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HIGH-POWER DC SCALING AMPLIFIER

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The author describes a scaling amplifier with an output power of 50 W into a 4-Ω load that has a level of spurious spectral components due to intermodulation of not more than −80 dB at frequencies of up to 20 kHz, a bandwidth of 1 MHz, an output-voltage rise rate of 100 V/μsec, and a vector error of 0.1% at a frequency of 20 kHz for a transmission coefficient of 5 V/V.

High-power dc scaling amplifiers with amplitude, frequency, and phase responses of maximum linearity at frequencies to 20 kHz and minimal spurious spectral components in processing of complex signals are often required in electroacoustic, electromechanical, and other experiments. The settling time of the output voltage with a given error of the steady-state value when the load current varies abruptly or the input voltage of amplifier is pulsed is of great importance.

On the basis of earlier studies [1, 2], it can be said that the principal problem that arises in the design of high-power scaling amplifiers of increased accuracy is that of suppression of intermodulation distortion, which is determined by the presence in the amplifier output signal of spurious spectral components – products of signal multiplication and harmonics.

According to its origin, intermodulation distortion can be divided into the following groups: 1) thermal; 2) that produced by nonlinearity of the amplitude response of an open-loop (i.e., without negative feedback (NFB)) amplifier; and 3) that produced by loss of dynamic linearity [3] in a closed-loop (i.e., with NFB) amplifier.

If the condition of dynamic linearity

$$V_m \geqslant 2\pi U_{\text{out max}} f_{\text{co nfb}},$$

where V_m is the output-voltage rise rate, $U_{\text{out max}}$ is the maximum output voltage, and $f_{\text{co nfb}}$ is the cutoff frequency of the amplifier for the main NFB loop is satisfied, the unwanted spectral component in the output signal of an amplifier whose final stage is operating in class B or AB can be estimated by the expression

$$U_{\text{s}} \approx \frac{2\pi U_{\text{out}} f\,(U_0 + 2I_\ell r_e)}{V_m (K_{f_{\text{nfb}}} + 1)},$$

where U_{out} is the output voltage, U_0 is the dead zone, f is the frequency of the input sine-wave signal, $K_{f\ \text{nfb}}$ is the degree of NFB at the given frequency, I_ℓ is the load current, and r_e is the equivalent resistance in the emitter circuit of the power transistor.

If the condition of dynamic linearity is not satisfied, the degree of NFB in the denominator of the above expression should be made equal to zero.

In addition, intermodulation of the second type arises in the following cases: in the input differential stage of the amplifier, which (for bipolar transistors) has a nonlinear transfer characteristic of cubical form [1]; when the amplifier has series NFB, due to a nonlinear dependence of the common-mode attenuation factor on the common-mode voltage and frequency; and in any voltage-amplification stage, since the modulation of the collector current with variation of the voltage on the junction is in a first approximation cubical in nature, and the capacitance of the collector junction is a function of the voltage on in it in a power of ½.

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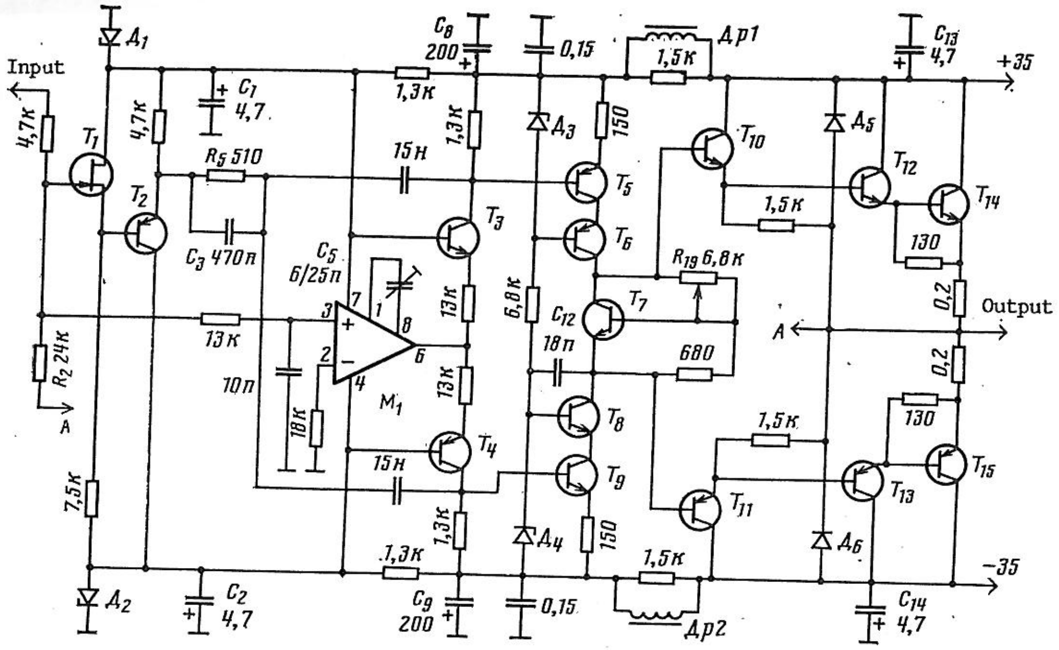

Fig. 1. Schematic diagram of amplifier: M_1) K153UD2; T_1) KP307A; T_2) KT326B; T_3, T_7, T_9) KT315G; T_4, T_5) KT361G; T_6) KT639D; T_8) KT630A; T_{10}) KT3102A; T_{11}) KT313A; T_{12}) KT602BM; T_{13}) KT626V; T_{14}) KT819GM; T_{15}) KT818GM; D_1, D_2) KS515A; D_3, D_4) KS139A; D_5, D_6) KD212A; C_1, C_2, C_{13}, C_{14}) K53-16; C_8, C_9) K50-16; C_5) KPK-M; R_{19}) SP5-2; chokes) DMO, 5-500 μH.

Some of the following Russian abbreviations may be found in the figure: *Л* = tube, *Д* = diode, *Тр* = transformer, *Др* or *др* = choke, *Вк* = switch, *в* = V, *М* = MΩ, *к* = kΩ, *мк* = μF or μH, *п* = pF or pH, and *н* = nF or nH. For individual tube designations see Appendix to Issue No. 1 of this year.

In the high-voltage (penultimate) voltage-amplification stages, these effects are made worse by the thermal inertia of the transistors [1, 2] and also by the appearance of a corresponding additional phase shift at frequencies of 40-150 Hz.

Thermal effects are especially striking when transients in dc scaling amplifiers (of any power) are studied, since they cause considerable (by a factor of 5-10, as compared with the calculated value) prolongation of the settling time in the zone of errors of less than 1% of the steady-state value of the output voltage. It was established that for an optimum – for a high-power scaling amplifier – "two-pole" amplitude–frequency characteristic (AFC), the output-voltage settling time with an error of 1% of the steady-state value must not exceed $5f_{\text{co}}^{-1}{}_{\text{nfb}}$ in the absence of ringing or sharp breaks in the transient response (the readjustment is determined by the shape of the AFC).

As a result of the studies, the following practical recommendations were developed for the design of high-power scaling amplifiers of increased accuracy to minimize the level of unwanted spectral components caused by intermodulation and to realize an acceptable settling time in the small-error zone: an inverting circuit with a parallel high-frequency amplification channel [4] is best for a high-power scaling amplifier; since any variation in the amplifier supply voltage is an unwanted common-mode signal for the voltage-amplification stages, such variations should be suppressed by means of regulating and filtering elements in all current-limiting circuits; and all amplifier stages and current generators must have cascade circuits.

Several versions of high-power scaling amplifiers have been built with allowance for the above considerations and recommendations. The UMVT50-84 amplifier has a nominal output of ±20 V, a nominal load resistance of 4 Ω, a bandwidth of 1 MHz, a bandwidth of 100 kHz at nominal output power, an output-voltage rise rate (as measured with a pulse signal with an off-duty factor of 5) of 100 V/μsec, a settling time of 2 μsec for a 10-V output with an error of 1%, a vector error of not over 0.1% at a frequency of 20 kHz for an output power of 27 W, a level of unwanted spectral components of all types of not more than –80 dB (the coefficients of harmonic and intermodulation distortion are <0.01%), a transmission coefficient of ~5 V/V, a peak-to-peak noise voltage of 50 μV in the band of 0.01-10 kHz, and static parameters that are determined by the K153UD2 integrated circuit.

The UMVT50-84 amplifier, a schematic diagram of which is shown in Fig. 1, has a parallel high-frequency amplification channel. The high-frequency channel uses two series-connected wideband voltage followers: a source follower (T_1) and an emitter follower (T_2). This channel must have a low output impedance to ensure fast recharging of the capacitors at the points of signal summation of the parallel channels (the bases of transistors T_5 and T_9) and, therefore, a high rise rate. Capacitors C_6 and C_7 provide dc isolation of the parallel channels.

The low-frequency channel employs an operational amplifier (OA) M_1 and transistors T_3 and T_4 in a common-base circuit. A trimmer capacitor in the OA correction network sets the maximum possible NFB for the entire amplifier at a frequency of 20 kHz, but the settling time must satisfy the relation given above. The output-voltage swing of M_1 does not exceed 0.5 V at a frequency of 20 kHz at nominal output power; therefore, the rise rate of the OA output voltage is not limited. Common NFB through resistor R_2 provides dc regulation for M_1. The offset voltage and drift of the amplifier are fully determined by the corresponding parameters of M_1, since the gate current of T_1 is smaller than the input current of the given OA by more than an order of magnitude.

The phase shift introduced by M_1 at high frequencies is compensated for by the high-frequency amplification channel; therefore, the AFC of the amplifier in the area of the cutoff frequency is chiefly determined by the single wideband voltage-amplification stage (T_5, T_6, T_8, and T_9), which provided the required margin of amplifier stability and a high cutoff frequency for the main NFB loop. The high-voltage amplifier has a cascode circuit, which provides considerable gain ($>10^3$) over a wide range of frequencies and suppression of intermodulation. Capacitor C_{12}, along with the collector-base capacitances of transistors T_6 and T_8, the input capacitance of the current amplifier, and the forcing network of R_5 and C_3 in the high-frequency channel, forms the AFC of the amplifier at high frequencies.

The supply circuit of the voltage-amplifier stages has fairly effective passive filters and parametric voltage regulators for the high-frequency channel and the OA and for clamping of the base voltages of T_3 and T_4. The low-frequency ripple in the supply circuit of the high-voltage stage is not suppressed by the filters, but it does not change the operating point (or, therefore, cause intermodulation), since the operating conditions of the stage are stabilized by the low-frequency amplification channel.

The output current amplifier (T_{10}-T_{15}) has a conventional circuit. The initial current of the output transistors (<50 mA) is set by variable resistor R_{19}. An isolating choke (10-30 μH) shunted by a resistor of 5-20 Ω must be connected in series with the amplifier output for operation into a load with a capacitive component of >300 pF. Transistors T_{12} and T_{13} have heat sinks with areas of 20 cm^2; the heat sinks of T_{14} and T_{15} have areas of 800 cm^2. The amplifier is mounted on a fabric-glass-laminate printed-circuit board of 160 × 100 mm. Getinax and textolite boards are unsuitable, owing to the unsatisfactory dielectric properties of these materials.

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