A RIAA Equalized Preamplifier

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Reproduction of RIAA-standard gramophone recordings has generated much debate from early 20th century up to now in the audio community. This debate is justified since the large variety of RIAA equalizer circuits have been presented and their performance often did not correspond to exerted efforts. This is the result of the strictly defined operating conditions:

* low-level input voltage with frequency-increased spectral density,

* predominantly inductive signal source impedance,

* accurate RIAA replay frequency response [1,23],

* matching with given cartridge (this includes specified resistance, capacitance and gain),

* considerable scattering in peak recorded velocities.

Several review articles on this topic represented a survey of circuit schematics [30,31]. We try to offer a broader outlook on RIAA preamplifier design by way of a more systematic approach; we restricted our examination preamplifiers to moving-magnet MM pickups with voltage gain at 1 kHz of 30 to 40 dB (wide-spread value is 34 dB).

Moreover, we did not consider old dual- and triple-transistor circuits, because the performance level that can be obtained from them is not adequate to high-quality reproduction. The cited references are mostly in English, but for lack of similar circuit examples the other sources are involved. In addition, we offer a medium-cost preamplifier with relatively high performance specifications.

Design principles

Possible RIAA-correction circuit arrangements are presented in Table 1. The tubed preamp with inter-stage frequency dependent networks (Table 1a) was the first preamp topology in gramophone record playback history. Audiophiles and engineers attention to this topology increased again towards the end of 20 century [2,3] due to absence of transient intermodulation distortion (TIM), which is originated with overall feedback loop that formed the standard RIAA frequency response.

The modern passively equalized design requires two low-noise gain blocks with flat response and overall gain of 40 to 60 dB. The second amplifying block is fed from the output node of the frequency dependent attenuator/divider with rather low signal. Thus the second gain stage with satisfactorily high input resistance would be optimized for noise matching with capacitive output impedance of the RIAA passive network. Let us assume that the discrete op amp [11-14] or modern integrated op amp (for example, NE5534) with unity-gain frequency around 5 MHz is employed in this circuitry. If the stage gains are nearly equal, then their loop gain at HF is of 18 to 40 dB. This amount is sufficient for quality performance if the open loop distortion is lower than 0.1% The modem techniques based en nonlinearities indirect compensation [7] are suitable for decreasing the open loop distortion. As examples one might be referred to KLEIN+HUMMEL Preamp model SA11 [4], which utilizes NE5534 and preamplifier [5], which uses simplified, symmetrical, low-distortion gain blocks [6]. We also would like mention here one of the well-sounding tube-based circuit [8].

Shunt feedback amplifier (Table 1b) has the merits of accuracy of frequency and transient responses and negligibly small distortion. Non-linearity due to a common mode input signal is avoided in this circuitry. The input resistor 47 kOhm is connected in series with the pickup coils, and the overall signal-to-noise ratio is degraded by 13 dB compared with series feedback connection, even with ideal noise-free amplifier [9]. So this configuration is seldom employed. The origin of noise degradation lies in the thermal noise voltage of the input resistor. This drawback can be easily eliminated by an input buffer stage (Table 1c).

In the RIAA preamp [10] the input stage is realized as a conventional instrumental differential amplifier based on three TL071. The appropriate setting of stage gains allows for approximately equal feedback amounts at HF. The overall gain can be varied easily in accordance with peak-recorded velocity to ensure safety overload margin.

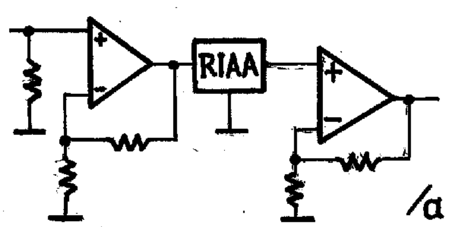
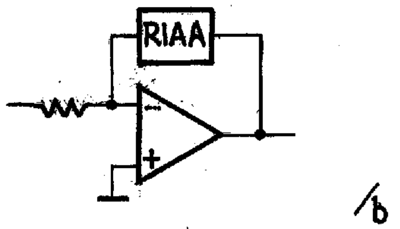
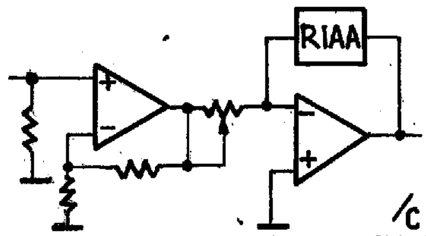
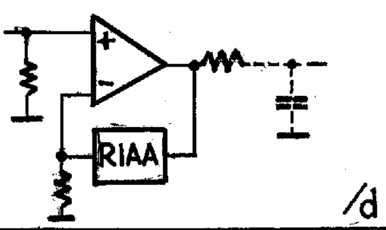
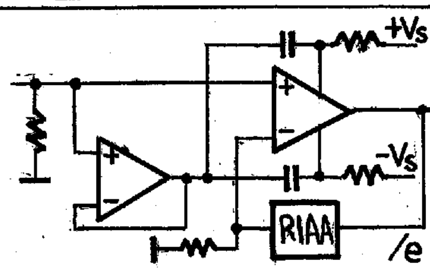
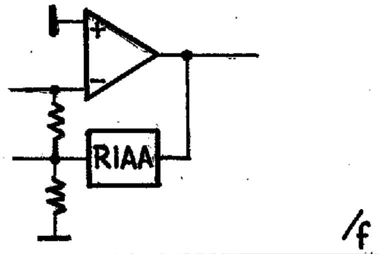
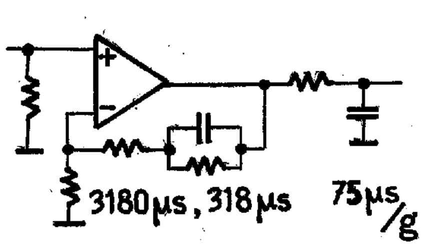
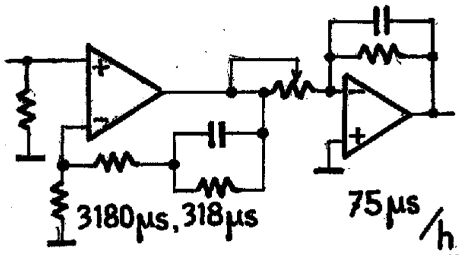
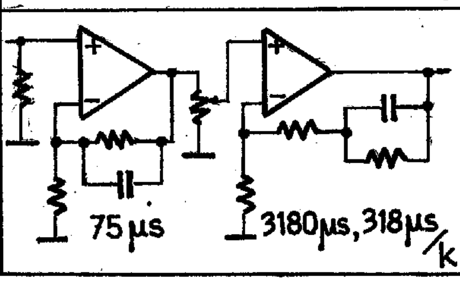
Series feedback connection (Table 1d) is widespread and well known. The almost universal adoption of this circuit resulted from its simplicity and satisfactory performance. The RIAA preamps [11-13] with discrete op amps in forward path show relatively good results on oscilloscope and in THD tests but not in listening. The loop gain is of 28 to 48 dB at HF and of 40 to 60 dB at LF with op amp cutoff frequency of 5 MHz. We should note that that LF feedback value is obtainable when the amplifier DC open loop gain reaches 100 dB and forward path dominant pole lies near 50 Hz. While these heavy amounts seem to be enough for low distortion reproduction, series configuration suffers from additional sources of distortion not removed by overall feedback:

* The long-tailed pair quiescent current source is shunted by small parasitic capacitance (between base and collector or gate and drain). A common mode input signal up to hundred milliVolts at HF leads to tail current and transconductance modulation and thus to even-order distortion.

* The base-collector junction (or gate-drain) capacitance of active amplifying transistor has been transferred in input network multiplied by gain value due to Miller effect. This capacitance is voltage-dependent therefore the input current is a nonlinear function of the input voltage. As the input signal spectral density and source impedance rise with frequency the result is relatively high in proportion to distortion products at HF [14].

* A dominant pole with large time constant in open-loop frequency response can lead to slew-rate limiting and other forms of TIM.

RIAA equalizer configurations Table 1.

		basic configuration	ref. No	advanced version	ref. No
with frequency-dependent attenuation network			2,3, 4,5, 8		
with overall frequency-dependent feedback	shunt feedback				10
	series feedback		11,12, 13;14		19,20
	floating source		14		
RIAA time constans are formed separately in two stages			24		25,26
			27		

There are many ways to overcome these problems. One is cascode (common-emitter with common-base) configuration in the input differential stage. For example, in GRIMS0N ELEKTRIK Preamp Model CPR1S51 [18] the serial cascode with fixed base potential is used. Cascode configuration with flying base potential as well as common-mode local feedback loops can also be used in prestigious units. These techniques are well documented [15-17].

TIM may be eliminated by increase in cut-off frequency and first stage clipper threshold levels. These methods are treated elsewhere [17]. Low-noise FETs are a convenient choice for the first stage due to high threshold voltage, as signal-to-noise ratio is worsened with emitter degeneration (local current feedback).

The method [19] for substantial improving of series feedback connection performance is realized in JVC phonograph preamps [20]. It is called FFPS and applies a floating power supply with feedforward and bootstrapping (Table 1e).

At HF the feedback network has low impedance and the RIAA preamp becomes a voltage follower. The zero in close-loop transfer function lies at the frequencies of 20 to 200 kHz and the deviation from standard response may be considerable. The RC integrating network on the output [18,22] (Table 1d, the dashed lines) with the same time constant removes the zero from overall transfer function.

The feedback resistor between inverting input and ground is chosen to be several times lower than MM cartridge dc resistance to meet the low noise requirements. There are limited possibilities of gain control in series-feedback connection, and power supply voltage is chosen frequently to be high enough to ensure a certain margin for output-stage clipping. This is the reason for output current capability enhancement. Low crossover distortion is obtainable with a class-A output stage and in other complicated designs as 'Class AA' in TECHNICS SU-A200 Equalizer Amplifier [22]. The object of method [21] is to perform the task of making a low-impedance load appear as high impedance to the preamp output. The circuit contains a subsidiary amplifier with a resistance bridge.

The floating source connection is shown in Table 1f. An input signal is introduced in the feedback network so that the subtraction process is performed external to the preamplifier. This circuit [14] has the overload capability of the shunt feedback connection but retains the signal-to-noise ratio of the series feedback configuration. It is easy to introduce the floating input to a conventional amplifier as a reference input. It is convenient to compare performance specifications for both inputs in order to define nonlinearity associated with common mode signal.

In configurations listed above the RIAA two-node circuit forms frequency response. At least, two circuit element values are mutually dependent so they are obtained by iteration procedure. Moreover, one component value strictly fixes others.

The RIAA equalizers with time constants formed separately in several stages are more versatile. We can build them proceeding from independent design principles. In circuit, shown in Table 1g, the higher time constants are formed by series feedback connection while 75 us time constant is formed in the output integrating RC network. The amplifier output is suffered from high-level HF contents, and the gain control cannot be realized. The supplementary amplifier [25,26] with 75 us time constant in shunt feedback network (Table 1h) contributes to overcoming these difficulties.

There is another configuration [27], with 75 us tine constant formed in the first stage by feedback network and 3180 us and 318 us - in the second one (Table 1k). The circuit has an advantage that second stage feeds from source with natural HF distribution. This fact together with sufficient feedback in both stages determines the lowest distortion figures. The same behavior is observed in case of preamplifiers [28,29] with 75 us time constant provided by integrating RC network between first and second gain blocks. Within the scope of the authors' experience the latter connections achieve the closer agreement between measurement data and subjective testing.

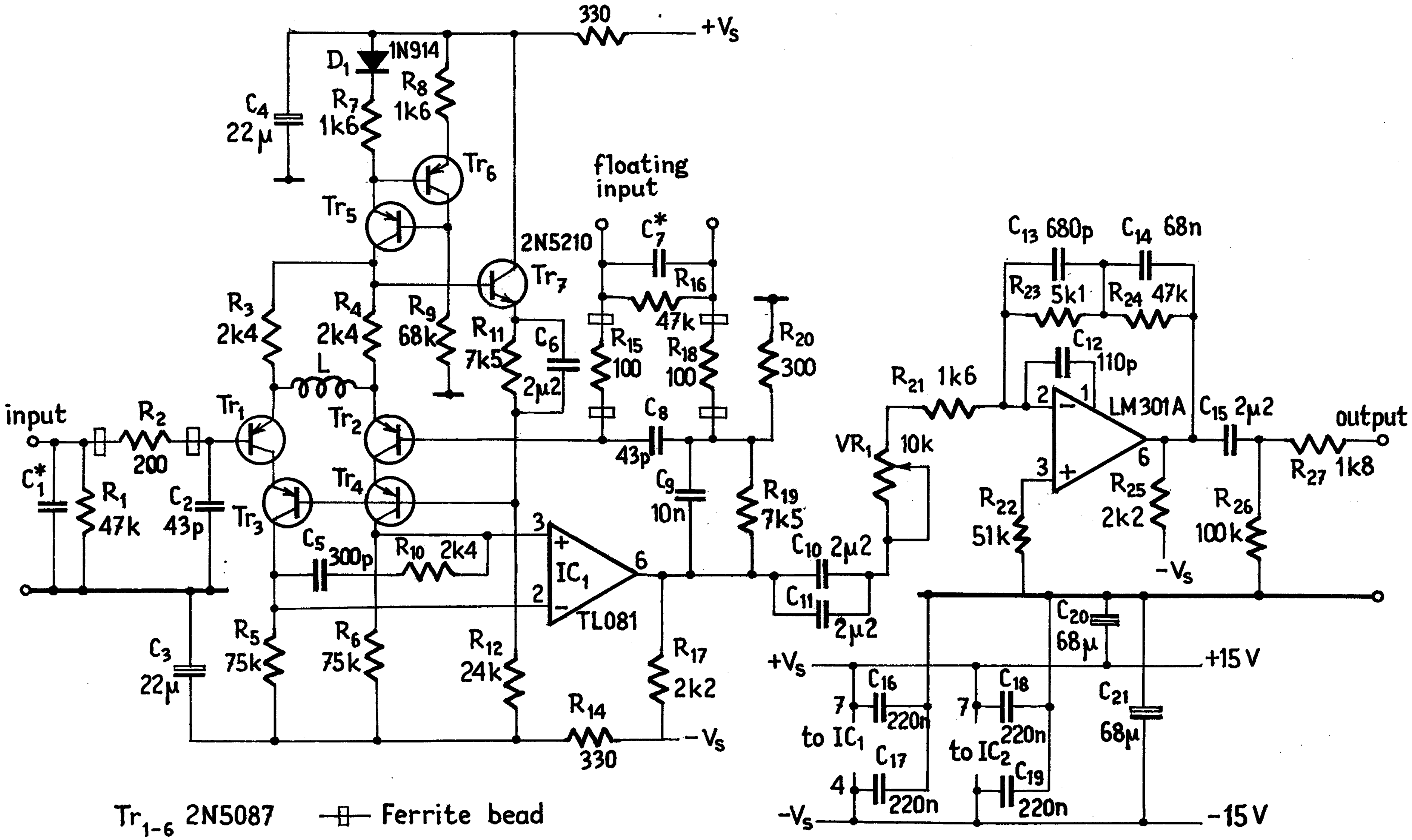


Fig. 1. Complete circuit diagram of the preamplifier. C_1 and C_7 are adjusted for given cartridge. Inductance of coil L is 32 μH.

Circuit description

The preamp has been purposely designed as a controversy to discrete fifty-transistor monsters; we have tried to obtain good results from low-cost and easy-available components. The complete circuit diagram is shown on Fig 1. It follows Table 1k configuration, but with shunt feedback connection in the second section. First stage with LF351A op amp forms the 75 us time constant. The second one with LM301A op amp forms the 3180 us and 318 us time constants.

The first stage

The first stage consists of the long-tailed pair input stage and the op amp. This approach [31,32] has the merit that an input stage can be optimized for low noise and interface distortion. Several simple configurations are shown in Fig 2. Time constant of the feedback network is R19C9=75 us, the DC gain is R19/R20+1=26 and gain at 1 kHz is equal to 23.5. The time constant determined by R19C19/(R19/R20+1)=2.9us corresponds to the zero in first section close-loop transfer function. The extra breaking point is compensated in the second section.

There are neither a series capacitor in the feedback network nor a decoupling one in the input circuit. We choose DC coupled connection with cartridge to avoid excess noise at LF. An input decoupling capacitor and DC cartridge resistance form relatively a small time constant with typical frequency about 500 Hz. Below this frequency the source impedance becomes predominantly capacitive, and the input resistance thermal noise source is no longer shunted by low cartridge DC resistance (about 500 Ohm). A high value electrolytic capacitor in feedback network exerts a strong effect on sound quality, so we decided to reject it. Output offset voltage is caused by input transistor mismatch and its value does not exceed 100 mV.

The floating input is introduced for performance comparison. If the shielding and guarding are properly applied there will be no problems with hum and instability even with 5 ft long input cable. If the floating input is not used, it must be short-circuited (or elements R15, R16, R18, C7, C8 may be omitted at all and C8 is short-circuited).

To avoid RFI the input low-pass filters R2C2 and RL5R18C8 are introduced so are the ferrite beads. The capacitors C1, C7 are chosen equal to nominal value for given cartridge's less cable and preamp input capacitance. The preamp input capacitance is equal to 50pF and a cable capacitance is about 40pF/ft usually.

The op amp output stage works in class A with the quiescent current 6.8mA flowing through R17. With the input stage an additional pole in open-loop transfer function appears, formed by R5, R7 and the op amp input capacitance. Frequency compensation network [33] R10, C5 produces a phase lead around unity-gain frequency. The transient response overshoot is about 30% and the phase margin is about 40 degrees.

Tr5, Tr6 current source set the input transistors quiescent current to be equal to 100uA [9]. During testing of the simple long-tailed pair (Fig 2a) it was found that the differential-mode nonlinearity and the nonlinear input current are the main distortion sources. Thus an input stage was consistently rebuilt (Fig 2b, c). The emitter degeneration resistors R3, R4 increased the clipping threshold ten times, but their thermal noise decreased the signal-to-noise ratio on 3.5 dB. We introduced the coil L in local feedback loop formed by R3, R4, L. The coil impedance at 20 kHz is equal to 4 kOhm, so the loop acts at HF only. The input stage clipping threshold level becomes frequency-dependent and corresponds to differential input spectra [34]. The measurement details are given in Table 2.

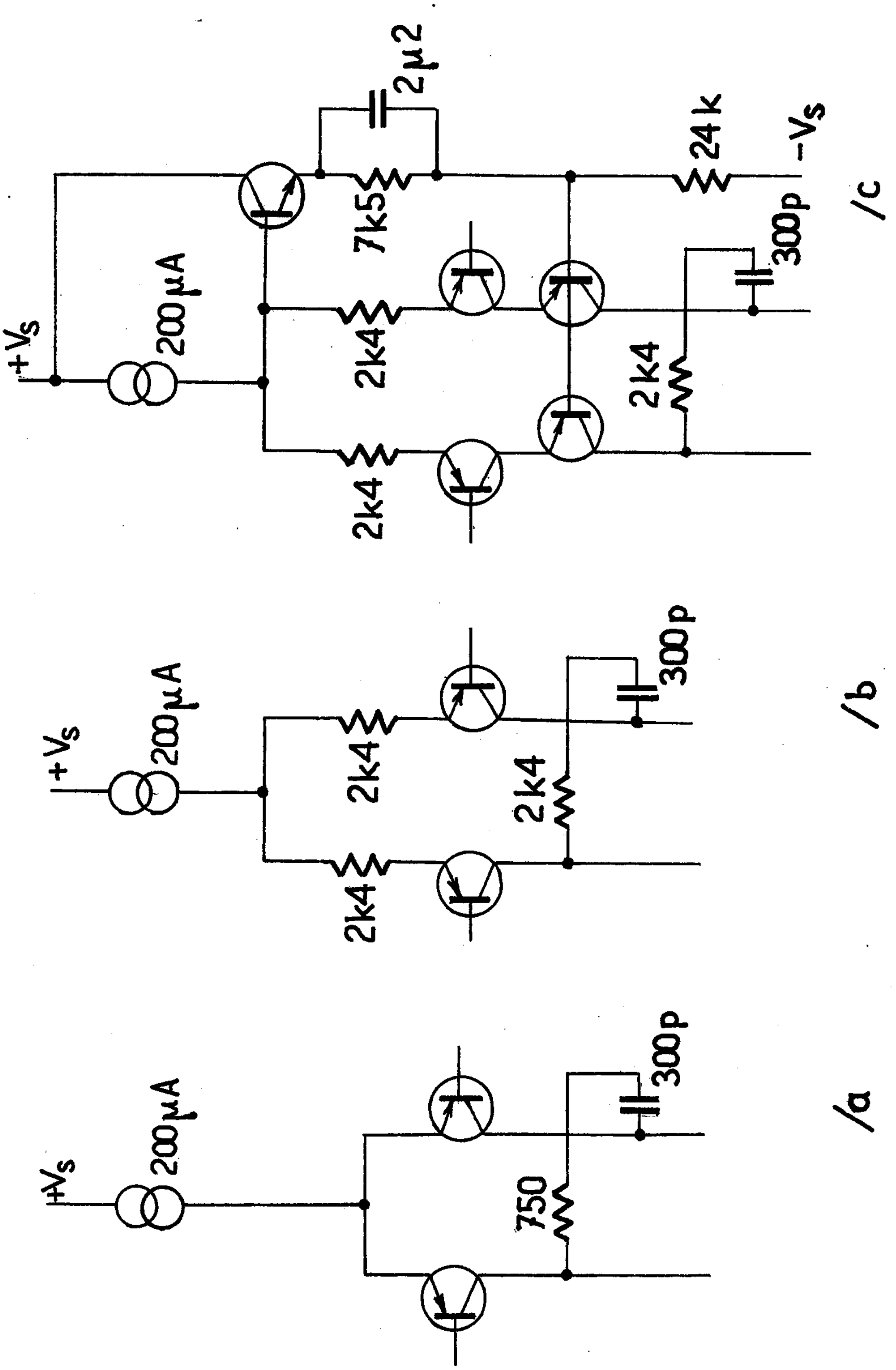


Fig. 2. Some input-stage configurations.

The second stage

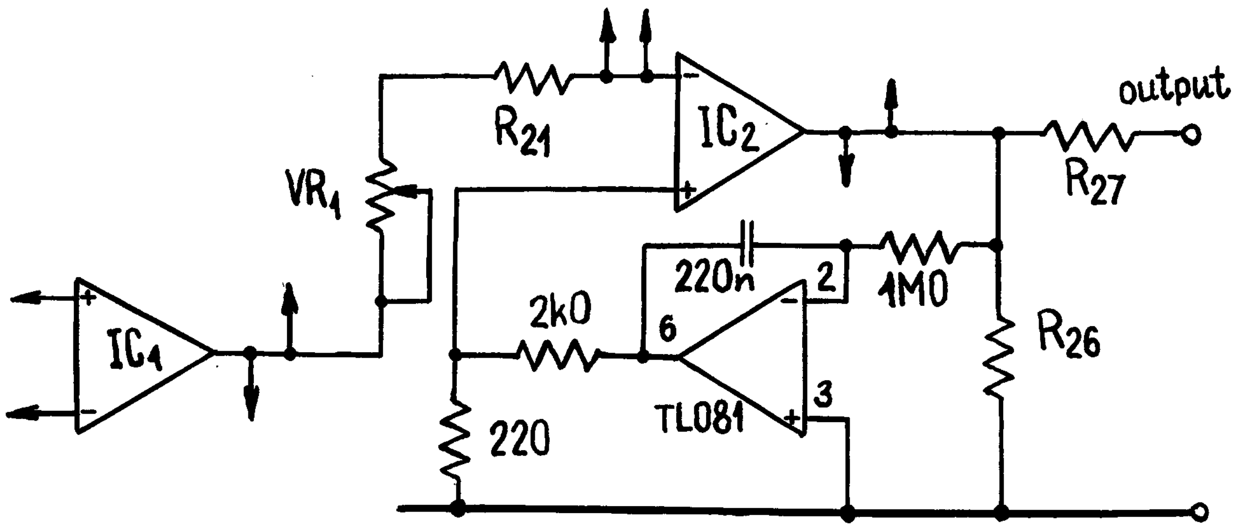
The second section is the ordinary shunt feedback amplifier. The feedback network time constants are:

R24C14=3180 us;

(C13+C14)R23R24/(R23+R24)=318 us;

R23C13=2.9 us.

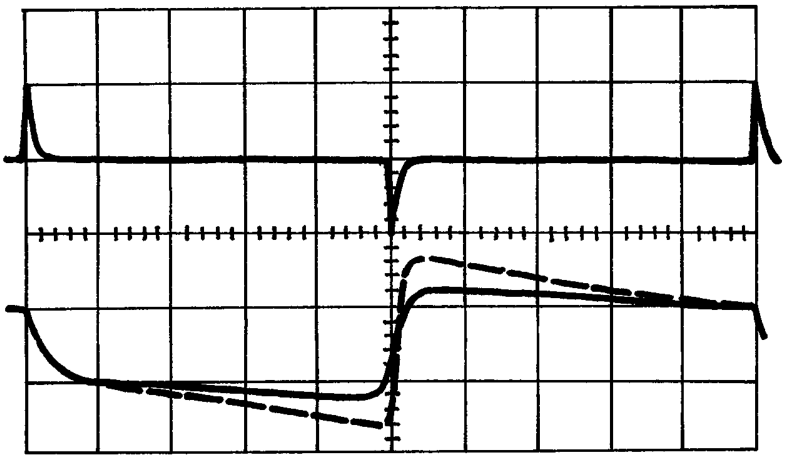
The section gain at 1 kHz varied from 0.5 to 3.5 by VR1 gain control pot. The op amp is feedforward compensated and its output stage works in class A, too. We chose decoupling polyester capacitors C10, C11, C15 ignoring 7950 us RIAA-78 zero time constant. This time constant is formed precisely in VRL maximum gain position only. We are sure that excess phase lead at LF, produced by a number of decoupling networks can really lead to worse and unpleasant low-end reproduction. A simple dc servo loop around the second section can easily be introduced. The part of decoupled version is presented in Fig 3. Turntable vibrations and acoustic breakthrough from associated loudspeakers at LF may be eliminated by low-cut filter in main amp. Another approach is M-S (middle-side) matrixing with LF attenuating on the side signal only.



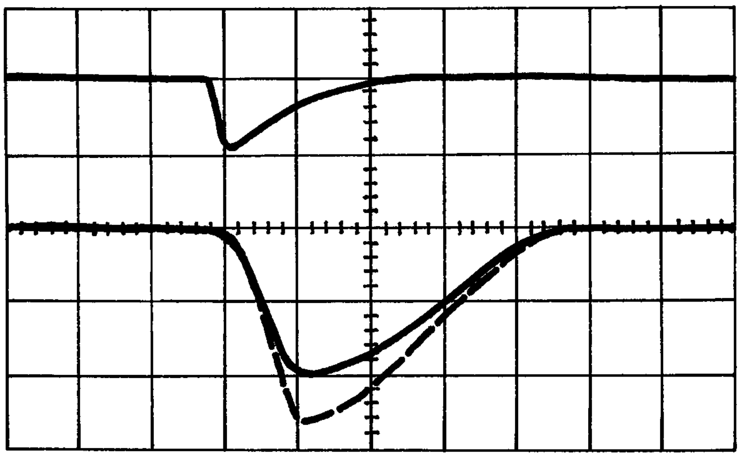
Testing

The main specifications are given in Table 2. Signal generator residuals are -110 dB for the second and -120 dB for the third harmonics. THD was measured with a frequency analyzer with a twin-T input notch filter tuned on the fundamental for increasing the equipment dynamic range. Input resistance of the twin-T notch filter is equal to 1.8 kOhm at resonance frequency. The fundamental frequency attenuation is more than 54 dB and the second harmonic attenuation is equal to 4.5 dB. THD figures exceed the generator residual only at HF.

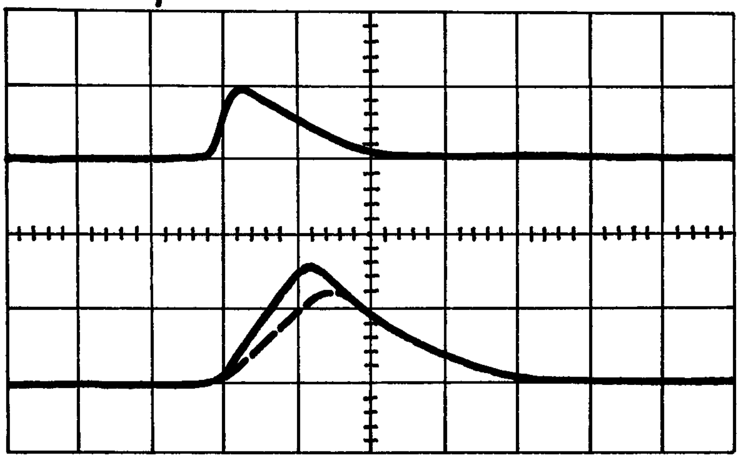
Intermediation distortion was measured with the active inverse RIAA circuit and is below the measurement floor.



a) The preamp input (upper trace, 5 V/div) and output (bottom trace, 2 V/div) signals, time-base- 10 µs/div.



b) The preamp input (upper trace, 5 V/div) and input amplifier feedback (bottom trace, 2V/div) signals, time-base- 1 µs/div.



c) The same signals as on (b), but polarity is reversed. The preamp response with RIAA-preemphasized 10 kHz square wave. Dotted lines – Tr_7 base-collector junction is shunted by 20 pF capacitor.

Low distortion results from large feedback in both amplification stages. The loop gain is higher than 40 dB at 20 kHz. The addition of a differential stage with symmetrical outputs permits the common-mode rejection increase by the gain of a differential stage [15]. A further improvement is achieved by decreasing the input stage output admittance. It is attainable with a common base stage [15,16] (Fig 2c), while the common-mode range is degraded on 3 V (approximately the base potential of a common base stage) compared to Fig 2a,b. The distortion figures for the ordinary and floating inputs are nearly the same and we give THD data for regular input in Table 2.

The stability test on the RIAA equalizer preamp was performed by feeding from a square-wave source via the inverse RIAA circuit [35]. The waveform on the output of an inverse RIAA circuit contains large spikes; such as never delivered by a pickup. Nevertheless, raising the test level to a high value simultaneously stresses both level and slew-rate abilities.

This simple test detects a whole variety of nonlinearities, specifically, connected with an input stage. The high signal slope initiates a slew-induced distortion mechanism and a true slew-rate data is obtainable. The pre-emphasis differentiation bandwidth and signal slope may be limited by a second-order filter.

The bandwidth in stability test was limited by a frequency of 200 kHz - the differentiation network breaking point. There are visible distortions in output square waveform with a peak input voltage of 5 V. It was found that the small asymmetrical tilt (Fig 4a) is caused by Tr5 and Tr7 base-collector junction capacitance. The details are given in Fig 4b, c where the input (base Tr1) and feedback signals (base Tr2) are presented. Feedback signal observation gives a multitude of details that are shadowed in the output signal by RIAA de-emphasis integration process. We used an oscilloscope with 50 Ohm input resistance and 1:30 input probe divider for low capacitance inserted in the feedback point during measurements. The negative voltage spike makes Tr1 turn on (Fig 4b) and its collector current is not limited. The current source capacitance charges rapidly. The positive voltage spike makes Tr1 turn off (Fig 4c) and its collector current is reduced to zero. The current source capacitance discharges slowly by tail current. So the feedback signal spikes are not equal, and integration process transfers this inequality in square wave tilt. The same response was measured with 20pF additional capacitor, connected in parallel with Tr7 base-collector junction. It is shown by the dotted lines in Fig 4.

This problem can be solved easily in Fig 2a, b input stages by inserting 15 kOhm resistor in series with current source. There are only few RAA preamps, to the authors' knowledge, which can deal with the same input signals without gross distortion.

Subjectively, the preamplifier has considerable amounts of detail, sharp transients such as a fortissimo cymbal clash are reproduced exactly and sibilants are not emphasized.

Acknowledgments

Fruitful discussion with Reg Williamson and Stanley Lipshitz are gratefully acknowledged, we want to thank also Ms. Z.V.Novitskaya for assistance in developing this paper.

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Please note we do not make any pretensions of completeness in the references cited. We have cited what we hope to be a selection of the relevant sources.

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Main specifications Table 2

<u>Gain</u>

Gain at 1 kHz	12x... 82x C21 dB... 38 dB)
Maximum output voltage (at the output of IC2)	+12 V, -15 V, peak, open circuit, +8 V, -12 V, peak, 1.8 kOhm loaded

<u>THD</u>

All measurements were made at the output of IC22, with 1.8 kOhm load. An output level 6.5 V peak, 2 dB below clipping, VRL in maximum gain position. Ordinary input.

The preamp input is fed directly from signal generator.

An input stage	10 kHz	20 kHz
Fig.2a	2nd 0.0004% (-108 dB)	2nd 0.0013% (-98 dB)
Fig 2b		
Fig 1		

* below the measurement floor (-110 dB for the second and -120 dB for the third harmonic)

The preamp input connected with cartridge equivalent (600 mH and 500 Ohm in series)

An input stage	10 kHz	20 kHz
Fig.2a	2nd 0.002% (-94 dB)	2nd 0.006% (-89 dB) 3rd 0.0004% (-108 dB)
Fig 2b	2nd 0.0008% (-108 dB)	2ndi 0.003% (-90 dB)
Fig 1		

* below the measurement floor

<u>Signal-to-noise ratio</u>

unweighted, 50 Hz - 20 kHz, referred to 5 mV rms input, with cartridge equivalent (600 mH and 500 Ohm in series)

An input stage	S/N
Fig 2a	- 76 dB
Fig 2b	- 72.9 dB
Fig 1	-79.9dB

Parts list Table 3

<u>Resistors</u>

R1, 16, 24	47 k
R2	200 5%
R3, 4, 10	2.4 k
R5, 6	75 k
R7, 8, 21	1.6 k
R9	68 k, 5%

R11	7.5 k, 5%
R12	24 k, 5%
R13, 14	330, 5%
R15.18	100, 5%
R17, 25	2.2 k, 5%
R19	7.5 k
R20	300
R22	51 k, 5%
R23	5.1 k
R26	100 k, 5%
R27	1,8 k, 5%

All resistors 1/4 to 1% unless otherwise specified.

Capacitors

C1, 7	see text	polystyrene
C2, 8	43 pF 10%	Ceramic
C3, 4	22 uF 25V	Electrolytic
C5	300 pF 10%	ceramic
C6, 10, 11, 15	2.2 uF, 5%, 63V	polyester
C9	10 nF 1%	polystyrene
C12	110 pF 10%	Ceramic
C13	390 pF 1%	polystyrene
C14	68 nF	polystyrene
C16, 17, 18, 19	220 nF	ceramic
C20, 21	68 uF 25V	electrolytic

Diode

DI 1N914

Transistors

Tr1,2,3,4,5,6,	2N5087
Tr7	2N5210

Integrated Circuits

IC1	LF351A (TL071C)
IC2	LM310A

Miscellaneous

VR1	10 k, linear pot
L	32 mH, wounded on small (1/2 inch) ferrite anchor ring with permeability of 1500
Ferrite beads	