmitting the message (the conversation) with a probability of $1-\alpha$. If $\alpha = 1$, then the subscriber is termed absolutely persistent.

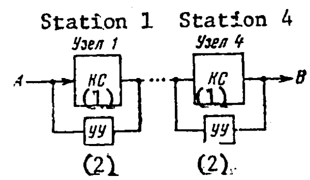

Figure 1.

Key: 1. KS [switching circuit];
2. UU [controller]

We shall introduce the rest of the symbols: m is the number of traffic stations on the call connection path; q and π_a are the probabilities that the attempt will be blocked at an individual station and in the called subscriber circuit (the subscriber is busy or does not answer) respectively; t and t_a are the average duration of time for making a call connection at one station and in the called subscriber circuit respectively; T is the average duration of a conversation.

We shall calculated the load intensity* approximately, which the KS and UU of the first station handle with the arrival of the call from subscriber A to subscriber B. If the duration of time needed for making the call connection is neglected and the subscriber is considered to be absolutely persistent, then the load serviced by the KS is equal to T in Erlangs, while the load serviced by the UU is equal to zero, something which does not agree with measurement data for the network.

We shall cite the formulas for the network design. The probability for the unsuccessful attempt to make a call connection via the path from A to B is:

$$Q = 1 - (1-q)^m \, (1-\pi_a). \tag{1}$$

We shall subsequently make use of the geometric distribution, in accordance with which the number of attempts until the first success with a success probability p is equal to 1/p. The average number of attempts per call M is determined by the geometric distribution with a parameter of $p = 1 - \alpha Q$

* In following the tradition of literature on telephone traffic theory, we shall employ the concept of "load" in the following instead of the concept of "load intensity" (which is measured in Erlangs). This does not lead to a misunderstanding, since the remaining load parameters will not be treated as a random process.

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and therefore:
$$M = \frac{1}{1 - \alpha\ [1 - (1 - q)^m\ (1 - \pi_a)]}. \qquad (2)$$

The call loss probability is:
$$P_c = MQ\ (1 - \alpha). \qquad (3)$$

The overall load handled by the switching circuit of the first station is:
$$A = M\ [tq + 2tq\ (1-q) + 3tq\ (1-q)^2 + \\ + 4tq\ (1-q)^3 + (4t + t_a)\ (1-q)^4 \pi_a + \\ + (4t + t_a + T)\ (1-q)^4\ (1 - \pi_a)]. \quad (4) \qquad (4)$$

The useful (paid) load is: $A_\pi = T(1 - P_c)$ (5)

and the load serviced by the UU of the first station is:
$$Y = tM \qquad (6)$$

In order to make a transition from the load per call to the load per unit time, the quantities A, A_π and U must be multiplied by λ - the average number of calls per unit time. In this case λ should be chosen so that the probability of blocking an attempt is equal to q.

TABLE

q	Q	M	P_c	A	A_π	Y	λ	λA	λA_π
0,01	0,52	1,88	0,11	1,15	0,89	0,02	39	45	34,6
0,1	0,67	2,52	0,17	1,08	0,83	0,03	44,4	48	36,5
0,5	0,97	7,75	0,75	0.44	0,25	0,08	111	49	27,8

The results of calculations using formulas (1) - (6) are shown in the Table for the case where: $m = 4$; $\alpha = 0.9$; $\pi_a = 0.5$; $T = 1$; $t = 0.01$ and $t_a = 0.1$. The column of values for λ was taken in approximate manner by working from the assumptions that the number of channels in the trunk group is $V = 50$, the repitition rate is $\nu = 10$ and the servicing time has one or two phases: the first is the time for making the call connection and the second is the duration of the conversation [2]. The unsuccessful attempt probability is equal to $\pi_a^* = 1 - (1-q)^3(1-\pi_a)$, and in this case, the servcing time contains one phase.

The data of the table show that in step with an increase in the unsuccessful attempt probability of making a call Q, the number of repeat attempts M rises rapidly and exponentially, something which leads to a corresponding exponential growth in the load Y on the controllers; in step with a rise in the call flow rate λ, the overall serviced load λA tends to v (i.e., to a 100 percent busy circuit), while the useful load λA_π first increases and then falls off.

The resulting data are approximate. A precise solution of the problem is apparently accessible only by means of statistical modeling. An

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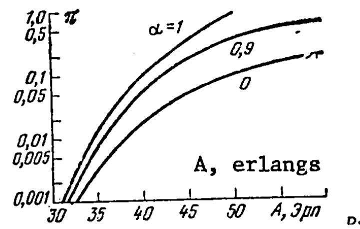

Figure 2.

analytical solution is given in [3] for a lossy fully accessible trunk group which handles a poisson load. Curves for the probability of an unsuccessful attempt at making a call π in a 50 channel trunk with a repetition rate of $\nu = 10$ are shown in Figure 2 for values of $\alpha = 0.9$, $\alpha = 1$ (an absolutely persistent subscriber) and $\alpha = 0$ (the last curve coincides with the curve for Erlang's first formula). A comparison of the three curves shows that Erlang's formula, which has up until now been universally used in calculations, cannot serve as the basis for the determination of the carrying capacity of trunk groups, especially in the case of large losses.

Two unsolved problems follow from the example considered here: 1) For the design of the controller, it is necessary to study systems with waiting during the admission of the load, during which, in step with the increase in occupancy of the system, a secondary additional flow of repeat attempts appears with an exponentially increasing rate (its own kind of positive feedback effect); 2) For an approximate calculation of the carrying capacity of a network, it is necessary to study systems with repeat attempts in the case of two-phase servicing, which can serve as a model of a network element.

An analysis of the carrying capacity. The efficiency of a long distance telephone network is determined by the income from payments for conversations which have taken place, which form the useful (paid) load. It is always less than the serviced load. As the data of the table show, the ratio A_π/A amounts to 77 percent when $Q = 0.01$ and 55.5 percent overall when $Q = 0.5$.

The results of an analysis of the carrying capacity of a channel trunk group, a switching station and controller are shown in Figure 3. It can be seen in Figure 3a how the serviced load A and the useful load A_π change as a function of the incoming load A_0 (V is the number of channels). We shall call the difference $A - A_\pi$ the service load A_{s1}, and it is composed of the occupancy intervals in making the call connection.

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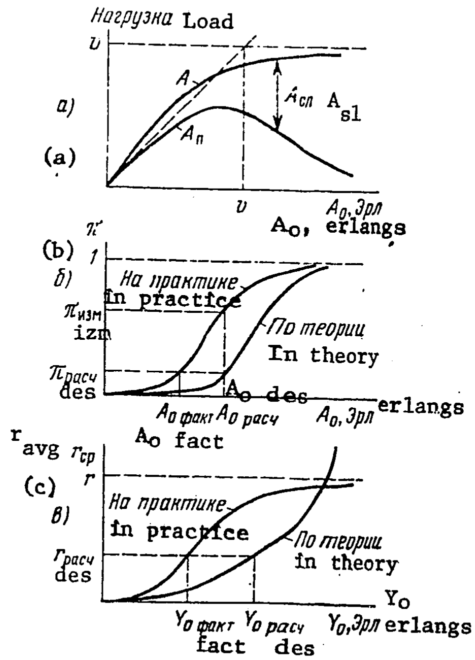

Figure 3.

The number of channels is usually chosen during the design stage so as to assure the permissible loss possibility π_{des}. However, because of the inaccurate calculation of the serviced load A, the loss probability π_{izm} which is observed in practice exceeds π_{des} (Figure 3b).

We will note that it is desirable to take an estimate of the timewise losses (the probability of a loss with the first attempt) as the loss probability, and not the estimate of losses based on calls (the loss probability for the attempts), since the first estimate has a lower statistical scatter than the second. It would be even more justified to take the call loss probability here P_c, but unfortunately, there are as yet no simple methods of measuring it.

The lack of agreement between the theoretical calculation of the average number of busy waiting positions, r_{avg}, in the controller and the results of observations in practice is shown in Figure 3c. This lack of agreement is due to two factors: the presence of an additional load because of repeat attempts and the finite number of waiting positions, r, (in contrast to the calculation where an unlimited number of them is usually assumed). Because of these factors, the permissible mean number of occupied waiting positions, r_{des}, is achieved not at the design load $Y_{0\ des}$, but rather at $Y_{0\ fact}$.

The "expenditures--income" relationship for a network. It is important for designers when planning and expanding a network to determine its economic efficiency S. This indicator can be approximately expressed in terms of the received income, i.e.:

$$S = D - (B+C) \tag{7}$$

where D is the payment for conversations which have taken place (income); B is the specific capital investments; C are the expenditures for the ongoing operation and repair and restoration work for a specified equipment readiness.

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Figure 4a illustrates the terms of expression (7) where the curve D is derived by multiplying A_{π} (Figure 3a) by the payment per conversation unit. Based on formula (7), one can judge the disadvantageousness of the communications system operation in the case of overloads (the range $A_0 > A_0^2$ in Figure 4b) as well as the economic efficiency of various measures to change the capital investments (a change in the curve B), operational expenditures (curve C) or to change the amount of payment per conversation (curve D). The income S is shown in Figure 5a as a function of the increase in the expenditures for repair and restoration work (a transition from curve C_1 to C_2). The latter increases the equipment readiness indicator and can increase the useful load (the transition from D_1 to D_2). And then we obtain two curves for the incomes S_1 and S_2 from formula (7) (Figure 5b), from an analysis of which it can be seen that the expedience of a decision to increase the outlays for repair and restoration work depends on the size of the incoming load A_0: if $A_0 < A_0^*$, then the transition from C_1 to C_2 is not advantageous, and in the case of $A_0^* < A_0$, it is advantageous.

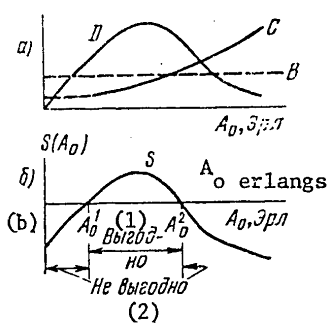

Figure 4.
Key: 1. Advantageous;
2. Not advantageous.

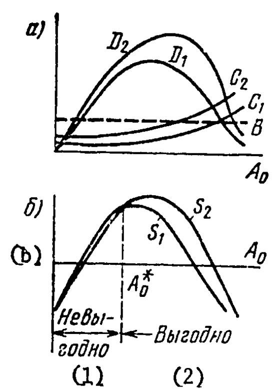

Figure 5.
Key: 1. Non advantageous;
2. Advantageous.

The substantiation of standards for losses. The major indicator of the carrying capacity of a network is the loss probability during the peak load hour. Along with this, work is being done to refine the standards set for communications quality, for example, not only the average losses are taken into account in paper [4], but also the losses during periods of statistical overshoots. For this, it is proposed in [5] that the average value of the losses among 30 and 7 of the maximum peak load hour values over a year of observations be standardized. Similar considerations are also taken as the basis for a design load procedure treated in [4].

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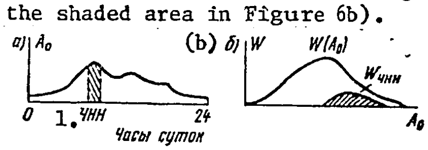
However, why can't the entire distribution curve for the values of the incoming load, A_0, be used and then be limited by one value of this load in the peak load hour? In fact, the incoming load, A_0, and, correspondingly, the serviced load, A, vary during the course of a 24 hour day (Figure 6a). If a histogram of the load values $W(A_0)$ is plotted for hourly intervals over the course of a year, it then turns out that the values of the load over one selected hour during the period of greatest occupancy (which is also understood as the values in the peak load hours) fall in a range which is not completely within the region of greatest load values (indicated by the shaded area in Figure 6b).



Hours of the day

Figure 6.

Key: 1. Peak load hour.

If the distribution density of the load values $W(A_0)$ and the income function S $S(A_0)$ are known, then the average income can be calculated:

$$\bar{S} = \int_0^\infty S(A_0)\, W(A_0)\, dA_0. \tag{8}$$

This quantity provides information on the economic efficiency of a trunk group of channels or a network as a whole, and can serve as the basis for setting carrying capacity standards. On analogy with (8), one can define the average losses:

$$\bar{\pi} = \int_0^\infty \pi(A_0)\, W(A_0)\, dA_0, \tag{9}$$

where various kinds of losses (timewise, call losses, etc.) can be taken as $\pi(A_0)$. Expression (9) is closer to the standards adopted at the present time than the economic characteristic of (8). A universal changeover in standards documents to characteristics of the type of (8) is hardly expedient, but they should be used for the economic substantiation of network expansion variants.

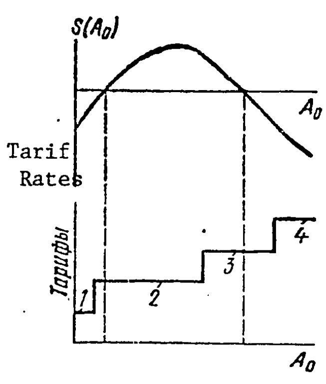


Figure 7.

Network control. It is shown in Figure 4b that it is economically disadvantageous to operate a network if the incoming load goes beyond the limits of the interval (A^1, A_0^2). In network operation, it is desirable to avoid disadvantageous operational modes, and this especially applies to network overloads. It is possible to eliminate them by means of the following:

1. The selection of the amount of payment for a completed call (the tariff). This measure is the most powerful tool of load control for the purpose of increasing it during the night hours and reducing it during overload periods. The interesting problem of choosing a model for the behavior of a subscriber in the situation of changing

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tariffs arises in this respect. For the needs of automated service, it is necessary to develop a theory of self-adapting tariffs [7]. The payment per current connection and call is established based on the results of observations of trunk operation during the preceding period (the choice of the length of this period is also to be substantiated).

One of the possible variants for the solution of this problem is given in Figure 7, where 1 is a reduced tariff, 2 is the normal tariff, 3 is a doubled tariff and 4 is a tripled tariff. The selection of the boundaries of the transitions from one tariff to another can be solved as an optimal control problem.

2. Control of bypass routes. The generation of signals which limit or inhibit the utilization of bypass routes is one of the problems confronting the CCITT [3]. There are no theoretical prerequisites in existence today for its solution. Questions of the calculation of the efficiency of bypass routes have now been worked out for conditions where Erlang's first formula is applicable [6]. A similar problem, including the incorporation of economic relationships for the cost of direct and bypass routes must be solved for a system with repeated call attempts.

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DIGITAL TRANSMISSION SYSTEM SIGNALS AND CODES

Moscow ELEKTROSVYAZ' in Russian No 1, 1980 pp 33-37 manuscript received 5 Jun 79

[Article by O.N. Porokhov]

[Text] *Introduction.* The present stage of the introduction of digital transmission systems (TsSP) is characterized by the offering of analog and digital channels to subscribers for data exchange, where the channels are formed by the series insertion of hierarchical series transmission systems with differing transmit rates and different means of routing. The development of uniform requirements for the design principles of transmission systems is one of the most important problems confronting digital transmission system designers. Along with the traditional questions in digital communications - improving noise immunity and carrying capacity - of no less importance is the synchronization of a large number of transmission systems which are tied together in series.

A great diversity of data transmission methods is characteristic of every transmission system (cable, optical, radio relay, etc.). However, the selection of the optimum method and the development of general recommendations for a specific system run up against difficulties because of the specific features of the system, the lack of a uniform listing of the parameters and estimates of their maximum values, as well as inaccuracies in individual terms and definitions.

An attempt is made in the following to resolve these problems based on recommendations from general communications theory.

Specific features of digital information transmission methods. The methods of information transmission via cable digital transmission systems with their specific characteristic features have proved to be unacceptable for transmission systems with other means of routing. Thus, the requirement for the balancing of a digital signal with respect to the DC component in cable transmission systems [1, 2] has led to its substantial redundancy, because of which these methods have proved to be unacceptable in optical and radio systems. On the other hand, the well-known methods of digital

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information transmission by means of multiposition modulation of radio signals are unsuited to cable systems, in particular, because of the fact that they do not meet the requirements for synchronization of the transmission system.

Methods of signal reception are known for cable transmission systems are known which provide for a slight dependence of the noise immunity on the properties of an unbalanced signal [3, 4]. Because of this, a general approach to the design of transmission methods used in verious transmission systems is possible.

The difficulty of developing a general approach to digital information transmission methods consists in the fact that in cable transmission systems, the concept "coding" in cable transmission systems is frequently understood to be some mathematical conversion of the binary information being transmitted and simultaneously, the signal waveform, transmitted via the communications line. We would recall that . . . "in the literature on information theory, the term 'coding' is used in different senses. In the broad sense of the word, coding is the term for the conversion of a message to a signal. In a narrow sense, coding is defined as the mapping of discrete messages by a sequence of preselected symbols" [5], with the additional explanation that the new symbols change (modulate) the parameters of the line signal. Thus, the code designation 4B3T [13] reflects the concept of coding in the broad sense of the word: the conversion of binary symbols to ternary symbols and the transmission of the symbols by a specific video pulse signal (DC pulses of different polarity and a passive pause).

Although such a representation of the coding operation is permissible, it still has certain drawbacks, since an analysis of the code properties is needed in conjunction with the parameters of any of the various transmission system signals. By using the concept of coding in the narrow sense of the word, we shall assume that binary signals, generated by any digital information sources and referenced to one of the hierarchical rates of a digital transmission system, regardless of their type, are converted to a digital signal in two stages (Figure 1): the coding operation (not obligatory) - the conversion of the binary symbols to a sequence of new symbols (code symbols); and the modulation operation - the change of the corresponding parameters of the carrier as a function of the code symbol. The vehicles for code symbols in cable systems can be the direct current parameters (polarity, amplitude, video pulse width), in radio systems - the parameters of the carrier (amplitude, frequency, phase), and in optical systems - the number of quanta. In analyzing transmission methods in optical systems, it is expedient to consider the signals transmitted via these systems as video pulse signals, because of the difficulties in processing the adopted signals.

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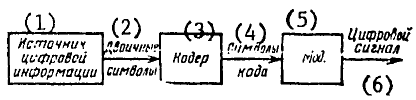

Figure 1.

Key: 1. Digital information source;
2. Binary symbols;
3. Coder;
4. Code symbols;
5. Modulator;
6. Digital signal.

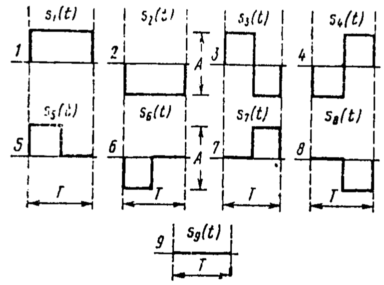

Figure 2.

The signals of digital transmission systems. In contrast to radio signals, it is difficult to make a breakdown of video pulse signals according to kinds of modulation. We shall consider a video pulse element to be any combination of video pulses and pauses within the clock interval T set aside for the transmission of one information symbol.

We shall assume the following (Figure 2) in order to limit the infinite set of video pulse signal elements: the pulses are square waves and their width if T or T/2; the leading edge of the pulses coincides with either the edge or the center of the clock interval; the amplitude of the video pulses (and later on, also the radio pulses) is the same regardless of the number of elements (we shall designate it as A/2).

Under these conditions, the number of video pulse signal elements is equal to nine. A video pulse signal can be generated from the elements of Figure 2 by either an absolute or relative modulation method [6]. In the first case, a correspondence is established between the transmitted symbols and the signal elements. Then the number of binary video pulse signals composed of the elements of Figure 2 is:

$$C_9^2 = 9!/2!\ 7! = 36$$

In the case of the relative modulation method, the binary information symbol 1 is transmitted by means of alternating two elements of the signal, while a zero is transmitted by repeating the element corresponding to the transmission of the last symbol 1. For this reason, the number of relative binary signals is likewise equal to 36. But the diversity of video pulse signals is not limited to these figures. Multi-element video pulse signals are used in transmission systems (several elements are used for the transmission of one symbol), as well as multiposition modulation of the individual elements, etc. It is expedient to limit ourselves here to a treatment of absolute binary and few multi-element

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TABLE 1

1. Элементы сигналов рис. 2 при передаче: 1	0	2. Число уровней	$P_э$ Pe	$K_п$, дБ dB	$p_т$	$K_т$
*1	2	2	A^2	0	0,5	0
1	3	2	$A^2/2$	−3	0,5	0
1	4	2	$A^2/2$	−3	0,5	0
1	5	2	$A^2/8$	−9	0,5	0
1	6	3	$5A^2/8$	−2	0,5	0
1	7	3	$A^2/8$	−9	0,5	0
1	8	2	$5A^2/8$	−2	0,5	0
1	9	3	$A^2/4$	−6	0,25	0
2	3	2	$A^2/2$	−3	0,5	0
2	4	2	$A^2/2$	−3	0,5	0
2	5	2	$5A^2/8$	−2	0,5	0
2	6	2	$A^2/8$	−9	0,5	0
2	7	3	$5A^2/8$	−2	0,5	0
2	8	2	$A^2/8$	−9	0,5	0
2	9	2	$A^2/4$	−6	0,25	0
**3	4	2	A^2	0	1	1
3	5	3	$A^2/8$	−9	1	1
3	6	3	$5A^2/8$	−2	1	1
3	7	3	$5A^2/8$	−2	1	1
3	8	3	$A^2/8$	−9	1	1
***3	9	2	$A^2/4$	−6	0,5	0
4	5	3	$5A^2/8$	−2	1	1
4	6	3	$A^2/8$	−9	1	1
4	7	3	$A^2/8$	−9	1	1
4	8	3	$5A^2/8$	−2	1	1
4	9	2	$A^2/4$	−6	0,5	0
5	6	3	$A^2/2$	−3	1	1
5	7	2	$A^2/4$	−6	1	1
5	8	3	$A^2/4$	−6	1	1
5	9	2	$A^2/8$	−9	0,5	1
6	7	3	$A^2/4$	−6	1	1
6	8	2	$A^2/4$	−6	1	1
6	9	2	$A^2/8$	−9	0,5	0
7	8	3	$A^2/2$	−3	1	1
7	9	2	$A^2/3$	−9	0,5	0
8	9	2	$A^2/8$	−9	0,5	0

Note: *This signal has been given the designation "monopulse" [11]; **This one is called "bipulse" [12]; ***Pulse doublet coding [17].

Key: 1. Elements of the signals of Figure 2 when transmitting a:;
2. Number of levels;

video pulse signals, a comparison of them and to compare radio signals based on generalized parameters, which rather simply and completely characterize the estimate of the possibility of their use in a digital transmission system.

The parameters treated in the following are summarized in the tables: 1. For all (36) absolute binary video pulse signals; 2. For some multi-element video pulse signals; 3. For binary radio signals.

The potential noise immunity of a digital signal. An important parameter which assesses the quality of digital information transmission is the error factor of a transmission system (which depends on the noise immunity of each repeater), the determination of which is difficult. Therefore, to compare any digital signals, it is proposed that their potential noise immunity be found under ideal conditions. As is well known [5], the potential noise immunity of a digital signal depends on the "equivalent" power P_e, of its elements, $s_i(t)$, $s_j(t)$, i.e.

$$P_э = \frac{1}{T}\int_0^T [s_i(t) - s_j(t)]^2\,dt, \qquad (1)$$

in which case, it is proposed that the potential noise immunity be determined for two elements of a signal with the lowest level of P_e for a comparison of multi-element and multiposition signals (the applicability of formula (1) for the signals of cable systems requires additional research).

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TABLE 2

Таблица 2

(1) Наименование сигналов	(2) Элементы сигналов рис. 2 при передаче: 1	0	(3) Число уровней	$P_э$ P_e	$K_п$, дБ dB	$p_т$	$K_т$
AMI [7]	5—6	9	3	$A^2/8$	—9	0,5	0
CMI [8]	1—2	4	2	$A^2/2$	—3	0,62	0,5
MPDC [9]	3—4	9	3	$A^2/4$	—6	0,5	0
Miller Coding [10]	3—4	1—2	2	$A^2/2$	—3	0,75	0,5

Key: 1. Designation of the signals;
2. Elements of the signals of Figure 2 when transmitting a:;
3. Number of levels.

TABLE 3

Type of Вид модуляции Modulation	$P_э$ P_e	$K_п$, дБ dB	$p_т$	$K_т$
Фазовая Phase	$A^2/2$	—3	0,5	0
Частотная FM	$A^2/4$	—6	0,5	0
Амплитудная AM	$A^2/8$	—9	0,5	0

Signals, the elements of which are opposite to each other and satisfy the equality $s_i(t) = s_j(t)$ have the maximum noise immunity [5]. For example, for the signals shown in Figure 2, this condition is met for only two kinds: with elements of $s_1(t)$, $s_2(t)$ and $s_3(t)$, $s_4(t)$. The equivalent power (1) of each pair reflects the maximum potential noise immunity of the digital signals, $P_{e.max} = A^2$.

The relative noise immunity coefficient shows by how much the potential noise immunity of a digital signal differs from the maximum noise immunity: $K_p = 10 \log (P_e/P_{e.max})$.

The minimum upper frequency of the signal spectrum, $f_{v.min}$, is a conditional frequency which is equal to the first harmonic of the periodic sequence of those signal elements for which it is a maximum. It is expedient to express this parameter in terms of the clock frequency of the digital signal, f_T. With the exception of a monopulse signal (Table 1) and an AMI signal (Table 2) for which $f_{v.min} = f_T/2$, the remaining video pulse signals have $f_{v.min} = f_T$.

The probability of a change in the parameter being modulated, p_T. This indicator is introduced to estimate the mean value of the clock frequency in the digital signal for the case of the equal probability of the transmission of the binary symbols. The presence of attributes of the clock frequency in a signal depends on the presence of leading edges of pulses (transitions between levels) in video pulse signals and on the change in the modulated parameters in radio signals.

Since the elements of the video pulse signals of Figure 2 have transitions at the boundaries and in the center of the clock intervals, taken into account in the determination of r_T (see Tables 1 and 2) are the probabilities of the transitions, the density of the appearance of which depends to a lesser extent on the variation in the statistical characteristics of the information being transmitted.

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The stability factor of the clock frequency attributes characterizes the stability of the change in the dependence of the modulated parameter of a digital signal on the statistical properties of the information being transmitted and is defined in terms of the maximum $p_{T.max}$ and the minimum $p_{T.min}$ probabilities of a change in the modulated parameters in the clock interval, i.e., $K_T = p_{T.min}/p_{T.max}$.

Besides the enumerated parameters, also given in Tables 1 and 2 are the numbers of the signal elements in accordance with Figure 2, the number of levels as well as the well-known designations of the signals.

We shall explain the rules for the generation of multi-element video pulse signals shown in Table 2. Despite the presence of the term "code", it is proposed that they be considered multi-element signals in the majority of the designations. In AMI (alternative mark inversion) [7], the binary symbol 1 is transmitted by alternating the elements $s_5(t)$ and $s_6(t)$ while 0 is transmitted by the passive pause $s_9(t)$. In CMI (coding mark inversion) [8], the symbol 1 is transmitted by the alternation of the elements $s_1(t)$ and $s_2(t)$, while 0 is transmitted by $s_4(t)$; in MPDC (modified pulse doublet coding) [9], the symbol 1 is transmitted by the alternation $s_3(t)$, $s_4(t)$, while 0 is transmitted by $s_9(t)$. In Miller coding [10], two elements each of the signal are used for each binary symbol: $s_1(t)$, $s_2(t)$ and $s_3(t)$, $s_4(t)$.

The following conclusions can be drawn from an analysis of the digital binary signals based on the listing of the parameters cited here: the change in the potential noise immunity of digital signals amounts to 9 dB; monopulse [11] and bipulse [12] signals have the maximum potential noise immunity; the potential noise immunity of phase keyed signals is 3 dB below the maximum for video pulse signals, something which is explained by the differing power levels of the square wave video pulse and sinusoidal signals in the case where their amplitude and widths are equal; the use of radio signals and the majority of video pulse signals for which $K_T = 0$ in digital transmission systems (in the case of arbitrary statistical properties of the digital information) requires the steps be taken to increase the stability of the attributes of the clock frequency.

In this case, the multiposition modulation methods with a base of 2^l (where l is any integer, greater than one) which are well known from radio systems, because of the lack of redundancy in the signal do not eliminate this drawback, and as has been noted, are not used for video signals. An exception is the bipulse signal. Because of the balanced nature of each of each of its elements $s_3(t)$ and $s_4(t)$, the maximum noise immunity and the maximum value of the stability factor of the clock frequency attributes, multiposition modulation of the elements of a bipulse signal makes it possible to compensate for the drawback: an increased minimum upper frequency of the spectrum of this signal (as compared to certain other signals).

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Digital transmission system codes. Additional conversion of the binary information (prior to the modulation of the carrier parameters) by means of varying its statistical properties is necessary for the purpose of increasing the stability of the clock frequency attributes. For this, scrambling and coding is accomplished prior to the insertion of the information in the digital transmission system.

Questions of scrambling (randomization of the statistical properties of digital information) are of independent interest and not considered here. We will note only that scrambling and coding are not mutually exclusive, but are operations which complement one another, however, it is necessary to take into account a drawback to randomization: a certain probability of maintaining extremal statistical situations.

In classifying codes, it is expedient to break them down into nonalphabetic and alphabetic codes. In the first case, changing the statistical properties of the original binary information (while preserving its clock frequency) occurs only under definite conditions: a specified number of continuous zeros, and sometimes also ones of the binary information. Alphabetic coding consists in dividing the sequence of binary symbols into groups with a constant number of clock intervals and into groups of code symbols with a new base of numeration in their subsequent conversion according to a definite alphabet, and predominantly with a new number of clock intervals. Thus, in the case of alphabetic coding, agreement is established between each binary group of symbols and the group of code symbols (or the groups in balanced codes [1, 2]). When the clock frequency is changed, it is necessary to transmit attributes which are sufficient during decoding to restore the boundaries (the frequency) of the groups of code symbols. Alphabetic coding is characterized by redundancy of the conversion of the binary information.

A common drawback to scrambling and coding is the possibility of error multiplication. The appearance of a single error during code symbol reception can cause errors in the binary information reproduced by the decoder over the extent of several clock intervals.

Nonalphabetic codes include the BnZS, HDBn and CHDBn [2], in which a sequence of binary zeros of specified length (equal to n + 1) is transmitted by a specified combination of pulses and pauses in the signal. The utilization of three elements of an AMI signal (see Table 2) is a common feature of the codes enumerated here. However, reducing the number of elements which do not contain clock frequency attributes [a signal element with a passive pause $s_9(t)$], as compared to AMI, violates the condition for the alternation of pulse polarity and increases the minimum upper frequency of the signal spectrum up to values of 0.6 f_T to 0.7 f_T [16]. The maximum noise immunity of signals of these codes corresponds to the quantity obtained for AMI.

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Alphabetic codes, besides improving the stability of the clock frequency attributes, increase the carrying capacity of a transmission system and lower its clock frequency. The following rule exists for the designation of alphabetic codes. The first number in the name of the code type indicates the number of symbols n, in the binary group being encoded. A binary numeration base is indicated by the Latin letter B (binary). The second number indicates the number of symbols, k, in a code group, while the last letter (or letters) reflects the new numeration base, M: T is ternary, Q is quaternary, QI is quinary, S is sextenary, H is heptanary, etc. [1, 2]. Thus, 4B3T code indicates that each group of four binary symbols is converted into a group of three ternary symbols.

It is obvious that the transmission of code symbols with an increased numeration base (M > 2) requires increasing the distinctive attributes, i.e., a multiposition signal. For example, to transmit ternary symbols, one can use any three elements of the video pulse signals depicted in Figure 2, or one, but with three amplitude values, or a mixture of elements and their amplitudes; for radio signals, one can use three position modulation (phase, frequency or amplitude), as well as mixed kinds of modulation.

It is sufficient to base a comparison of alphabetic codes on the following parameters: the number of groups of binary symbols, 2^n and code symbols, M^k (these parameters characterize the complexity of the realization of the coding and decoding operations); the coefficient of the change in the clock frequency, $K_M = n/k$; the code redundancy [1] is:

$$r = (k/n \log_2 M - 1)\ 100\%, \qquad (2)$$

and the maximum coefficient for the reduction in the clock frequency (r = 0) for any numeration base M, which, taking formula (2) into account is defined as:

$$K_{M\ max} = K_{M\ \text{макс}} = \lim_{r \to 0} n/k = \log_2 M.$$

In compiling a list of alphabetic codes, it is necessary to observe the condition for independent transmission of groups of binary symbols by combination of code symbols, i.e., the condition $2^n < M^k$. In the case of the limitation of an infinite set of code types within reasonable bounds – to tens of symbols in binary groups (code types up to 10BnM, for the realization of which, it is necessary to discriminate up to 10^{10} = 1,024 binary groups) and to a base seven numeration system – the list of alphabetic codes presented in Table 4 was compiled.

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The term (code type) adopted here is understood as a general algorithm for the conversion of binary symbols to code symbols. In Table 4, the 4B3T code is understood to be not only the 4B3T code [13], but also variants of it: MS43 [14] and FOMOT (four mode ternary) [16], and the L742 code [15] belongs to the 4B2H type code, etc.

The total number of code tables (a table showing the correspondence of the groups of binary symbols to the code symbol groups) is extremely large for every type of code. For example, for one of the simplest code types, 3B2T, the number of code tables is equal to the number of permutations of $2^3 = 8$ or $P_8 = 8! = 40,320$.

TABLE 4

Тип кода (1)	3^k	K_M	r, %	Тип кода (1)	4^k	K_M	r, %	Тип кода (1)	5^k	K_M	r, %	Тип кода (1)	6^k	K_M	r, %	Тип кода (1)	7^k	K_M	r, %
1B1T	3	1,00	58,50	1B1Q	4	1,00	100	1B1QI	5	1,00	132,19	1B1S	6	1,00	158,50	1B1H	7	1,00	180,73
2B2T	9	1,00	58,50	2B1Q	4	1,00	100	2B1QI	5	2,00	16,10	2B1S	6	2,00	29,25	2B1H	7	2,00	40,37
3B2T	9	1,50	5,66	3B2Q	16	1,50	33	3B2QI	25	1,50	54,60	3B2S	36	1,50	72,33	3B2H	49	1,50	87,16
4B3T	27	1,33	18,87	4B3Q	64	1,33	50	4B2QI	25	2,00	16,10	4B2S	36	2,00	29,25	4B2H	49	2,00	40,37
5B4T	81	1,25	26,80	5B4Q	256	1,25	60	5B3QI	125	1,66	39,32	5B2S	36	2,50	3,40	5B2H	49	2,50	12,29
6B4T	81	1,50	5,66	6B4Q	256	1,50	33	6B3QI	125	2,00	16,10	6B3S	216	2,00	29,25	6B3H	343	2,00	40,37
7B5T	243	1,40	13,21	7B5Q	1024	1,40	43	7B4QI	625	1,75	32,68	7B3S	216	2,33	10,78	7B3H	343	2,33	20,31
8B6T	729	1,33	18,87	8B6Q	2048	1,33	50	8B4QI	625	2,00	16,10	8B4S	1296	2,00	29,25	8B3H	343	2,66	5,27
9B6T	729	1,50	5,66	9B6Q	2048	1,50	33	9B4QI	625	2,25	3,20	9B4S	1296	2,25	14,89	9B4H	2401	2,25	24,78
10B7T	2187	1,43	10,95	10B7Q	8192	1,43	40	10B5QI	3125	2,00	16,10	10B4S	1296	2,50	3,40	10B4H	2401	2,50	12,29
$K_{3макс} = 1.58$				$K_{4макс} = 2,00$				$K_{5макс} = 2,32$				$K_{6макс} = 2,58$				$K_{7макс} = 2,81$			

Key: 1. Type of code.

A criterion is proposed for the comparison of codes and code tables of one type: the multiplication factor for single errors, which result during decoding. Although minimization with respect to the indicated criterion does not lead to a unique code table, it nonetheless substantially reduces the number of them. Thus, for a 3B2T code, it is reduced to eight with a minimum single error multiplication factor of 1.375.

The following conclusions can be drawn from an analysis of the list of alphabetic codes based on the proposed parameters: the following codes have an advantage in terms of closeness to the corresponding maximum coefficients of reduction in the clock frequency and code redundancy (with the clock frequency change coefficient indicated in parentheses): 3B2T (1.5); 9B4QI (2.25); 5B2S (2.5) and 8B3H (2.66), the small redundancy of which is sufficient to preclude extreme situations during the transmission of digital information (which lead to a loss of synchronization in transmission systems); if no substantial reduction in the transmit rate is required in the transmission systems, then the best code is the 3B2T based on the

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aggregate of parameters (redundancy, complexity in the realization of the coding and decoding operations and the error multiplication factor).

Conclusion. The utilization of the coding concept in the narrow sense of the word and the breakdown of methods of digital information transmission into coding and modulation operations make it possible to more completely consider the specific features of each operation and evaluate their potential capabilities. The subsequent selection of the set of operations for specific transmission systems should satisfy the requirements placed on digital communications networks. In this case, because of the strong dependence of some digital signal parameters on the statistical properties of the digital information being transmitted, they are to be determined for worst case conditions to obtain a qualitative comparison of the parameters.

It is expedient to combine the codes cited above with monopulse, bipulse and phase keyed signals to obtain the maximum noise immunity of repeater sections and the carrying capacity of digital transmission systems, as well as to meet the requirements for digital transmission system synchronization given any statistical characteristics of the digital information.

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DIGITAL COUPLING DEVICES (TRANSMULTIPLEXERS) IN COMMUNICATIONS SYSTEMS

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[Article by L.M. Gol'denberg, B.D. Matyushkin and M.N. Polyak]

[Text] The stage of the widescale introduction of digital transmission systems (TsSP) which use the principles of pulse-code modulation (IKM), as well as electronic digital switching systems, has started on the communication networks [1, 2]. Digital transmission systems are universal and can be used to organize a specified number of telephone channels (depending on the type of system) using time division of the channels (IKM VRK) or a group signal with frequency division of the channels (IKM ChRK). The principles of electronic switching of IKM VRK signals have been taken as the basis for the design of the automatic switching centers (UAK) and automatic electronic long distance telephone exchanges (EAMTS).

At the same time, the basis of the existing communications network is the multichannel analog systems with frequency division multiplexing [2]. For this reason, when organizing a long distance communications network, automatic switching centers should operate during a specific, rather long period of time using a combination system for the organization of transmission service (analog frequency division systems, time division PCM and frequency division PCM). Consequently, the construction of a communications network is impossible without the design of interfaces for systems with frequency and time division multiplexing, as well as devices for coupling frequency division systems to automatic switching centers (or EAMTS's). These coupling devices have been given the name "transmultiplexers" (TM).

This article is the first part of a paper devoted to the presentation of the design theory and principles for transmultiplexers using the modern component base, which are of considerable importance to other fields of communications engineering, for example, in the construction of digital modems, PCM to binary PCM signal converters, etc. Transmultiplexer circuits will be treated in the second part of the paper and a comparative analysis will be made of them.

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The functions of transmultiplexers. Transmultiplexers have been called upon to solve two basic problems: in the first place, the segregation of individual channel signals from a group frequency division signal for transmission to the switching field of an automatic switching center, and secondly, the generation of a group frequency division signal from the individual channel signals incoming from the switching field of an automatic switching center, for subsequent transmission in a frequency division system. These problems could in principle be solved using the set of standard terminal equipment of frequency division analog systems and PCM systems (for example, the K-60 and IKM-30 equipment for 60-channel frequency division systems). However, such a solution is hardly acceptable since it involves the use of comparatively expensive analog equipment, which is additionally of considerable size.

In step with the development of future digital transmission systems, when an integrated network comes to replace the existing communications network, in which signal transmission and switching is accomplished only in digital form, the functions of transmultiplexers will change to a certain extent. They will serve for interfacing wideband PCM system channels to time division PCM systems as well as to interface frequency division PCM systems and automatic switching centers (or EAMTS's).

In connection with what was said above, a number of requirements must be placed on the principles for the design and realization of transmultiplexers:

A TM should take the form of a digital device, the design principles of which are in line with the design principles for future digital transmission systems and couple to contemporary transmission systems by means of the simple addition of an analog-digital (or digital-analog) converter at the input (or output);

A TM should be designed around a digital component base with rather strict limitations on size and cost, something which is due to the need for a considerable number of interface equipment sets for coupling frequency division analog systems to one EAMTS;

The digital signal processors included in the complement of a TM should operate at high signal digitization frequencies (on the order of hundreds of kilohertz, even in the case of a 60 channel group signal).

The latter requirement leads to the use of special design methods for the devices and the development of procedures which assure the optimization of specific TM parameters (for example, the minimization of the standardized number of operations per unit time or equipment volume), as well as the newest series of integrated circuits which contain high speed immediate access (OZU) and read-only (PZU) memories, arithmetic-logic devices, registers and other functional assemblies.

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Transmultiplexer realization using a digital component base requires the application of digital methods of filtering, modulation and demodulation of the signals, which differ in many respects from the traditional analog methods. TM variants are treated in [3 - 7], the practical realization of which proved to be rather complicated because of their considerable volume.

TM structures were proposed in [8, 9] for 60-channel transmission systems which lead to a substantial reduction in the equipment volume through the standardization of the digital filters, the TsF, included in the structure. The standardization of the digital filters made it possible in the realization of TM's to use the latest achievements of the theory and practice of digital device synthesis (the construction of the digital filters using read-only memories and multichannel signal processing).

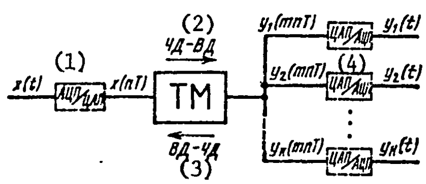

Figure 1.

Key: 1. Analog-digital converter/digital-analog converter;
2. Frequency division--time division;
3. Time division--frequency division;
4. Digital-analog converter/ /analog-digital converter.

As was noted above, a TM serves to interface systems with frequency and time division multiplexing (Figure 1). The frequency division and time division systems can be both digital and analog. In the latter case, analog-digital converters, ATsP, and digital-analog converters, TsAP, are incorporated in the transmultiplexer. The following symbols are adopted in Figure 1: x(t) and x(nT) are the group analog and digital signals from the frequency division, respectively; $y_i(m\,n\,T)$ and $y_i(t)$ are the digital and analog signals of the i-th channel ($i = 1, 2, \ldots k$).

The group k-channel analog signal with x(t) frequency division of the channels has a spectrum of:

$$X(i\omega)=\sum_{i=1}^{k} X_i(i\omega),$$

where $X_i(i\omega)$ are the spectra of the channel signals, each of which occupies a bandwidth of $\Delta\omega_i = \Delta\omega = \text{const}$. The group signal takes up a bandwidth of $[\omega_{min}, \omega_{max}]$, where $\omega_{min} = l\Delta\omega$, $\omega_{max} = (l + k)\Delta\omega$. The spectra of the individual channel signals, $y_i(t)$, in a time division system are located in the low frequency range $[0, \Delta\omega]$ (Figure 2)*.

* The absolute values of the signal spectra are depicted in Figure 2 and all the subsequent figures. However, for the sake of brevity in the following, we will be speaking of signal spectra when referring to the figures.

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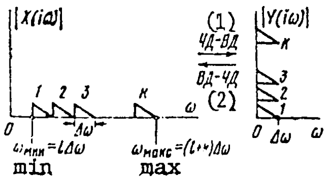

Figure 2.

Key: 1. Frequency division--
--time division;

2. Time division--frequency division.

The spectrum $X^*(e^{i\omega T})$ of the group digital signal $x(nT)$ is periodic with a digitization frequency of ω_0 [10]:

$$X^*(e^{i\omega T}) = \frac{1}{T}\sum_{n=-\infty}^{\infty} X(i\omega + in\omega_0) = $$
$$= \frac{1}{T}\sum_{i=1}^{k}\sum_{n=-\infty}^{\infty} X_i(i\omega + in\omega_0),$$

where the quantity ω_0 is chosen from the condition of [1]: $\omega_{max} \leq \omega_0 \leq$ $\leq \omega_{min}$ when $\omega_{max} < 2\omega_{min}$; $2\omega_{max} \leq \omega_0$ when $\omega_{max} > 2\omega_{min}$, $\omega_0 = \mu 2\Delta\omega$, $\mu = 1, 2, \ldots$, while the main spectrum (when $n = 0$) occupies a bandwidth of $|\omega| \in [\omega_{min}, \omega_{max}]$.

The spectrum $Y_i^*(e^{i\omega T'})$ of the digital channel signal, $y_i(m\,n\,T)$ is periodic with a digitization frequency of $\omega_0' = \omega_0/m = 2\Delta\omega$:

$$Y_i^*(e^{i\omega T'}) = \frac{1}{mT}\sum_{n=-\infty}^{\infty} X_i(i\omega + in\omega_0') =$$
$$= \frac{1}{mT}\sum_{n=-\infty}^{\infty} X_i^+(i\omega + in\omega_0') +$$
$$+ \frac{1}{mT}\sum_{n=-\infty}^{\infty} X_i^-(i\omega + in\omega_0'), \qquad (1)$$

(1)

where $X_i^+(i\omega)$ and $X_i^-(i\omega)$ correspond to the upper ($\omega > 0$) and lower ($\omega < 0$) bands of the spectrum of the i-th channel signal.

The main spectrum of the $y_i(m\,n\,T)$ signal occupies a bandwidth of $|\omega| \in [0, \Delta\omega]$.

A transmultiplexer should provide for the segregation of the $y_i(m\,n\,T)$ channel signals from the $x(n\,T)$ frequency division group signal with the shift of the spectra of the segregated channel signals into the low frequency range (direct conversion) and the generation of the group signal $x(n\,T)$ from the individual channel signals $y_i(m\,n\,T)$ (back conversion).

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The basic types of operations in a transmultiplexer. The main operations in a TM in the case of direct conversion are: digital filtering of the group signal (to segregate the spectra of individual channels); the shifting of the spectrum of a segregated channel signal to the low frequency range and the reduction of the digitization frequency of the channel signals (decimation) down to the quantity ω_0'; the inversion of the spectrum of specific channel signals.

The main operations in a TM in the case of back conversion are: digital filtering of the individual channel signals for the purpose of boosting the digitization frequency (interpolation); shifting the psectrum of each channel signal to the requisite frequency band; preliminary inversion of the spectrum of individual channel signals.

The segregation of a channel signal from the group signal with simultaneous conversion of the segregated spectrum to the low frequency region. This problem is solved by means of the circuit shown in Figure 3, which was described in detail in [14].

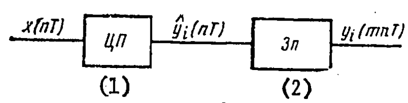

Figure 3.

Key: 1. TsP [probable misprint for TsF = digital filter];
2. 3n [inhibit gate].

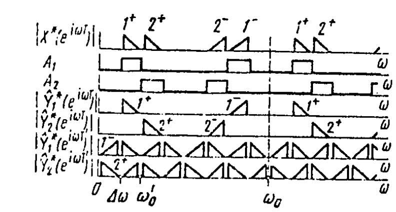

Figure 4.

The segregation of an individual channel is accomplished by means of a digital bandpass filter TsF, while the shift of the spectrum of the segregated signal to the low frequency region is realized by means of a simple reduction in the digitization frequency of the output signal of the digital filter by a factor of m, using an inhibit gate, 3n. The principle of the segregation of a channel signal from a group to channel signal ($l = 1$, $k = 2$, see Figure 2) with the shift of the spectrum to the low frequency region, where $X^*(e^{i\omega T}$ is the spectrum of the two channel group signal x(n T) is illustrated in Figure 4. A_1 and A_2 are the idealized amplitude-frequency responses (AChKh) of the digital filters for the segregation of the first and second channels; $\hat{Y}_1^*(e^{i\omega T})$ and $\hat{Y}_2^*(e^{i\omega T})$ are the spectra of the $\hat{y}_1(nT)$ and $\hat{y}_2(nT)$ signals at the outputs of the

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filters which segregate the first and second channels; $Y_1^*(e^{i\omega T'})$ and $Y_2^*(e^{i\omega T'})$ are the spectra of the $y_1(m\cdot n\ T)$ and $Y_2(m\ n\ T)$ signals following the reduction of the digitization frequency. As can be seen from Figure 4, either the direct spectrum of the channel being segregated (for the second channel) or the inverse spectrum (for the first channel) is positioned in the low frequency region.

We shall note the following circumstances: The idealized amplitude-frequency response of filter A_i for the segregation of the i-th channel in the passband $|\omega| \in [0, \omega_0/2]$ is defined by the relationship:

$$A_i = |H^*(e^{i\omega T})| = \begin{cases} 1 \text{ при } \omega \in [(l+i-1)\Delta\omega,\ (l+i)\Delta\omega], & (2) \\ 0 \text{ при } \omega \overline{\in} [(l+i-1)\Delta\omega,\ (l+i)\Delta\omega], \end{cases}$$

where

$$H^*(e^{i\omega T}) = H^*(z)\big|_{z=e^{i\omega T}},\ H^*(z)$$

is the transfer function of the digital filter.

In a real digital filter, expression (2) is approximated with a specific degree of accuracy, $A_i \neq 0$ when $\omega \in [(l+i-1)\Delta\omega,\ (l+i)\Delta\omega]$, and with the reduction of the digitization frequency, the phenomenon of the superposition of the spectra takes place, something which leads to signal distortions. Because of this, rather severe requirements are placed on the parameters of the amplitude-frequency response of a channel filter in the top band.

If one additionally takes into account the fact that the ratio of the sum of the transition bands (the region between the passband and the stopband) to the sum of the passbands and stopbands is very small (even for a 12 channel group, it is approximately 0.033), considerable difficulties arise in the realization of the channel filters.

Moreover, it can be seen from (2), the transfer functions of the filters for the segregation of the various channels are different, and for this reason, the design of a coupling device based on this principle requires the utilization of a set of bandpass filters.

Another solution of the problem exists where one can use the same digital low pass filter for the segregation of any channel (Figure 5). Figure 6 illustrates the operational principle of the circuit. The input group signal $x(n\ T)$ with a spectrum of $X^*(e^{i\omega T})$ is multiplied by the carrier so that the spectrum $\hat{X}_i^*(e^{i\omega T})$ of the channel being isolated is positioned in the frequency range $[-\Delta\omega/2,\ \Delta\omega/2]$ (Figure 6b). Then the requisite

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channel is isolated by means of a low pass filter (Figure 6c, d), the digitization frequency is reduced (Figure 6e) and the back conversion of the spectrum to the frequency range $[0, \Delta\omega]$ is realized (Figure 6f). It should be noted that the input and output signals of the low pass filter are complex. This can also be seen from Figures 6b-d, since the spectra of the signals are not symmetrical relative to the frequency $\omega = 0$.

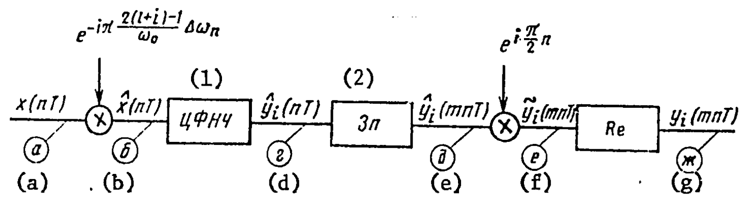

Figure 5.

Key: 1. TsFNCh [digital low pass filter]; 2. Зn [inhibit gate].

The processing of the complex signal is accomplished by a "complex" circuit. This means that there are separate branches for the processing of the real $\hat{x}_1(n\ T)$ and imaginary $\hat{x}(n\ T)$ components of the signal $\hat{x}(n\ T)$. The operational principle of the "complex" circuit of Figure 5 is illustrated in detail in Figure 7. The signal with the requisite spectrum (1) is derived through the segregation of the real component of the signal $\tilde{y}_i(m\ n\ T)$ (the element Re in Figure 5).

The inversion of the spectrum of a signal located in the low frequency range. If the primary signal spectrum $z(n\ T)$ occupies a bandwidth of $|\omega| \in [0, \Delta\omega]$, while the digitization frequency is $\omega_0 = 2\Delta\omega$, the spectrum inversion is accomplished by means of a simple change in the sign of every second signal readout $\hat{z}(n\ T)$: $\hat{z}(n\ T) = (-1)^n z(n\ T)$, $n = 0, 1, \ldots$, where $\hat{z}(n\ T)$ is the signal, the spectrum of which is inverted with respect to the $z(n\ T)$ signal spectrum. The relevant explanations can be found, for example, in [8].

Signal interpolation by means of a digital filter. In pure form, the interpolation problem consists in boosting the digitization frequency of a signal having a finite spectrum. The analysis of the interpolation process in the frequency range [12] shows that this process is essentially one of linear digital filtering of the signal being interpolated.

If $z(n\ T)$ is the signal being interpolated and having a spectrum $Z^*(e^{i\omega T})$, and $f([n/m]T)$ is the signal following interpolation (increasing the digitization frequency by a factor of m) having a spectrum $F^*(e^{i\omega T})$, then as can be seen from Figure 8 ($m = 3$), the interpolation process consists in suppressing the components of the spectrum of the signal $z(n\ T)$ which fall in a range of $[\pi/T, \pi/T']$. This problem can be solved by means of a digital low pass filter, to the input of which the sequence of readouts is fed:

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$$\tilde{z}(nT') = \begin{cases} z\left(\frac{n}{m}T\right) & \text{for (при) } n = 0,\ m,\ 2m,\ \dots, \\ 0 & \text{for other n. (при других } n\text{.)} \end{cases}$$

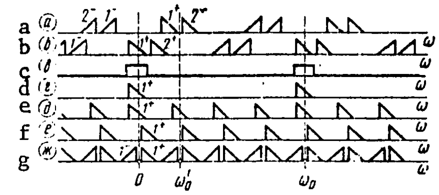

Figure 6.

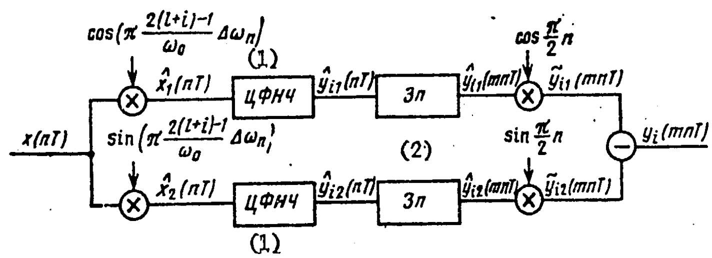

Figure 7.

Key: 1. Digital low pass filters;
2. Inhibit gates.

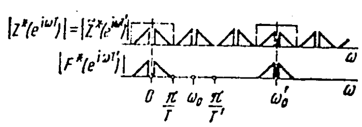

Figure 8.

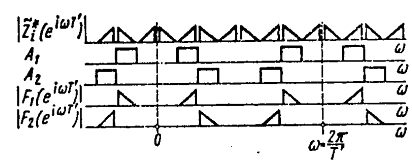

Figure 9.

In other words, the sequence z(n T') is derived from the sequence z(n T) by means of inserting (m-1) of the zero readout between two information readout sequences. It was shown in [12] that the spectrum $\tilde{Z}^*(e^{i\omega T'})$ of the signal $\tilde{z}(n\ T')$ coincides with the spectrum $Z^*(e^{i\omega T})$ of the signal z(n T) (see Figure 8), while the signal with a spectrum of $F^*(e^{i\omega T'})$ is attained by means of processing the signal $\tilde{z}(n\ T')$ with a digital low pass filter operating at a digitization frequency of $\tilde{\omega}_0' = 2\pi/T'$:

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$$F^{*}(e^{i\omega T'}) = H^{*}(e^{i\omega T'})\tilde{Z}^{*}(e^{i\omega T'}).$$

The interpolation process can be combined with the shifting of the spectrum of the signal being interpolated to the requisite frequency band. For this, it is sufficient to use a digital bandpass filter as the filter-interpolator. Figure 9 illustrates the process of interpolation with simultaneous spectrum conversion. The following symbols have been adopted in this figure: $Z_i^*(e^{i\omega T'})$ is the spectrum of the input signal to the i-th digital filter; A_1 and A_2 are the amplitude-frequency responses of the digital bandpass filter-interpolators which accomplish the simultaneous conversion of the primary spectrum of the signal being interpolated to the frequency ranges $[\Delta\omega, 2\Delta\omega]$ and $[2\Delta\omega, 3\Delta\omega]$ respectively; $F_1 e^{i\omega T'})$ and $F_2(e^{i\omega T'})$ are the signal spectra at the digital filter output. In comparing Figure 4 and Figure 9, it can be noted that when interpolating signals, digital filters are used having the same transfer functions as in the case of the segregation of individual channels from the group signal with simultaneous conversion of the spectrum to the low frequency range. For this reason, the same drawbacks are inherent in the interpolation method described here with spectrum conversion as in the corresponding method of isolation of individual channels.

The generation of a single sideband signal with subsequent conversion of the spectrum. The process of signal interpolation and spectrum conversion to the requisite frequency band can be carried out in two stages using the circuit shown in Figure 10, where the digital filter-interpolator is designated TsFI. Figure 11 explains this process. We note that the signals $\tilde{z}(n\ T)$, $\hat{z}(n\ T)$, $\hat{f}(n\ T')$ and $\tilde{f}(n\ T')$ are complex and the circuit of Figure 11 contains two branches for processing the real and imaginary components of the complex signal, just as the circuit of Figure 5.

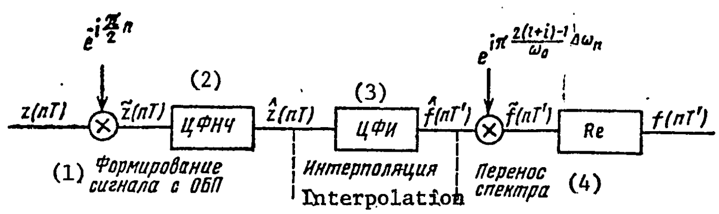

Figure 10.

Key:
1. Generation of the SSB signal;
2. Digital low pass filter;
3. Digital interpolation filter;
4. Spectrum conversion.

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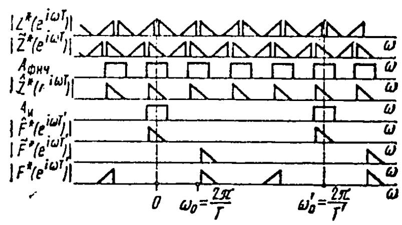

Figure 11.

To be included among the merits of this circuit is the fact that the operations of generating the single sideband signal (SSB) is accomplished at the low frequency, and in the case of signal interpolation, there is considerable spacing between adjacent components of the spectrum (see Figure 11). This greatly simplifies the transfer functions of the digital filters used as well as their realization. The generation of a SSB signal at the low frequency can also be realized by means of the converter treated in [13].

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THE SERVICE QUALITY OF AUTOMATIC LONG DISTANCE TELEPHONE SERVICE

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[Article by A.B. Klibaner and N.V. Pevtsov]

[Text] *Introduction.* The fast rate of growth in the number of long distance telephone conversations and the placement of new types of AMTS [automatic long distance telephone exchanges] in service have created a number of serious scientific and engineering problems related to the requirement that the service quality for subscribers be increased [1]. One of the major characteristics of service quality in the case of automatic long distance telephone service is the losses.

Trunk groups of expensive channels are organized between various AMTS's, where these channels must be utilized to the maximum possible extent. This can be achieved with a reduction in service quality, i.e., increasing the amount of losses [2], something which is unacceptable. The capacity of a trunk group of channels, in accordance with CCITT recommendations, should assure losses of no more than one percent. Optimal solutions must be found which tie together the requirements of increasing both the level of channel usage and communications quality.

Various factors have an influence on the quality of automatic long distance service, and among them we shall single out the following main ones: the variation in the load on the equipment and the line; the perceptible flow of repeat calls; and the lack of operationally timely monitoring of communications traffic.

Moreover, reasons for poor service quality can be inadequate knowledge of the nature of the load incoming to an AMTS, methods of calculating it which are not always satisfactory, the lack of the technical capability of controlling the incoming load and distributing it. We shall consider the three major influencing factors which we have singled out.

Fluctuations in the long distance load. The long distance load, Y_M, produced by call sources is a random quantity and subject to fluctuations.

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These are random fluctuations in the load intensity from long distance message sources, as well as seasonal fluctuations.

The design procedure for AMTS's takes into account the nonuniformity of the long distance load as a function of the seasons of the year. The average number of incoming calls per 24 hours is determined by means of dividing the annual level by 300 days instead of 365 so that the load is increased by approximately 20 percent. A similar approach in the area of large loads can lead to a substantial overstatement of the planned equipment volume. If, let us assume, with losses of one percent and an incoming load of 10 Erlangs, 18 channels are actually required, then taking this 20 percent into account, there will be more than 20 channels; the volume of channel equipment will be increased by 10 percent. Given the same losses, an increase in the load of 100 to 120 Erlangs leads to an increase in the volume of channel equipment by more than 20 percent.

Consequently, it is insufficient to characterize the random quantity Y_M with only the mean value. Deviations in the values, which can be treated as a random quantity, can be so great that the quality of a long distance message can prove to be unsatisfactory.

Because of this, the calculation of AMTS equipment volume must be based on that level of the incoming load at which there is the probability ω of assuring the requisite service quality to subscribers with a specified standard norm for the losses. Such a level has come to be called the design level.

The theory of the design load for municipal telephone exchanges is presented in [2], where a relationship is found between the average Y and design Y_p load levels when $\omega = 0.75$:

$$Y_{\text{p}} = Y + 0{,}674\sqrt{\bar{Y}}. \qquad (1)$$

Taking into account the specific features of fluctuations in the incoming long distance load, the design long distance load, $Y_{p.m}$, can be determined from the formula proposed in [5]:

$$Y_{\text{p.м}} = \bar{Y}_{\text{м}} + \varphi_1(\bar{Y}_{\text{м}}) + \varphi_2(\bar{Y}_{\text{м}}), \qquad (2)$$

where $\bar{Y}_M$ is the average value of the long distance load; $\phi_1(\bar{Y}_M)$ is a function which takes into account the random fluctuations in its sources; $\phi_2(\bar{Y}_M)$ is a function which takes into account the seasonal fluctuations in the load.

We will note that in accordance with the existing design procedure of [1]:

$$\varphi_1(\bar{Y}_{\text{м}}) = 0, \quad \varphi_2(\bar{Y}_{\text{м}}) = 0{,}2\,\bar{Y}_{\text{м}}. \qquad (3)$$

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and only the seasonal fluctuations are reflected in it, something which is clearly inadequate for the characteristics of a long distance load.

A detailed analysis of the statistical data from materials of the Leningrad AMTS and a hypothesis concerning the very simple nature of call flow and an exponential servicing time with an average value of τ = 0.05 hours has made it possible to derive the following formulas, taking (1) and (3) into account:

$$\varphi_1(\bar{Y}_м) = 0,215\sqrt{\bar{Y}_м}, \tag{4}$$

$$\varphi_2(\bar{Y}_м) = \bar{Y}_м \exp[-(1,6 + 0,01\bar{Y}_м)]. \tag{5}$$

Formula (2) assumes the following form in this case:

$$Y_{р.м} = \bar{Y}_м\left[1 + e^{-(1,6+0,01\bar{Y}_м)}\right] + 0,215\sqrt{\bar{Y}_м}. \tag{6}$$

An analysis of formula (6) shows that the greater the value of $\bar{Y}_M$. the less the design value Y_{pM} differs from it. Table 1 illustrates what has been said here. The physical essence of the fluctuations is reflected here: the greater the load, the smaller the fluctuations to which it is subjected.

TABLE 1

$\bar{Y}_м$, Эрл	$Y_{рм}$, Эрл	$\frac{Y_{рм} - Y_м}{\bar{Y}_м} \times 100\%$
0,1	0,188	88
1	1,415	42
10	12,49	24
100	102,15	2,1

In the planning of an AMTS, the design value of the long distance load is used only to determine the number of junction and control devices. All operations to combine and split the flows of long distance traffic at various selector stages of the AMTS and the transition to the calculated design values are realized in accordance with the generally well known algorithm of [2].

The design long distance load, determined from formula (6), is justified for the calculation of the number of channels or devices, which are inserted in a fully accessible manner. Studies show that the design long distance loads, in the case of incompletely accessible insertion, with a small amount of error can also be taken as equal to $Y_{p.M}$.

Since load fluctuations lead to an increase in the average traffic losses planning an AMTS with the use of the design load method makes it possible to specify regulations for the mean level of losses on a communications route.

Setting loss standards. As has already been mentioned, the main characteristic of service quality is the losses. In accordance with [1], losses are standardized for a switching section of a long distance telephone

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TABLE 2

Item No.	Route	Peak Load Hours
1	Vil'nyus	20-21
2	Tallin	20-21
3	Arkhangel'sk	20-21
4	Volgograd	20-21
5	Riga	10-11
6	Kalinin	10-11
7	Novgorod	9-10
8	Voronezh	14-15
9	L'vov	15-16
10	Moscow	15-16
11	Intrazonal routes	10-11

network, i.e., the trunk group channels between AMTS's, without taking into account the switching equipment of the AMTS's; a norm of one percent has also been adopted there. At the same time, the operational instructions for an ARM-20 AMTS sets the standard for failures in an exchange at a level of two percent, taking into account failures on the traffic routes, leaving one percent for losses due to the channels being busy. Such an approach is correct, however, it is useful to specify the two percent level more precisely. In fact, losses at an AMTS are also due to the occupancy of the channels, instruments as well as blockages in the switch field; losses can occur because of row overflow and the limited waiting time in the control devices which service the calls in a system having waiting. For this reason, it is more expedient to understand the switched section of a long distance telephone network to be the aggregate of the AMTS equipment and the long distance channel, where the input to the section is the input to the switching equipment, while the output is the sets of terminal equipment of the channels or lines.

The question of setting loss standards for AMTS's requires careful study. First, the overall standard for losses from subscriber to subscriber in the case of a long distance call must be determined, and then this quantity must be distributed over the switched sections. The loss standard for a section will depend on the type of AMTS, something which will make it possible to reflect the actual cost relationships on the network. In this case, it is necessary that the loss standard for a trunk group of channels not exceed one percent in accordance with modern requirements.

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In the case of loss distribution, it is necessary to take into account the specific features of each particular communications route. It was shown in [4] that routes can exist, for which a degradation of service quality is permissible from an economic point of view. This question can be resolved in precisely specifying the network structure and the existing equipment. The principles for the distribution of the overall loss standard for the switched sections of a long distance network are the topic of a separate article.

The organization of quality control for AMTS service. The organization of observations of service quality and the measurement of the parameters of long distance telephone messages are complex technical problems, the solution of which generates a number of definite difficulties for operational services. One of them is the lack of agreement in the peak load hours on different communications routes.

Data on the peak load hours on trunk line and intrazonal routes of the Leningrad AMTS are shown in Table 2.

The main load peaks come at time intervals of 10 - 11, 14 - 15 and 20 - 22 hours [Moscow time]. A system of statistical measurements makes it possible record the data of the meters every hour. It is natural that the lack of agreement in the peak load hours for the routes creates difficulties in determining the overall exchange peak load hour. The latter is determined during the period of maximum current consumption at the station, and for the Leningrad AMTS, is the interval from 10 to 11 hours. For this reason, all of the statistical data on the load and the service quality are recorded specifically during this period. A similar picture is also observed at other of the nation's AMTS's.

It is expedient to measure the telephone traffic parameters and monitor the service quality only during the peak load hours on the communications routes. This increases the value of the statistical data for planning. The volume of statistical data does not increase though as compared to the volume obtained using existing procedures.

It is expedient to utilize check calls for the purpose of obtaining subjective information from a subscriber on how the long distance telephone network is functioning. Such a method has been used for a long time now at the Leningrad GTS [municipal telephone exchange]. Check calls are made systematically during the peak load hours of the communications routes. It is desirable to automate this process, because of which it is necessary to develop special receiving and transmitting test equipment. The monthly generalization of the observation results will permit a more complete representation of the situation on the network, while the daily data can be used by the AMTS operational services as monitor information.

The combined observation of long distance load fluctuations and the results of check calls will make it possible to come up with a real statistical picture of the quality of long distance telephone traffic.

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Conclusion. The problems considered here are related to improving the quality of long distance telephone service and the necessity of more complete accounting for the specific features of long distance telephone service, for which the following are recommended:

—The calculation of AMTS equipment volume based on the design level of the long distance load;

—Determine the set standard for the losses on a switched section of a long distance network, taking into account AMTS losses;

—Make the measurements of the telephone traffic parameters and monitor service quality during the peak load hours of the communications routes;

—Make systematic check calls at an AMTS.

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CONVERTERS, INVERTERS, TRANSDUCERS

UDC 534.232

BROAD-BAND MATCHING OF PIEZOTRANSDUCERS OF ACOUSTO-OPTICAL DEVICES

Kiev IZVESTIYA VUZOV SSSR - RADIOELEKTRONIKA in Russian No 3, 1980 pp 98-101

[Article by A. V. Yurchenko, manuscript received 14 Feb 79; after revision, 15 June 79]

[Text] The solving of the problem of optimal broad-band matching of arbitrary impedance amounts to finding the function of the reflection coefficient Γ_1 which must satisfy a number of integral relations obtained from the conditions of physical feasibility of the matching circuit and for which the deviation from zero in the prescribed frequency band is minimal $(|\Gamma_1| \leq |\Gamma_1|_{max})$. This solution turns out to be sufficiently simple only for the case of a load consisting of not more than two reactive elements [1]. Broad-band matching of acousto-optical devices presents considerable difficulties since even the simplest equivalent circuit of a loaded piezotransducer shown in Figure 1a [2] contains three reactive elements C_1'-L_1'-C_2', and the application of the general theory [1] is difficult. Therefore, the dependence of the resistance of the dynamic branch on the frequency is usually disgarded in the matching band, and the combination of the frequency-independent effective resistance and the static capacitance of the piezoplate is considered as a load being matched [3,4]. However, as will be shown below, such simplification leads in many instances to a substantial difference of the calculated matching circuit from the optimal one, and it becomes necessary to take the dynamic branch into consideration.

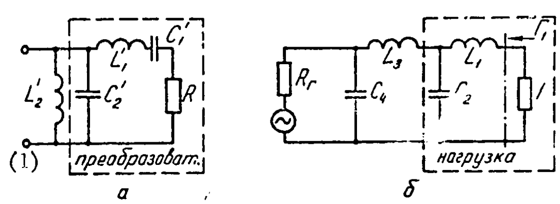

Figure 1

Key: 1. Transducer
2. Load

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The purpose of this work is to design a matching circuit for a load approximated by a circuit (Figure 1a, "transducer") in which the reactances of the dynamic branch are taken into consideration. In order to evaluate the degree of the influence of these reactances on the parameters of the coupling circuit, let us examine an equivalent circuit of the transducer together with the external compensating inductance L_2'. If the value of L_2' is selected in such a way that the oscillation circuit formed by L_2' and C_2' is tuned to the resonance frequency of the dynamic branch ω_0, then the entire circuit shown in Figure 1a which has the structure of a band filter can be replaced with its low-frequency prototype (Figure 1b, "load"). By normalizing the reactive elements of the load, it is always possible to reduce the value of the terminal effective resistance R to unity. For such two-element (by the number of the reactive elements) load, it is easy to synthesize an optimal matching circuit [1] with a cutoff frequency of ω_c which is then recalculated to the band frequency with a passband numerically equal to the passband of a low-frequency circuit. The function Γ_1 of such a circuit must satisfy two equations obtained from the condition of physical feasibility [1]:

$$\int_0^\infty \ln 1/|\Gamma_1|\, d\omega = \pi\left(A_1 - 2\sum_i p_{ri}\right)/2,$$

$$\int_0^\infty \omega^2 \ln 1/|\Gamma_1|\, d\omega = -\pi\left(A_3 - \frac{2}{3}\sum_i p_{ri}^3\right)/2.$$

Here, A_1 and A_3 are the coefficients of the Taylor expansion of function ln $1/|\Gamma_1|$ in the neighborhood of zeros of the transmission coefficient of the quadripole L_1-C_2 (Figure 1b), p_{ri} are the zeros of Γ_1 lying in the right half-plane of the complex frequency p. A_1 is defined only by L_1, A_3-L_1 and C_2. Depending on the relation of L_1 and C_2, the value of the integral in the second equation can be greater or smaller than $\pi A_3/2$. If it is greater, than the optimal tolerance is achieved only by the introduction of zeros of Γ_1 in the right half-plane, i.e., by giving up the simple structure of the matching circuit which is easily realized on the superhigh frequency. If it is smaller, then the optimal tolerance is achieved by using the matching circuit with an input shunting capacitance, which increases the capacitance C_2. This case is the most interesting for use, since the structure of the matching circuit is very simple.

The limitations imposed on L_1 and C_2 in accomplishing the second case are found directly from the condition

$$\int_0^\infty \omega^2 \ln 1/|\Gamma_1|\, d\omega \leqslant -\pi A_3/2 \qquad (1)$$

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for any type of function Γ_1. However, representation of these limitations in an analytical form is simple only when there is an ideal rectangular form of the dependence of $|\Gamma_1|$ on the frequency. In approximating the rectangular characteristic by functions permitting physical realization of the matching circuit with a finite number of elements, the value of the integral will depend on the kind and properties of the approximating function; therefore, it is not practical to obtain analytical expressions for some concrete approximation. In order to find the domain of values of L_1 and C_2 where (1) is fulfilled, let us use the curves for finding the values of the elements of a matching circuit with the Tschebyscheff characteristic for a load with one reactance of L_1^I given in [5]. For this, let us examine the value of $L_1C_2 = Q_H/Q$, where $Q_H = \omega_0 L_1$ and $Q = \omega_0 C_2$ are the quality factors of the reactive elements of the load on the resonance frequency. It is evident that, if for the load in question $Q_H/Q \geqslant A$, where the value of $A = L_1^I/C_2^I$ is calculated by the curves plotted for a load with one reactance, then condition (1) will be fulfilled. Having plotted the dependence of $A(\delta)$, where $\delta = 1/\omega_c L_1$ is the decrement of the load being matched on the critical frequency of the passband ω_c, we shall find the domain being sought. The curves of such a dependence for the number of reactive elements n = 4 (Figure 1b) are shown in Figure 2. By using these curves for a prescribed impedance with a two-element low-frequency prototype, it is possible to find the minimum value of δ_{min} for which it is possible to construct an optimal coupling circuit with a simple structure of a band filter. For smaller δ, it is necessary to accomplish circuits having zeros of Γ_1 in the right half-plane.

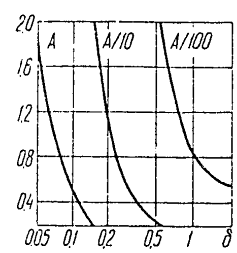

Figure 2

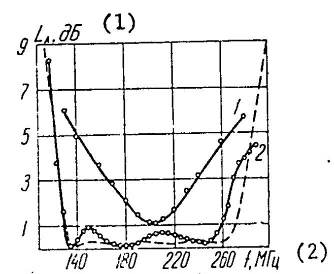

Figure 3
Key: 1. dB
2. MHz

From the analysis of the input characteristics of the loaded piezoplate, it is easy to establish that the values of Q_H and Q do not depend either on the resonance frequency ω_0 or on the area of the transducer, but are only a function of the material constants of the piezoelectric and acoustic line. This makes it possible to find a relative cutoff frequency band W_{max} for this transducer -- acoustic line combination in which this device can be matched by means of a simple optimal matching circuit:

$$W_{max} = \omega_{c\,max}/\omega_0 = 1/\delta_{min} Q_n .$$

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For example, for a converter of longitudinal waves made of a 35 degree yz $LiNbO_3$ cut using fused quartz, the calculated values of $Q_H = 3.13$; $Q_H/Q = 3.0$; $\delta_{min} = 0.42$, which corresponds to $W_{max} = 0.76$. The values of the elements of the matching circuit are calculated either by formulas [1], or by the curves of [5] for a one-element load. It is evident that, if the desired matching band is smaller than the critical one ($\delta > \delta_{min}$), then the matching circuit can be calculated by the same method. The maximum value of reflection losses $L_{A\ max} = 10 \lg 1/(1-|\Gamma_1|^2 max)$ and the value of the pulsations of the Tschebyscheff characteristic $L_{A\ r}$ are found by the curves given in [5].

The above method was used to calculate the matching circuit for an acousto-optical modulator using fused quartz with an $LiNbO_3$ longitudinal wave converter welded to the acoustic line by the method of diffusion vacuum welding. The type of the frequency dependence of the reflection losses L_A of such a device is shown in Figure 3 (broken curve). The equivalent circuit constructed by the experimental input characteristics had the following parameters: $Q_H=2.94$; $R=43.4$ Ohm; $C_2'=18.1$ picofarads; $L_1'=108$ nanohenries; $f_0=188$ MHz; $Q_H/Q=3.18$. The matching band $W=0.68$ (and, respectively, $L_{A\ max} = 0.33$ dB) were selected in such a way as to eliminate the necessity of using an ideal transducer in the matching circuit for recalculating the resistance of a standard 75-Ohm tract. After converting the low-frequency prototype into a band structure, the calculated circuit was designed and constructed with the use of elements with lumped parameters. In this case, in order to obtain optimal matching, it was necessary to include an output shunting capacitance increasing the value of the input capacitance of the load being matched C_2'. The calculated passband $f_c=128$ MHz, critical frequencies of the passband $f_{-1}=-f_c/2+\sqrt{f_c^2/4+f_0^2}=134{,}5$ MHz, $f_1=f_{-1}+f_c=262.5$ MHz, pulsations of the Tschebyscheff characteristic $L_{Ar} \approx 0.1$ dB.

The measured values of reflection losses L_A for an acousto-optical cell with a matching circuit (curve 2) and without it (curve 1) are shown in Figure 3. Some difference of the measured characteristic from the calculated characteristic can, evidently, be explained by the inaccurate representation of the real arrangement of the equivalent circuit, a slight deviation of the values of the elements of the matching circuit from the calculated values, and their final quality factor, as well as by the presence of parasitic elements which were not taken into consideration in calculations.

The above results indicate that the use of the described method makes it possible to design simple matching circuits which widen substantially the band of acousto-optical devices and decreased the value of reflection losses.

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INSTRUMENTS, MEASURING DEVICES AND TESTERS, METHODS OF MEASURING, GENERAL EXPERIMENTAL TECHNIQUES

UDC 538.57

OPTICAL HETERODYNE METHOD FOR STUDYING SURFACE ACOUSTIC WAVES

Kiev IZVESTIYA VUZOV SSSR - RADIOELEKTRONIKA in Russian No 3, 1980 pp 94-96

[Article by Yu. I. Posudin and V. G. Gritz, submitted 5 Feb 1979]

[Text] One of the main problems in studying surface acoustic waves (PAV) is precise determination of the main parameters of PAV: propagation speed, amplitude, attenuation, losses, etc.

This article describes theoretical and experimental studies on PAV by the method of optical heterodyne detection which was proposed for the first time in [1]. The essence of this method is as follows: a laser beam is passed through a polished crystal on whose surface a PAV is excited. As a result of periodic perturbations of the surface and the refraction index of the material of the crystal, the laser beam defracts and its frequency increases or decreases by the value $m\Omega$, where m is the order of the diffraction; Ω is the PAV frequency. The difficulty is that the laser beam diffracted on the PAV is by several orders smaller in intensity in comparison with an undefracted beam. Therefore, registration of PAV with a rather small amplitude is a difficult problem. The heterodyne method makes it possible to avoid these difficulties due to the fact that the laser beam diffracted on the PAV combines with the reference beam which has a rather large amplitude. As a result of this, the photodetector registers a signal with a differential frequency Ω and an amplitude proportional to the product of the amplitudes of the diffracted and reference waves. The signal carries information about the parameters of the PAV.

Let us designate the electric field and the time-averaged optical power of the diffracted light wave on the photodetector as

$$E_{д} = E_1 \sin(\omega + \Omega)t; \quad P_1 = 0{,}5\sqrt{\varepsilon/\mu}E_1^2, \tag{1}$$

where ω -- optical frequency; $\sqrt{\varepsilon/\mu}$ -- parameter characterizing the optical absorption of the medium. For the reference wave we have:

$$E_{оп} = E_0 \sin(\omega t + \varphi_0); \quad P_0 = 0{,}5\sqrt{\varepsilon/\mu}E_0, \tag{2}$$

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where φ_0 -- certain relative optical phase.

Since the photoelectric process is characterized by a linear dependence between the number of incident photons and the number of the carriers of free charges, the current of the photodetector can be represented as

$$I_p = \eta e P/h\nu, \tag{3}$$

where P -- total optical power at the detector input; hv -- energy of one photon; e -- electrical charge; η -- quantum effectiveness.

For our case of two waves, the optical power is

$$P = 0{,}5\sqrt{\varepsilon/\mu} < |E_{\text{оп}} + E_{\text{д}}| \tag{4}$$

(here, the averaging is done over optical frequency).

Substituting (1) and (2) in (4) and averaging, we obtain

$$P = 0{,}5\sqrt{\varepsilon/\mu}\,[E_0^2 + 2E_0E_1\cos(\Omega t - \varphi) + E_1^2]. \tag{5}$$

Thus, the current of the photodetector $I_{\text{Д}}$ contains a component I_Ω with the frequency of the PAV whose amplitude depends on φ_0

$$I_\Omega = \frac{\eta_e}{h\nu}\sqrt{\frac{\varepsilon}{\mu}}\,E_0E_1 = \frac{2\eta_e}{h\nu}\sqrt{P_0P_1}. \tag{6}$$

Expression (6) shows clearly the advantage of the heterodyne method ($I_\Omega = E_0E_1$) in comparison with the method of direct optical detection [2] ($I_{\text{h}} \equiv E_1^2$).

The above expressions were obtained for a traveling acoustic wave. In practice, there take place various reflections from the boundaries of the crystal, therefore it is necessary to consider the nature of the PAV propagating along the crystal. As a result of this, the light wave diffracted on such a PAV, will have the following form:

$$E_{\text{д}} = E_1^+\sin(\omega + \Omega)t + E_1^-\sin(\omega - \Omega)t, \tag{7}$$

where E_1^+ and E_1^- -- forward and backward waves, respectively.

The reference wave will be described by expression (2). In this case, the average optical power on the photodetector is

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$$P = 0{,}5\sqrt{\varepsilon/\mu}\,[E_0^2 + E_1^2 + E_1^{-2} + 2E_1^+E_1^- \cos 2\Omega t + 2E_0E_1^+ \cos(\Omega t - \varphi) + \\ + 2E_0E_1^- \cos(\Omega t + \varphi)]. \tag{8}$$

Oscillations with a frequency of 2Ω can be disregarded due to their small amplitudes. The main contribution is made by the components with a frequency of Ω. It follows from the analysis of expression (8) that, depending on the phase φ, the amplitude of the photocurrent will be limited by two limits:

$$I = (\eta_e/h\nu)\sqrt{\varepsilon/\mu}E_0 \begin{cases} |E_-^{\pm} + E_1^-| & \text{at } \varphi = m\pi, \\ |E_1^+ - E_1^-| & \text{at } \varphi = (m+0{,}5)\pi. \end{cases} \tag{9}$$

In the absence of the reflected wave E_1^-, expression (9) is transformed to (6).

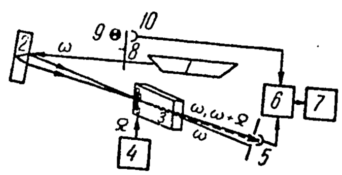

Figure 1

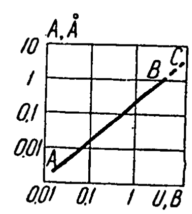

Figure 2

A block diagram of the experimental unit is shown in Figure 1. The laser beam 1 is split with the aid of the plate 2 into two beams one of which was used for diffraction, the other as a reference beam. Lithium niobate polished on two sides was used as the crystal 3 on which the interaction of the laser beam with the PAV took place. Stripline transducers calculated for 27 MHz where sprayed on one of the sides. A GZ-41-type generator 4 was used for PAV excitation. Diffraction occurred under the effect of the laser light from the PAV, and with the aid of plate 2, the trajectories of the reference beam and the diffracted beam were brought into coincidence and sent to the FEU-28-type photomultiplier. In order to register the differential signal, we used a synchronous detection system consisting of a synchronous detector 6, modulator 8, tube 9, photodiode 10, and recorder 7. The work of the synchronous detector is described in [3].

The method of measuring PAV amplitudes was as follows. In the work [3], the method of measuring PAV amplitudes by the method of direct optical probing was described. The authors obtained the dependence of the amplitude A of the surface wave on the excitation voltage U_v of an ultrasonic generator for the amplitude area of above 3Å (Figure 2, section BC -- direct method). This

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dependence made it possible to connect PAV amplitudes in angstroms with the values of voltage U_V in the heterodyne method both for amplitudes above 3Å (section BC in Figure 2), and for amplitudes smaller than 3Å (section AB in Figure 2) which could not be measured by the direct method.

Thus, the heterodyne method has definitely greater sensitivity than the direct method and can be used for precision measurements of the field distribution of surface acoustic waves.

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PUBLICATIONS, INCLUDING COLLECTIONS OF ABSTRACTS

ABSTRACTS OF PAPERS ON DYNAMIC PROCESS AND SIGNAL IDENTIFICATION AND RECOGNITION

Frunze IDENTIFIKATSKIYA I RASPOZNAVANIYE DINAMICHESKIKH PROTSESSOV I SIGNALOV; MEZHVUZOVSKIY TEMATICHESKIY SBORNIK NAUCHNYKH TRUDOV in Russian No 108, 1978 pp 1-2, 104-107

[Annotation and abstracts of papers from the collection edited by Zh. Sh. Sharshenaliyev (editor-in-chief), A.M. Tsykunov, G.A. Tyunayev and N.A. Petrovskaya, Frunze Polytechnical Institute, 1978, 108 pages]

[Text] The Identification and Recognition of Dynamic Processes and Signals

Intervuz Topical Collection of Scientific Papers

The papers included in this collection were reported to the professorial and teaching conference in 1978 at the "Engineering Cybernetics" subsection at Frunze Polytechnical Institute. The papers are devoted to theoretical and applied questions of engineering cybernetics, in particular, the problem of identifying and recognizing dynamic processes and signals.

ABSTRACTS

UDC 621.3.015

THE ESTIMATION AND RECOGNITION OF PULSES WITH SEPARABLE MOTIONS IN PROBLEMS STUDYING THE ENERGY SPECTRA OF CHARGED AND NEUTRAL PARTICLES

[Abstract of paper by Zh.Sh. Sharshenaliyev and N.A. Petrovskaya]

[Text] Some of the most efficient methods of postdetector signal discrimination from the viewpoint of the possibilities of using the signals to estimate and recognize pulses which are close in their waveform and consist of fast and slow components are analyzed in this paper.

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APPROVED FOR RELEASE: 2007/02/08: CIA-RDP82-00850R000300020012-3

UDC 621.391.1

A METHOD OF EVALUATING QUASI-STEADY STATE CONTROLLABLE PROCESSOR IN NONSEARCHING SELF-ADAPTIVE SYSTEMS

[Abstract of paper by N.A. Petrovskaya]

[Text] One method of computing the inverse transfer function of a control object in terms of the exponential spectra of the input and output signals is analyzed, functional schematics of the spectrum analyzer are proposed, where the analyzer is designed around a multichannel orthogonal filter, and a controllable equalizing device section having a transfer function of $W_0^{-1}(j\omega)$ is also proposed.

UDC 621.396.6

A SIMULATOR OF POSTDETECTOR SIGNALS WITH CONTROLLABLE PARAMETERS

[Abstract of paper by Ye.P. Kazantsev and K.E. Eraliyev]

[Text] A simulator of exponential postdetector signals with controllable parameters is designed which is intended for the alignment and experimental testing of various types of signal waveform discriminators.

UDC 681.327.12

A METHOD OF FILTERING OUT INTERFERENCE FROM HALF-SELECTED CORES IN THE MAINFRAME MEMORY OF COMPUTERS

[Abstract of paper by A.K. Musulmankulova and S.V. Kuz'menko]

[Text] The application of an orthogonal linear filtering algorithm to the development of a high speed pulse recognition device, where the pulses are read out from the main memory of a computer is analyzed.

UDC 62-50

ON THE IDENTIFICATION OF ONE CLASS OF NONLINEAR DYNAMIC OBJECTS

[Abstract of paper by I.G. Ten and A.M. Tsykunov]

[Text] The problem of the parametric identification of the controlled units of irrigation systems is analyzed, where the units are described by nonlinear hyperbolic partial differential equations with complex boundary conditions and which are subject to the action of random perturbations.

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UDC 621.316.7

THE IDENTIFICATION OF INDUSTRIAL FACILITIES IN AN SELF-OSCILLATING CONTROL MODE

[Abstract of paper by V.V. Burlyayev and A.V. Netushil]

[Text] The control of a self-oscillating mode is proposed by means of a relay regulator with a variable hysteresis loop width, which is easily made in the form of an independent unit or automated on a digital control computer.

UDC 621.391.1

THE IDENTIFICATION OF STRUCTURAL ELEMENTS IN THE DESIGN OF A SPEECH CONTROL SYSTEM

[Abstract of paper by V.N. Plotnikov, V.A. Sukhanov, Yu.N. Zhigulevtsev]

[Text] A method of solving a teaching—identification problem for the nonderivative (structural) components of speech signals is treated in this paper, where the procedure consists in the complete automation of the identification procedure based on the use of automatic segmentation algorithms, the calculation of normalized spectral descriptions of the speech signal segments which are singled out and the automated classification of the resulting set of descriptions.

UDC 621.391

HYBRID (DIGITAL-ANALOG) MATCHED FILTERS FOR DISCRETE FREQUENCY MODULATED DIGNALS

[Abstract of paper by N. Zhayloobayev]

[Text] A hybrid (digital-analog) matched filter is proposed for discrete frequency modulated signals, which combine precision and stability of the digital circuitry with the computational speed of analog circuitry.

UDC 621.311.338.658.57

ESTIMATING THE PARAMETERS OF THE MAJOR INDICATORS OF ENTERPRISES AND PREDICTING THEM

[Abstract of paper by M.A. Suyerkulov]

[Text] Equations are derived from a statistical analysis which describe the dynamic series of the main indicators of enterprises. The parameters are estimated. A prognosis is made of the parameters of the indicators for 1978 and 1979.

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UDC 621.376.5

A PULSE WIDTH MODULATED SIGNAL ALGORITHM BASED ON A PRECISION CRITERION

[Abstract of paper by M. K. Masimov]

[Text] An optimal pulse-width modulated signal algorithm with respect to precision and noise immunity criteria is proposed. The algorithm derived is also optimal for pulse-phase modulated and pulse-frequency modulated signals.

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ABSTRACTS OF PAPERS ON ELECTROMAGNETIC WAVE AND SIGNAL CONVERSION AND TRANSMISSION

Moscow TRUDY MOSKOVSKOGO ORDENA LENINA ENERGETICHESKOGO INSTITUTA; TEMATICHESKIY SBORNIK. VOPROSY PEREDACHI I PREOBRAZOVANIYA ELEKTROMAGNITNYKH VOLN I SIGNALOV in Russian No 433, 1979 signed to press 14 Dec 79 pp 1-2, 91-97

[Annotation and abstracts from the book edited by M.Sh. Kulakhmetova: "The Proceedings of the Moscow Order of Lenin Power Engineering Institute. Topical Collected Papers. Questions of the Transmission and Conversion of Electromagnetic Signals", 97 pages, 500 copies]

[Text] This topical collection of papers is devoted to urgent questions in the transmission and conversion of electromagnetic waves and signals.

The collection contains articles on modeling nonlinear microwave effects in a gas plasma by means of semiconductors, on a study of a CO_2 laser using an infrared spectrometer and a study of gas discharge microwave detectors. This series of experimental and theoretical papers belongs to the field of laser and plasma engineering. Mathematical models of the circuit for the effect of noise on a phase locked loop, metal-insulator-metal tunnel structures and models for the scattering of electromagnetic waves at a cylinder which is inhomogeneous in two dimensions are treated in a number of the papers.

Integral equations are used in the entire series of papers for the solution of practical problems of diffraction at dielectric solids in free space, in resonators, in a magnetically active medium and in charged particle beams. The performance of the relevant numerical experiments makes it possible to analyze rather complex physical phenomena which are difficult to subject to a conventional experimental check.

Also treated in the collection are examples of the realization of VHF radio communications by means of polar aurora scattering.

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ABSTRACTS

UDC 621.373.826

MONITORING THE TRANSVERSE MODES OF A CO_2 BY MEANS OF AN INFRARED SPECTROMETER

[Abstract of article by Drugov, L. V., pp 3-4]

[Text] A method of monitoring the modal composition of CO_2 laser radiation by means of recording the change in the intensity of one of the output lines with an infrared spectrometers while slowly tuning the resonator length is described. The tuning is accomplished by cutting off the water cooling and turning on the discharge for brief periods of time.

UDC 621.376.239:533.915

A STUDY OF A MICROWAVE GAS DISCHARGE DETECTOR OPERATING IN A STEADY-STATE MICROWAVE MODE

[Abstract of article by Kalinin, V. A. and Lobov, G. D., pp 5-9]

[Text] The characteristics of a gas discharge detector are computed on the basis of an analysis of the equilibrium equations for charged particles and their energy, nonlinear effects in the layers near the electrodes and a high frequency equivalent circuit. Theoretical results are compared with experimental data. It is ascertained that the major mechanism for detection in the steady-state mode is detection at the nonlinear conductance of the layers near the electrodes.

UDC 621.376.239:533.915

TRANSIENT PROCESSES IN A MICROWAVE GAS DISCHARGE DETECTOR OPERATING IN THE MICROWAVE DISCHARGE MODE

[Abstract of article by Kalinin, V. A. and Lobov, G. D., pp 9-12]

[Text] The output signal rise time with the action of a microwave field pulse on the detector is analyzed. The results of calculating the output voltage settling time are compared with the experimental data. The output signal waveform and the reflected power pulse are qualitatively explained on the basis of a solution of non-steady-state particle and energy equilibrium equations.

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UDC 621.391.244:621.3.018.2

THE ANALYSIS OF SIGNAL SPECTRA BY MEANS OF A MULTIFUNCTION PARAMETRIC DEVICE

[Abstract of article by Ivanov, Yu. V. and Shtykov, V. V., pp 13-16]

[Text] Based on a comparison of the transmission factors of a parametric device and a device which utilizes the spin echo effect, the possibility of spectrum analysis in the spin device is demonstrated in this paper. The spin device is understood to be a multifunction parametric device.

It is shown that the spin device, which operates using a dual pulse procedure, can be equated with a signal spectrum analyzer under certain conditions.

UDC 621.376.239

THE INFLUENCE OF LOCALIZED STATES ON THE PROPERTIES OF METAL-INSULATOR-METAL TUNNEL STRUCTURES

[Abstract of article by Zhgun, S. A. and Shtykov, V. V., pp 16-19]

[Text] An accounting is made of the influence of localized states on the proper ties of the volt-ampere characteristic of metal-insulator-metal structures. By using a method of expansion in a Maclaurin series, it is shown that some experimentally observed features of the coefficients of the series can be due to impurities and defects in the crystalline structure of the dielectric.

UDC 621.371:551.510.535

A NUMERICAL SOLUTION OF THE PROBLEM OF RADIO WAVE PROPAGATION IN AN INHOMOGENEOUS MAGNETICALLY ACTIVE MEDIUM

[Abstract of article by Kramm, M. N. and Filonov, S. V., pp 20-23]

[Text] A transmission matrix apparatus is used to determine the wave which has passed through an inhomogeneous layer. The results of the calculations are represented in the form of curves for the transmittance factor as a function of the magnitude and direction of the external permanent magnetic field and the layer thickness.

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UDC 621.384.6

A FUNDAMENTAL SOLUTION OF THE INTEGRAL EQUATION FOR THE DENSITY DISTRIBUTION OF A SPACE CHARGE IN THE PROBLEM OF THE INTERACTION OF A STRONGLY MODULATED CHARGED PARTICLE BEAM AND THE SYNCHRONOUS HARMONICS OF A MICROWAVE FIELD

[Abstract of article by Balashov, V. N., Gavich, V. T. and Zhileyko, G. I., pp 24-26]

[Text] A method of solving the integral equation which relates the density distribution of a space charge to the external electrical field is given. Special cases of the solution are also given.

UDC 621.384.6

A COMPARISON OF ANALYTICAL AND NUMERICAL INTEGRATIONS OF PHASE EQUATIONS FOR THE PROCESS OF THE INTERACTION OF CHARGES WITH AN ELECTROMAGNETIC FIELD

[Abstract of article by Azizbekyan, G. V., Sveshnikova, N. N. and Zhileyko, G. I., pp 27-29]

[Text] It is shown that the errors in calculations using analytical formulas and a numerical method for the process of the interaction of clusters of electrons with an electromagnetic field, taking into account the space charge, fall in a range of 10 percent.

UDC 621.376.019.4

OPTIMAL NONLINEAR CONVERSION IN A PHASE KEYING RECORDING AND CARRIER RESTORATION CIRCUIT

[Abstract of article by Bezuglyy, V. V. and Zhukov, V. P., pp 29-34]

[Text] Characteristics of nonlinear elements which are optimal in terms of a signal to noise ratio criterion are determined. The maximum signal to noise ratios in circuits with optimal characteristics are computed.

UDC 621.396.668

THE STATISTICAL SYNTHESIS OF THE PARAMETERS OF A DUAL CHANNEL ITERATION PHASE LOCKED LOOP SYSTEM

[Abstract of article by Samoylenko, V. N., pp 34-38]

[Text] The optimal parameters of a dual channel phase locked loop system are determined on the basis of the criterion of a minimum mean square error.

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UDC 621.391.22/23

THE DETERMINATION OF THE PERMISSIBLE PRECISION IN THE EXECUTION OF AN ALGORITHM FOR AN OPTIMUM RECEIVER OF BINARY SIGNALS

[Abstract of article by Arapov, S. M., pp 38-42]

[Text] A two-step procedure is proposed for the determination of the permissible precision in the realization of an optimal receiver algorithm, where the procedure is based on the introduction of the concept of the equivalent shift in the threshold of a resolver. The permissible error probability and the requirement of minimal receiving system cost are used as the optimization constraints. An expression is derived for the determination of the relative tolerance of the receiver parameters.

UDC 621.396.677.3

THE EXCITATION OF A CYLINDRICAL RESONATOR WHERE A DIELECTRIC SOLID IS PRESENT IN IT

[Abstract of article by Vasil'yev, Ye. N. and Kapustin, Yu. G., pp 43-47]

[Text] A system of integral equations for a dielectric axially symmetric solid in free space relative to the surface electrical and magnetic currents is generalized for the case where the dielectric solid is placed inside a cylindrical resonator. The nuclei of the new integral equation differ from the original ones in the existence of an additional term, which has no singularities.

UDC 621.396.677.3

AN INTEGRAL EQUATION FOR THE PROBLEM OF INTERNAL AXIALLY SYMMETRIC DIFFRACTION AT A DIELECTRIC SOLID

[Abstract of article by Kapustin, Yu. G. and Kovalenko, A. I., pp 47-52]

[Text] The problem of determining the fields of a resonator perturbed by a dielectric, axially symmetric solid is treated. A planar non-self-adjoint boundary problem is formulated with respect to the Abraham potential for the E mode. A boundary problem, formulated using Green's theorem, which adjoint to the other boundary problem, is defined. A system of integral Friedholm equations of the second kind is written. An expression is given for Green's function for the problem of a circular cylindrical resonator.

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UDC 621.396

THE DIFFRACTION OF A PLANE WAVE AT A DIELECTRIC CYLINDER

[Abstract of article by Sedel'nikova, Z. V., pp 52-55]

[Text] The problem of electromagnetic wave diffraction at a dielectric cylinder is reduced to an integral equation for the density of the surface electrical and magnetic currents, which can be solved numerically. An example of the calculation of the excitation of a dielectric cylinder by an electrical dipole perpendicular to its axis is given.

UDC 538.569:621.396.1

A NEW APPROACH TO THE CALCULATION OF THE CHARACTERISTICS OF ELECTROMAGNETIC WAVE SCATTERING AT A TWO-DIMENSIONALLY INHOMOGENEOUS CYLINDER WITH AN ARBITRARY CROSS-SECTION

[Abstract of article by Kuvayev, V. M. and Permyakov, V. A., pp 56-60]

[Text] A new method of determining electromagnetic waves scattered by a two dimensionally inhomogeneous cylinder is proposed with utilizes a method of straight lines for calculating the fields inside the inhomogeneous solid and an analytical extension of the external field at the surface of the solid.

UDC 538.569:621.315.592

ON USING SEMICONDUCTORS TO MODEL NONLINEAR MICROWAVE EFFECTS IN A GAS PLASMA

[Abstract of article by Korneyeva, T. M., Permyakov, V. A., Slezkin, V. G., Steshenko, A. G. and Fedesov, G. Ye., pp 61-66]

[Text] The possibility of using semiconductors to model the self-stress of a high power microwave field and the generation of the third harmonic, which arise in a gas collision plasma, is treated. The normal incidence of a plane microwave pulse on a plasma is described in the framework of elementary theory. The results of waveguide experiments with thin plates of silicon and modes where they are arranged transversely are given [sic].

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UDC 538.569

THE COMPUTER CALCULATION OF THE PARAMETERS OF A DISPERSING MEDIUM USING BREMERMAN'S METHOD

[Abstract of article by Finat'yev, Yu. P. and Tsel'sov, Yu. G., pp 67-68]

[Text] The determination of the equilibrium chemical composition of the medium is reduced to the solution of a system of nonlinear algebraic equations. This system is solved by means of minimizing a function of several variables. It is shown that this is accomplished most efficiently using Bremerman's iterations.

UDC 621.371.332.4.001.5

SPECIAL FEATURES OF AURORAL COMMUNICATIONS LINKS

[Abstract of article by Bubennikov, S. V. and Volovskiy, V. N., pp 69-72]

[Text] The special features of auroral communications links at frequency of 144 MHz during the years of the "quiet" sun are treated on the basis of processing experimental material. The range of reflection for moderate and strong auroras is plotted and examples are given of two-way radio communications where one of the stations is located in Moscow.

UDC 538.574.6

THE DIFFRACTION OF A PLANE E-WAVE AT A SLOT BETWEEN TWO SEMI-INFINITE DIELECTRIC PLATES

[Abstract of article by Vasil'yev, Ye. N., Polynkin, A. V. and Solodukhov, V. V., pp 73-79]

[Text] The problem of the excitation of a plane electromagnetic wave of a slot between two semi-infinite dielectric plates is solved numerically using integral equations with respect to the equivalent surface currents. The amplitudes of the excited natural waves of the plate were studied as a function of the slot size and angle of incidence of the plane wave.

UDC 621.371 : 551.510.535

ON THE DEPOLARIZATION OF ELECTROMAGNETIC WAVES WHEN REFLECTED FROM A MAGNETICALLY ACTIVE LAYER OF AN IONIZED MEDIUM WITH A METAL BASE

[Abstract of article by Kramm, M. N. and Goncharov, V. L., pp 80-86]

[Text] A numerical method of computing the backscatter matrix elements is proposed for the case of an electromagnetic wave which is normally incident to a plane layered magnetically active ionized medium. Attention is turned

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to the process of the transformation of wave polarization when reflected from the magnetically active ionized medium, which is characterized by the coefficients of the backscattering matrix, which are located outside the main diagonal (by the depolarization coefficients). The results of the calculations are presented in the form of the depolarization coefficients expressed as a function of the ratio of the gyro-frequency of the electrons to the wave frequency.

UDC 621.382.323

SIGNAL QUANTIZATION NOISE IN A CHARGE COUPLED DEVICE WITH OPTICAL INPUT

[Abstract of article by Zhurikhin, A. V., pp 86-88]

[Text] For the case of the optical input of a signal into a CCD (charge coupled device), formulas are derived for the calculation of the signal quantization noise dispersion based on the known energy spectrum of this signal.

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UDC 621.382

CONTROL IN MICROELECTRONIC TECHNOLOGY

Minsk KONTROL' V TEKHNOLOGII MIKROELEKTRONIKI [Control in Microelectronic Technology] in Russian 1979 signed to press 20 Jun 79 p 2, 311-312

[Annotation and table of contents from book by V. M. Koleshko, P. P. Goydenko, and L. D. Buyko, Nauka i tekhnika, 1475 copies, 312 pages]

[Text] The authors describe modern problems of the theory and practice of methods for controlling the defects of crystal lattices and electrophysical properties of multilayer structures during various stages of manufacturing semiconductor devices and integrated microcircuits. They give the data and practical recommendations for controlling the surfaces, structure, and properties of thin films and phase interfaces of the metal-semiconductor and metal-dielectric-semiconductor contacts with the use of modern physical research methods.

They examine physical methods of quality control of microwelding and soldering of integrated microcircuits, as well as the problems of optimization and automation of the control of technological processes with the use of information on the quality of multilayer structures.

The book is intended for broad sections of scientists, engineers and technicians of enterprises and design and research organizations of the radio and electronic industry, and can be useful for instructors and graduate and undergraduate students specializing in microelectronics.

Tables -- 23, figures -- 108, bibliography -- 736 items.

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LASER BEAMS, COLLECTION OF ARTICLES ANNOUNCED

Gor'kiy IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY, RADIOFIZIKA in Russian No 4, 1980 pp 507-508

[Annotation, table of contents and supplementary information for inter-VUZ collection of articles "Lazernyye puchki" [Laser Beams] edited by N.K. Berger, Izdatel'stvo Khabarovskogo Politekhnicheskogo Instituta, 1979]

[Text] In this collection are published articles devoted to investigation of the space-time structure of laser radiation and to the formation of this structure by means of active resonators and optical systems. Special attention is paid to questions relating to the active control of the space-time structure of laser radiation, i.e., to adaptive optics, wave front reversal, etc.

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LINES OF COMMUNICATIONS

Moscow LINII SVYAZI in Russian 1980 signed to press 25 Dec 79, p 2, 437-440

[Annotation and table of contents from book by Igor' Izmaylovich Grodnev and Nikolay Dmitriyevich Kurbatov, Izdatel'stvo "Svyaz," 20,000 copies, 439 pages]

Annotation

[Text] This book examines the principles of constructing mainline, zonal and local communications networks. It cites the designs, characteristics and electrical parameters of various guide systems (symmetrical and coaxial cables, waveguides, light guides, optical and superconducting cables, etc.). It outlines the theory of transmission through guide systems and also the theory of the reciprocal and external influences and measures for shielding from them. The third edition was published in 1974. New material is devoted to the planning, construction and operation of the communications facilities.

The text book is for students in communications institutions of higher learning, who are studying the specialities of multichannel and automatic electric communications.

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PRODUCTION TECHNOLOGY FOR MICROELECTRONIC DEVICES

Moscow TEKHNOLOGIYA PROIZVODSTVA MIKROELEKTRONNYKH USTROYSTV in Russian 1980 signed to press 15 Feb 80 pp 442-447

[Title, annotation, and table of contents of a book by Ideya Aleksandrova Malysheva, Energiya, 15,000 copies, 448 pages]

[Text] This book give a general description of the production of microelectronic devices, and over-all production requirements are formulated. The basic technological methods and processes described include mechanical working, cleaning the substrate surface, photolithography, the free mask method, x-ray and electronic lithography, film deposition methods, epitaxy, diffusion, and ion doping. Typical manufacturing processes for bipolar and MDS microcircuits and for thin-film and thick-film microcircuits are analyzed. Assembly and sealing processes, problems connected with ensuring quality and efficiency, and trends in the future development of microelectronics are considered.

This book is designed as a textbook for students in technical schools specializing in the production of microelectronic devices.

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SCIENTIFIC ARTICLES ON ANTENNAS

Moscow ANTENNY: SBORNIK NAUCHNYKH STATEY (Antennas: Collection of Scientific Articles) in Russian No 27, 1979 signed to press 9 June 79 pp 2-3

[Annotation and table of contents from book edited by A. A. Pistol'kors et al., Izdatel'stvo Svyaz', 5000 copies, 184 pages]

[Text] This volume addresses a number of important problems of modern antenna technology: investigation of high-frequency breakdown in radar antennas, synthesis of director antennas by directive gain maximum, holographic methods of measuring antenna radiation patterns, modernization of the radiation pattern control system of the North-South antenna of the DKR-1000 radio telescope of the Physics Institute imeni P. N. Lebedev of the USSR Academy of Sciences, and theoretical investigation of the noise temperature of the RATAN-600 radio telescope.

The authors devote attention to the problem of optimal synthesis of linear antenna arrays, including unequally spaced, and calculation of the reflection factor from the open end of a microstrip line.

This volume is intended for scientists and radio engineers working in the area of antennas.

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SMALL SIGNAL MODULATORS

Leningrad MODULYATORY MALYKH SIGNALOV in Russian, 2d edition 1980 signed to press 4 Dec 79 p 2, 200

/Title, annotation, and table of contents of a book by Boris Andreyevich Kalinchuk and Oleg Aleksandrovich Pichugin, Energiya, 7800 copies, 200 pages/

[Text] This book discusses the conversion of small signals with a dc low-frequency electric current into ac signals.

The second edition differs from the first, which was published in 1972, in that a number of new methods for constructing modulators that have been developed recently are considered; the sections on the calculation and design of photo- and transistor modulators, along with those on the utilization of modulators in technology, are expanded.

The book is designed for specialists studying the enhancement of small signals with dc and low-frequency currents; it will also be useful for graduate students and students at institutions of higher education.

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SPACE-TIME PROCESSING OF SIGNALS, COLLECTION OF ARTICLES ANNOUNCED

Gor'kiy IZVESTIYA VYSSHIKH UCHEBNYKH ZAVEDENIY, RADIOFIZIKA in Russian
No 4, 1980 pp 505-506

[Annotation and table of contents of inter-VUZ collection of articles "Prostranstvenno-vremennaya obrabotka signalov, vypusk 2" [Space-Time Processing of Signals, No 2] edited by I.Ya. Kremer, Izdatel'stvo Voronezhskogo Gosudarstvennogo Universiteta, scheduled for first quarter of 1980, 8.5 quires]

[Text] In this collection are discussed questions relating to the creation and analysis of systems for processing space-time, time and vector-time signals against a background of noise. An investigation is made of the characteristics of systems for the space-time processing of signals and of the advisable principles of their design. The generalized characteristics of processing systems as space-time filters are discussed, along with questions relating to the creation of such filters. A study is made of the optimal space-time filtering of random fields and of an evaluation of their parameters. The results are given of the solution to problems in detecting space-time and time signals both with known and partly unknown noise statistics. An analysis is made of optimal and quasi-optimal algorithms for the detection of signals with unknown parameters and of non-Gaussian signals. An investigation is made of the characteristics of acousto-optical equipment for processing radio signals.

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QUANTUM ELECTRONICS

UDC 621.378.325

CALCULATION OF ANGULAR CHARACTERISTICS OF LASERS WITH UNSTABLE RESONATORS

Kiev IZVESTIYA VUZOV SSSR - RADIOELEKTRONIKA in Russian No 3, 1980 pp 93-94

Article by N. Ye. Sklyarov, Yu. A. Timofeyev, N. N. Petrov, and V. A. Patsayeva, submitted 23 Jan 1979]

[Text] In order to calculate angular characteristics of unstable resonators [1], it is possible to use Tang and Statts' multiwave generation model for parallel-plate resonators. In this case, the equivalent length $L_{equivalent}$ of the parallel-plate system is determined for each type of unstable resonators which are being examined under the condition of the equality of their diffraction losses.

The criterion of the applicability of the parallel-plate resonator theory can be written in the following form

$$(\Delta L/\lambda) < 1/N,$$

where ΔL -- maximum difference of the optical path in the resonator; λ -- radiation wave length; N -- number of Fresnel zones.

Tberefore, the Tang-Statts model satisfactorily describes time-averaged angular characteristics of laser radiation in the case of perfect active models and resonators.

For a telescopic resonator (Figure 1), $L_{equivalent}$ can be determined in the following way. The value of diffraction losses of the resonators is determined by the number of Fresnel zones. For a telescopic resonator,

$$N = 2h/\lambda,$$

where h -- sag of the output mirror. In the case of a parallel-plate resonator,

$$N = D^2/4L^{(1)}_{экв}\lambda,$$

Key: 1. Equivalent

where D -- diameter of the active element; $L_{Equivalent}$ -- length of the resonator.

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From the equality of the Fresnel numbers of two resonators, we obtain

$$L_{экв} = D^2/8h.$$

By using the condition of the cofocal nature of the mirrors $R_1 = -(2L+R_2)$ and the expression for the enlargement factor of the telescopic resonator $M = |R_1|/R_2$, we obtain

$$L_{экв} = \frac{D^2(M-1)}{16L\left(1-\sqrt{1-\frac{D^2(M-1)^2}{16M^2L^2}}\right)}, \quad (1)$$

where L -- geometrical length of the telescopic resonator.

When performing computations, L in formula (1) should be replaced with

$$L_{эф} = L - l_а + \frac{l_а}{n},$$
(1)

Key: 1. Effective

where l_a -- length of the active element; n -- refractive index of the active substance.

The divergence θ in the case of the multiwave approximation [2]

$$\theta_m = m\varphi_д,$$

where m -- number of excited angular types of oscillations;

$$\varphi_д \approx \lambda/D.$$

The number of the transverse types of oscillation under the condition $\sigma_1^д \ll \sigma_0$ is defined by the expression [2]

$$\frac{4}{3}m^3 + 2m^2 \approx \frac{\sigma_0}{\sigma_1^д}\left(1-\frac{1}{n_p}\right),$$

where $\sigma_0 = 2 \ln M$ -- unselective losses of the resonator; $\sigma_1^д = 2\left(\frac{\sqrt{\lambda L_{экв}}}{D}\right)^3$ -- diffraction losses of the first transverse type of oscillations; n_p -- excess over the pumping power threshold.

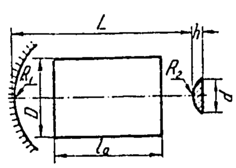

Figure 1

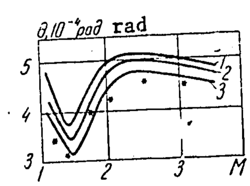

Figure 2

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Considering the annular extraction of radiation from a telescopic resonator, the final expression for the value of the angular divergence θ with respect to the half-intensity level can be written in the form of

$$\theta = \varphi_{д}\left(m + \frac{2M}{M-1}\right).$$

Figure 2 shows the curves of the computed dependence of the angular divergence of radiation from the coefficient M for various levels of pumping: 1 -- n_p=4; 2 -- n_p=3; 3 -- n_p=2; * -- experimental values of θ for n_p=2. The computation of divergence was done for a laser using neodymium glass with l_a=62 cm, D-4.5 cm, L-100 cm. As can be seen from Figure 2, there exists an optimal value of M at which angular divergence is minimal.

Experimental checking of this laser showed that the value of angular divergence corresponds satisfactorily to the computed values (Figure 2).

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