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Perkins Electro-Acoustic Research Lab, Inc. ❦ Engineering and Intuition Serving the Soul of Music

Part I

A MOVING COIL PREAMP

BY ERNO BORBELY
Contributing Editor

FACING THE DANGER of being attacked by MM-pickup fans, I must admit I have been an MC fan all my life. I guess it all started in the mid-1960s with the old Ortofon mono MC-pickups we used exclusively at the Norwegian Broadcasting Corp. Thanks to Dave Hafler, I was introduced in the early 1970s to one of the earliest Ortofon stereo MCs. I later graduated to some Denons, and now live with the best MC I ever heard, a Clearaudio PSO.

This early exposure to MCs raised my interest for the difficult task of designing step-up devices. Not being a transformer designer, I concentrated on active devices, and through the years, I have designed nearly a dozen of them. One went into production as the pre-pre for the Hafler DH101.

During my work on the EB-585 preamp, my highest priorities were considerations for MC-input. I first concentrated on offering high gain RIAA inputs that could take MCs directly. I soon found, however, that covering the whole range of low-to-high output MCs called for unacceptable compromises, so I decided against it. I also found a high gain RIAA stage was a relatively easy way to take care of what I call the medium-to-high output pickups. This is one of the approaches I will present in this article. For low-to-medium output pickups, or if you cannot change the gain in the normal RIAA stage, I propose a separate pre-pre. In both cases, I believe the sound quality justifies the investment.

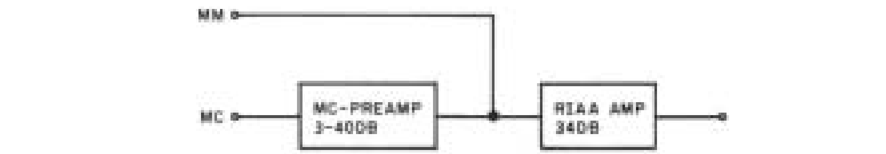

FIGURE 1: To cover the entire range of MCs, you need 3-40dB gain ahead of the MM input.

MC-preamp considerations

If you look at the annual directory published by *Audio* (October 1985), you will find the lowest output MC is the Ortofon MC2000, which has an output of 0.05mV at 5cm/S RMS lateral velocity. The one with the highest output appears to be the Dynavector DV-20B2, with 3.6mV at the same groove velocity. While the latter one has an output which is close to the MM-pickups (5mV), and needs only a couple of decibels extra gain, the Ortofon MC2000 has nearly 40dB less output than the MMs. Although, to my knowledge, no clear definition exists, I consider the 0.5mV output to be a medium one and will refer to it accordingly in the remainder of this article.

To handle all MCs, you will need anywhere from 3 to 40dB gain in front of the normal MM-input (*Fig. 1*). This must be accompanied by very low noise to preserve the system's dynamic range. Obviously, if you need only a couple of extra decibels to cover the high output MCs, you can use the normal RIAA amp. After all, all preamps should have enough reserve gain to cover such demands.

It is becoming a bit more difficult to cover the whole range of medium-to-high output MCs because this requires up to 20dB extra gain in front of the RIAA stage. The most convenient approach would be to permanently incorporate these extra 20dBs into the RIAA amp. Unfortunately, it would spell disaster with normal MMs; they would overload the input. The best solution is to allow gain selection in the RIAA amp in, say, 6 or 10dB increments, between the traditional 34dB and the desired gain of 54dB (*Fig. 2*). As you will see later, this is the approach I will propose for the EB-585 preamp.

We can go one step further and say that even low-to-medium output pickups can be handled by adding yet another 20dB gain to the RIAA stage. Although I don't intend to cover the whole 40dB range (most of us operate with sensitivities in the 0.1 to 5mV range), you will still need approximately 32dB extra gain in front of the RIAA amp. In most cases, this will stretch the capabilities of most designs beyond their limits. As a result, severe compromises must be accepted, notably in dynamic range, THD, and most likely in RIAA-accuracy. For such pickups, a separate step-up device is the answer.

For small signal amplification, which is what an MC-preamp is sup-

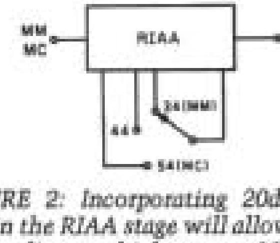

FIGURE 2: Incorporating 20dB extra gain in the RIAA stage will allow you to use medium-to-high output MCs.

posed to provide, I consider the following points to be most important for sound quality:

- Coupling the signal source to the input
- Dynamic range of the amplifier
- Input non-linearities of the amplifier, and
- Interference from power supplies.

These might seem obvious to you, but I would like to comment on some of them. Concerning coupling the signal source to the input, it is important that you maximize the input power to the amplifier. You can do this with impedance matching, which is what you do for microphones and for MC-pickups with transformers. In addition to providing power matching, transformers can also be considered noise-free. They will, therefore, always give superior performance as far as signal-to-noise ratio is concerned. Transformers, however, have disadvantages as well. It is difficult to design them with high overload capability and without resonances when the turns ratio is high. Also, they are very susceptible to hum pickup from surrounding magnetic fields. Well-designed transformers with double shielding tend, therefore, to be very expensive. I believe the best compromise is a hybrid design: use a transformer with low turns ratio which, with relative ease, you can make for wide bandwidth and high overload capability. Then, provide the rest of the gain in active form. I considered this for the present design, but due to the high cost of quality transformers, I decided to use an all-active approach.

An all-active design, however, requires a careful consideration of the amplifier's dynamic range. More specifically, you must pay special attention to the lower limit of the dynamic range, i.e., the amplifier's input noise. I am going to spend some time on the theoretical side of this subject, so if you are only interested in the actual design, skip this section and turn to the description of the MC-preamp.

Basic Noise Theory

As with distortion, noise can be considered anything which, when added to the signal, reduces or changes its information content[1]. It is, therefore, of paramount importance that you reduce this noise level so that it no longer influences the information content. In our case, information stands for music.

There are four main types of audio amplifier noise mechanisms: thermal noise, shot noise, 1/f or flicker noise, and popcorn noise. Thermal noise, which results from thermal agigation of electrons in a conductor (e.g., a resistor), was first observed by Johnson of Bell Labs in 1927 and was analyzed by Nyquist in 1928. Because of their work, thermal noise is sometimes called Johnson or Nyquist noise. Thermal noise is given by the following formula:

$$e_n = \sqrt{4kTR\,\Delta f} \qquad (1)$$

where:

e_n is the RMS noise voltage in volts
k is the Boltzmann's constant (1.38×10^{-23} Joule/degree K)
Δf is the noise bandwidth in Hz
R is the resistance in Ω.

Thermal noise is frequency-independent over a broad range of frequencies and is, analogous with white light, often called white noise. The generation of thermal noise is not affected by the flow of current through the resistor R. In fact, even if you suspend it on a silk thread in free space, it will still generate a noise voltage because the resistor picks up thermal energy from the ambient heat sources. Ideal inductors and capacitors do not generate thermal noise. In the case of complex impedances of the form $Z = R(\Omega) + jX(\Omega)$, such as an MM-pickup, the real part of the impedance is responsible for the noise and is frequency-independent[2].

Shot noise results when a DC current flows through an electronic device. Since electronic current is composed of discrete charge carriers, fluctuations are always present when the current crosses a barrier because the carriers pass independently of one another. A typical example is a PN junction in a transistor, where the passage takes place by diffusion. These fluctuations generate a noise called shot noise, which is given by the formula:

$$i_n = \sqrt{2q\,I_o\,\Delta f} \qquad (2)$$

where:

i_n is the RMS noise current in Amp.
q is the charge of the electron (1.6×10^{-19} Coulomb)
I_o is the DC current flowing through the junction (Amp)
Δf is the noise bandwidth in Hz.

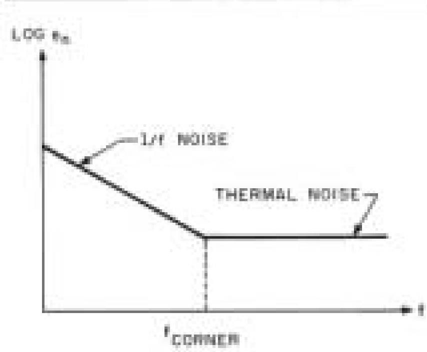

FIGURE 3: 1/f noise appears in most conductors, including semiconductors, vacuum tubes, and composition carbon resistors.

Again, the shot noise is frequency-independent over a broad frequency range. It is important to note that shot noise is not generated in electrical conductors (e.g., resistors) due to the long-range correlation between charge carriers.

Most conducting materials, including semiconductors and vacuum tubes, also have an additional noise component that is inversely proportional to frequency (*Fig. 3*). This noise is referred to as flicker, 1/f, or pink noise. 1/f noise also appears in composition carbon resistors made up of squeezed-together carbon granules. Current tends to flow unevenly through these granules, and this behavior gives rise to a noise which is proportional to 1/f. 1/f noise in semiconductors is caused mainly by surface problems, with important contributing factors being generation and recombination of carriers in surface energy states, and the density of

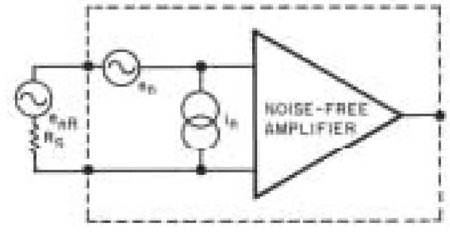

FIGURE 4: Amplifier noise model.

surface states[3]. Improved wafer fabrication and surface passivation techniques have significantly decreased 1/f noise in the last few years.

1/f noise is especially troublesome in amplifiers with bass boost, such as RIAA amps. The situation is further aggravated when DC coupling is used, because below approximately 1Hz, it becomes almost impossible to distinguish between 1/f noise and DC drift effects. In RIAA amplifiers, I always make two noise measurements: one over the normal 20-20kHz range, and one over the 200-20kHz range. Not considering hum here, the difference between the two RMS readings should be as small as possible, preferably not more than 6dB.

Popcorn noise, or burst noise, appears in all semiconductors and certain resistors. When fed to a loudspeaker, it sounds like corn popping. Popcorn noise is not white noise. Its spectral density varies with $1/f^{\alpha}$, with α often being 2 ($1/f^2$). I have seldom experienced popcorn noise in good, discrete devices, but I had lots of problems with it in the early days of IC op amps. Again, with today's high quality processing techniques, this is an almost non-existent problem for designers.

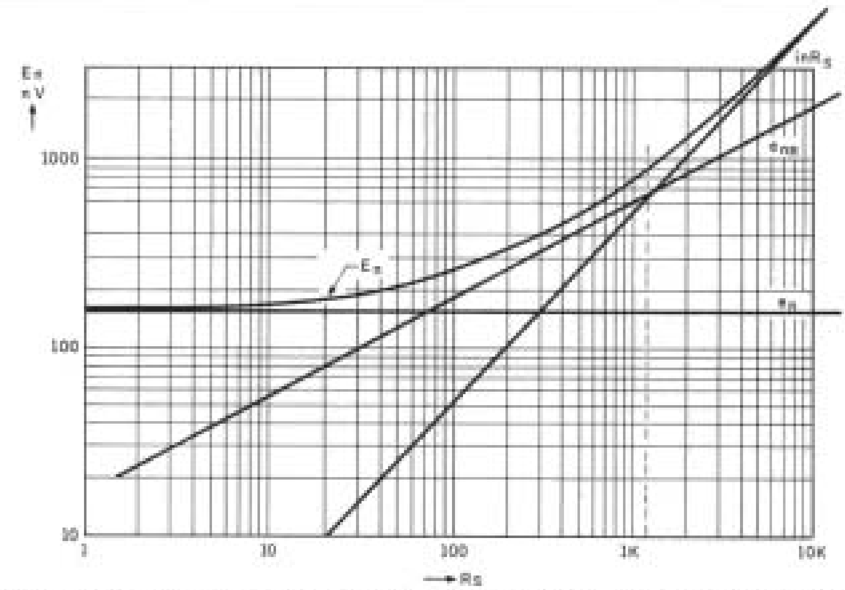

FIGURE 5: Equivalent input noise (E_n) versus R_s for a hypothetical amplifier (20-20kHz, flat, RMS).

Noise in Audio Amplifiers

As an audio engineer, I am interested in getting an acceptable signal-to-noise ratio at the amplifier's output. This means I must design my system for a minimum equivalent input noise. But what is equivalent input noise?

An amplifier's noise performance is usually described by modeling the noise sources as a series voltage generator: e_n, representing the thermal noise, and shunt current generator, i_n, representing the shot noise (*Fig. 4*). The source resistance noise (your MC pickup) is represented by one more series voltage generator: e_nR.

The equivalent input noise of the amplifier, resulting from all three noise sources, is given in the formula:

$$E_n = \sqrt{e_n^2 + e_nR^2 + i_n^2R_s^2} \qquad (3)$$

Notice that we have three different terms here. The first term, e_n, is independent of the source resistor R_s. The second term increases with R_s, and the third term increases with R_s^2.

Figure 5 shows the equivalent input noise versus the source resistance for a hypothetical amplifier, represented by three asymptotes corresponding to the three terms in formula 3. This is a general curve, which applies to all types of active devices. Of course, the levels and the intersection points between the three asymptotes will be different for the different amplifiers. As you can see, e_n alone is important at very low values of R_s, and is called the amplifier's equivalent short circuit input noise. This is a key issue with MC-preamps, where the source impedence is usually in the range of 2-50Ω. Consequently, MC-preamps must have very low e_n.

You can calculate the noise contribution (e_nR) from the source resistance (R_s) by using Johnson's formula. For example, a 1kΩ resistor will, over a 20kHz bandwidth and at room temperature (300°K), generate:

$$e_nR = \sqrt{4x1.38x10^{-23}x300x2000} = 0.575\mu V \qquad (4)$$

I have calculated the noise for resistors between 1Ω and 10kΩ for the same conditions and plotted the results in *Fig. 5*. Although you will invariably end up with different values for the first and third terms of the equivalent input noise for different amplifiers, the term e_nR will always be the same, and you can use it as a universal graph to determine the noise contribution of different source resistances. For example, my Clearaudio pickup, with approximately 50Ω, generates 128nV or 0.128μV of noise in the audio frequency range. More about this later.

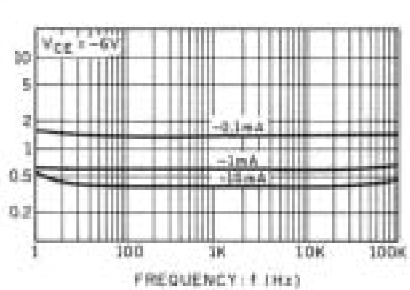

FIGURE 6a: Noise voltage versus frequency for 2SB737. h_{FE} = 270-560 (S-Group).

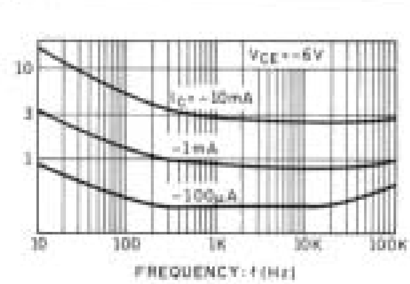

FIGURE 6b: Noise current versus frequency for 2SB737. h_{FE} = 270-560 (S-Group).

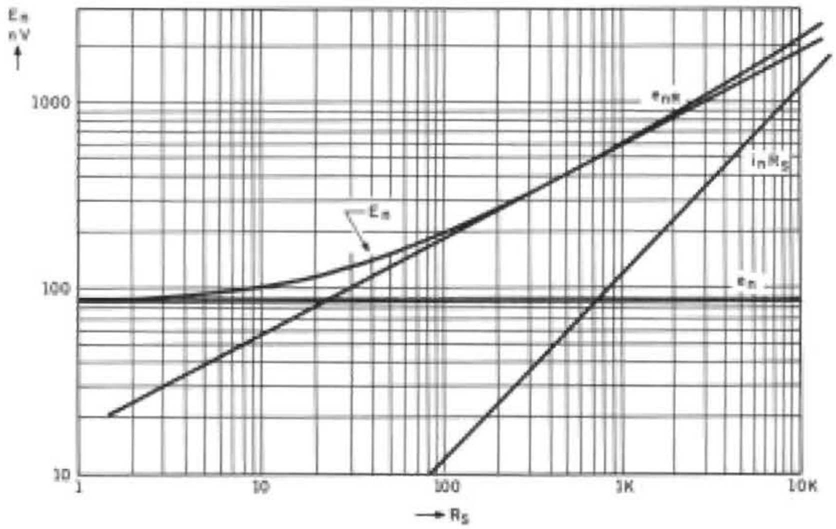

FIGURE 7: Equivalent input noise (E_n) versus R_s for an amplifier using 2SB737.

When you use high source impedances, it is important to consider the contribution of term three: shot noise (in formula 3). This is especially important with bipolar transistors, as we will see in a moment. With FETs and vacuum tubes, shot noise usually doesn't make any contribution in audio applications, and the equivalent input noise is determined by e_n and e_nR.

The curve for E_n is composed of all three terms in formula 3, and at the intersection points of two asymptotes, both contribute equally. The resultant mean square voltage, however, is the sum of the two mean square voltages, so the increase is 3dB at this point (contrary to 6dB when you add two sine waves with the same frequency and amplitude).

Let's now look at the noise sources of the active devices we use in audio amplifiers: bipolar and field-effect transistors.

Bipolar Transistors

The two equivalent noise sources for a bipolar transistor are given by:

$$i_n = \sqrt{2q\, I_B\, \Delta f} = \sqrt{2q\, \frac{I_E}{h_{fe}}\, \Delta f} \qquad (5)$$

and

$$e_n = \sqrt{4kT\left(r_{bb}' + \frac{r_e}{2}\right)\Delta f}$$

where

$$r_e = \frac{kT}{qI_E} \qquad (6)$$

I_B and I_E are the base and emitter currents, and r_{bb}' and r_e are the spreading base resistance and the emitter small signal resistance respectively. Current noise is shot noise caused by the base current, while voltage noise corresponds to the thermal or Johnson noise of the base resistance, plus one half of the small signal emitter resistance in series.

The noise sources listed above are independent of the collector voltage as long as the leakage current is negligible. They are also independent of the transistor configuration: common base (CB), common emitter (CE), or common collector (CC). Due to its unity voltage gain, the common collector (or emitter follower) configuration is a bad choice for low-noise applications. Though you can use both CB and CE configuration in low-noise audio amplifiers, one might be better than the other in a particular application due to differences in voltage gain, current gain, and input impedance. Both configurations have been used extensively in MC-preamps, while the CE configuration is the natural choice for MM-pickups.

From the above mentioned equations, you can see that a transistor with high h_{FE} and low r_{bb}' is generally best for minimum noise. With low source impedance, you can optimize the emitter current for the given source resistance (R_s). You must, however, select a transistor with low r_{bb}' to begin with. In the early days of MCs, we spent much time selecting transistors for low r_{bb}'. I believe John Curl was the first to find 2N4401/4403 switching transistors were suitable for low impedance applications. Subsequently, these were used in many MC-preamps.

Typical values of r_{bb}' for general purpose transistors vary from several tens to several hundred ohm. You can reduce r_{bb}' by paralleling transistors, and reduce it further with special geometry transistors. To my knowledge, the lowest r_{bb}' devices on the market are the 2SB737(PNP) and 2SD786(NPN) from ROHM, having typically 2 and 4Ω. Input noise voltage and input noise current versus frequency for the 2SB737(PNP) transistor is shown in *Fig. 6a* and *6b*. Typical e_n and i_n values, read from these figures for a collector current of 1mA and at 1kHz, are: 0.6nV/$\sqrt{\text{Hz}}$ and 0.85pA/$\sqrt{\text{Hz}}$. Over a 20kHz bandwidth, this corresponds to: $e_n = 84$nV and $i_n = 120$pA. Using these values, you can construct a curve for the equivalent input noise versus R_s using, for example, one 2SB737 bipolar transistor (*Fig. 7*). Up to about 20Ω, the thermal noise (e_n) of the transistor will dominate. From 20Ω to about 10kΩ, the contribution from the source resistance

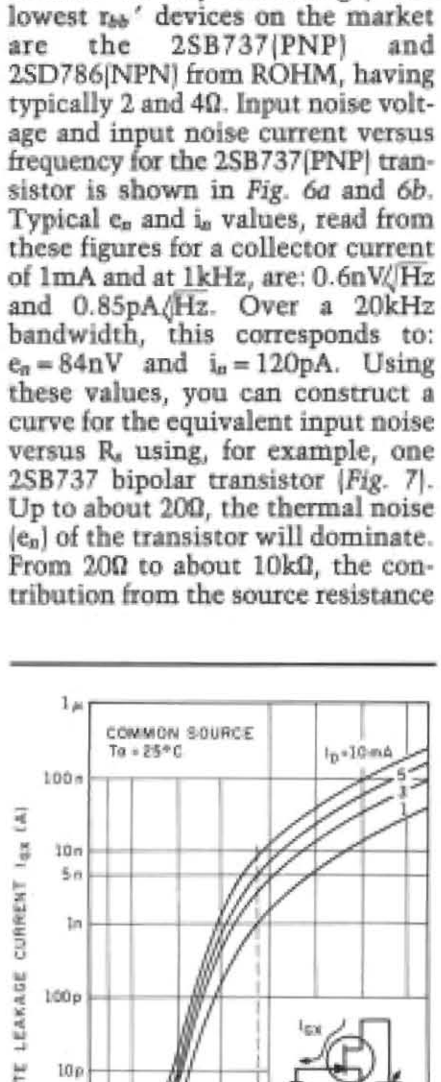

FIGURE 8: Gate leakage current for 2SK147.

(R_s) will determine the total noise. From 10k and up, the shot noise will take over. Paralleling N transistors will reduce e_n by the square root of N, but will increase i_n by the square root of N. It is, therefore, an advantage to parallel bipolar transistors for applications where R_s is low, but it is a disadvantage when the source impedance is high.

The E_n versus R_s values in *Fig. 7* are theoretical, and they are quite difficult to obtain in a practical circuit. We will look at some of the problems associated with practical circuits in the next chapter.

Field-Effect Transistors

If you are familiar with my previous articles, you probably expect me to talk about FETs because they are my favorite devices for audio use. The two noise generators, e_n and i_n, are expressed by the following formulas:

$$e_n = \sqrt{0.7 \times 4kT\,\Delta f \frac{1}{g_m}} \qquad (7)$$

where g_m is the transconductance,

$$i_n = \sqrt{2qI_G\,\Delta f} \qquad (8)$$

where I_G is the gate leakage current.

I have found the values calculated from formula 7 yield results that are too low when compared to the data sheet e_n values, or those measured in a real device. I attribute this to the source terminal's bulk resistance. This discrepency also exists with the Toshiba FETs you read about in my preamp article[4]. For our calculations, I will, therefore, use the values given in the datasheet. Reading e_n off from *Fig. 9* in the preamp article for 2SK147 at a drain current of 5mA, you get $0.75\text{nV}/\sqrt{\text{Hz}}$. For 20kHz, this results in $e_n = 106\text{nV}$. This is the same amount of noise produced by a 34Ω resistor. Therefore, the 2SK147 has an equivalent noise resistance of 34Ω. You can calculate the contribution by reading the value for I_G from *Fig. 8*: at a drain-source voltage of 15V, $I_G = 5\text{nA}$, which produces an $i_n = 5.6\times10^{-12}\text{A}$ over the 20kHz audio bandwidth. To compare the noise performance of a 2SK147 FET to the 2SB737 bipolar transistor, I have drawn the equivalent input noise versus R_s in *Fig. 9* for an amplifier

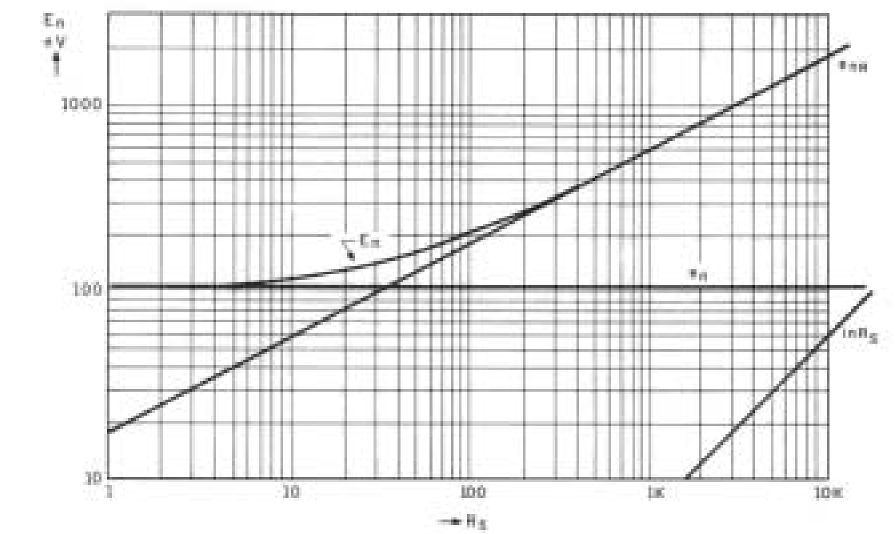

FIGURE 9: Equivalent input noise (E_n) versus R_s for an amplifier using 2SK147 FET.

using the 2SK147. e_n, being a bit higher for the FET, means the short circuit noise will be about 2dB higher than with the bipolar. Shot noise, however, will not make any contribution at all with source resistances lower than 100kΩ, and this is significantly better than the bipolar transistor.

Paralleling FETs will produce the same reduction in e_n as with bipolars, but having a much lower shot noise means it will still not contribute to the overall noise at high source impedances. Therefore, the overall noise performance of an amplifier with paralleled FETs is working with a wide range of source resistors is better than with bipolars.

Circuit Design Problems

As I mentioned earlier, my noise calculations were theoretical. Due to circuit design problems, it is very difficult to get close to these figures in practical amplifier circuits. Even if you are not an expert on circuit design, you can see that the two circuits shown in *Fig. 10* will not conduct at all; they require a base-emitter voltage of approximately 0.65V, or a base current to cause a collector current to flow. Even if you manage to get a collector current, you must make sure it is independent of variations of h_{FE}, temperature, and so on. In other words, you must add some bias network.

The same applies to the FET shown in *Fig. 10b*. As shown, it will conduct I_{DSS}, which varies greatly for a given FET type. Again, you must add a bias network that will reduce and stabilize the drain current to the value you want in your application.

Most practical transistor circuits include some emitter/source regeneration (local DC feedback) to achieve the above. In other words, the two circuits in *Fig. 10* would look more like the ones shown in

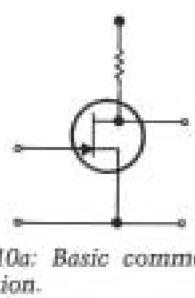

FIGURE 10a: Basic common emitter configuration.

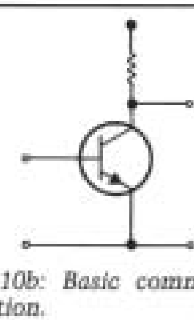

FIGURE 10b: Basic common source configuration.

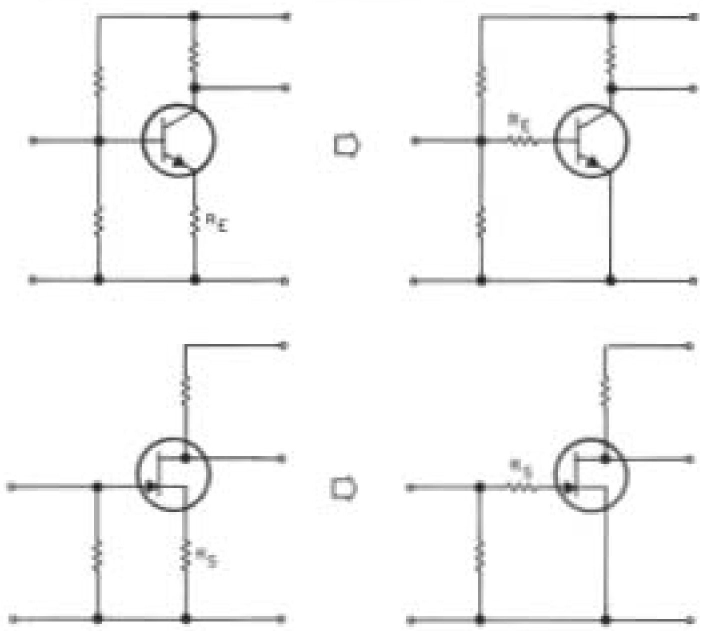

FIGURE 11: Biasing the transistors with RE/RS will increase the noise.

Fig. 11. It is also clear, however, that adding resistors in the emitter/source will increase the circuit noise. These resistors are connected in series with the input, as shown in the simplified equivalent noise circuits in *Fig. 11*. In other words, they act as though you have increased the source resistance.

To avoid this noise increase, you can decouple the resistors with capacitors. Unfortunately, these resistors are usually on the order of several tens to several hundred ohm, and an effective decoupling requires very large capacitors. Now, I don't want to get into an argument about capacitors, but I don't think I must convince you it is a bad idea to have large electrolytics in the input stage of an MC-preamp. With bipolar devices in the input, there is practically no way to avoid large electrolytics. With FETs, however, at least with the Toshibas, you can avoid electrolytics. More about this later.

Similar to the undecoupled emitter/source resistors, the overall feedback network also contributes to amplifier noise. *Figure 12* shows that, from the point of view of noise, the value of the two feedback resistors in parallel can be considered in series with the input. To minimize the noise contribution from the feedback network, R1‖R2 should be smaller than the source impedance.

MC-Preamplifier

As I mentioned earlier, you must have very low input noise in the MC-preamp to handle the low-level signals from moving coil pickups.

I have again chosen the topology I used for the RIAA-1 stage in my preamp (see *Fig. 8*, p. 11, *TAA* 4/85). In addition to being symmetrical, one of the advantages of this topology is wide bandwidth. The other one is low noise. The two input FETs appear to be connected in parallel, thus reducing the equivalent noise resistance by two, and the thermal noise (e_n) by $\sqrt{2} = 1.41$. Although this feature has not been fully exploited, the RIAA-1 amplifier is a very low-noise amplifier in its own right. It has an equivalent short circuit input noise of 200nV, and an equivalent input noise of 0.6μV with a 1k source resistance (20-20kHz, flat, RMS). To use this topology for an MC-preamp, however, you must fully capitalize on its low-noise capabilities. This requires the following modifications:

1. Paralleling FETs. As I have shown, paralleling transistors reduce the input noise voltage by the square root of N, where N is the number of transistors connected in parallel. Paralleling FETs, however, is a bit different from paralleling bipolar transistors. FETs usually operate at significantly higher currents than bipolars (5-10mA, compared to 0.5-2mA), so paralleling several devices will increase the total drain current to a very high level. Assuming we have 5mA drain current, as I used in the RIAA-1 stage, paralleling four will result in 20mA total drain current. To avoid a too high voltage drop across the drain resistor, adjust its value accordingly. Also, you must make special arrangements in the cascode amplifier so it can handle such a high drain current.

The very high input capacitance is another problem of the Toshiba FETs, since paralleling them will increase the capacitance. Four pairs of FETs will have an input capacitance of approximately 250pF, compared to the 60pF of the RIAA-1 amp,

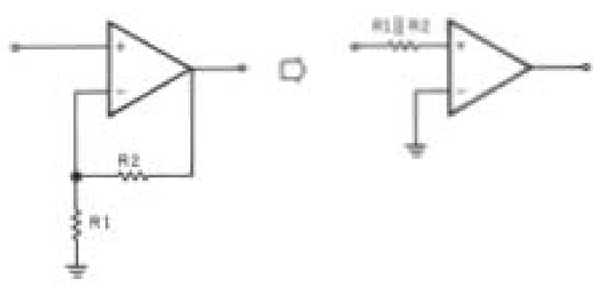

FIGURE 12: The overall feedback network also contributes to the noise.

which uses one pair. This is of no concern as long as you use the amplifier exclusively for MC-pickups. You must be aware of this problem, however, if it is used for both MM and MC types.

2. Feedback network. Because the feedback network also contributes to the total input noise, you must reduce it as much as possible. Ideally, the value of the two resistors in parallel should be lower than the lowest impedance pickup you are likely to use. Since the lowest impedance pickup is around 2Ω, it should ideally be 1Ω or so, as shown in *Fig. 13*. With a closed-loop gain of 10x or 20dB, the amplifier output must drive a 10Ω load, which is almost equivalent to a loudspeaker load. Clearly, you will run into problems driving such a load if you need, say, 10V RMS at the output. Fortunately, this is not necessary. Most normal RIAA inputs have an input overload capability of only a few hundred mV, which is what the MC-preamp must deliver. (You may recall the RIAA-1 overloads at 1V RMS, but the combined RIAA-1/RIAA-2 has 240mV overload capability at 1kHz.) Also, the minimum gain I am proposing here is not 20dB, but 26dB, which further alleviates the loading problem.

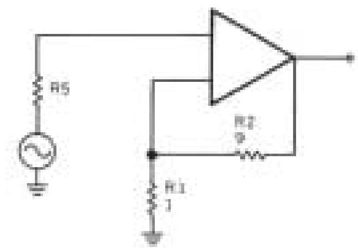

FIGURE 13: For low noise, R2||R1 should be lower than R_s.

Another problem with such a low impedance feedback network is the servo circuit must work across this 1Ω shunt resistor to correct the amplifier's input offset. If you have 30mV of offset, you will need 30mA through the 1Ω resistor to correct it. As you probably know, there aren't many opamps on the market that can deliver this much current, so you must compromise somewhat in terms of feedback impedance. □

In Part II, Mr. Borbely will discuss the MC-preamp's circuitry and power supply. A complete component list will also be included.

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4. Borbely, E., "The Borbely Preamp," *The Audio Amateur* (4/85, 1/86).

THE BORBELY PREAMP

Setup Procedure

General Comments

If possible, test each preamplifier module separately before installing it in the chassis. This simplifies measurements, adjustment and, if necessary, component changes. If you have access to a scope, connect it to the output of the module and check whether radio frequency (RF) oscillations are present. If you have complete audio instrumentation in your workshop, perform the usual gain, frequency response, noise, total harmonic distortion (THD), intermodulation distortion (IM) measurements. Inputs should be shorted under DC measurements/adjustments.

RIAA-1

Before switching on the module, set P1 to minimum position (CCW). If Q5, the servo amp is socketed, don't insert it just yet. Connect the plus and minus (±) 24V supplies to the module and make the following measurements/adjustments:

1. Check the two zener voltages, D1 and D2. They should be 15V. Check the output offset. It should be less than 1V.

2. Check the current in the input stage by measuring the voltage drop across R4/R14. It should read 2.8V, ±20%, corresponding to a current of 5mA. If it is outside the tolerance, you might have to change R7/R8. Before you do that, however, do a brief check of the rest of the circuitry, as described below.

When everything checks OK, you can change the values of R7/R8: if the voltage drop is too great, increase the values (both at the same time) to 27 or 33Ω. If the voltage drop is too small, then decrease them to 18 or 15Ω.

3. Second stage current should be approximately 5.5mA. Check this by measuring the voltage drop across R18-R19 (or R21-22). It should be about 2.1V. Again, a 20% tolerance is acceptable. If it is outside these limits, go back to the first stage and correct the current there. Second stage current should not be too low, because you will not be able to adjust the output stage bias.

4. If you are using MPSA transistors in the output stage, connect your VOM/DVM across resistors R24-R26, and adjust the voltage drop to 1V with P2. This corresponds to the quiescent current of 15mA. Don't forget to put a snap-on heatsink on Q8 and Q9—they tend to run hot.

If you are using MOSFET output devices, change P1 to 500Ω, and short out D3, D4, R24 and R26. With MOSFETs you cannot directly measure the quiescent current. You can, however, check the total current consumption of the module by inserting a milliammeter in series with one of the supply lines. You should adjust the total current to approximately 40mA, which will give you approximately 20mA quiescent current in the output stage. If you don't have enough adjustment range, either increase the value of R20 (221 of 332Ω) or replace P1 with a 1kΩ trimpot.

5. Finally, (with the power turned OFF and the filter caps discharged through a 1-2kΩ resistor) insert Q5 in the socket, turn on the supply, and re-check the output offset. After about a minute, it should be less than 2mV.

RIAA-2

Before connecting the power supply, set P1 to mid-position and P2 to minimum (CCW). Again, if the servo amp Q9 is socketed, don't insert it as yet. Connect the ± 28V unregulated voltages to the module and carry out the following measurements/adjustments:

1. Check the supply voltage after the regulators (for example across C13/C14), which should be 24V. Check the zener voltages. They should be 15V. Check the DC offset at the output and, using P1, adjust for zero volts.

2. Check the current in the input stage by measuring the voltage drop across R4 and R5 (or R15 and R16). It should be close to 2.8V, which is equivalent to 5mA in each of the input devices. If this voltage drop is off by more than 10%, check the current sources very carefully.

3. The current in the second stage is controlled by the voltage drops mentioned above and, consequently, you need not make checks here.

4. When using the MPSA output transistors, connect your VOM/DVM across R29/R30 and adjust the voltage drop to 1V with P2. This corresponds to a quiescent current of 15mA. Again, don't forget the snap-on heatsinks for Q12 and Q15.

If you are using MOSFETs, change P2 to 500Ω and short circuit D7, D8, R29 and R30. The quiescent current cannot be checked directly, but you can check total current consumption. Insert a milliammeter in series with one of the supply lines and adjust the total current to 55–60mA. This will give you approximately 20mA in the output stage. Although not absolutely necessary, if you have a couple of small heatsinks handy, put them on the MOSFETs. This applies to all modules where you are using MOSFETs.

5. Now (with the power turned off and capacitors discharged through a 1-2kΩ resistor) you can insert Q9, turn on the power again, and re-check the offset at the output. It should be less than 2mV after a minute or so.

6. Since the RIAA compensation is distributed in the two RIAA modules, you can check its accuracy only when the two are connected together. I usually check it after installing them in the case. You will need a stable oscillator and an accurate inverse RIAA network for this measurement. (Old Colony KL-3C is one of the few inverse RIAA networks available.) Your only adjustment here is trimming R25. If the gain is lower at 10kHz than at the lower frequencies, parallel R25 with several hundred kΩ. If it is higher, connect a small resistor in series with R25. Using the ±1% components recommended (and supplied by Old Colony), the tolerance on RIAA accuracy is approximately 0.2dB, so the necessary adjustment is indeed very small.

Line Amplifier

This module is essentially the same as RIAA-2. Consequently the setup procedure is the same. When testing the line amp module separately, you can replace the balance control with a 250Ω resistor. Connect the ± 24V supply to the module and do the measurements/adjustments described in RIAA-2.

Tape Buffer

Set P1 to mid-position before switching on the module. Connect ±22V (probably from the line amp) to the supply terminals and do the following measurements/adjustments:

1. Check the current in the input stage by measuring the voltage drop across R4/R7. It should be approximately 5V, which is equivalent to 5mA. If it is more than 5.5V, change R5/R6 to a higher value. If it is lower than 4.5V, short out R5/R6. If it is still lower than 4.5V, replace P1 with a 20–25Ω trimpot.
2. Second stage current should be about 15mA. Check this by measuring the voltage drop across R9/R12, which should be around 4.3V. Second stage current is controlled by the voltage drop across R4/R7 in the first stage, and if it is significantly different from 15mA, you must go back to the first stage to correct it.
3. Adjust output offset to zero volts with P1.

NOTE:

R2 must be 221Ω as indicated in the parts list (22.1Ω on the schematic is wrong!).

This completes the setup of the four modules in the EB-585 preamplifier.

Please report any serious discrepancies or difficulties in the setup process to Old Colony Sound Lab, P.O. Box 243, Peterborough NH 03458. They will be shared with the author and, if necessary, corrective action will be taken. Necessary updates, if any, will be published in *Audio Amateur*.

Part II

A MOVING COIL PREAMP

BY ERNO BORBELY
Contributing Editor

LET'S LOOK AT the complete MC-preamp schematic (*Fig. 1*). I have connected four pairs of FETs in parallel, which become a good compromise as far as total drain current, cost and input noise are concerned. As I promised in the noise overview, you can do without electrolytics in the input stage by using only the lowest I_{DSS} group (GR: 5–10mA) of the Toshiba FETs. You can use four pairs of the single devices (2SK147GR/2SJ72GR) I described in the preamp article or, if you have access to them, use two pairs of dual FETs: 2SK146GR/2SJ73GR. The dual FETs have matched devices in one common case (*Photo 4*), but they only offer an advantage if the N-channel types are matched to the P-channel types. If not, the single devices will work just as well in this circuit. Whether you use singles or duals is really a question of what you can find.

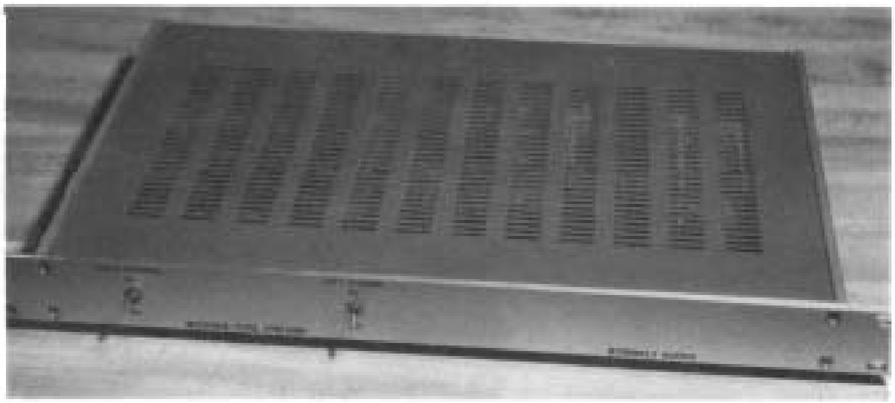


PHOTO 1: Borbely's pre-preamp is housed in a Swiss-made modular cabinet made by ELMA. The only controls are two gain selector toggle switches on the front panel.

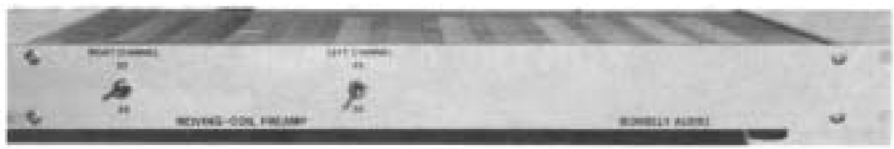


PHOTO 2: Front panel closeup.

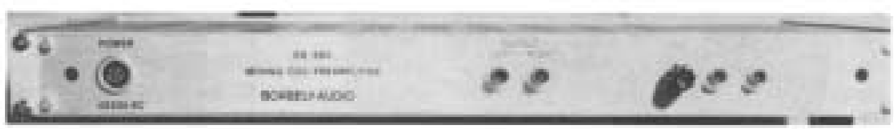


PHOTO 3: The rear panel contains the power connector (also by ELMA), isolated input and output jacks, and a separate grounding post.

The GR group is specified at 5–10mA, and to get very low noise, they should be operated at a minimum of 5mA. By connecting four in parallel without source resistors, you might get, in the worst case, 20-40mA total drain current. This is too much variation, and to handle the drain current, you must get closer to the low-end than the high-end. Ideally, you should select devices with I_{DSS} = 5–6mA and connect these in parallel without source resistors. You might, however, need to buy many devices to find four within this range.

Instead, I have chosen a source resistor of 6.8Ω, which slightly increases the input noise but lets you use practically all devices from the GR group. Although this should assure proper input stage operation, check the voltage drop across drain resistors R11 and R26. It should be 2–2.5V (20–25mA total drain current) but being 10% outside these limits doesn't seem to hurt the amp's operation.

To avoid excessively high current (and power dissipation) in the cascode transistors, connect two of the input devices' drains and feed separate cascode transistors (Q9 and Q10 on the N-channel side, and Q11 and Q12 on the P-channel side). For minimum noise, I have used extra filtering for the reference voltage of the cascodes. Actually, I should also have reduced the voltage across the input FETs to avoid excessively high power dissipation. High temperature will reduce the effective g_m and increase the FETs' gate leakage current, both of which tend to increase noise. The 15V seems to have an insignificant effect on the noise, yet helps keep the input devices in the linear range of their operation, so I opted to stay with the cascodes' original reference voltage.

The amplifier's second stage is the

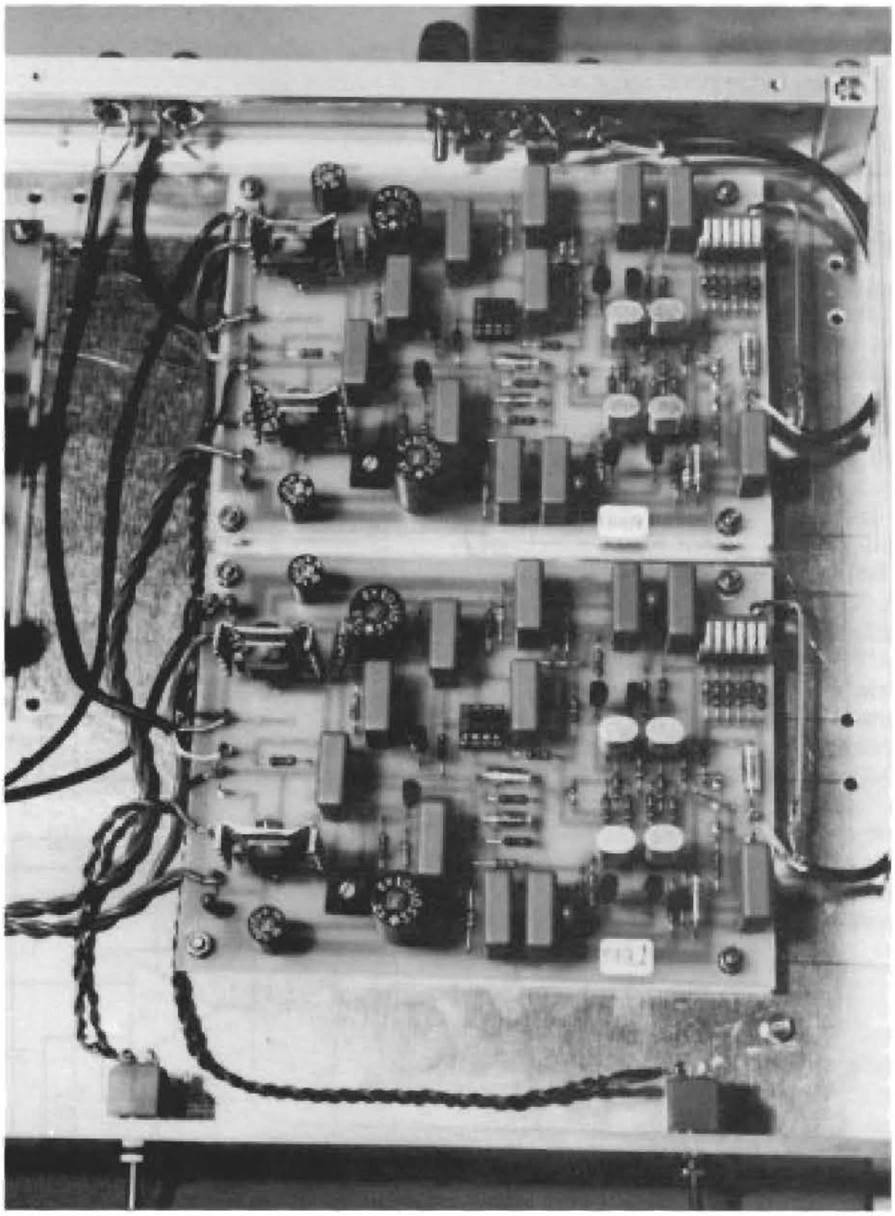


PHOTO 4: The moving coil preamp boards.

same as in the RIAA-1 amp (see the preamp article for a detailed description). The output stage uses the Hitachi MOSFETs (*Photo 4*), and these cannot be substituted by the bipolars used in the RIAA-1 amp. Adjust the bias with P1 to approximately 20mA. For long-term stability, P1 should be a Cermet trimpot. A heatsink on the TO-220 MOSFETs is a must.

For pickup load switching, I added a DIL switch with up to six poles at the amp's input. I usually use the resistors shown in *Fig. 1*, but you can change these values to suit your MCs. You can also leave all switches in the "off" position if you need a high impedance load for your pickup! In this case, the load will equal R8, which I have chosen to be 47.5kΩ but which you can select as you wish. If you change your MCs frequently and don't want to go inside the amp to change the load resistor, add a small switch (gold-plated ELMA 01 or equivalent) on the back panel and arrange the load resistors on this switch. You can also add a second set of input connectors, connected in parallel with the main inputs and in which you can plug the necessary load resistors.

I have chosen a 0.0022μF cap for the input. This might not be optimum for all MCs, so check the manufacturer's recommendations and change the value accordingly. The capacitor must be high-quality film, preferably polypropylene. The same goes for cap C, which connects the ground side of the input connector to the chassis. I am using a 0.01μF polypropylene (type WIMA FKP 2) in this position.

As I mentioned earlier, I compromised on the value of the feedback resistor (R31 = 2.21Ω). This will allow you to use a 43Ω series feedback resistor for a gain of 26dB, and an 86Ω for 32dB. These values solve the output stage loading problems, but the offset problem remains. There is no practical way to adjust the offset by putting a trimpot in the source circuit. (You will notice there is a place for such a trimpot on the layout, and I use it in the higher impedance RIAA-1 version 2 [V.2] amplifier. I don't think you can find a 3Ω/10 turn Cermet trimpot needed for this job.) So, you must rely on the servo circuit. Using the LF411 here, with a minimum recommended load of 2kΩ, you can easily calculate the amount of offset the servo can compensate:

$$V_o \max = \frac{V_{out}^{\ max}}{2k\Omega} \times 2.21\Omega = 13mV \qquad (9)$$

where $V_{OUT}^{\ max}$ is the maximum output swing of the LF411, here assumed to be 12V. This doesn't sound like much, but I never had problems tracking the offset when the input FETs were from the same group. I am afraid if the offset is higher, you must find either an op amp with higher current capability, or get FETs with better matching.

The equivalent noise circuit for the MC-preamp's input stage is shown in *Fig. 2*. The FETs are represented by their equivalent noise resistance, which, you might recall from our noise theory, is 34Ω. The equivalent noise resistance of the entire input stage is calculated to be 9.5Ω, that is, the input stage is generating the same amount of noise with shorted input as a 9.5Ω resistor.

The equivalent input noise versus source resistance (R_s), as measured on several prototypes, is shown in *Fig. 3*. Notice the curve is very close to the theoretical one, going asymptotically to e_n = 56mV, which is equivalent to the noise of a 10Ω resistor.

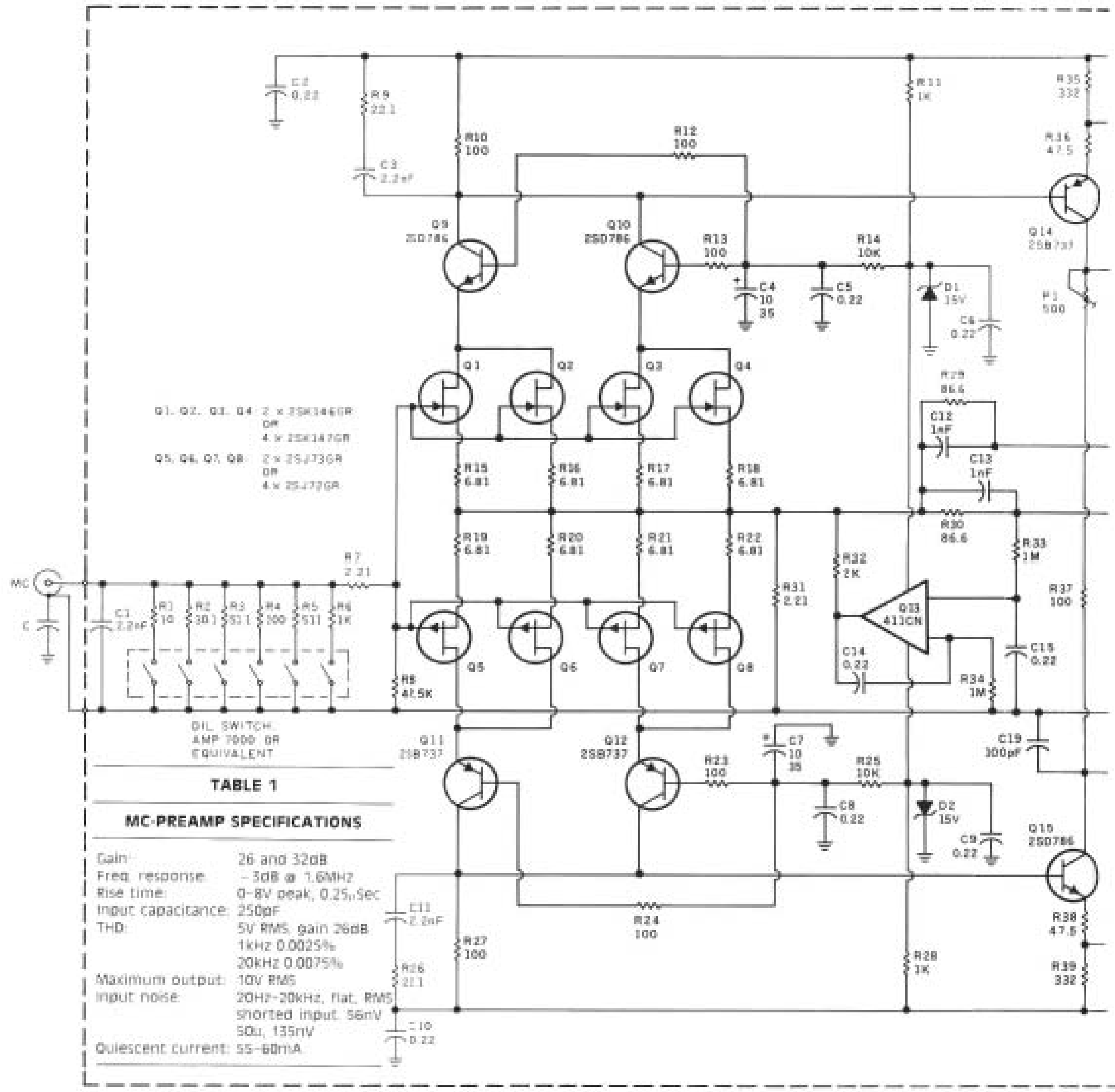
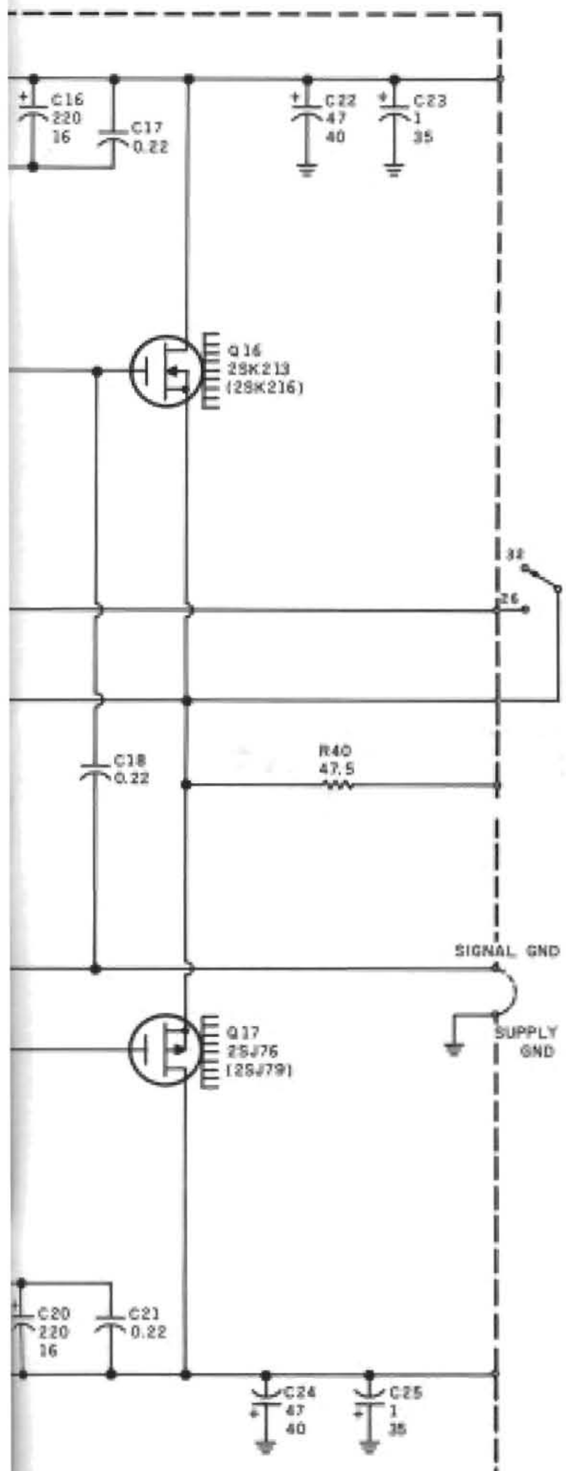


TABLE 1

MC-PREAMP SPECIFICATIONS

Gain:	26 and 32dB
Freq. response:	−3dB @ 1.6MHz
Rise time:	0–8V peak, 0.25μSec
Input capacitance:	250pF
THD:	5V RMS, gain 26dB 1kHz 0.0025% 20kHz 0.0075%
Maximum output:	10V RMS
Input noise:	20Hz–20kHz, flat, RMS shorted input: 56nV 50Ω, 135nV
Quiescent current:	55–60mA

FIGURE 1: MC-preamplifier schematic



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Although not shown, the noise at 47.5kΩ is what is expected from a resistor of that value and with no sign of shot noise. I'm not claiming this is the lowest noise MC-preamp on the market, but it does offer excellent noise performance for pickups with sensitivities down to about 0.2mV.

Figure 4 shows the MC-preamp layout. In accordance with the wishes of the editor, I have changed the resistor spacing to 0.5". This will allow you to use Resista's MK-3-type metalfilm resistors which Old Colony plans to standardize for future designs. Note the "clean" signal ground on the board. You can connect this separately to the regulator board, but you must connect a 0.01μF/160V polypropylene capacitor directly on the board between the signal ground and the power supply ground. This will avoid problems at high frequencies. As an alternative, you can connect the two grounds on the board, as shown on the stuffing guide (*Fig. 5*). When stuffing the board, be careful when you insert the input FETs. The stuffing guide shows the layout for the dual FETs. When using the single FETs, carefully consult the pinout in *Fig. 14*.

Power Supply

As mentioned earlier, one of the important factors in low-noise amplification is interference from power supplies. When you deal with extremely low levels of an MC-pickup, there should be practically no interference at all. Consequently, only the highest quality power supply is suitable for MC-preamps.

When I set out to design my power supply, I looked through my old supply designs. Instead of trying to update these, however, I decided to use one of the excellent Sulzer designs in *TAA*. The final circuit is shown in *Fig. 6*, a modified version of the Sulzer circuit published by J. Breakall *et al* in *TAA* 1/83.[2]

It consists of a preregulator using the LM317/337 three-terminal IC regulators (*Photo 5*), and the op amp regulator using the NE5534. Unregulated input to the preregulators should be about 3V higher than their output, and the input to the op amp regulator should be about 3V higher than the final output. To build in some safety, I adjusted the preregulators for an output of approximately 28V for a final output of 24V. The rectified DC voltage of 32–33V comes from a transformer with 2x24V RMS secondary. Just as with the preamplifier, I used independent power supplies for the two channels.

Here are the formulas to calculate the necessary resistor values should

TABLE 2

PARTS LIST

Resistors*	
R1	10Ω
R2	30.1Ω
R3	51.1Ω
R4, 10, 12, 13, 23, 24, 27, 37	100Ω
R5	511Ω
R6, 11, 28	1k
R7, 31	2.21Ω
R8	47.5k
R9, 26	22.1Ω
R14, 25	10k
R15-22	6.81Ω
R29, 30	86.6Ω
R32	2k
R33, 34	1MΩ
R35, 39	332Ω
R36, 38, 40	47.5Ω
*all ¼W ±1% metalfilm, Resista MK-2/equiv.	
Trimpotentiometer	
P1	500Ω Cermet, Dale 101T/equiv.
Capacitors	
C	0.01μF/63V/20% PP WIMA FKP 2 /equiv
C1	0.0022μF/160V PP WIMA FKP 2 /equiv
C2, 5, 6, 8, 9, 10, 14, 15, 17, 18, 21	0.22μF/160V/20% PP WIMA MKP-10/equiv.
C3, 11	2200pF/160V/2.5% PP Siemens B33063/equiv.
C4, 7	10μF/35V TA Roederstein ETPW/equiv.
C12, 13	1nF/160V PP Siemens B33063/equiv.
C16, 20	220μF/16V EL Roederstein EK/equiv.
C19	100pF/630V/2.5% PS Siemens B31063/equiv.
C22, 24	47μF/40V EL Roederstein, EK/equiv.
C23, 25	1μF/35V TA Roederstein ETPW/equiv.
Semiconductors	
Q1-4	2SK146GR 2x or 2SK147GR 4x Toshiba
Q5-8	2SJ73GR 2x or 2SJ72GR 4x Toshiba
Q9, 10, 15	2SD786 ROHM
Q11, 12, 14	2SB737 ROHM
Q13	LF411CN National
Q16	2SK213 or 2SK216 Hitachi
Q17	2SJ76 or 2SJ79 Hitachi
D1, D2	15V zener, 0.5W

you choose voltages different from the ones shown in *Fig. 6*:

$$R1(8) = \frac{V_{out}1 - 1.45}{1.25} \times 121\Omega \qquad (10)$$

For a $V_{out}1$ = 28V R1, and R8 is 2.55kΩ:

$$R4(11) = \frac{V_{out}2 - 5}{5} \text{k}\Omega \qquad (11)$$

Formula 11 is valid when R3 and R10 are 1kΩ and the reference voltage is 5V. For a $V_{out}2$ = 24V R4, and R11 is 3.8kΩ.

The op amp regulators' reference voltage is supplied by the LM336Z-5.0 reference diodes. These "diodes" have a tolerance of ±4%, so the absolute value of the output voltage can be 4% off the nominal value: 24V ± 1V. If you are concerned with the absolute value and/or with the plus and minus sides being equal, you'll have to trim resistors R3 and R10.

Layout for the regulator is shown in *Fig. 7*. You will need two of these boards for an MC-preamp. The stuffing guide is shown in *Fig. 8*. Note I have used a common heatsink (*Photo 6*) for the preregulator and the series pass transistor on each side (*Fig. 9*). Make sure you mount these with mica and nylon screws, and use a generous amount of silicone grease on both sides of the mica insulator.

No fuses are shown in connection with the power supply. I recommend, however, putting a fuse in the plus and minus leads, between the power supply and the MC-preamp. The quiescent current consumption of the MC-preamp is 55–60mA, but will rise to approximately 100mA with full drive. You should, therefore, fuse the circuit with 200mA medium-blow fuses.

Naturally, you can use this supply for other projects, such as instruments and crossovers, by scaling the voltages for your particular application. Before you install the power supply, test it with a load resistor, a voltmeter, and preferably a scope. The load resistor should reflect your application's final current demand. I also measured the power supply noise, using a low-noise preamp and a 20Hz–20kHz noise filter. The RMS noise output was 1–2μV. This should satisfy just about all audio application requirements.

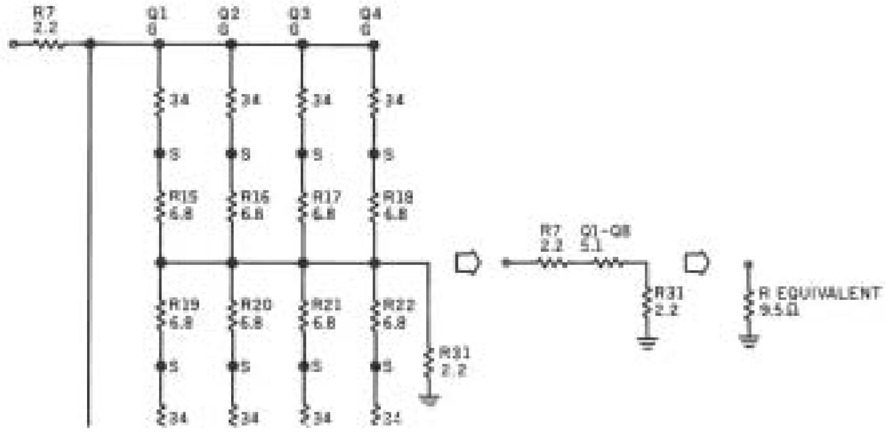

FIGURE 2: Equivalent noise resistance is 9.5Ω.

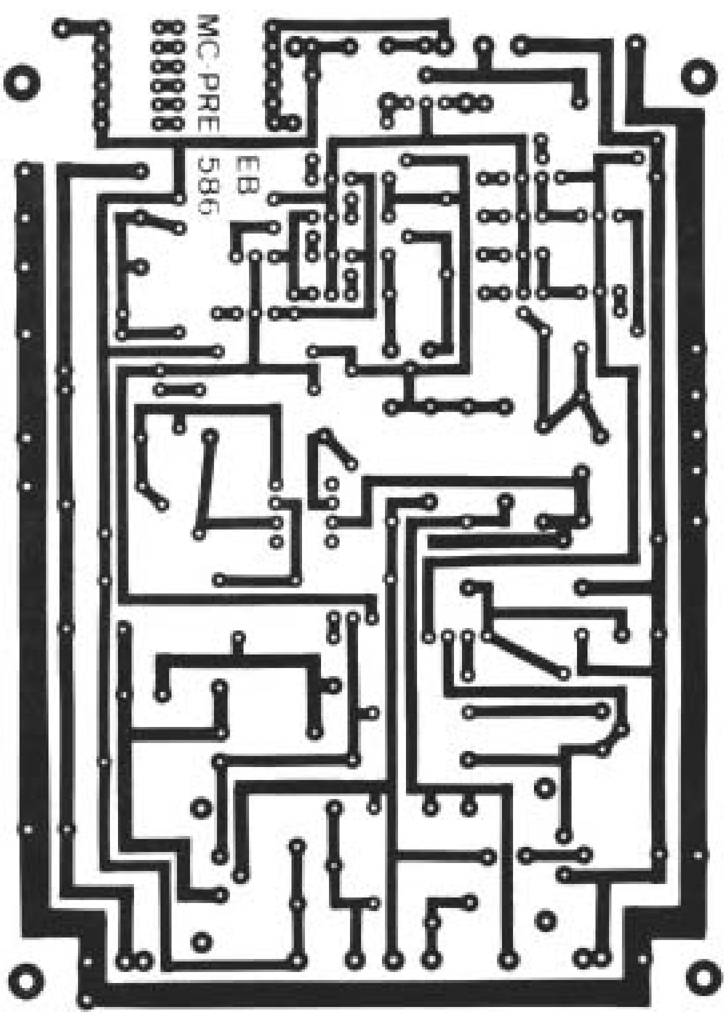

FIGURE 4: MC-preamp layout.

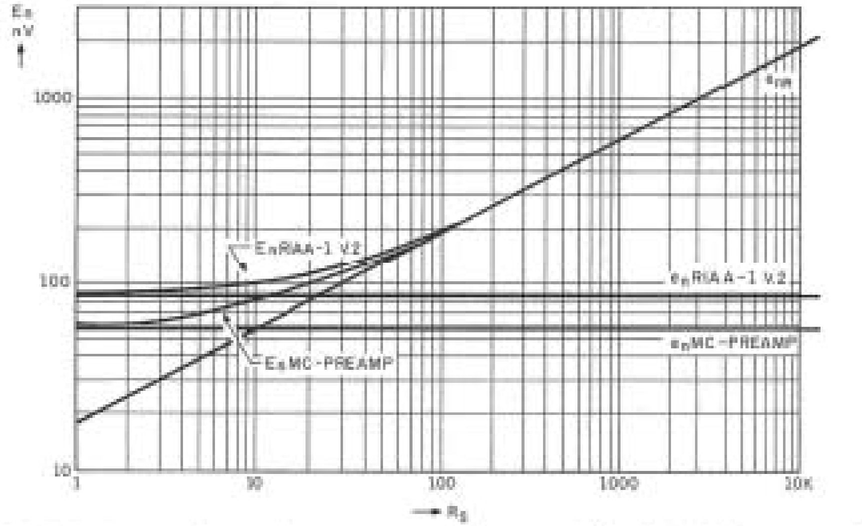

FIGURE 3: Measured equivalent input noise E_n versus R_s for the MC-preamp and for RIAA-1 V.2 (20Hz–20kHz, flat, RMS).

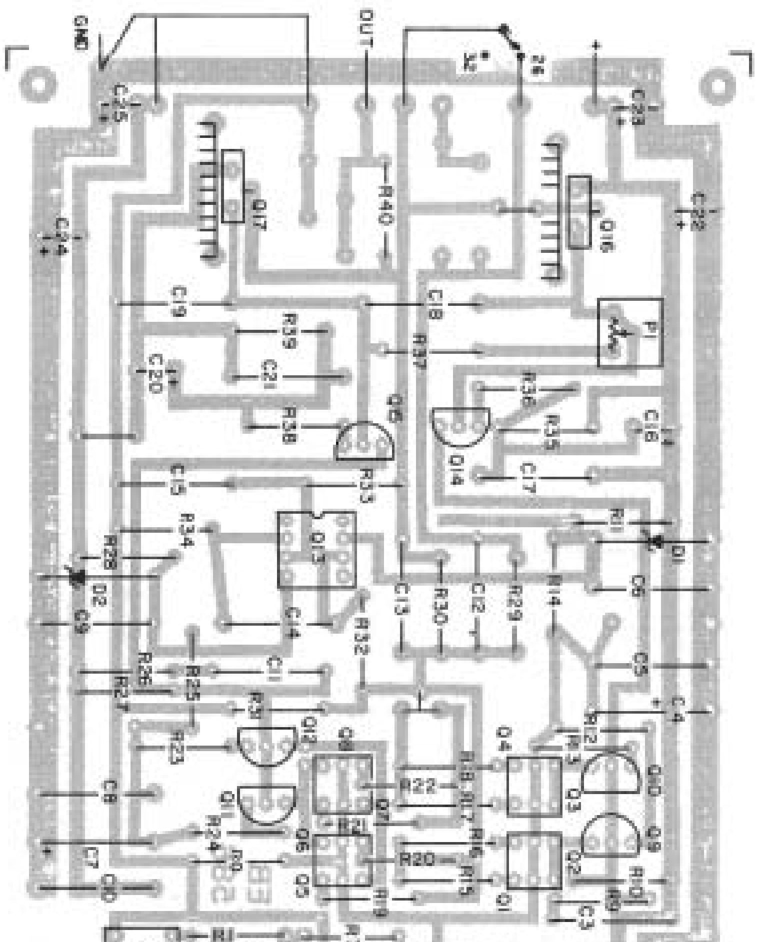

FIGURE 5: Stuffing guide for the MC-preamp.

The wiring diagram for the MC-preamp is shown in *Fig. 10*. I have been using a one-unit high, 260mm deep, 19" box for the pre-pre, and have considerable space to spare. I put the gain switches on the front, and gold-plated phono connectors for input and output on the back panel. The transformer, like the preamp, is in a separate box, interconnected with a 6-pin LEMO connector (*Photo 7*) to the pre-pre.

Medium Output MCs

Although the previously-described MC-preamp can take care of just about all MCs, you do not need it when you have a medium-to-high output pickup and an EB-585 preamp. A medium output has been defined as, you will recall, 0.5mV @ 5cm/sec RMS lateral velocity. With these pickups, you need approximately 20dB extra gain in front of the RIAA stage or, as I propose here, built into the RIAA stage.

In theory, if you were only interested in MCs, you could use the MC-preamp in *Fig. 11* (*TAA* 4/86) as an RIAA-1 stage. If you are like me and prefer to have both MM and MC input, you can't use the MC-preamp without modifications. Four pairs of FETs have an input capacitance of approximately 250pF, and this might be too high for some MM-pickups. Also, the MC-preamp can't drive the feedback network in pure class A beyond approximately 1V RMS, which is not enough for the RIAA-1 stage used for MM pickups. For MMs, the single pair is the best solution, as used in the RIAA-1 stage of the preamplifier. For MCs, the four-pair input circuit is the best, as shown in the MC-preamp. For a circuit that must work with both, a two-pair input is the best compromise.

Figure 11 shows the complete schematic of the high-gain version of the RIAA-1 amplifier. Notice the input stage uses two pairs of single FETs (2SK147/2SJ72) or one pair of dual FETs (2SK146/2SJ73). Since the MC-input is for medium-to-high output pickups, it is not necessary to use very low values for the feedback and source resistors. The relaxation of the first makes it easier to drive the feedback network and to track the input offset. The second gives you a wider choice of input devices. Although I recommend you stick to the BL group for best low-level linearity, you can mix BLs with GRs in your circuit. I have also mixed two BLs on the N-channel side with one GR and one V on the P-channel side, but this is a worst case situation and should be avoided.

The feedback network, which uses a 6.8Ω shunt resistor, allows you to select 20, 30 or 40dB gain. The lowest gain position is for MM-pickups and for high output MCs, and the highest gain position is for medium output MCs. With such a closed-loop gain (40dB), you might be wondering about the performance of the RIAA-1 V.2. Naturally, the THD of the RIAA-1 V.2 deteriorates somewhat at the highest gain setting because of reduced feedback. It's still only 0.0035% at 5V RMS and 1kHz, which I consider negligible. More importantly, I confirmed with extensive listening tests that there was no sound quality deterioration.

The RIAA-1 V.2's noise performance is indicated in *Fig. 3*. The equivalent noise resistance of the input stage equals 22Ω, with a theoretical value of $e_n = 84nV$. Actual measurements, as shown in *Fig. 3*, come very close to confirming this value.

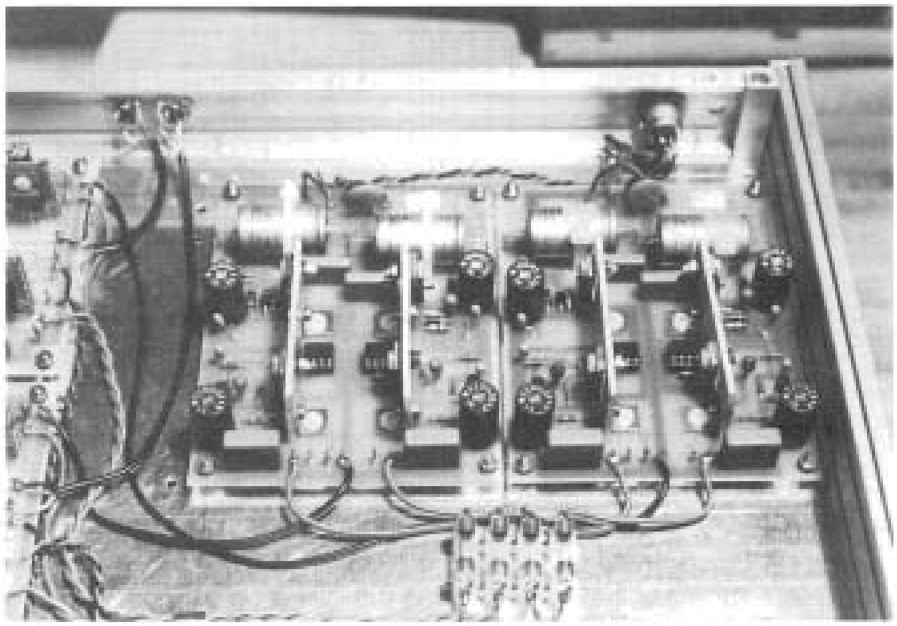

PHOTO 5: Separate power supplies for each channel are provided with full regulation and individual fusing for each of the four supplies.

The input can be switched between MM and MC with an on-board switch (gold-plated, 0.1" spacing), or with a switch you can put on the front panel and wire to the board with a three-wire shielded cable (connect the shield to the input ground on the board). The MM input has 47.5kΩ load, installed permanently. The MC input, like the MC-preamp, can be

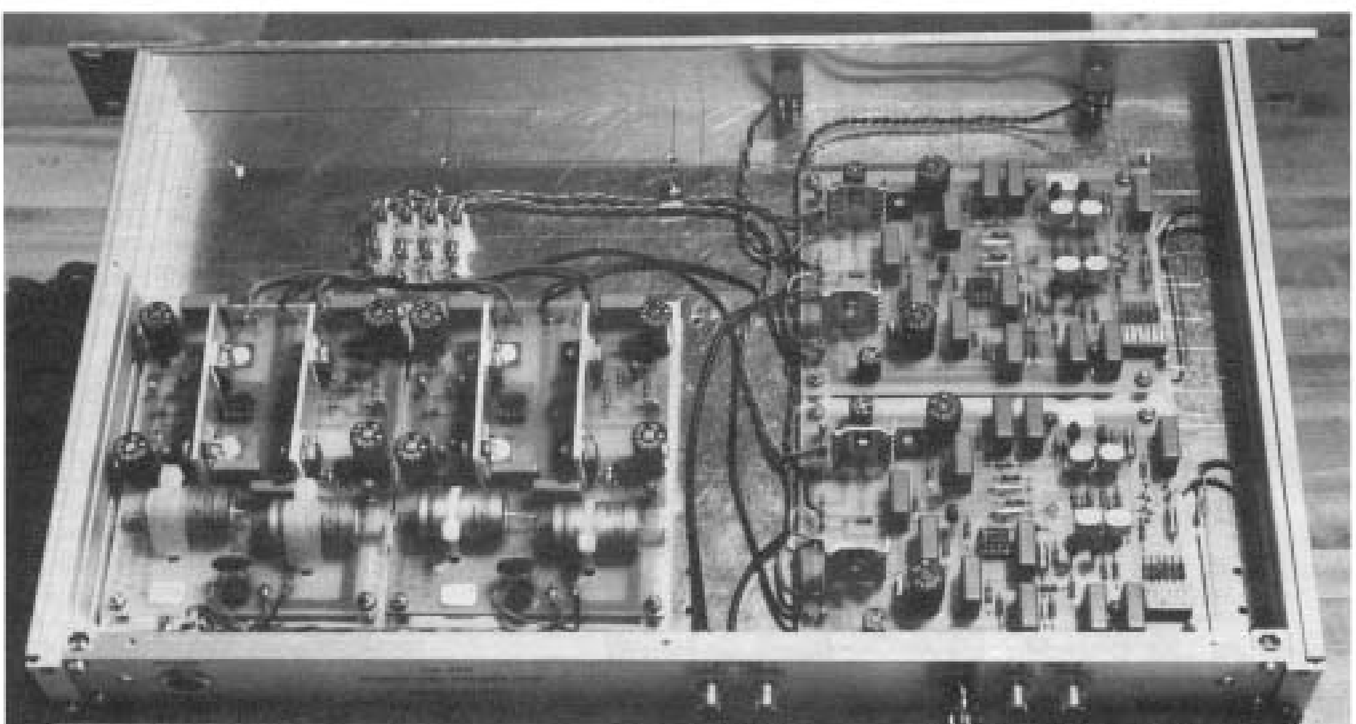

PHOTO 6: Full view of the chassis and the four boards. The unit is standard rack width (19"), 1¾" high by 10½" deep.

PHOTO 7: The raw DC power supply is housed in a separate enclosure with a six-pin locking connector, a modular AC power connector with its on/off switch on the front end.

loaded with up to six different resistors. Note that capacitor C1 is equal to C_{load}-C_{in}, where C_{load} is the recommended load capacitor for your MM-pickup and C_{in} is the input capacitance of the amplifier (C_{in} = 120pF). C2 is again 0.0022μF, but you should adjust it according to the manufacturer's recommendations. Both capacitors must be high-quality film types.

The 75μsec high-frequency rolloff is provided by R36 and C26. I reduced this network's impedance to reduce the source impedance as seen by the RIAA-2 amplifier, which in turn reduces the noise in that stage. As you did with the original preamp, I suggest you check the RIAA accuracy with an inverse RIAA network when you install this high-gain ver-

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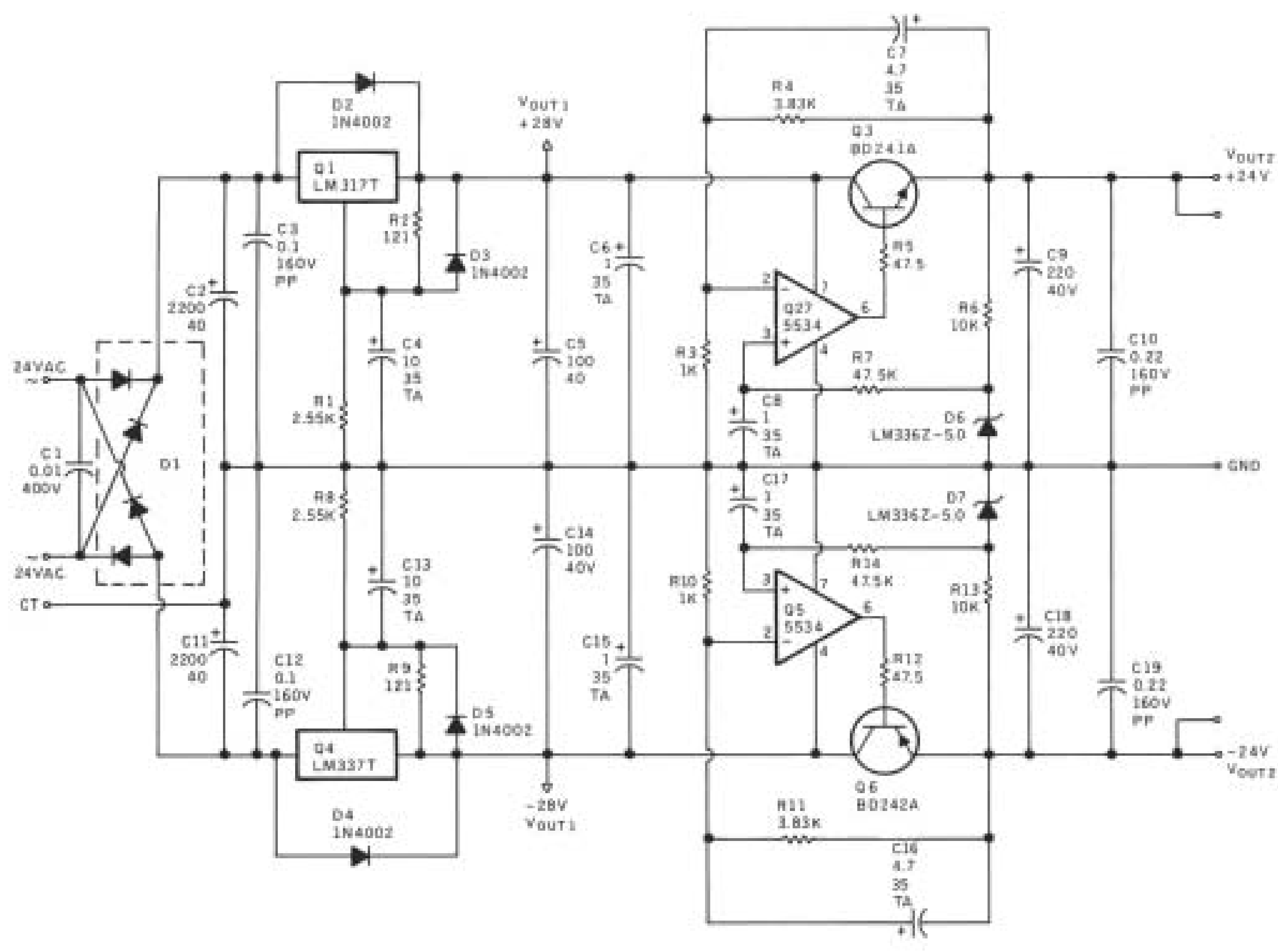


FIGURE 6: The MC-preamp's high-quality power supply.

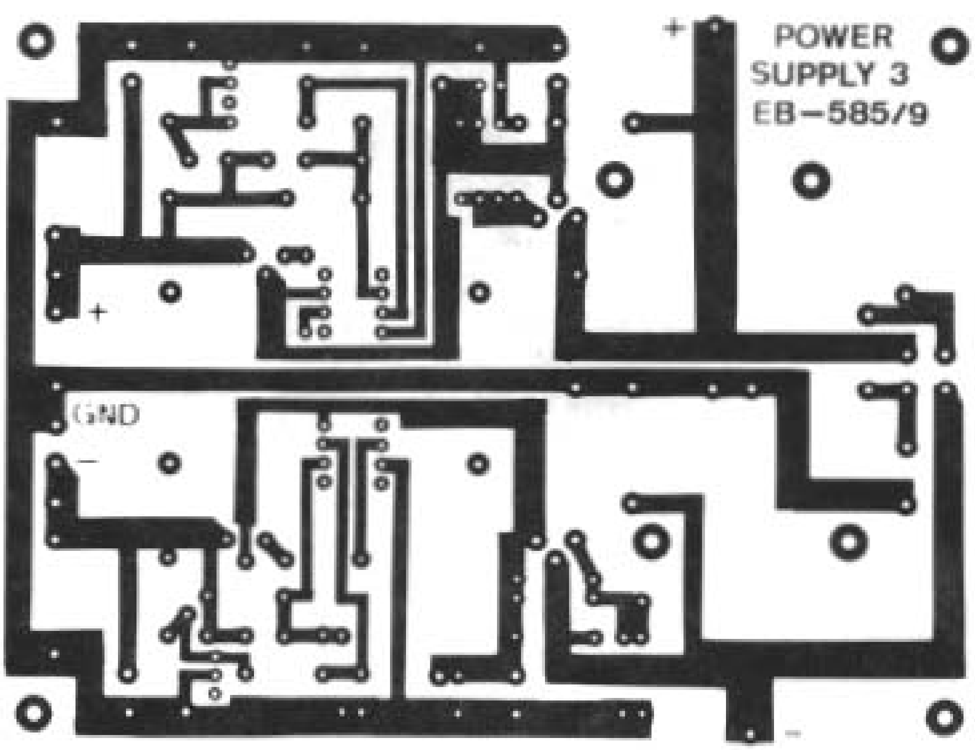


FIGURE 7: Copper side layout for the preamp's power supply.

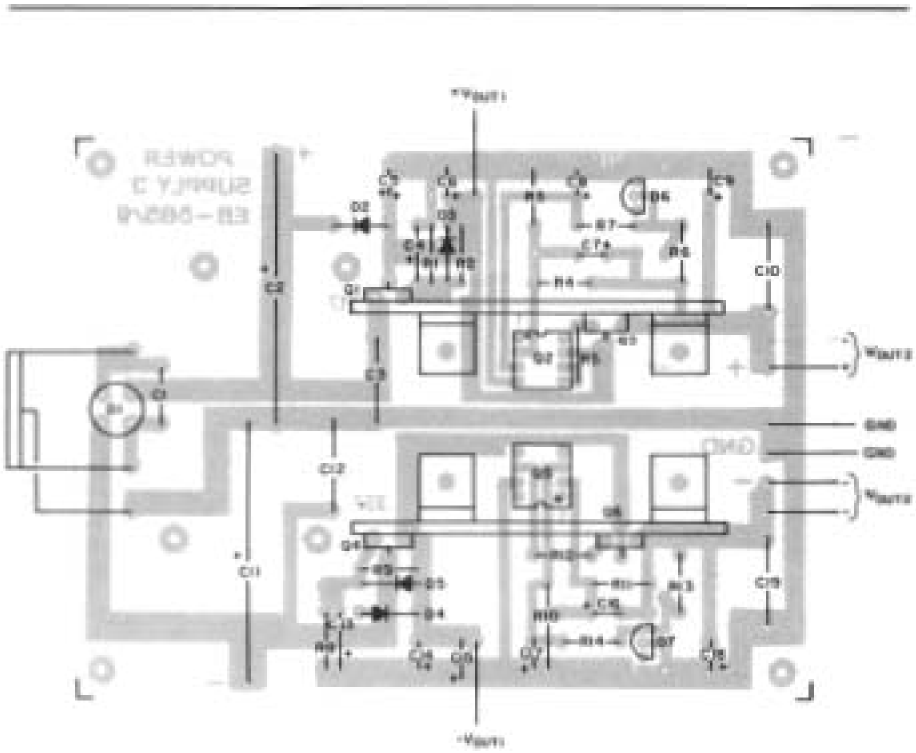


FIGURE 8: Power supply stuffing guide.

sion of the RIAA-1 amplifier. If necessary, trim the accuracy with R36.

Layout for the RIAA-1 V.2 is the same as the MC-preamp (*Fig. 4*). The stuffing guide is shown in *Fig. 12*. Again, the stuffing guide shows dual FETs at the input. If you use single FETs, carefully study the pinout diagram in *Fig. 14*.

Unfortunately, using the MC-preamp layout for the RIAA-1 V.2 has a drawback: its size. Because it is more than an inch longer than the original RIAA-1, you might have problems replacing it with this larger board in an existing box. Two possible solutions immediately come to mind when you use the box and mechanical layout indicated in my preamp article. (If you don't need the tape buffers, the problem doesn't exist; the RIAA-1 V.2 and the RIAA-2 boards will fit nicely in the available space.)

If you need the tape buffers and want to use the RIAA-1 V.2 boards, consider moving the power supply boards out of the preamp box and into the separate transformer box. This will free up an entire column of space

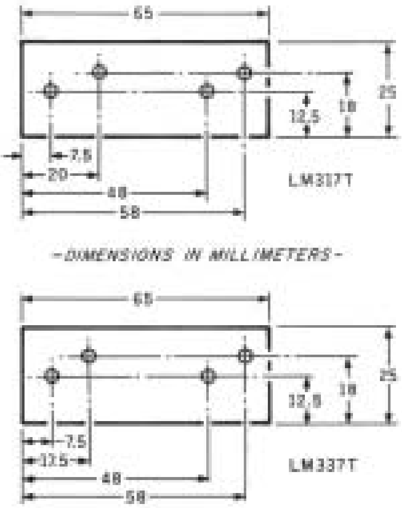


FIGURE 9: Heatsinks for preregulator and series pass transistor.

in this box size and allow you to put in many extra features, including the tape buffers. As an alternative, consider making a special layout for the RIAA-1 V.2 that fits into the place of the original RIAA-1 board. You won't have a problem if you have a dual-

Continued on page 41

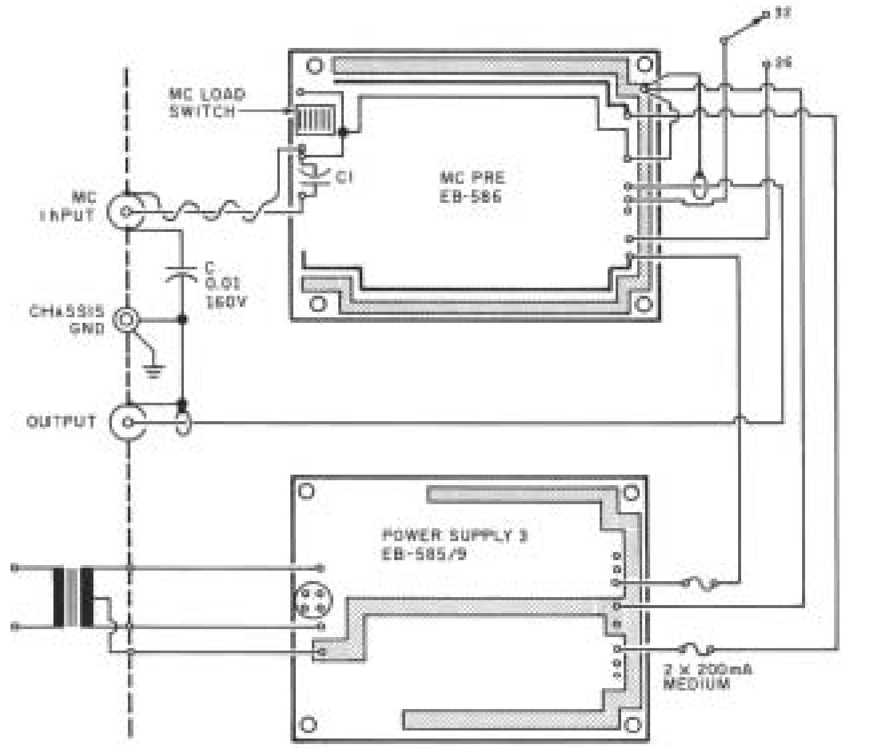


FIGURE 10: The preamp's wiring diagram.

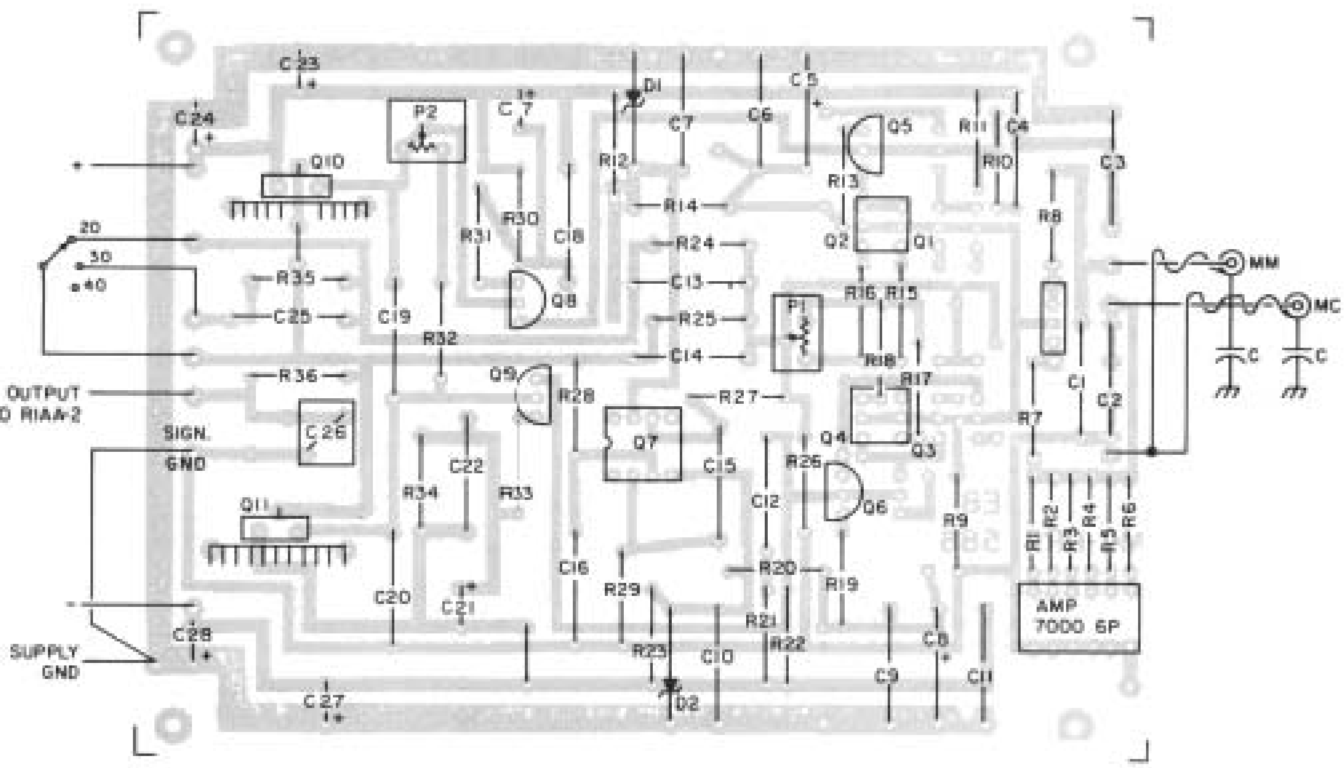


FIGURE 12: Stuffing guide for the RIAA-1 V.2.

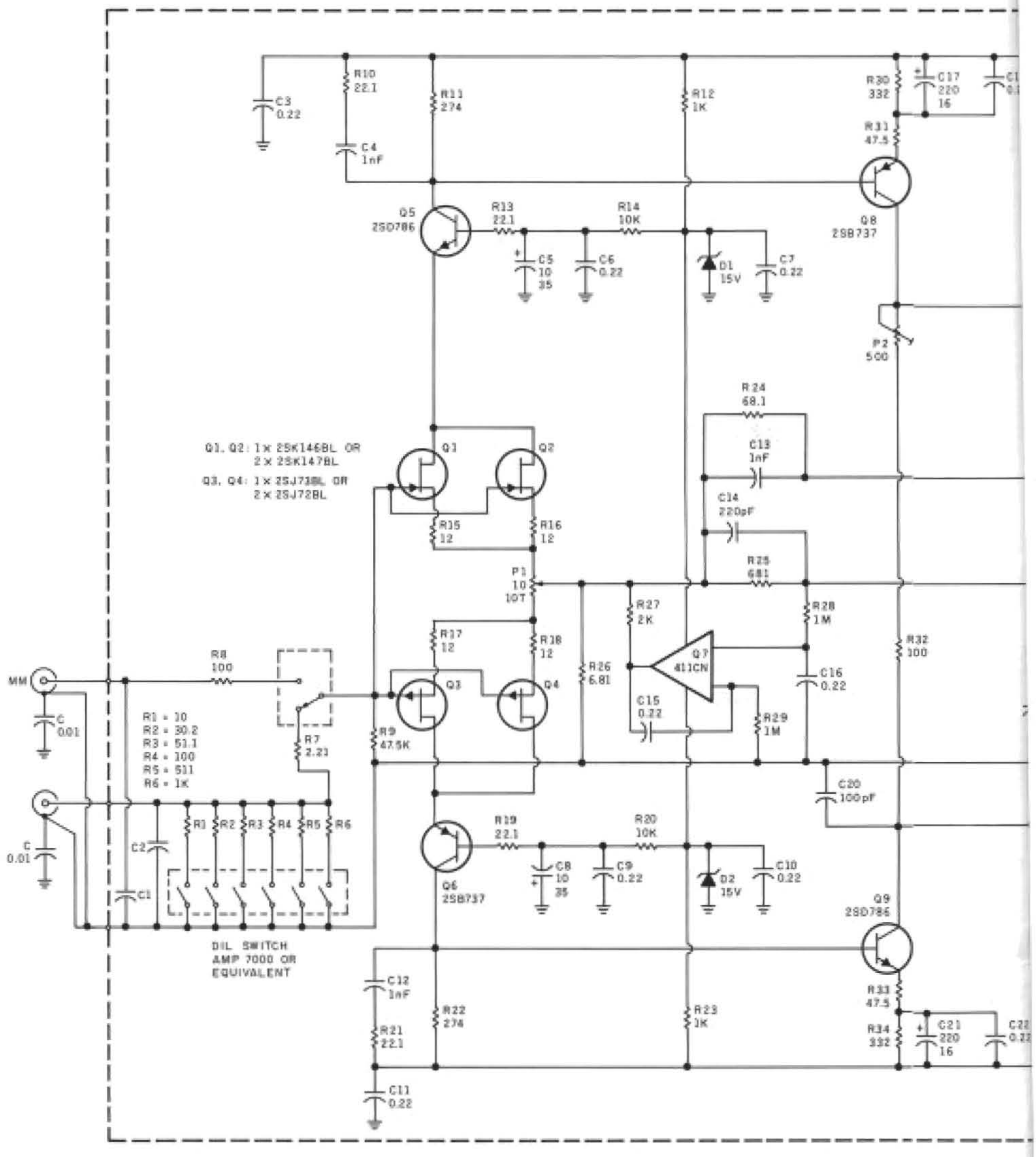


FIGURE 11: Schematic for the RIAA-1 V. 2 (high-gain version of the EB-585 preamp).

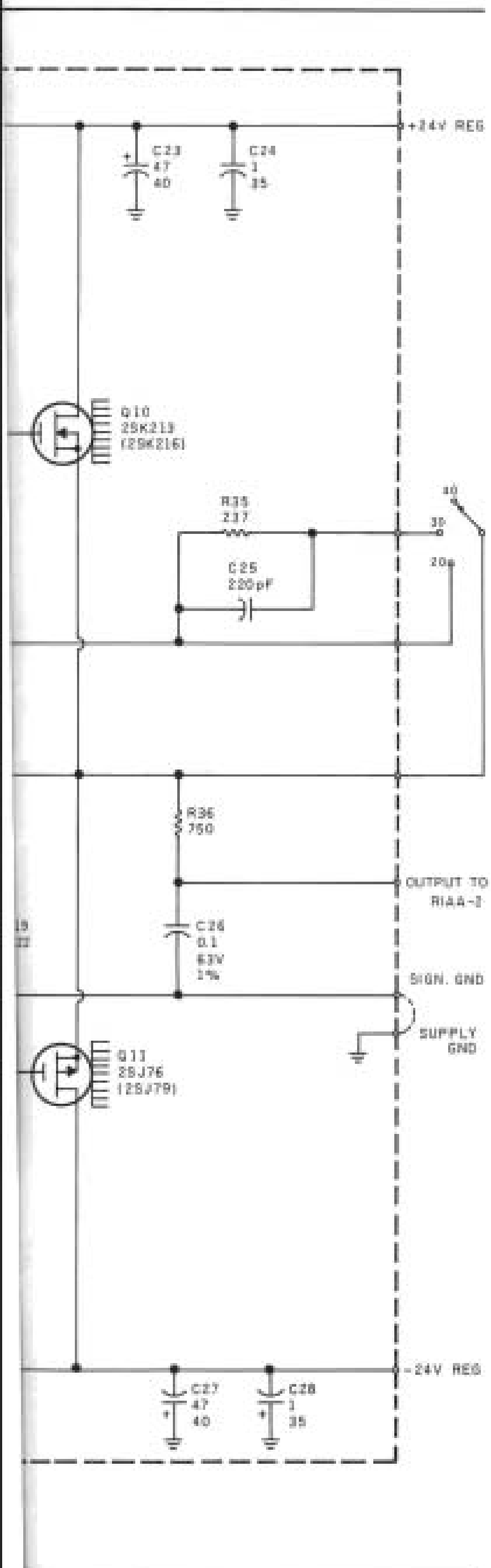


TABLE 4

RIAA-1 V.2 COMPONENT LIST

Resistors*	
R1	10Ω
R2	30.1Ω
R3	51.1Ω
R4, 8, 32	100Ω
R5	511Ω
R6, 12, 23	1k
R7	2.21Ω
R9	47.5k
R10, 13, 19, 21	22.1Ω
R11, 22	274Ω
R14, 20	10k
R15-18	12Ω
R24	68.1Ω
R25	681Ω
R26	6.81Ω
R27	2k
R28, 29	1M
R30, 34	332Ω
R31, 33	47.5Ω
R35	237Ω
R36	750Ω

*¼W 1% metalfilm, Resista MK-2/equiv.

Trimpotentiometer	
P1	10Ω Cermet, 10 turns
P2	500Ω Cermet, Dale 101T/equiv.

Capacitors	
C	0.01μF/63V/20% PP WIMA FKP 2/equiv.
C1	See text
C2	0.0022μF/63V/20% PP WIMA FKP 2/equiv.
C3, 6, 7, 9, 10, 11, 15, 16, 18, 19, 22	0.22μF/160V/20% PP WIMA MKP-10/equiv.
C4, 12, 13	1000pF/160V/2.5% PP Siemens B33063/equiv.
C5, 8	10μF/35V TA Roederstein ETPW/equiv.
C14, 25	220pF/630V/2.5% PS Siemens B31063/equiv.
C17, 21	220μF/16V EL Roederstein EK/equiv.
C20	100pF/630V/2.5% PS Siemens B31063/equiv.
C23, 27	47μF/40V EL Roederstein EK/equiv.
C24, 28	1μF/35V TA Roederstein ETPW/equiv.
C26	0.1μF/63V/1% PP RIFA PHE425/equiv.

Semiconductors	
Q1, 2	2SK146BL 1x or 2SK147BL 2x Toshiba
Q3, 4	2SJ73BL 1x or 2SJ72BL 2x Toshiba
Q5, 9	2SD786 ROHM
Q6, 8	2SB737 ROHM
Q7	LF411CN National
Q10	2SK213 or 2SK216 Hitachi
Q11	2SJ76 or 2SJ79 Hitachi
D1, 2	15V zener, 0.5W

TABLE 3

POWER SUPPLY COMPONENT LIST

Resistors	
R1, 8	2.55k
R2, 9	121Ω
R3, 10	1k
R4, 11	3.83k
R5, 12	47.5Ω
R6, 13	10k
R7, 14	47.5k
Capacitors	
C1	0.01μF/400V Ceramic
C2, 11	2200μF/40V EL Roederstein EG/equiv.
C3, 12	0.1μF/160V PP WIMA MKP-10/equiv.
C4, 13	10μF/35V TA Roederstein ETPW/equiv.
C5, 14	100μF/40V EL Roederstein EK/equiv.
C6, 8, 15, 17	1μF/35V TA Roederstein ETPW/equiv.
C7, 16	4.7μF/35V TA Roederstein ETPW/equiv.
C9, 18	220μF/40V EL Roederstein EK/equiv.
C10, 19	0.22μF/160V PP WIMA MKP-10/equiv.
Semiconductors	
Q1	LM317T National
Q2, 5	NE5534 Signetics
Q3	BD241A Motorola, SGS
Q4	LM337T National
Q6	BD242A Motorola, SGS
D1	1A/250V Bridge
D2-5	1N4002
D6, 7	LM336Z-5.0 National

Continued from page 39

height box because you can stack two boards on top of each other.

Refer to *Fig. 13* before wiring the RIAA-1 V.2 board into the EB-585 preamp. I included some changes in this wiring diagram that also apply to the original preamp, including wiring of the RIAA-2 stage output, the 1kΩ resistor in series with the volume control and the twisted pair of wiring to the line amp input.

Conclusion

I apologize if you have already built the regular RIAA-1 input stage when you really needed the high-gain version. At the time I submitted the preamp article, however, I hadn't decided to offer this high-gain version.

I believe the approach presented here is the best compromise for sound quality, which is, after all, what this hobby is all about. Build-

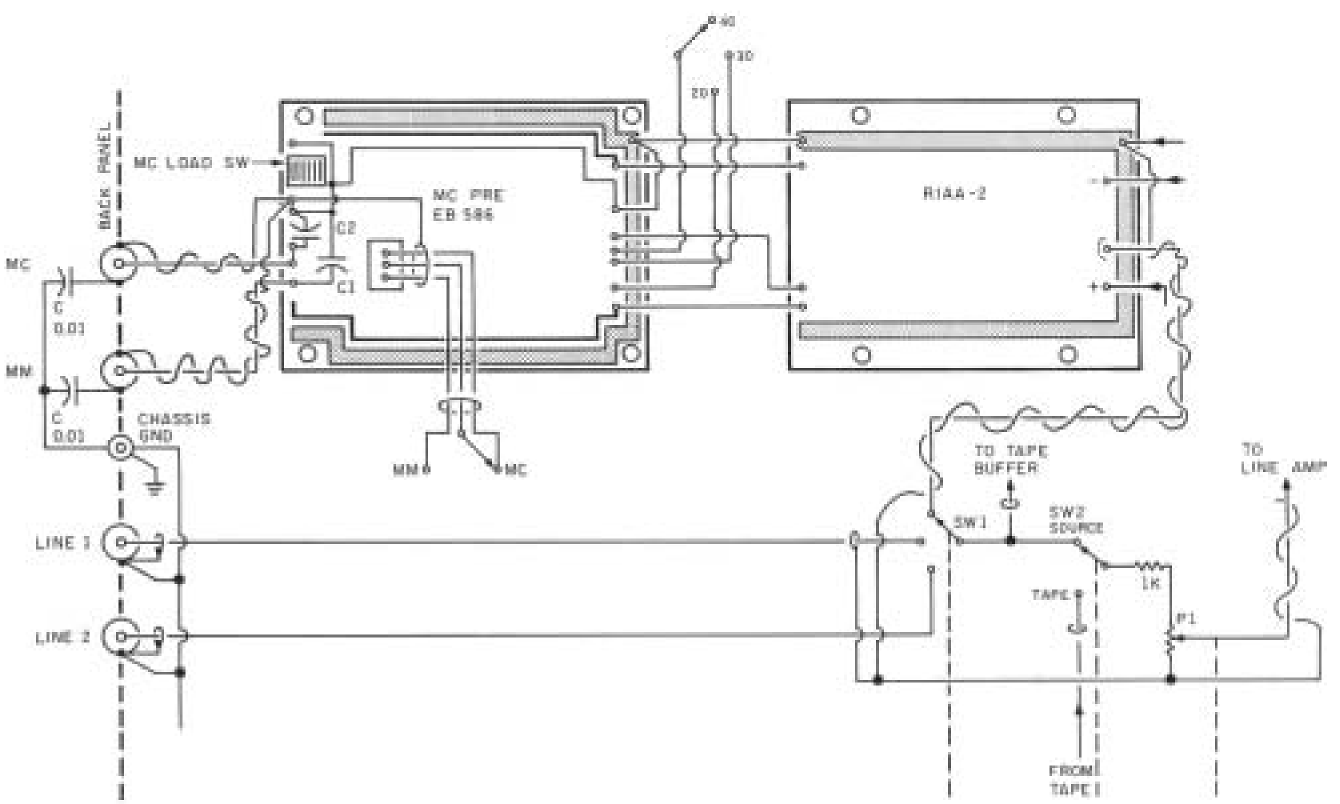


FIGURE 13: Wiring diagram for the high-gain version of the EB-585.

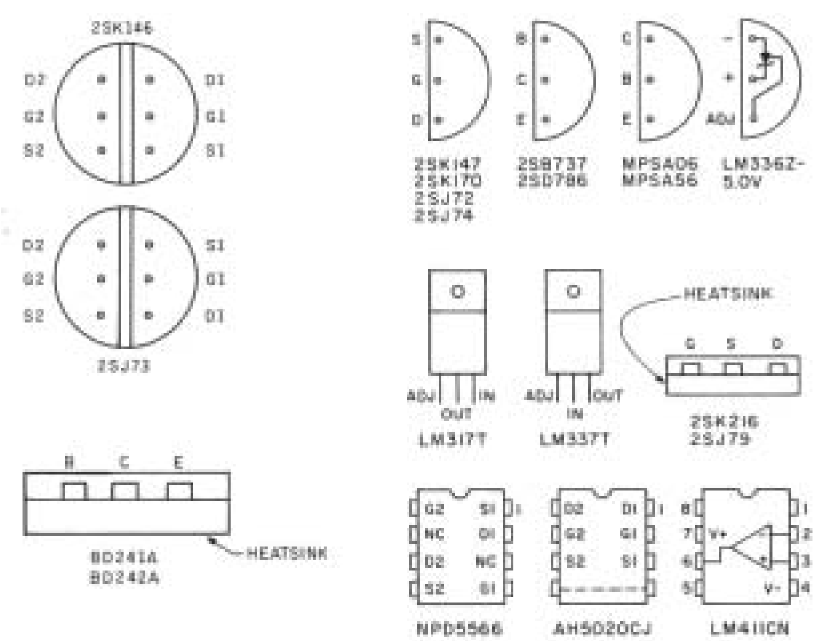


FIGURE 14: Pinout devices used in the MC-preamp, the power supply, and RIAA-1 V.2. Except for the 317/337 regulators, all devices are shown from the bottom view.

Continued from page 41

ing the preamp with or without the MC-preamp, as dictated by your particular pickup(s), will, I believe, give you some of the best sound quality available today. Good luck with your project. □

Acknowledgment

Dr. Kalman Molnar, consultant, reviewed the chapter on basic noise theory and suggested many excellent improvements. I greatly appreciate his contribution.

References

1. TOM, "The Koetsu Rosewood Signature Review," *The Absolute Sound*, Vol. 10, Issue 40, Winter 1986.
2. Breakall, J. *et al*, "Measuring Power Supply Output Impedance," *TAA* 1/83.

MC-preamplifier schematic

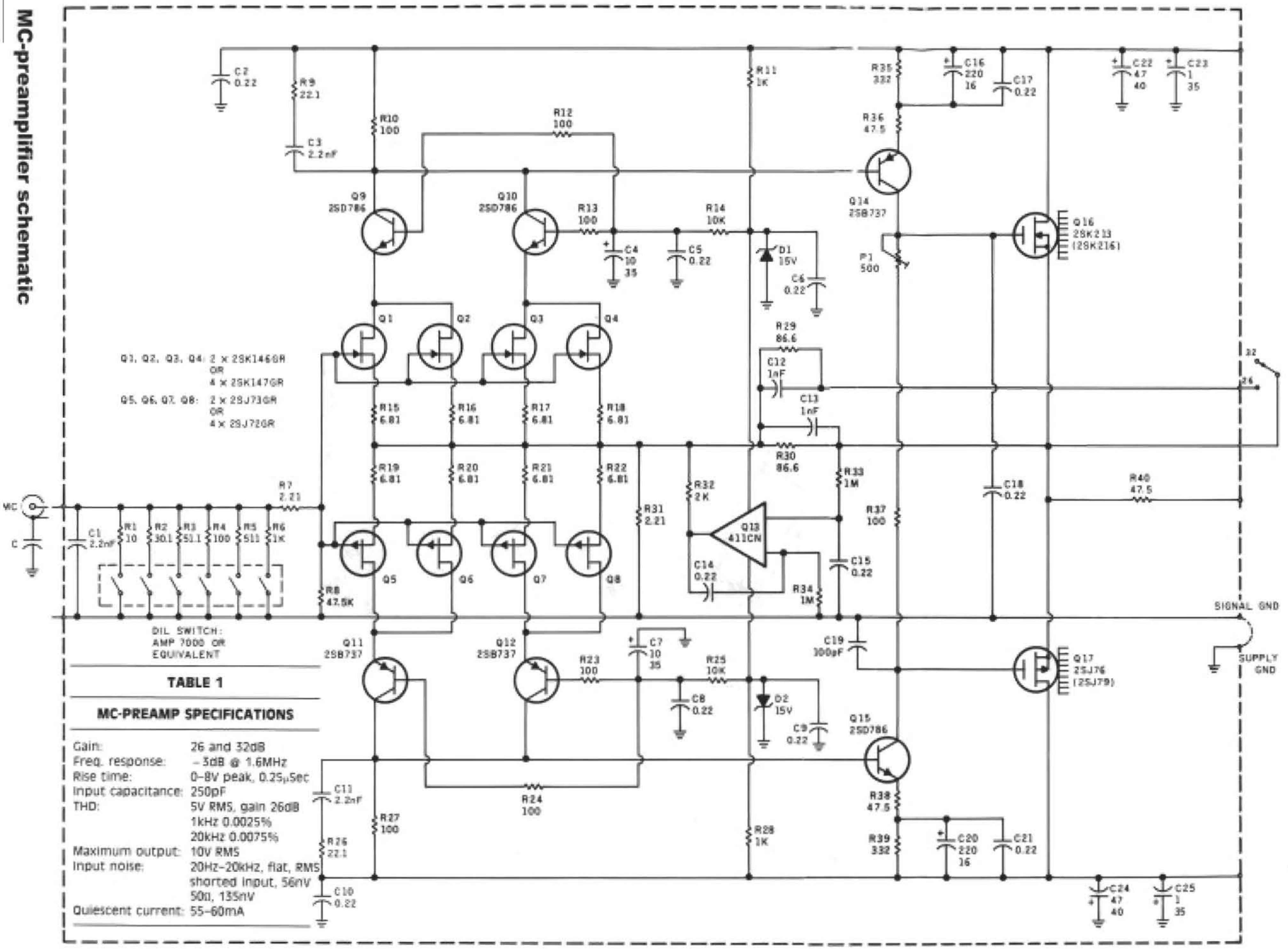
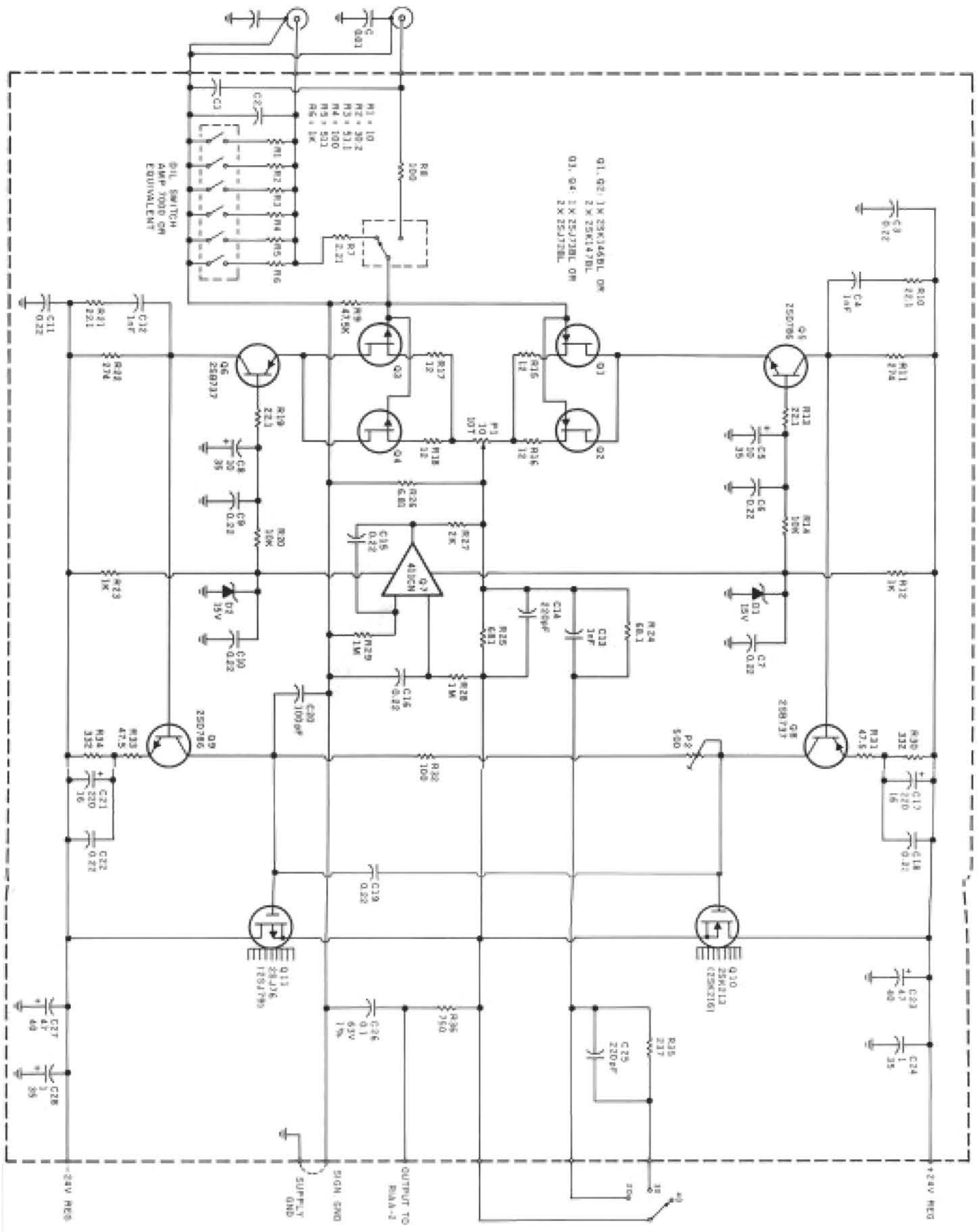


TABLE 1

MC-PREAMP SPECIFICATIONS

Gain:	26 and 32dB
Freq. response:	−3dB @ 1.6MHz
Rise time:	0–8V peak, 0.25μSec
Input capacitance:	250pF
THD:	5V RMS, gain 26dB 1kHz 0.0025% 20kHz 0.0075%
Maximum output:	10V RMS
Input noise:	20Hz–20kHz, flat, RMS shorted input, 56nV 50Ω, 135nV
Quiescent current:	55–60mA



Schematic for the RIAA-1 V.2 (high gain version of the EB-585 preamp

AN MC/MM PREAMPLIFIER

BY ERNO BORBELY

When I first published my semiconductor phono preamp some ten years ago,[1] I never thought I would try to make a tube version of it. It is hard enough to deal with the necessary low noise levels for MC (moving-coil) pickups using semiconductors, but with a pure tube MC phono, it is practically impossible to achieve the same quality of noise performance that exists in the best semiconductor models. Even if you use tubes having the lowest noise–such as the EC86 with an equivalent noise resistance of 250Ω–and parallel a number of them, it will always sound noisy compared to semiconductors.

This was not an issue in the golden days of tube audio, because the pickups at that time, mostly MM (moving-magnet), had much higher outputs than the current MCs. So when the MCs appeared, people had to use input transformers. Some of these were of high quality, providing very satisfactory performance, but they tended to be expensive, and not everyone could afford them.

The solution is, of course, to combine old and new technology: use low-noise JFETs to manage the input-noise problem, and let tubes handle the large signals. Then you will get the best of both worlds: low noise and high overload capability. And it sounds good, too.

MC-Preamp Considerations

I don't believe much has happened in terms of pickup sensitivity in the last several years. Most MCs fall in what I call the medium output category, which means an output of approximately 0.5mV at 5cm/s RMS lateral velocity. This is about 20dB lower than the normal MM output of 5mV. Naturally, there are lower output pickups than 0.5mV, but they need an additional gain. The high-output MCs are very close to the MMs and need no extra gain. Overall, an extra gain of up to 30dB can be necessary in front of the MM input to handle all MCs on the market.[1]

Two advantages of tubes are that they work with very high supply voltages and that their signal-handling capability is significantly better than that of transistors. It is easy to envisage an output voltage of 30–40V RMS from a tube stage, compared to the usual 5–10V of transistor amplifiers. So the upper limit of the amplifier's "dynamic range" is very good with tubes.

But what about the lower limit, which is normally restricted by noise? Here, semiconductors are much better than tubes. Suppose you want an 80dB signal-to-noise ratio in an MM system. Referred to 5mV, the input noise of a phono stage must be less than 0.5μV. This is just about the limit of what is possible with simple tube circuitry. However, if you want the same signal-to-noise ratio in an MC system with 0.5mV output, the input noise must be less than 50nV, which would be very difficult with tubes. In fact, such noise levels are not easy even with semiconductors.

Clearly, a compromise is necessary, and the best one I have found is to use semiconductors to handle the low noise and tubes to work with the large signals. This is done in a cascode circuit, where the lower part of the cascode is a low-noise, dual JFET, and the upper part is a low-noise dual triode.

The MC/MM Phono Preamp

The EB-1195/221 MC/MM phono preamp is a high-quality, two-stage tube preamplifier with approximately 63dB gain on the MC input and 44dB gain on the MM input. The circuit uses three tubes and one dual, low-noise JFET per channel, and works without feedback (*Fig. 1*).

The MC input stage is a low-noise, hybrid circuit made up of dual (or two matched) JFETs and a 6922 double tri-

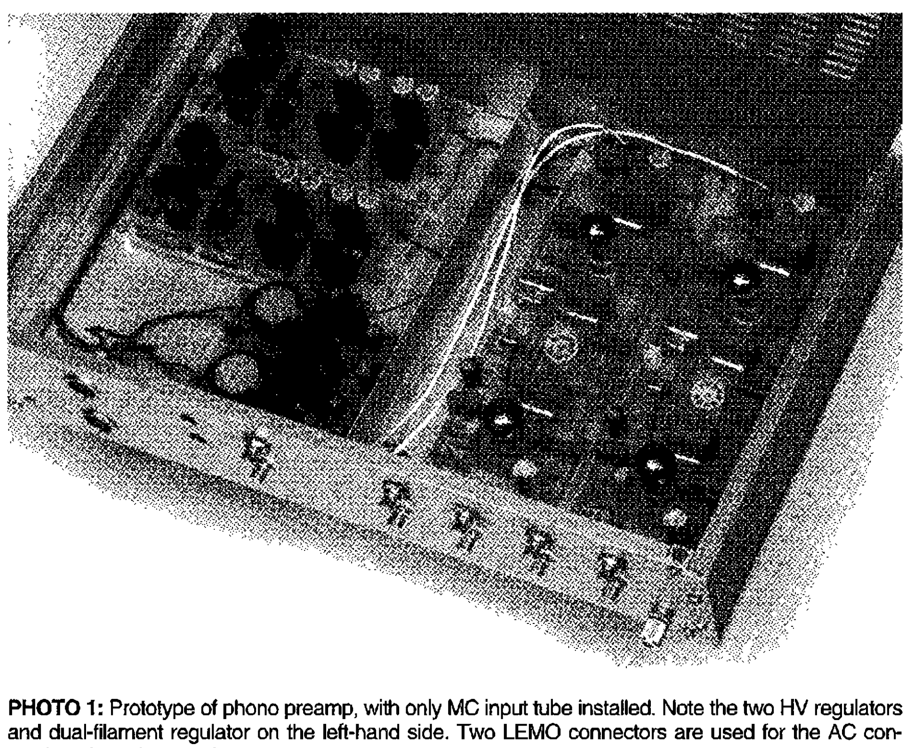


PHOTO 1: Prototype of phono preamp, with only MC input tube installed. Note the two HV regulators and dual-filament regulator on the left-hand side. Two LEMO connectors are used for the AC connections from the transformer.

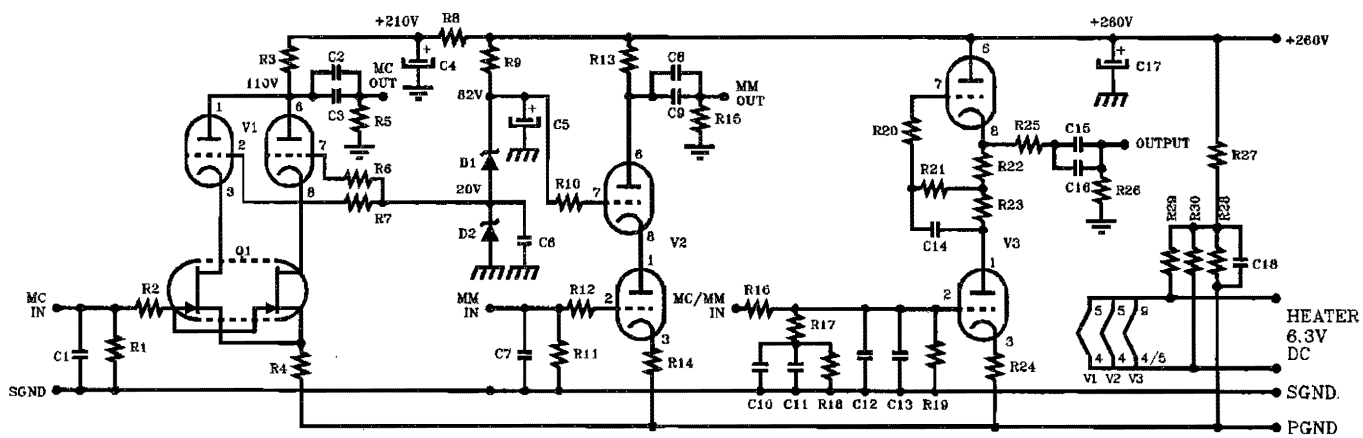


FIGURE 1: MC/MM phono preamplifier.

ode. The two JFETs and the two halves of the double triode are connected in parallel and then in cascode. The cascode circuit derives from high-frequency amplification, where low noise, wide bandwidth, and overload capability are the major requirements.

The current in the input stage, determined by the JFET source resistor R4, is set to about 10mA to optimize noise. The reference voltage for the triodes develops across a 20V zener diode. The voltage drop across V1 is 90V, the maximum permitted for the tube. The stage must be supplied from +210V, which can be derived from the main 260V supply through a 3.3k/4.5W resistor, or from a second regulated supply of 210V (see below).

Typical specifications for the MC input stage (all specs measured with 100k load):

Gain	50dB	
Frequency response	–3dB at 80kHz	
THD	1V	0.08%
	3V	0.25%
	10V	0.8%

ABOUT THE AUTHOR

Erno Borbely has been employed by National Semiconductor Europe for the last 17 years. He was manager of technical training and worked as a consultant in human resources development. He received an MSc degree in electronic engineering from the Institute of Technology, University of Norway in 1961 and worked seven years for the Norwegian Broadcasting Corporation designing professional audio equipment. For a time, he lived in the US and was director of engineering for Dynaco and The David Hafler Company. From 1973–1978 he worked for Motorola in Geneva, Switzerland, as senior applications engineer and applications manager. He is about to take an early retirement from National Semiconductor, and is looking for OEM customers for whom he can design high-end audio equipment. His E-mail address is: BorbelyAudio@t-online.de

The distortion is second harmonic.
Equivalent input noise: 120nV

MM Input Stage

The MM input stage is a cascode-connected 6922. The reference voltage for the cascode tube is 80V, developed across two zener diodes. The circuit works at around 9mA to get the best noise performance.

Typical specifications for the MM input stage (measured with a 100k load):

Gain	30dB	
Frequency response	–3dB at 80kHz	
Distortion	0.3V	0.027%
	1V	0.05%
	3V	0.15%
	10V	0.5%
	20V	1%

All distortion is second harmonic.
Equivalent input noise: 0.6μV

Both MC and MM inputs have an impedance of 47k. If you need other terminating resistors for your pickups, you can solder the appropriate resistor either on the PCB or on the input connector. Capacitors C1 and C7 are soldered on the PCB. Again, you might want to leave these off the board and solder the appropriate ones on the input terminal or connector. You can also place both the terminating resistor and capacitor on a back-panel switch.

Both input stages use the 6H23P-EB/6922 Russian military tube, which has the lowest noise I have ever measured on any tube. Equally important, it has practically no microphony. You can use other equivalent tubes, but, although they might provide better sound, they would probably degrade the circuit's noise performance. Our kits are deliv-

TABLE 1

PARTS LIST FOR PHONO PREAMPLIFIER

EB-1195/221

Resistors	
R1, R11**	47k5
R2	2.21Ω
R3, R13	10k, 1%, 2W, ROE MK-8
R4	12.1Ω, adjusted for 10mA
R5, R15, R21, R26**	1M
R6, R7, R10**	1k
R8*	3k3, 4.5W, ROE WK-8
R9	33k, 4.5W, ROE WK-8
R12, R20, R25**	47.5Ω
R14**	150Ω
R16**	75k
R17**	8.25k
R18	412k
R19	2M20
R22, R24**	825Ω
R23**	10k, 2.3W, ROE WK-5
R27	150k, 1.4W, ROE WK-4
R28	100k, 1.4W, ROE WK-4
R29, R30	470Ω, 1.1W ROE WK-2
All resistors are ½W, 1% metal film, unless otherwise noted.	
Capacitors	
C1, C11	2200pF, 160V or 630V, PP
C2, C8, C12, C15	10nF, 160V or 630V, PP
C3, C9, C14	0.22μF, 400V, WIMA MKP-10
C4, C17	10μF, 450V, ROE EKO
C5	47μF, 100V, ROE EKO
C6, C18	0.1μF, 400V, WIMA MKP-10
C7	100pF, 160V or 630V, PP
C10	39nF, 1%, 63V, RIFA PHE 425
C13	3.3nF, 160V or 630V, PP
C16	1μF, 250V, WIMA MKP-10
JFET, Tubes, Diodes	
Q1	2SK146BL, or 2x2SK147BL matched
V1, V2	6922/E88CC Russian
V3	ECC83/E83CC Tungsram
D1	62V, 1W zener, ZPY62
D2	20V, 1W zener, ZPY20
Miscellaneous	
6	9-pin ceramic sockets with gold-plated contacts
30	1.3mm solder pins
2	EB-1195/221 PCB

* Leave out R8 if MC input stage is supplied from separate, +210V regulator.
** A Tantalum resistor upgrade kit is available for these resistors (34 × ½W, 2 × 1W).

ered with the Russian tubes, but we can also provide them with Siemens E88CC or other equivalent tubes. Please see our price list for upgrades.

The Second Stage

The second stage, common to both MC and MM, contains the passive RIAA equalization and a mu stage. The RIAA network has an attenuation of 25dB and provides an RIAA accuracy of better than ±0.5dB across the audio band.

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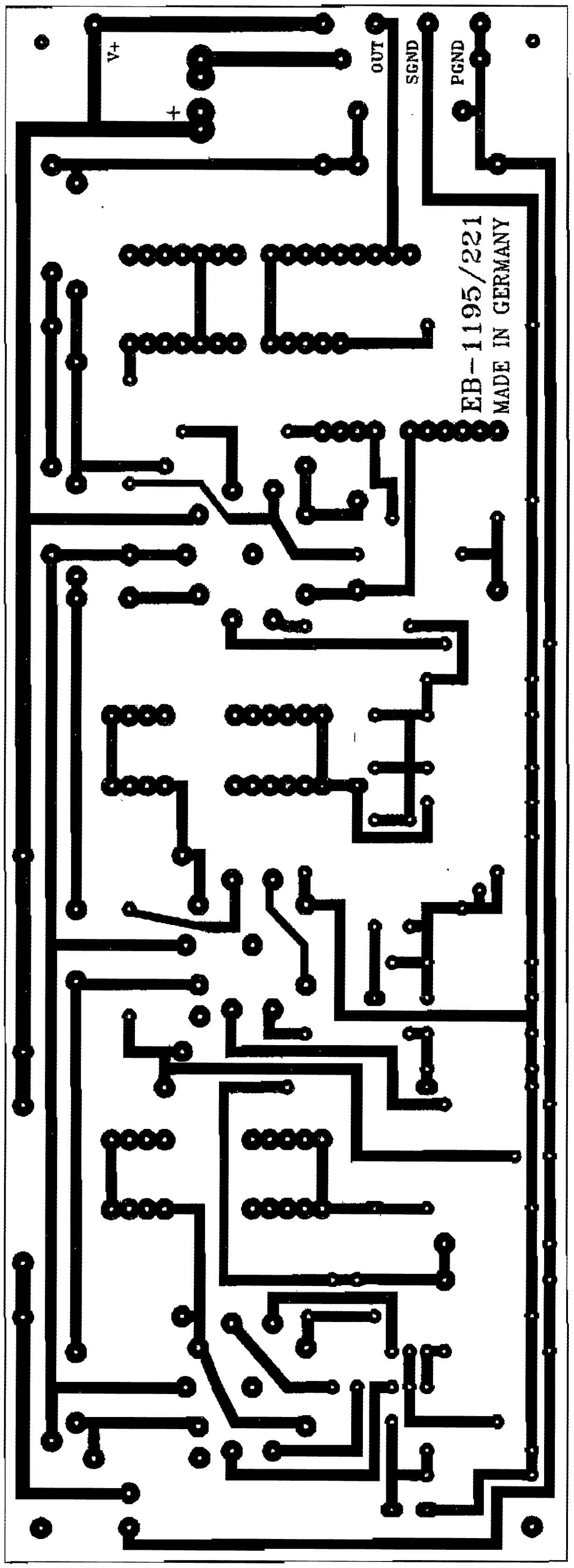

FIGURE 2: Preamp copper side (100%).

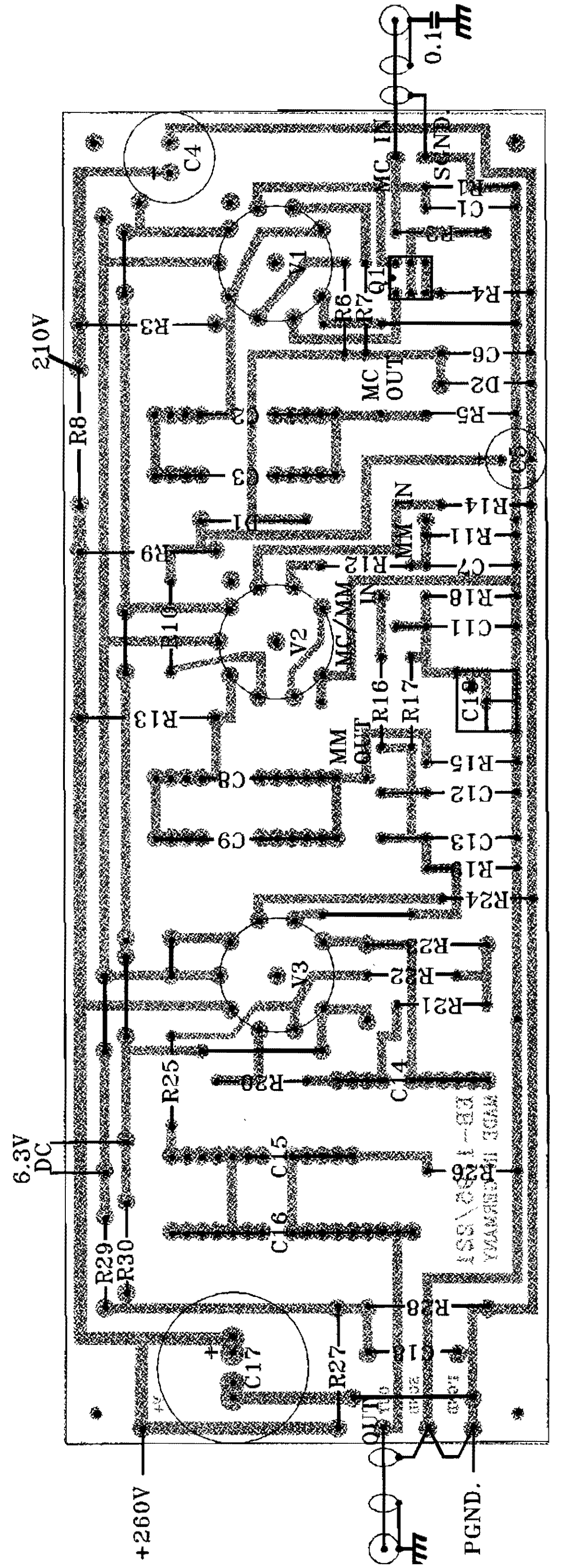

FIGURE 3: Preamp component side (85%).

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The mu stage uses an ECC83 and operates at approximately 1mA. The typical specifications are:

Gain	38dB	
Frequency response	–3dB at 75kHz	
THD	1V	0.012%
	3V	0.028%
	10V	0.095%
	20V	0.2%

The mu stage reaches 1% distortion at about 30V and saturates at over 40V. This stage determines the overall distortion of the phono preamp because it works with the highest signal levels. Although the output impedance is reasonably low (several kΩ), you can't load the stage very much without loss of amplitude and distortion. The recommended minimum load is 100k, but it can drive 50k without significant losses. A 50k or 100k volume control in the following lineamp works very well.

Note that all coupling capacitors consist of two capacitors in parallel. Although all the coupling caps are polypropylene, I use a smaller one in parallel with a larger. This appears to improve the midrange and the high end of the audio range. The layout accommodates the use of two caps. Feel free to experiment with different capacitors to get the best sound from your amplifier.

In case you wish to experiment with different types of tubes, the second-stage layout allows the use of both ECC83 and 6922 pinout tubes in this position. All that's necessary is to reconfigure the filaments: the 6922 needs 6.3V between pins 4 and 5; for the ECC83, connect pins 4 and 5 and make the 6.3V connection between pins 4/5 and 9. In case you use a 6922 here, you should connect pin 9 to ground.

Combining Stages

Only one of the input stages is operational at any one time, and only the appropriate tube is installed (V1 for MC, or V2 for MM). This limits the filament current to about 600mA per channel. If you have two pickups, one MC and one MM, you can operate both input stages and switch between them. The filament current will then be 900mA. The kit comes with all three tubes for each channel.

Due to the high gain and the RIAA bass boost, the circuit is very sensitive to hum pickup. Hum can come from power supplies or transformers. I recommend using regulated power supplies for both the high voltage and the filament. It is also good procedure to place the mains transformer and the regulators in a separate box for maximum hum protection.

For additional protection, you should

TABLE 2

PARTS LIST FOR TWO-OUTPUT MOSFET REGULATOR

EB-296/218A		
Resistors		
R1	1k, 4.5W	
R2, R5, R8	1k, 1.1W	
R3, R6, R9	270k, 1.4W	
R4, R7	10k, 1.1W	
Capacitors		
C1, C2	47µF, 450V	Radial, Siemens
C3, C4	0.1µF, 400V	WIMA MKP-10
C5, C6, C7, C8	10µF, 450V	Radial, ROE EKO, or Rubycon
Semiconductors		
Q1	2SA1156	400V, 0.5A, NEC
Q2, Q3	BUZ92	600V, 3A, Siemens
D1, D2, D3, D4	BYT11-1000	1000V, 1A fast rec.
D5	ZPD5.1	5.1, 0.5W, zener
D6	ZPY51	51V, 1W, zener
D7, D8	ZPY100	100V, 1W, zener
D9, D10	ZPY18	18V, 1W, zener
Miscellaneous		
1	EB-296/218A PCB	
10	1mm solder pins	
2	SK75 heatsink for Q2, Q3	
1	FK209 heatsink for Q1	
1	Fuse holder/fuse Wickman 19646, 19648 holder and cover	
1	5 × 20mm, 100mA, medium-fast fuse	

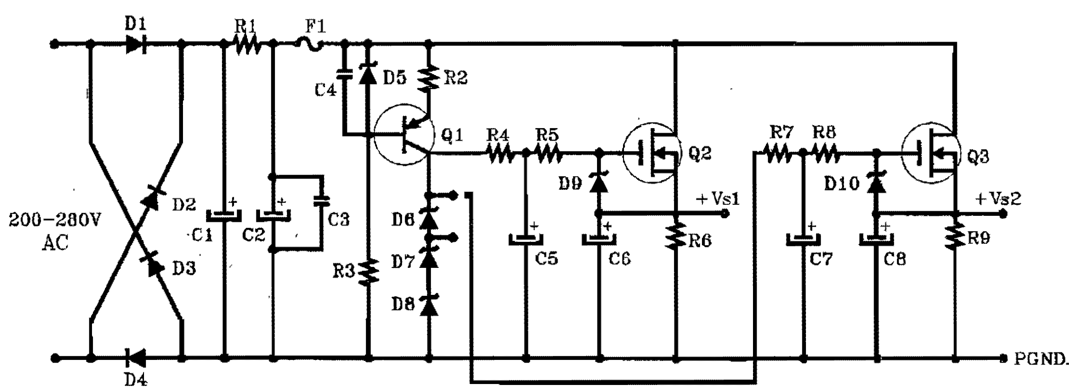


FIGURE 4: Two-output MOSFET regulator.

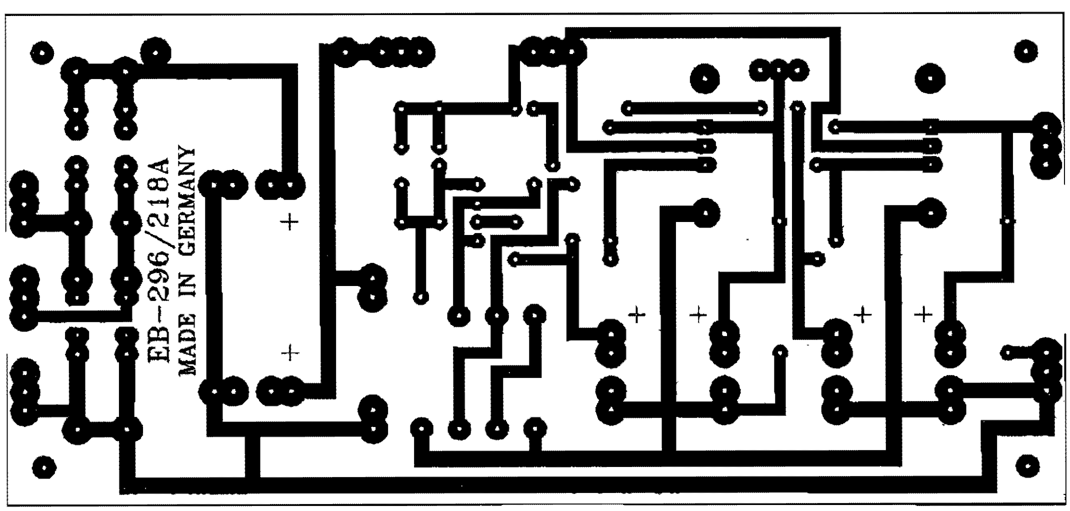


FIGURE 5: MOSET regulator copper side (100%).

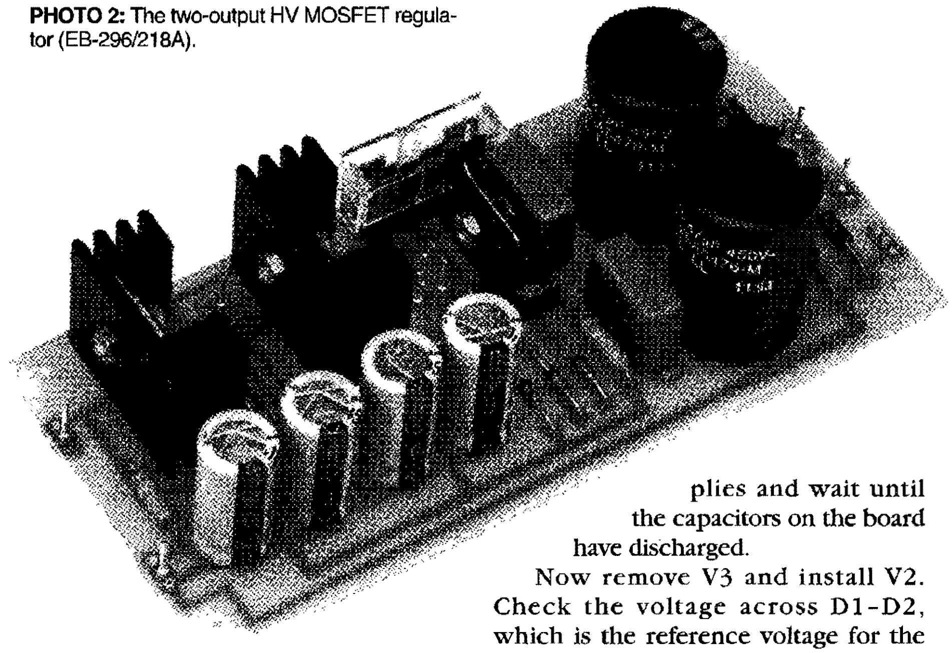
use shielded tube sockets for V1 and V2. Although shielded PCB sockets are not readily available (and are not supplied with the kit), you can make them from shielded chassis-type sockets by cutting off half of the pins with a pair of very sharp pliers, and then grounding the socket and the shield to a ground track on the PCB.

The filaments are biased to approximately +100V, which keeps the maximum voltage between cathode and filament within the limits given in the data sheet. This also helps reduce the tubes' hum susceptibility, which is very important in low-noise circuits like a phono preamp.

Phono-Preamp Setup Procedure

I recommend that you test the amplifier modules separately before building them into the chassis (*Figs. 2* and *3*). As a minimum, you will need a digital voltmeter (DVM) to set up the circuit. Use utmost caution in testing the circuit, for you are dealing with high voltages. If you have no experience with tube circuits, you should ask an experienced friend or an electronics technician to test your circuit for you.

PHOTO 2: The two-output HV MOSFET regulator (EB-296/218A).



First, short all three inputs to ground. Insert one tube at a time, starting with V3. Apply 260V and 6.3V regulated voltages to the board. Connect the DVM across R24 and check the voltage drop, which should be about 0.75V, indicating a current of 0.9mA. Switch off the supplies and wait until the capacitors on the board have discharged.

Now remove V3 and install V2. Check the voltage across D1-D2, which is the reference voltage for the cascode tube. It should be 82-84V. Connect the DVM across R14. The voltage drop is 135mV, indicating a current of 9mA in V2. Also check the anode voltage on pin 6 of V2; it should be close to 170V.

Repeat the procedure of switching off the supplies and discharging the

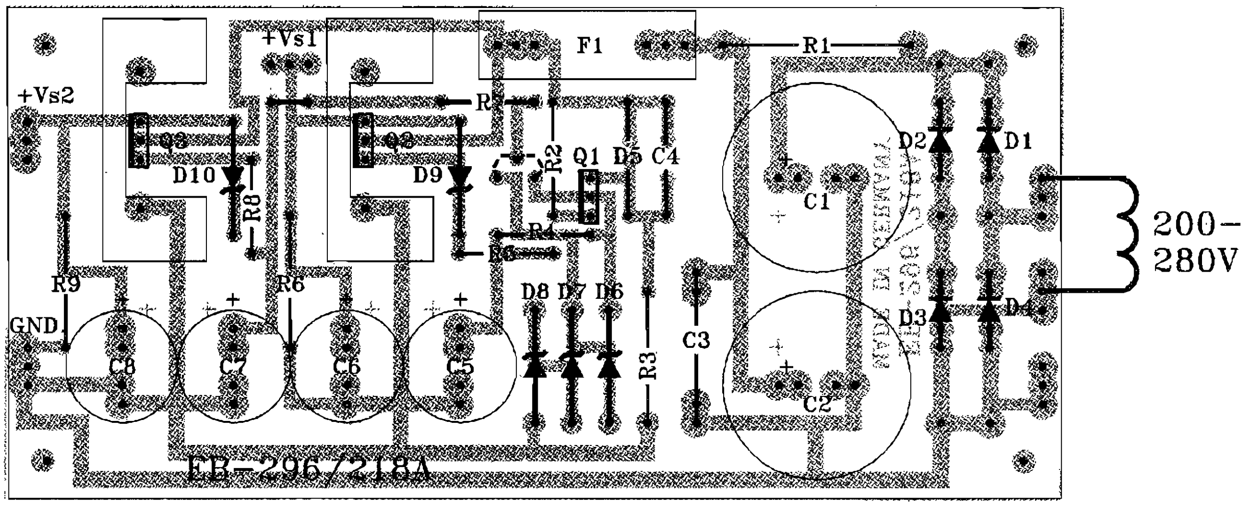


FIGURE 6: MOSFET regulator stuffing guide (100%).

capacitors on the board. Remove V2 and install V1. Then switch on the supplies and check the voltage across D2, which should be 20V. Connect the DVM across R4, where the voltage drop should be 120mV, indicating a current of 10mA. If the current is more than that, increase the value of R4; if it is less, decrease the value of R4. After adjusting the current to 10mA, check the anode voltage on pins 1-6 of V1, which should be 110V.

If you have audio instrumentation, I recommend you test each stage separately. Connect the audio oscillator to the MC input and measure its performance at the output of that stage. Similarly, connect the oscillator to the MM input and check the performance there. Finally, connect the oscillator to the second stage (V3) and check it through.

If you wish to check the whole circuit, you may need to connect a 40-50dB attenuator at the input of the amplifier

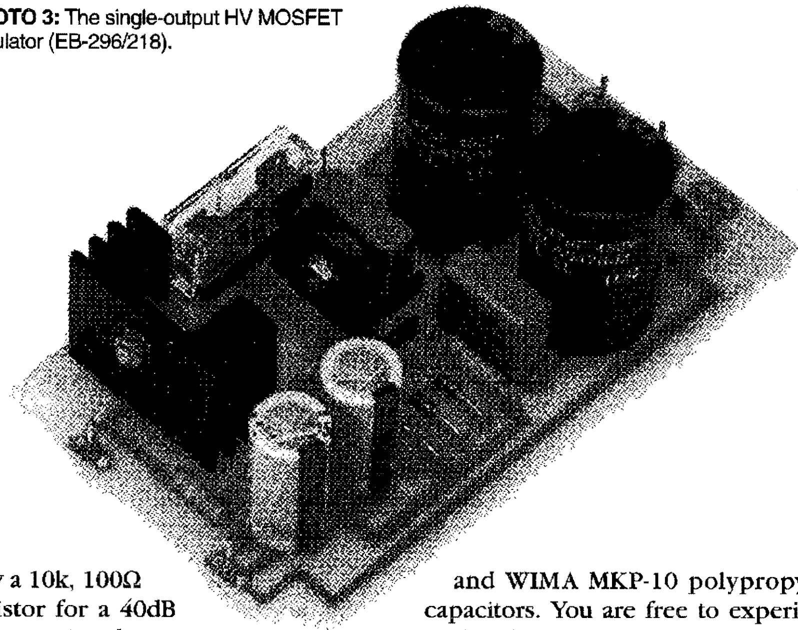


PHOTO 3: The single-output HV MOSFET regulator (EB-296/218).

(try a 10k, 100Ω resistor for a 40dB attenuator); otherwise, ground loops and hum will make the measurements difficult. To check the RIAA accuracy use an inverse RIAA circuit, such as the one described by Lipshitz and Jung.[2] The RIAA accuracy should be within ±0.5dB.

The phono-preamp kit comes with Roederstein MK-2 metal-film resistors and WIMA MKP-10 polypropylene capacitors. You are free to experiment with other components like tantalum resistors, other capacitors, and so on. I recommend using tantalum resistors in selected locations. These are available as an upgrade kit; please see our price list.

High-Voltage Regulator

In order to preserve the low-noise capability of the phono preamp, it is essential to feed it from a power supply having little ripple and noise. Older designs used capacitors and choke filters to reduce the ripple to an acceptable level. I have found that in addition to reducing the ripple, it is also an advantage to regulate the supply voltage, especially in low-level stages, because it tends to improve the imaging. This applies to both semiconductor and tube circuits.

The degree of regulation can vary from circuit to circuit, and in tube circuits you might get away with less regulation than with semiconductors. The important thing is, of course, that the regulator removes the ripple and adds very little wideband noise to the DC.

The EB-296/218A power supply (*Photo 2*; *Figs. 4–6*) combines a full-wave rectifier with fast-recovery diodes, high-quality capacitors, and two MOSFET source-follower regula-

REFERENCES

1. Erno Borbely, "A Moving Coil Preamp," Parts 1 and 2, *TAA* 4/86, 1/87.

2. Stanley P. Lipshitz and Walt Jung, "A High Accuracy Inverse RIAA Network," *TAA* 1/80.

tors. Using 450V capacitors, the maximum AC input should be limited to approximately 280V. With 400V caps, the AC input should be less than 250V.

R1 and C2 provide additional filtering after C1. You can also use R1 to drop the incoming DC voltage in case the required regulated voltage is significantly lower than the unregulated.

The regulated output voltage is defined by the zener string at the gate of the MOSFETs. This string is fed from a constant-current source to make sure that ripple is not transmitted to the zeners. The constant-current source, Q1, is again using a zener (5.1V) as a reference.

The 5.1V zener is fed through a 270k resistor, and both it and the string of high-voltage zeners are fed with a current of 1–2mA. The output of the zener string is further filtered with a 10kΩ resistor and a 10µF capacitor. The 1kΩ resistors in series with the gate, as well as the 18V zener between the gate and the source, provide protection for the MOSFET in case of overload.

Output +Vs1 is fed directly from the top of the three zener diodes; the second output, +Vs2, can be selected between the top and the second zener. You can therefore have two equal voltages, or one higher and one lower. In the case of the phono preamp, the higher voltage is 260V, and the lower one, which feeds the MC input stage, is 210V. By using different zeners, you can adjust the two voltages to your needs.

Naturally, you can use the two-output regulator in many other applications, such as lineamps, crossovers, and driver stages for power amplifiers, as well as for stabilizing the voltage for grid 2 of output pentodes.

You should heatsink Q1, Q2, and Q3. The heatsinks for Q2 and Q3 are mounted on the PCB, and that for Q1 is free-standing. All heatsinks should be mounted with mica or silicon rubber for insulation.

Regulator Setup Procedure

To test the regulator, connect 27k, 4.5W resistors to both outputs. Connect 250V AC to the AC input. Insert a 100mA medium fuse in F1. If you have a Variac, increase the mains slowly and monitor the output voltages with a DC voltmeter. The DC should stabilize close to the zener reference voltage(s) after a couple of minutes. If you have a scope or an AC mV meter, check the ripple at the output. It should be less than 5mV peak-to-peak.

Borbely Audio also offers a simplified version of the high-voltage regulator with only one output (type number EB-296/218) (*Photo 3*). All BA kits include epoxy PC boards, all resistors, capacitors and mechanical components that go on the board. Main transformers for 115/230V, 50/60Hz are also available on request. ❖

Erno Borbely

JFETS: THE NEW FRONTIER, PART 1

Welcome to a new era in audio amplification where JFETs rule. This noted designer champions their use to produce the best sound in your audio amp circuits.

As most of our customers know, I have been advocating the advantages of FETs in general and JFETs in particular, especially for low and medium level circuits. JFETs provide extremely high resolution, bringing out more details, sounding cleaner, clearer, and more natural than the best bipolar transistors such as the LM394, and even the best Telefunken tubes. Overall, I believe the JFETs offer the best sound in audio circuits.

I have been working with JFETs since the middle of the '70s, when I developed low-level amplifier modules with JFETs at Motorola. However, they were not competitive with the best bipolars at that time. In the early '80s came the first really low-noise, high-g_m devices on the market. I have used these devices in the input stages of practically all my designs since then. However, I use bipolar transistors in the second stages, mostly because they offer a fairly simple design. The output stages have always been MOSFETs, because of the relatively high current required in these stages.

In the ever-continuing quest for better sound, I have reviewed my designs regularly, improving the topology of the amplifiers and also using better components, thus bringing significant improvements. However, I first achieved a real breakthrough when I started to use mostly JFETs in the amps. It is my considered opinion that it would be best to use only JFETs in all stages of the audio chain. However, due to their limited power-handling capability, it is practically impossible to use them in output stages. Here, MOSFETs will rule for the foreseeable future.

In spite of their quadratic characteristics and relatively high input capacitance, JFETs are fairly simple to use in audio amplifiers, and you, as an amateur, can design most low-level stages in an audio chain yourself. Just like a single vacuum-tube triode or pentode, a single JFET can handle the task of a line amp, and it is significantly simpler to hook up. You can also build a single-ended (SE) phono stage with only two JFETs. The rest is up to your imagination. Suffice it to say that I hope the following introduction to JFETs will whet your appetite for the "new frontiers" in audio amplification.

About the Author

Erno Borbely has been employed by National Semiconductor Europe for the last 17 years. He was Manager of Technical Training and worked as a consultant in human-resources development. He received an MSc degree in electronic engineering from the Institute of Technology, University of Norway in 1961, and worked seven years for the Norwegian Broadcasting Corp. designing professional audio equipment. He lived in the US and was Director of Engineering for Dynaco and The David Hafler Co. From 1973–1978, he worked for Motorola in Geneva, Switzerland, as Senior Applications Designer and Applications Manager. He has now taken an early retirement from National and is looking for OEM customers for whom he can design high-end audio equipment.

JFETs

Field-effect transistors (FETs) have been around for a long time; in fact, they were invented, at least theoretically, before the bipolar transistors. The basic principle of the FET has been known since J.E. Lilienfeld's US patent in 1930, and Oscar Heil described the possibility of controlling the resistance in a semiconducting material with an electric field in a British patent in 1935. Several other researchers described similar mechanisms in the '40s and '50s, but not until the '60s did the advances in semiconductor technology allow practical realization of these devices.

The junction field-effect transistor, or JFET, consists of a channel of semiconducting material through which a current flows. This channel acts as a resistor, and the current through it is controlled by a voltage (electric field) applied to its gate. The gate is a pn junction, formed along the channel. This description implies the primary difference between a bipolar transistor and a JFET: the pn junction in a JFET is reverse-biased, so the gate current is zero, whereas the base of a bipolar transistor is forward-biased, and the base conducts a base current. The JFET is therefore an inherently high-input impedance device, and the bipolar transistor is comparatively low-impedance.

Depending on the doping of the semiconductor material, you get so-called N-type or P-type material, and these result in the N-channel or P-channel types of JFET. The symbol for an N-channel JFET is shown in *Fig. 1A*. The three "electrodes" are called G, D, and S, for gate, drain, and source. The output characteristic for the N-channel JFET with the gate shorted to source (i.e., $V_{GS} = 0$) is shown in *Fig. 1B*.

The characteristic field is divided into two regions, first a "resistive" region below the saturation voltage V_{SAT}, where an increase in V_{DS} results in a nearly linear increase in drain current I_D. Above

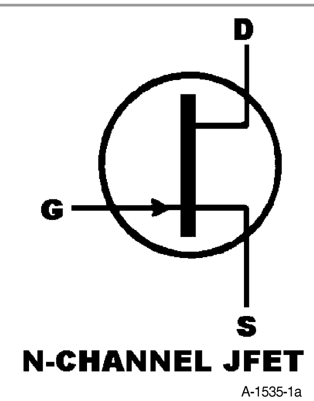


FIGURE 1A: Symbol for N-channel JFET.

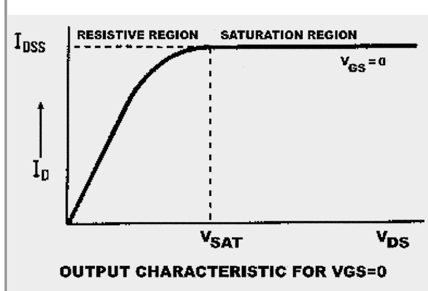


FIGURE1B: Output characteristic for V_{GS}=0V.

V_{SAT}, an increase in V_{DS} does not result in a further increase in I_D, and the characteristic flattens out, indicating the "saturation" region. Sometimes these two regions are also called "triode" and "pentode" regions.

You can use the JFET as a voltage-controlled resistor or a low-level switch in the triode region, and as an amplifier in the pentode region. As you see, the N-channel JFET conducts maximum current I_{DSS} with V_{GS} = 0V. If you apply a negative voltage to the gate, it reduces the current in the channel, and you get a family of output characteristics as shown in *Fig. 2A*. This device is called a "depletion" type of JFET.

In summary, the JFET consists of a channel of semiconducting material, along which a current can flow, and this flow is controlled by two voltages, V_{DS} and V_{GS}. When V_{DS} is greater than V_{SAT}, the current is controlled by V_{GS} alone, and because the V_{GS} is applied to a reverse-biased junction, the gate current is extremely small. In this respect, the N-channel JFET is analogous to a vacuum-tube pentode and, like a pentode, can be connected as an amplifier.

The P-channel JFETs behave in a similar manner, but with the direction of current flow and voltage polarities reversed. The P-channel JFET has no good analogy among vacuum tubes.

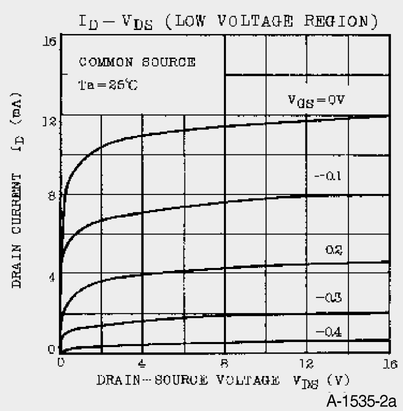


FIGURE 2A: Family of output characteristics for 2SK170.

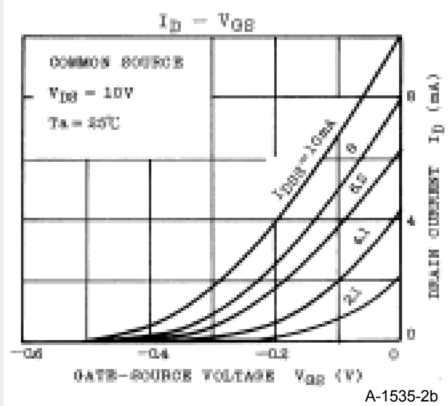


FIGURE 2B: Transconductance curves for 2SK170.

The Transconductance Curve

As mentioned previously, you can use the JFET as an amplifier in the pentode, or saturation, region. Here the V_{DS} has little effect on the output characteristics, and the gate voltage controls the channel current I_D. Because of this, it is easy to characterize the JFET in terms of the relationship between I_D and V_{GS}, that is, with the transconductance curve. *Figure 2B* shows the transconductance curves for a typical low-noise, high-g_m JFET, the 2SK170.

The drain current as a function of V_{GS} is given by the formula:

$$I_D = I_{DSS}\left[1 - \frac{V_{GS}}{V_P}\right]^2.$$

V_P is the gate pinch-off voltage, and is defined as the gate-source voltage that reduces I_D to a very low value, such as 0.1µA. The formula indicates that the transconductance curve has a square-law form. It also shows that if you know I_{DSS} and V_P, you can draw the transconductance curve for any JFET. The transconductance g_m, which is the slope of the transconductance curve, is found by differentiating I_D with respect to V_{GS}:

$$g_m = \frac{dI_D}{dV_{GS}} = -\frac{2I_{DSS}}{V_P}\left[1 - \frac{V_{GS}}{V_P}\right]$$

The transconductance g_m becomes $-2I_{DSS}/V_P$ where the transconductance curve meets the y-axis. This is the value you normally find given in the data sheets. Notice that there are five different transconductance curves given for the 2SK170 in *Fig. 2B*. This indicates there is a range of I_D curves for each JFET, due to manufacturing tolerances.

Also notice that the transconductance curve stops where it meets the y-axis. This is because the gate pn junction would be forward-biased if V_{GS} were made positive for N-channel and negative for P-channel JFETs, and gate current would flow. This is analogous to the condition of vacuum tubes when the grid is made positive. Of course, a silicon pn junction does not conduct before the forward voltage reaches 0.6–0.7V, so you can apply several hundred mV in the forward direction without ill effects. JFETs are often operated with both polarities of gate voltage—i.e., with gate current—in RF applications.

The change in the transconductance curve is not just a matter of tolerances due to manufacturing, but it also depends on the temperature, and this is due to two different effects. As the temperature increases, the mobility of the charge carriers in the channel decreases, which leads to an increasing channel resistance, and hence a reduction in I_D.

On the other hand, the barrier potential of the gate pn junction decreases about 2.2mV/°C, which causes the I_D to increase. There is a point on the transconductance curve where these two effects cancel one another, and the temperature coefficient (tempco) becomes zero. Obviously, if you need to design for low drift, then the JFET must be operated at this point.

You can calculate the zero tempco point with the following formula:

$$V_{GS} = V_P + 0.63V$$

Typical transconductance curves for two different JFETs are shown in *Figs. 3A* and *3B* for a high-V_P and a low-V_P JFET, respectively. It is obvious from the curves that the zero tempco point occurs at a lower I_D for high-V_P JFETs and at a higher I_D for low-V_P JFETs. If the V_P is close to 0.6V, then the zero tempco point is close to I_{DSS}.

The Bias Point

As shown in *Fig. 2B*, the JFETs have a relatively wide range of transconductance curves. In order to operate the JFET as a linear amplifier, you need to have a clearly defined operating point. A typical common-source amplifier stage is shown in *Fig. 4A*. Assume that the +Vs is 36V, and you have selected a load resistor R_L = 10k. What happens now if you insert a typical JFET, such as the 2SK170, for Q1?

Figure 4B shows five of the transconductance curves for the 2SK170, with I_{DSS} between 2.1mA and 10mA. If you take one of these at random and operate it without R_S, the actual drain current will be the I_{DSS} value. With 2.1mA, the voltage drop across R_L will be 21V; i.e., the drain (OUT) will be sitting at 36 − 21 = 15V. This might not be optimal from the point of view of maximum output or minimum THD, but it will work all right.

However, with I_{DSS} = 10mA, the voltage drop should be 100V, which is clearly impossible with Vs = 36V, and the

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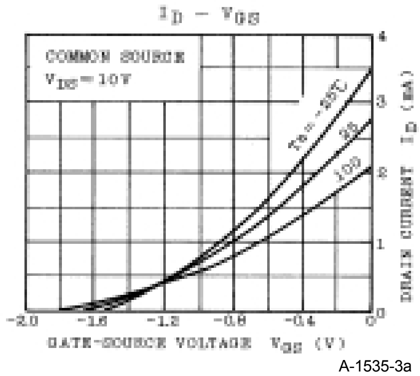

FIGURE 3A: The zero tempco point for 2SK246.

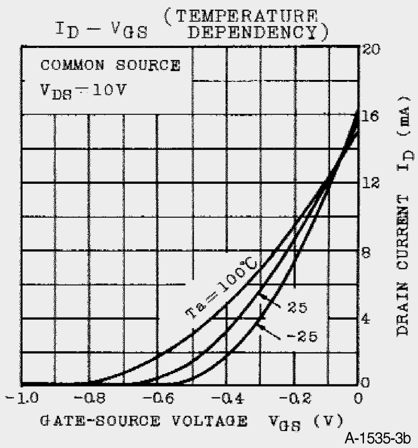

FIGURE 3B: The zero tempco point for 2SK170.

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amplifier goes into saturation. Obviously, if you wish to use any or all of these JFETs, you must reduce the effect of the wide range of I_{DSS}.

The solution is to use a source resistor R_S, similar to the biasing arrangements used in bipolar transistors or tube amplifiers. To illustrate the effect, I have drawn in the line for a 100Ω resistor in the transconductance characteristics. The range of drain currents is now limited between 1mA for the I_{DSS} = 2.1mA device, and about 2.6mA for the I_{DSS} = 10mA device. The drain voltages will be 36 – 10 = 26V and 36 – 26 = 10V, respectively. This is still too much variation from the point of view of THD and maximum output swing, but at least there is no saturation with any of these devices.

Fortunately, JFETs are sold with much narrower I_{DSS} ranges, which makes life easier in terms of proper biasing. The 2SK170 comes in three I_{DSS} groups: the "GR" group is 2.6–6.5mA, the "BL" group is 6–12mA, and the "V" group is 10–20mA. If you use a "GR" device with R_S = 100Ω, the I_D will vary between 1 and 2mA, which is almost acceptable.

The best solution, of course, is to select the devices for your particular application. Assume you wish to build a single-ended phono amp with JFETs and a passive RIAA correction network, and you decide to use the 2SK170 devices. In order to keep circuit noise to a minimum, you would use the 2SK170 without R_S, i.e., at I_{DSS}. Furthermore, you would need a relatively high current to be able to drive a passive RIAA correction network. If you choose, say, 5mA, you would need to select the devices from the "GR" group. But how? The selection is easy.

Testing JFETs

Figure 5 shows a simple circuit with which you can select JFETs and also match them if necessary. The tester feeds current into the source or connects the source to ground to measure the essential parameters of the device. In position 1 (switch in counterclockwise position), the source is connected to –10V through a 1M resistor. This feeds the source with approximately a 10μA current, which you can consider the cutoff point V_P for the JFET. (Data sheets specify lower values, but this gives you a more practical value for measurements.) The voltmeter now indicates the pinch-off voltage V_P for the device.

The next two positions measure the V_{GS} for the device at given drain currents. These positions give practical readings for design purposes, and you can choose the constant-current sources for the values you need. The push-button switch shorts the source to ground, and the mA meter measures I_{DSS}. If you wish to measure only V_P and I_{DSS}, you can permanently wire the source to –10V through the 1M resistor, which gives you V_P, and then short the source to ground with the push-button to read I_{DSS}.

If you test P-channel devices, you must reverse the supply voltages and the constant-current diodes. Normally, I test a large batch of devices (say 100 of each type) and sort them by I_{DSS}. The different devices are then used in different applications.

Some Practical Measurements

As mentioned previously, the transconductance curve has a quadratic form, and if you wish to use it to amplify audio signals, it will create harmonics. A true quadratic curve would generate only second harmonics; however, ideal devices are hard to come by, and practical devices also generate some higher harmonics. Again, in this respect there is a close similarity to vacuum tubes. Looking at the transconductance curve, you can easily see that it is more linear close to the y-axis than further down on the curve. From the point of view of linearity, it is therefore an advantage to operate the JFET with a higher I_D.

Figures 6A and *6B* show the transconductance characteristics for two JFETs I use in many of my amplifiers. The 2SK170 is a high-transconductance device with low V_P, and the 2SK246 is a low-transconductance JFET with a higher V_P.

I have selected a 2SK170 with I_{DSS} = 6.2mA and a 2SK246 with I_{DSS} = 5.6mA

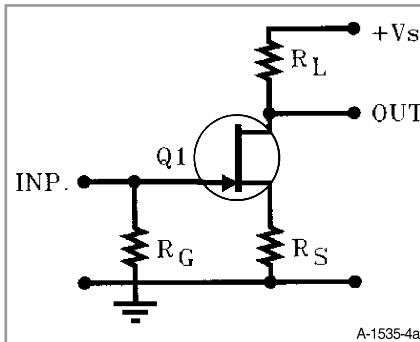


FIGURE 4A: A common-source amplifier.

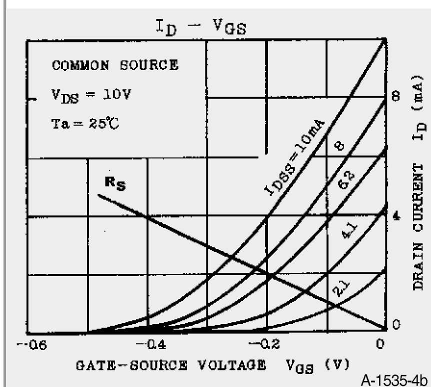


FIGURE 4B: The source resistor R_s stabilizes the bias point.

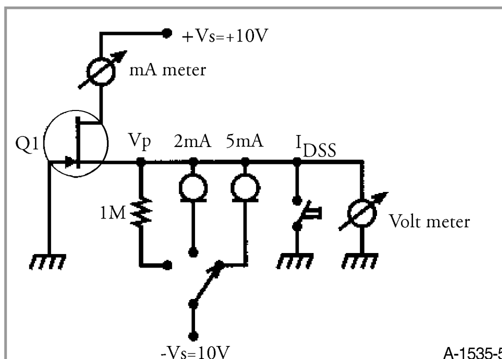
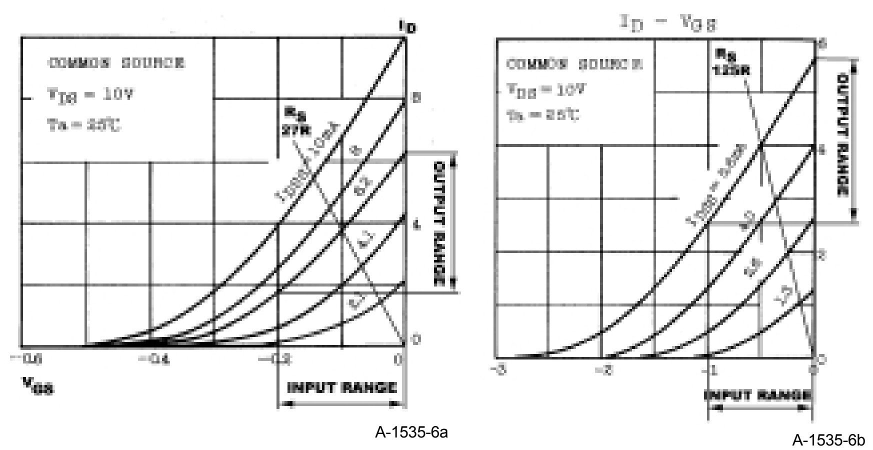


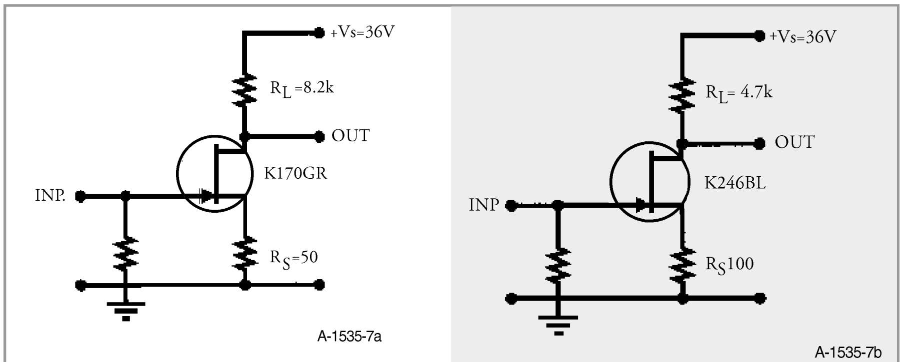
FIGURE 5: A simple test jig for N-channel JFETs.

to illustrate the difference of operation with very similar values of I_{DSS}. The gate pinch-off voltage is approximately 0.45V for the K170, and 2.75V for the K246. In order to operate them at the most linear part of the characteristic, I selected bias points at V_{GS} = 0.1V and I_D = 3.8mA for the K170, and V_{GS} = 0.5V and I_D = 4mA for the K246. These points are set with R_S = 27Ω and 125Ω, respectively.

The most obvious difference between the two JFETs is in the maximum input swing with which you can drive them. The K170 allows approximately ±0.1V peak before the gate goes positive, but the K246 has a range of ±0.5V! Naturally, I could move the working point further down on the transconductance curve in order to increase the input range, but eventually I would reach the other limiting point, where the gate cuts off at V_P. The thing to understand here is that a high-V_P JFET has a wider range of input swing than one with a low V_P.



FIGURES 6A/6B: Input/output range for 2SK170 and 2SK246, respectively.



FIGURES 7A/7B: Practical amplifiers with 2SK170 and 2SK246.

Other obvious differences involve the output range and the gain. With a ±0.1V gate voltage, the drain current varies between 1.8 and 6.2mA for the K170. With a drain resistor R_L = 4.7k, this results in an output swing of 29.14V – 8.46V = 20.68V pk-pk. The gain will then be 20.68/0.2 = 103.4, which is 40dB. The output range for the K246 is 2.5mA to 5.6mA. With the same drain resistor of 4.7k, the output-voltage swing will be 26.32 – 11.75 = 14.57V pk-pk. The gain is 14.57/1 = 14.57 times, which is 23.38dB. That is, the high-V_P device has lower gain than the low-V_P one.

When Higher Is Lower

Of course, this can be explained by the transconductance. The g_m for the K170 is $2I_{DSS}/V_P$ = 27.55mS. The gain is $g_m \times R_L$, which gives 127 times, a bit higher than the graphical analysis. The explanation for this is that this g_m is at the point where the curve crosses the y-axis, which is always higher than at the working point, and that the curve is not a straight line, making the output swing smaller than the theoretical value.

In any case, this quick calculation gives you a reasonable starting point from which to design the circuit. The corresponding g_m for the K246 is 4mS, so obviously the gain is also much smaller at 19.14, that is, 25.63dB. Again, this results in a higher value than the graphical analysis.

Now for some real circuits and THD measurements. *Figures 7A* and *7B* show two amplifiers with K170 and K246. The K170GR had an I_{DSS} of 5.5mA, and I operated it first with RS = 0 and R_L = 3.3k. This gave me a gain of 36.4dB and a frequency response of over 400kHz. The THD is shown in Column 1 of *Table 1*.

Column 2 shows the same K170GR device, but this time with R_S = 50Ω. This reduces the drain current to approximately 2.5mA, so I increased the drain resistor to 8.2k to have the same DC conditions as before. The THD is reduced by roughly 6dB. Column 3 shows the K246BL amp operating at I_D = 5.1mA, with R_S = 100Ω, and R_L = 4.7k. The output is now a bit lower than half of the supply voltage, and the maximum output is therefore limited. But the THD is quite low, again about 6dB lower than the previous circuit.

The K170GR circuit seems to be popular for phono input stages, and a number of these are circulating on the Internet. R_S is usually shorted to achieve minimum noise. However, even without R_S, the noise of a single K170 is not low enough for MC pickups. To achieve lower noise, you can parallel several of these devices. Doubling the JFETs with comparable g_m reduces the noise by approximately 3dB. I hooked up four K170s in parallel to see how it works (*Fig. 8*). Each device had an I_{DSS} of approximately 15mA, and the drain currents with R_S = 6R8 are 10mA each. With an R_L = 511Ω, the drain is sitting at 14.8V DC.

The gain is 34dB and the frequency response is 360kHz. The THD for this circuit is shown in Column 4 of *Table 1*. Remember that this circuit is working at very low levels, where THD is indeed low. The equivalent input noise is also reasonably low at approximately 100nV over a 20kHz bandwidth. Not bad for a simple circuit. Want to try it?

Input Capacitance

As mentioned before, the JFETs have a relatively high input capacitance, which can be an important design factor. Just like tubes and bipolar transistors, JFETs also have interelectrode capacitances that affect the frequency response of the JFET when it is used as an amplifier. The two capacitances, which are of importance for audio use, are the Ciss and Crss.

The Ciss is called the input capacitance and Crss the reverse transfer capacitance. Typical values for the Ciss are 30pF for the K170, and 9pF for the K246. The high-g_m devices have a much higher input capacitance than the low-g_m ones. The Crss is 6pF and 2.5pF, respectively. The Crss seems to be relatively low, but this is the one that dominates the input capacitance of an amplifier through the Miller-effect.

The input capacitance of a normal common-source JFET stage as shown in *Fig. 7*, but with $R_S = 0$, is given by the formula: $C_{in} = Ciss - A_V \times Crss$, where A_V is the voltage gain of the stage. Note that a common-source stage inverts the phase, so A_V is negative, making C_{in} a positive number. Since A_V can be a significantly large number, the input capacitance of the stage can be very high.

I have measured the input capacitance for the amplifier in *Fig. 7*, both with and without R_S. Without R_S, the capacitance was over 600pF! With $R_S = 100\Omega$, the input capacitance dropped to 127pF, because of the local feedback through R_S. To appreciate the significance of this, assume that you are driving the amplifier from a 100kΩ volume control. The amplifier will see a maximum "source impedance" of 25k when the volume control is in the middle. If you calculate the 3dB point of the low-pass filter formed by the volume control and the input capacitance of 600pF, you find that it is about 10kHz! If you use the K170 without R_S, you certainly must use a volume control, which is less than 100k.

Cascode to the Rescue

There is another way of reducing the input capacitance of the amplifier. Cascode connection of devices was invented in the tube era, but has also been used extensively with bipolar transistors. One of the advantages of cascoding, if you recall, is reduction of input capacitance, which makes it easier to design high-frequency amplifiers.

I have connected two circuits to test this (*Fig. 9*). The upper JFET needs a bias voltage, and it is easy to get this by connecting its gate to the source of the lower JFET. (Of course, you can also generate this bias from the supply voltage with a voltage divider, as you normally do with tube cascodes.) I am using a high-V_P JFET for the upper device, so that the lower JFET has enough voltage across it to operate in the saturation region.

The input capacitance of the circuit in *Fig. 9A* is approximately 160pF, so the cascoding indeed reduces the input capacitance. Further reduction is achieved by adding local feedback with R_s (*Fig. 9B*). The input capacitance is now reduced to 50pF. With such low input capacitance there is no longer any danger of creating a low-pass filter with the volume control.

As though the existence and size of the input capacitance were not enough, it is also voltage dependent, which might cause distortion in certain applications. *Figures 10A* and *10B* show the voltage dependence of Ciss and Crss, respectively, of the K170 JFET.

Depending on the excursion of the input/output signal, you get a capacitance modulation, and this can cause distortion of the audio signal. This shows up mostly when you drive the circuit from a high-source impedance. I have tested the circuit described in Column 1 and Column 2 of *Table 1* with different source impedances, and could not measure any significant increase in THD up to 50k source.

However, when the noncascode circuit was driven from 500k, the THD increased approximately 6dB. The cascoded circuit showed no significant increase at any source impedance up to 500k. To avoid capacitance modulation problems, I recommend that you use a volume control of no more than 50k. (Of course, you would probably use no more than 50k anyway, because of the increased noise with higher impedances.)

Note that in these circuits only two types of JFETs have been involved,

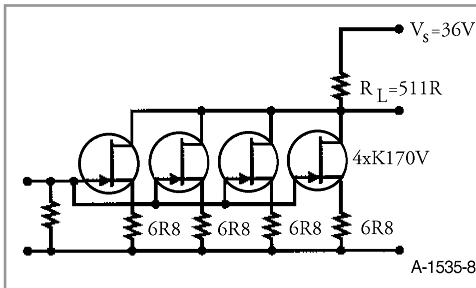
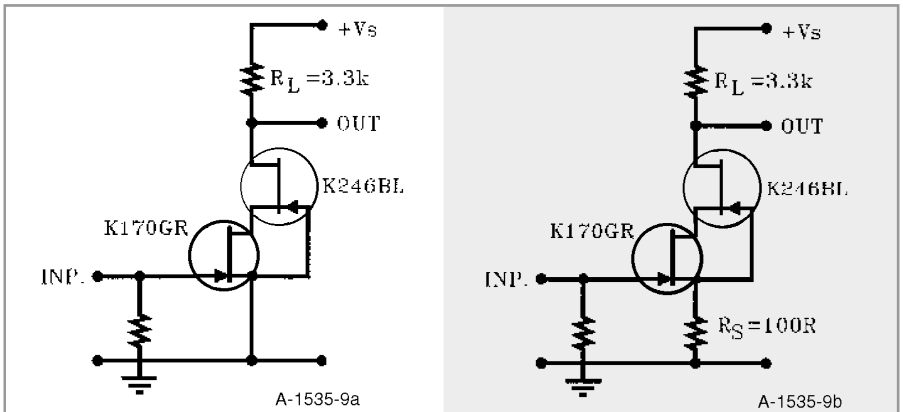


FIGURE 8: Paralleling JFETs reduces the noise.

TABLE 1

Output voltage, V RMS	Column 1 K170GR, $R_S = 0$, $R_L = 3.3k$	Column 2 K170GR, $R_S = 50$, $R_L = 8.2k$	Column 3 K246BL, $R_S = 100$, $R_L = 4.7k$	Column 4 4xK170V, $R_S = 6R8$ $R_L = 511R$
0.1V	0.095%	0.06%	0.02%	0.04%
0.3V	0.2%	0.1%	0.047%	0.1%
1V	0.6%	0.32%	0.15%	0.32%
2V	1.3%	0.65%	0.29%	0.67%
3V	1.9%	0.98%	0.4%	1%
5V	3.2%	1.7%		1.65%
10V	6%	3.4%		3.5%



FIGURES 9A/9B: Cascoding and local feedback with R_s reduces the input capacitance.

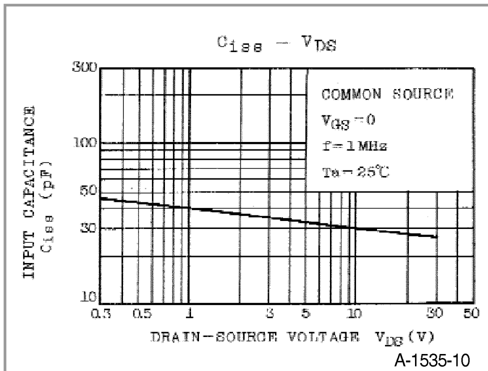


FIGURE 10: Voltage dependence of Ciss for 2SK170.

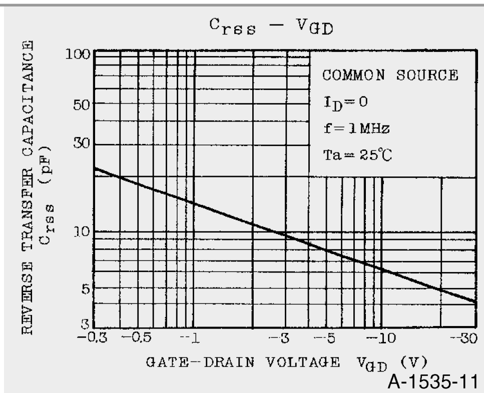


FIGURE 11: Voltage dependence of Crss for 2SK170.

whereas there are thousands of them on the market. Also, I have used them for illustration purposes only, and, although they work as described, I have made no attempt to optimize them for any particular application.

In Part 2 of this article, I will discuss the differential topologies. If you have questions, please don't hesitate to send me an e-mail or a fax (Borbely Audio, e-mail: borbelyaudio@t-online.de, FAX: +49/8232/903618, Web site: http://home.earthlink.net/~borbelyaudio). And, of course, if you wish to buy some JFETs to experiment with, we have tons of them in stock. For a little extra, we even do a selection for you. Have fun experiencing the "new frontier" in audio amplification. ■

Acknowledgements

My sincere thanks to Walt Jung of Analog Devices, who kindly read the manuscript and provided valuable comments and suggestions.

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Erno Borbely

JFETS: THE NEW FRONTIERS, PART 2

This noted design expert from Germany continues his series by showing how you can best take advantage of JFET performance.

In Part 1 of this article, I discussed the single-stage (or single-ended) amplifier operating in common-source mode. As these stages are usually limited in audio to AC signals, the inherent DC drift is of relatively little importance. You can even use them for DC signals if you select the working point carefully at zero temperature coefficient. However, if you remember the formula for zero tempco ($V_{GS} = V_P + 0.63V$), you realize that the condition is different from unit to unit, since V_P is different.

A better solution is to use a differential amplifier, where the drifts of two matched JFETs tend to cancel each other. The configuration is shown in *Fig. 12a*. If R_0 is large enough, then:

$$I_{D1} + I_{D2} = I_0.$$

Further, if I_{D1} changes ΔI_{D1}, then I_{D2} also changes by the same amount, but in the opposite direction, i.e.,

$$\Delta I_{D1} = -\Delta I_{D2}.$$

The differential gain of the stage is:

$$A_V(DD) = (V_{D1} - V_{D2})/(V_{GS1} - V_{GS2}) \\ = R_D \times g_m,$$

which is the same as the gain of a single common-source stage. For R_0 to be very large, $-V_S$ must also be very large. This is usually inconvenient, so instead of a resistor, you use a so-called constant-current source, which delivers I_0 independent of $-V_S$ (*Fig. 12b*).

Due to its symmetrical nature, you can also consider the differential amplifier as two symmetrically arranged "half-circuits," each with a JFET, a load resistor, and half of a current source, providing $I_0/2$.[1] This is shown in *Fig. 13*. If the two JFETs are "identical," then you can join the two half-circuits together at the sources without upsetting the DC operation. However, you now have balanced single-ended amplifiers.[2]

Seen from gate 1, JFET 1 operates as a common-source amplifier, except that the source is connected to the source of JFET 2, operating it with source input. Seen from gate 2, the same thing happens—JFET 2 is in common-source mode, driving JFET 1 in the source. There are a number of advantages to operating two JFETs in this way, and I will start here with the common-mode rejection.

Common-Mode Signals

A very important feature of the differential amplifier is its ability to reject common-mode signals. Common mode means that both gates are driven with the same polarity and equal amplitude signals. It is easy to see that if only gate 1 is driven positive, then I_{D1} increases and I_{D2} decreases. But if both gates are driven positive, then both I_{D1} *and* I_{D2} must increase, which is impossible because $I_{D1} + I_{D2} = I_0$; i.e., I_0 is constant. Consequently, the differential amplifier cannot amplify same-polarity or common-mode signals.

Just how good it is in rejecting common-mode signals is expressed with the common-mode gain:

$$A_V(CM) = -R_D/2r_0,$$

where r_0 is the output impedance of the constant-current source. In order to have low common-mode gain (i.e., good rejection), the output impedance of the current source must be very large.

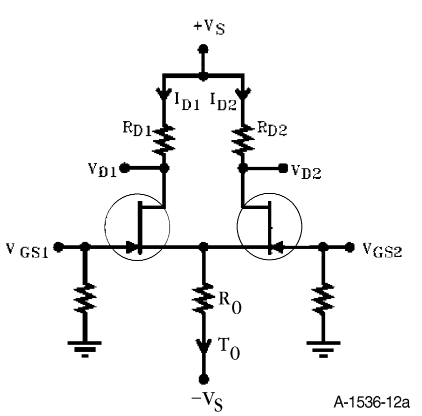


FIGURE 12A: Basic differential amplifier with JFETs.

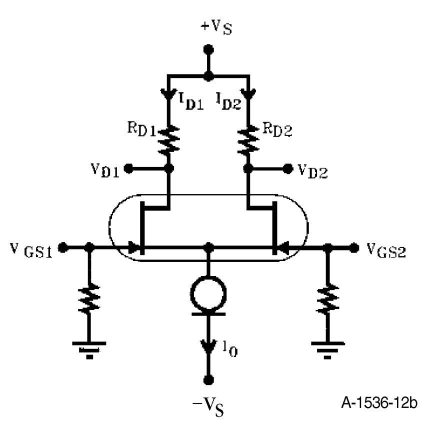


FIGURE 12B: Improved differential amplifier with constant-current source.

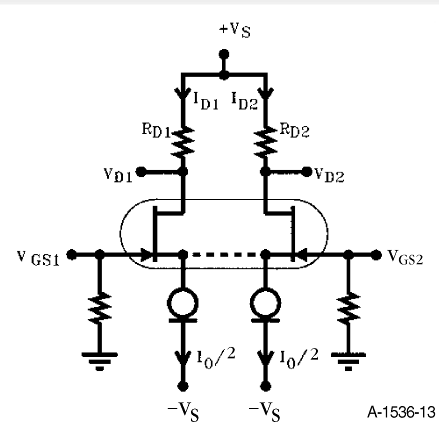


FIGURE 13: The differential amplifier represented with two symmetrically arranged "half-circuits."

The importance of low common-mode gain is closely related to the temperature drift, because changes in I_D, V_{GS}, and g_m can be considered common-mode signals if they are the same for both JFETs. Normally, a common-mode rejection ratio (CMRR) is specified for the differential amplifier. It is the ratio between the differential gain and the common-mode gain:

$$CMRR = A_V(DD)/A_V(CM) \approx 2g_m \times r_0.$$

Obviously the two JFETs must be closely matched to achieve good common-mode rejection. In fact, these two formulas are valid only if the two JFETs are perfectly matched. Although it is possible to select well-matched JFETs, the easier way is to use dual JFETs matched by the manufacturer, or even better, dual JFETs manufactured on the same silicon chip, i.e., monolithic duals.

I have been using the NPD 5566 dual N-channel and the AH 5020CJ dual P-channel JFETs.[3] However, these are not truly complementary types, as I pointed out in my article.[3] The first complementary types on the market were the 2SK240/2SJ74 medium g_m and the 2SK146/2SJ73 high g_m/low-noise types. These are closely matched single devices, mounted in a common aluminum case for good thermal tracking. Unfortunately, these devices are no longer in production.

Dual Monolithic JFETs

Although there are plenty of N-channel dual JFETs on the market, complementary dual monolithic JFETs are rare. In fact, I know of only one family, the 2SK389/2SJ109, made by Toshiba. These are still manufactured and available, so I use them in all my amps with differential input. Now I'll describe some practical differential circuits.

Figure 14a shows a simple differential amplifier with the 2SK389 dual monolithic JFET from I_{DSS} group V. I hooked it up with ±36V to operate the JFETs under conditions similar to those of the SE ones. The constant-current source is a J511 JFET delivering 4.7mA. In order to run the drains at roughly one-half the supply voltage (about 18V), I chose $R_{D1} = R_{D2}$ =10k.

First I tested the amplifier in single-ended mode, i.e., gate 2 connected to ground, with the measurements taken at V_{D2}. (V_{D2} has the same phase as V_{GS1}.) Although from the operational point of view it is single-ended, I think this mode is more appropriately called the unbalanced mode. Gain without local feedback ($R_{S1} = R_{S2} = 0$) is about 64 times, which is 36dB. Frequency response is 175kHz, and the input capacitance is 330pF. THD, measured at 1kHz, is shown in column 1 of *Table 2*.

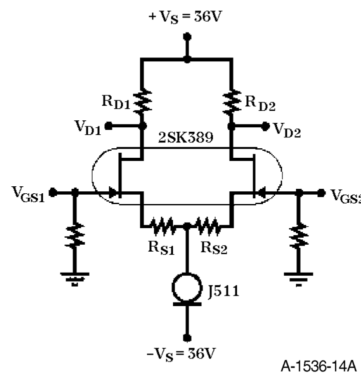


FIGURE 14A: A practical differential amplifier with the 2SK389V dual monolithic JFET.

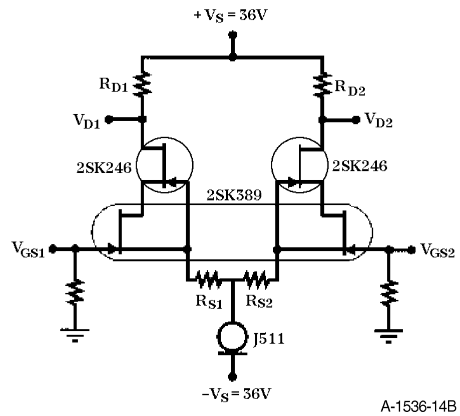


FIGURE 14B: The cascode connection reduces the input capacitance.

Next I inserted source resistors $R_{S1} = R_{S2}$ = 100R, and reran the measurements. Due to the local feedback, the gain dropped to approximately 28 times, and the input capacitance to 160pF. The THD also decreased by about 6dB.

In order to reduce the input capacitance further, I put 2SK246 cascodes in the circuit (*Fig. 14b*). The gain did not change significantly, but the input capacitance dropped to 50pF! THD also decreased, as shown in column 2, *Table 2*.

TABLE 2

Output in V RMS	Column 1 SE mode, $R_{S1} = R_{S2} = 0$	Column 2 SE/cascode mode, $R_{S1} = R_{S2} = 100R$	Column 3 Balanced mode, $R_{S1} = R_{S2} = 100R$
0.3	0.013%	0.006% (noise)	
1	0.035%	0.012%	
3	0.27%	0.1%	0.023%
5	0.85%	0.33%	0.06%
8	2.6%	1%	0.12%
10	4.7%	2.3%	0.17%

Balanced Mode

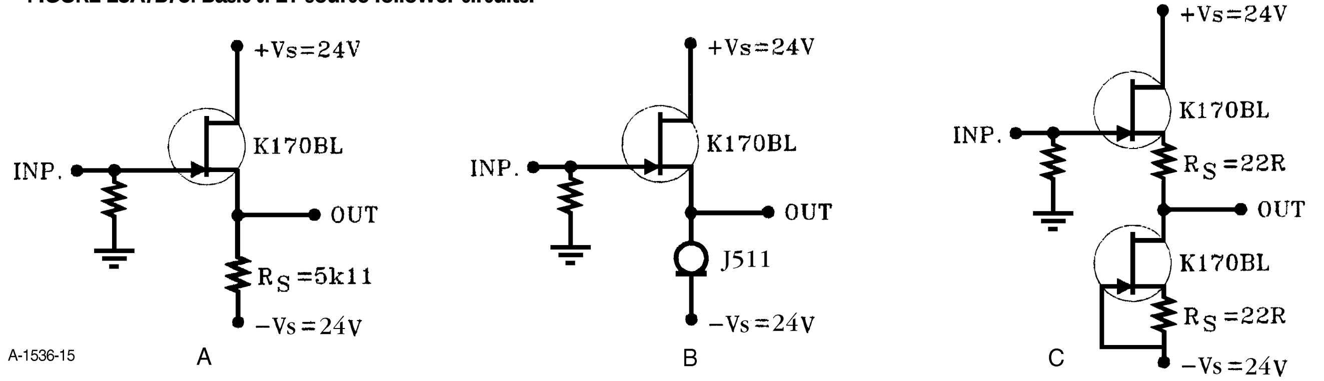
A couple of comments are in order concerning this circuit. According to the measurements, it is a very decent design, considering that it uses only a very small amount of local feedback. The gain is still fairly high, and you can reduce it further by increasing the source resistors, which in turn further reduces the THD. However, to fully take advantage of the symmetrical nature of this circuit, you should use it in balanced mode, which requires applying a balanced signal at the two gates and taking the balanced signal from the two drains.

I have made some rudimentary THD measurements in balanced mode, shown in column 3 of *Table 2*. Unfortunately, my oscillator and THD analyzer (HP339A) are unbalanced, so I needed to improvise the balanced operation with op amps, limiting the measurements to the levels shown in the table (lower levels were masked by noise). Nevertheless, it clearly indicates that the circuit thrives in balanced mode, having 10–20dB less THD compared to unbalanced mode. It also indicates the advantages of this circuit relative to the SE circuits discussed in Part 1.

The SE purists might naturally say that this is due to the cancellation of even-order harmonics in the balanced circuits, which is true. But as Nelson Pass points out on his homepage, in comparison to the SE stages, the balanced circuit does not give rise to odd-order distortion. There is simply not much distortion left in the balanced circuit.

As mentioned in Part 1, the input capacitance is voltage-dependent, which can cause THD when the amplifier is driven from high source impedances. I have tested the circuit described in column 2 of *Table 2* with 50k, 100k, and 500k sources. There was no measurable change in distortion up to 100k, but at

FIGURE 15A/B/C: Basic JFET source-follower circuits.



500k, I could see a slight increase.

Again, for noise reasons, you should probably keep the source impedance below 50k, so there is no problem with the capacitance modulation anyway. I also checked the CMRR by connecting the two gates together and driving them with a 3V RMS signal. The output, again in balanced mode, was down 87dB at 1kHz. The CMRR dropped to 70dB at 10kHz and 63.5dB at 20kHz, but even at 100kHz, it was 50dB!

The Output

I have now described two types of amplifier stages using JFETs, the common-source or single-ended stage, and the differential or balanced amplifier. You can use either of these to build audio amplifiers, depending on your preference for balanced or unbalanced operation. Personally, I prefer the differential circuit, because you can use it with balanced or unbalanced sources, and it can also feed balanced or unbalanced power amplifiers. Balanced operation gives a subjective impression of increased dynamics. It can also be an extremely useful interfacing consideration in breaking up ground loops.[4]

There are two issues to consider when talking about the SE and balanced amplifiers. First of all, the output does not sit at 0V DC, but at some 10–20V above ground. If you wish to connect it to, say, a DC-coupled power amplifier, you must block this DC voltage from reaching the power-amp input. This is easily done using a capacitor, and this problem is well known to all SE fans, whether of tube or semiconductor variety. I will therefore not spend much time on the subject.

A much more important question is whether these circuits can drive the input impedance of a power amplifier. The output impedance of the amps ex-

TABLE 3

Output V RMS	Column 1 R_S = 5.11k	Column 2 R_S = constant-current source	Column 3 R_S = JFET current source	Column 4 The “Borbely” source follower
0.3V	0.0025	0.0023	0.002	0.0025
1V	0.0033	0.0024	0.0018	0.0035
3V	0.011	0.0025	0.0016	0.0045
5V	0.02	0.003	0.0016	0.0074

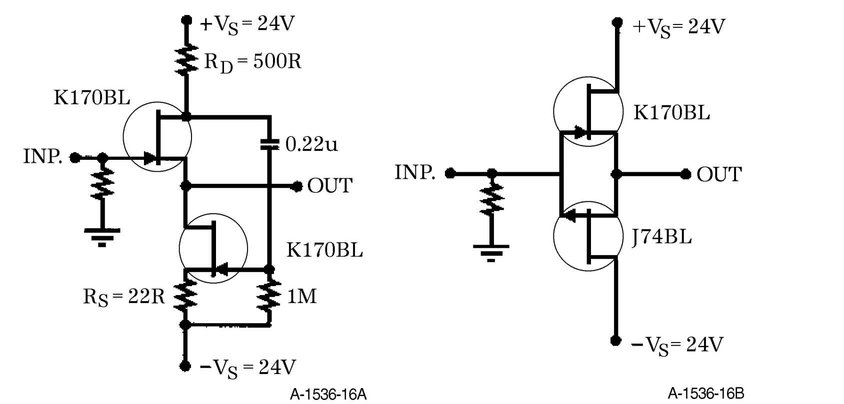


FIGURE 16A/B: These source-follower circuits can drive low-impedance loads with very low THD.

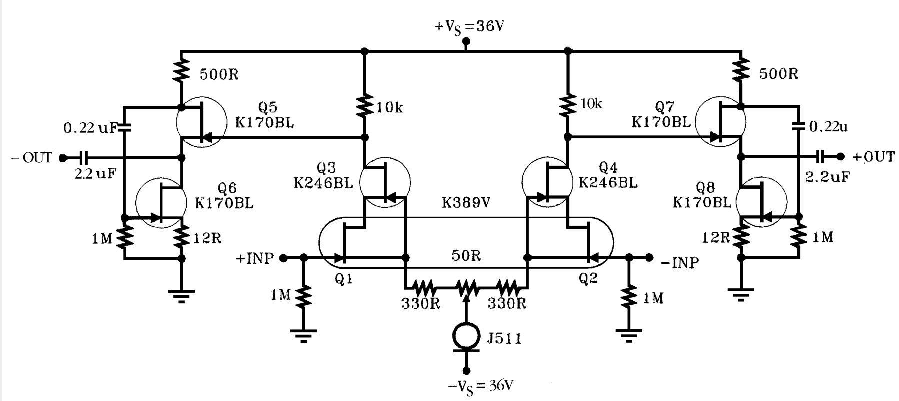


FIGURE 17: The all-JFET balanced SE line amp.

amined is basically equal to the drain resistor of the amplifier. If R_D = 10k, then the output impedance is also close to 10k. But if the input impedance of the power amp is also 10k, then you are certainly in trouble. First, you lose 6dB of gain by the voltage division between the 10k resistors; second, the 10k input will most likely load the output and cause a lot of THD. Even with 20k or 50k input impedance, you might run into problems. It is advisable to put an impedance transformer at the output to avoid this. Source followers to the rescue!

JFETs as Followers

Just like tubes and bipolar transistors, JFETs can also be operated as followers, more specifically source followers. The basic circuit is shown in *Fig. 15a*. The drain is AC-grounded, and the output signal is taken out across the source resistor, which means it operates with 100% local feedback. The gain of the source follower is:

$$A_V = g_m \times R_S/(1 + g_m \times R_S)$$

Two things become obvious from the formula: first, the source follower does not reverse the phase of the signal, and second, if $g_m \times R_S >> 1$, then the gain becomes approximately unity. In order to make R_S large, you can use a constant-current source with high output impedance (*Fig. 15b*). The linearity is also dependent on R_S (see the 1kHz THD measurements in columns 1 and 2 of *Table 3)*.

The input capacitance is low because it is not augmented by the Miller effect. I measured approximately 5pF for the circuits in *Fig. 15a* and *b*. The output impedance equals approximately $1/g_m$. With high-g_m devices, this will be fairly low. I measured 38Ω for the basic circuits in *Fig. 15a* and *b*.

The circuits in *15a* and *b* have a DC offset voltage at the output—the gate-source voltage at the given drain current. For the JFETs I used in the test setup, I measured a 0.2V offset. If you need zero DC output, you can use the circuit in *Fig. 15c*. Here the constant-current source is made with the same type of JFET as the follower.

If the two JFETs are matched and the two source resistors are equal, then the DC offset will be very small. With two matched K170BLs, I measured less than 1mV offset. (It would probably be even lower if you used here a dual monolithic JFET like the K389BL/V.) DC drifts tend to cancel out as well, because of the matched devices. The circuit also has very low THD (see column 3 of *Table 3*).

Follower Feature

One of the most important features of a follower is its ability to drive low-impedance loads. I checked all three circuits with 1k and 10k loads at 3V RMS output. With 1k they measured 1%, 1.7%, and 0.23%, respectively. Although its output impedance is actually higher than the circuits in *Figs. 15a* and *b*, that in *15c* is better in driving low-impedance loads. With a 10k load, the circuits in *15b* and *15c* didn't have many problems. THD was 0.004 and 0.0022%.

My choices of source followers are shown in *Fig. 16*. The circuit in *16a* is a JFET version of the tube White cathode follower.[5] Basically, the circuit is an extension of *Fig. 15c*, in that the follower is fed with a constant-current source, but in addition the drain current of the current source is modulated by the AC signal. When the output signal goes positive, the tail current decreases, and when it goes negative, the current increases. The result is a significant reduction of the output impedance and an apparent increase in drive capability.

The output impedance with the devices shown measured 2.3Ω. The input capacitance is about 5pF, the same as the previous source-follower circuits. The penalty for the drive capability is a slight increase of distortion (see column 4 in *Table 3*). The THD with a 1k load and 3V RMS is 0.0095%. The necessary gate drive voltage is derived from a small resistor in the source follower's drain circuit, and it is AC-coupled to the current source.

Power Dissipation

John Curl used the complementary JFET source follower shown in *Fig. 16b* in the JC-2 phono-preamp module.[6] The JFETs work in Class A as long as the peak load current is less than twice the bias current. After that, the circuit works in Class AB. I usually let the two matched devices work at I_{DSS}, to maintain as much Class A headroom as possible. However, you must watch the power dissipation. If I_{DSS} is such that the power dissipation is more than the maximum allowed, then you need to insert a source resistor to reduce the drain current or select a device with lower I_{DSS}.

I tested the circuit with K170/J74, both the BL and V types, and got excellent results. The THD is shown in column 4 of *Table 3*. Depending on the matching, the offset can be as low as 1mV. The output impedance is about 18Ω, and the input capacitance is 28pF. Most important, the circuit can drive low-impedance loads without distress—the 1k/3V RMS THD was 0.0078%. With a 10k load, there is no difference from the no-load results.

I also tested the source followers for THD caused by the voltage-dependence of the input capacitance. Since the voltage excursion is much larger at the input because of the unity gain, the circuits are also more susceptible to the distortion. There is no significant increase up to a 10k source; however, at 50k the THD is increasing by an order of magnitude. Normally this is no problem, because the source impedance is usually very low. However, in certain applications such as filters, this can cause distortion.

Of course, using any of these source-follower buffer circuits with the SE and differential amplifiers discussed previously solves only one of the problems stated at the start of this section—the drive-capability problem. The DC voltage is still there. Given the topology of these circuits, you must use a capacitor at the output to block the DC voltage. Naturally, you can also solve this problem by using level-shifting circuits, but it requires a bit more circuit design. For now, I'll look at an all-JFET balanced/SE all-FET line amp, using the circuits already developed.

The Balanced/SE All-JFET Line Amp

The schematic shown in *Fig. 17* consists of the differential amplifier Q1/Q2, cascoded with Q3/Q4, and the output buffers Q5/Q6 and Q7/Q8. The differential amplifier uses a dual monolithic K389V JFET. Each JFET operates at just over 2mA, this current supplied by the J511 constant-current source. The 330R source resistors provide local feedback and control the gain of the differential amplifier. The trimpot P1 cancels out small imbalances between the two JFETs, but it is normally unnecessary with monolithic duals, and you can leave it out.

The Cascode FETs are K246BLs. The output buffers are those I developed from the tube White Cathode Follower, shown in *Fig. 16a* ("modestly" called the "Borbely" source followers here). The supply voltage is ±36V. Of course, you can make the negative supply much less than 36V; the constant-current source requires only a couple of volts for proper operation. I made them both 36V to be able to try other configurations.

The output caps must be of highest quality in order to preserve the outstanding sound quality of this simple circuit. If you are likely to drive loads down to 1k, then the caps must be a minimum of 10μF. If you are driving normal 10k or higher loads, you can get away with a 1μF or 2.2μF cap. I tried the Hovland Musicaps, which are rather neutral, but there are plenty of good caps on the market you can try.

Normal oil caps are *not* for this circuit; they destroy the excellent resolution to a "nice" blurred mish-mash. (Don't get the idea that I don't like oil caps; I use them in some of my amps.) I would have liked to try some silver-foil caps, but, alas, the prices are more ridiculous than the cable prices, and I refuse to play that game. (If anyone knows of a reasonably priced silver-foil cap, please let me know.)

You can use the line amp with unbalanced or balanced sources, and you can feed power amps with balanced or unbalanced inputs. However, you should really take advantage of its superior performance in balanced operation, as I mentioned before. Should you use it with unbalanced sources, then you must short the –INP to ground. And in the unlikely event that you don't wish to take advantage of the balanced outputs, you can leave out the circuit around Q5/Q6, i.e., the negative output. I recommend a 10k or 20k ladder attenuator as a volume control. Good luck with the JFETs, the "New Frontiers" in audio amplification. ■

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Acknowledgments

Many thanks to Walt Jung of Analog Devices for his valuable comments and suggestions.

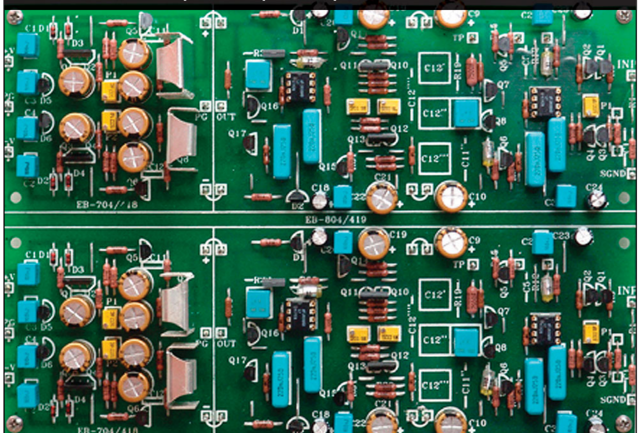
Starter Kits Series: ALL-JFET MM/MC Phono Preamp, Part 1

By Erno Borbely

This distinguished veteran of the audio design wars presents an interesting tutorial on phono preamps-followed in Part 2 by details of his latest high-quality, low-cost unit.

In the article "A Moving Coil Preamp, Part 1"[1] I covered basic noise theory and noise in audio amplifiers in detail. Please read that article, because I will be giving you only a simplified description here. The article has a detailed treatment of the noise issues in phono preamps.

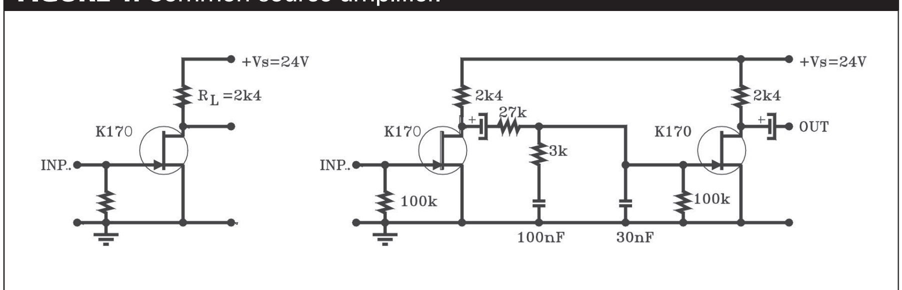
PHOTO 1: The phono preamp.



The majority of pickups fall into two categories: moving magnet (MM) and moving coil (MC). MM pickups have a typical output of 5mV, while the MCs feature 0.5mV at 5cm/sec lateral velocity. There are MCs with lower output. (I believe the old ORTOFON MC2000 was one of the lowest at 0.05mV!) The high-output MCs are close to the MMs in terms of output, and you can use the normal MM input for these. MMs usually need a gain of 30–40dB and normal MCs 20–30dB more.

Of course, the extra gain needed for the MCs can be supplied with a transformer that has the proper turns ratio. There are a number of suitable transformers on the market, ranging from about $60 to >$1000[2]. The transformer has the advantage of providing the gain without adding noise to the signal. However, the transformers also have disadvantages, mainly in terms of distortion and limited frequency range. I am not a transformer designer; consequently, I am always using active devices also for MCs.

FIGURE 1: Common source amplifier.



The major challenge in designing phono preamps is the dynamic range, which is the difference between the lower limit—i.e., noise floor of the amplifier—and the upper limit, which is the clipping level of the amplifier. In a well-designed amplifier the noise is determined by the noise of the input stage and is usually specified as the equivalent short circuit input noise.

If you want this noise to be, say, 80dB below the normal output of 5mV of an MM pickup, then the equivalent short input noise must be 5mV/10000 = 0.5µV. This is usually easy to achieve, even with tubes. However, in the case of an MC pickup with 0.5mV output, the equivalent short circuit input noise must be 0.5mV/10000 = 0.05µV (50nV!). This is much more difficult to achieve, as we will see later.

Of course, it's not just the amplifier that generates noise; all resistors are doing the same. In fact, the pickup itself can be considered a resistor that generates its own noise. The noise voltage generated by a resistor is given by the formula:

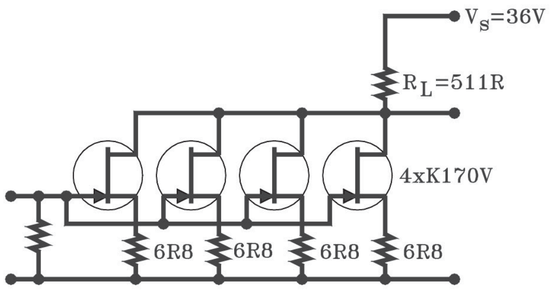
$$e_n = \sqrt{(4kTR\Delta f)}$$

where:

e_n is the RMS noise voltage in volts
k is the Boltzmann constant (1.38×10^{-23} joule/degree K)
T is the room temperature in kelvin (300°K)
Δf is the noise bandwidth in Hz (20000Hz)
R is the resistance in ohms.

If you have an MC pickup with an ohmic resistance of, say, 50Ω (I have chosen the 50Ω arbitrarily, MCs come with 10–50Ω), then this pickup will generate a noise voltage of 0.128µV, or 128nV, in the audio frequency range. So you see that both the source—i.e., the pickup—and the amplifier contribute to the overall noise you hear.

FIGURE 2: Paralleling JFETs reduces the input noise.



Assume that you connect this 50Ω pickup to an amplifier that has an equivalent short circuit input noise of 125nV. The total input noise will be $\sqrt{(128^2 + 125^2)}$ = 179nV, and the signal-to-noise ratio is 0.5mV/179nV = 2793x, which is 69dB. There is nothing you can do about the first term, the noise generated by the pickup, but you can certainly do something about the noise of the amplifier by designing a lower noise input stage! The lower its noise, the less it contributes to the overall noise.

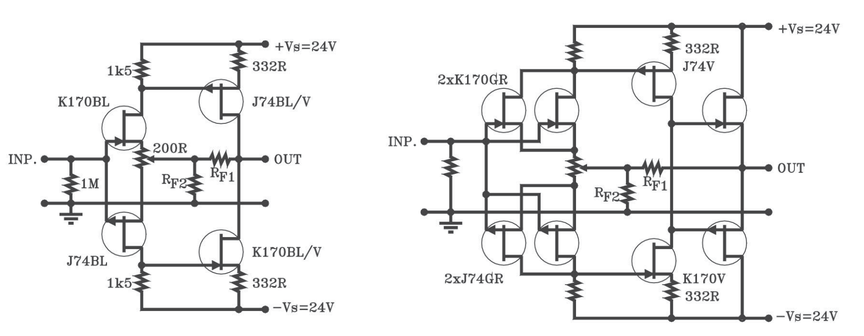
If you look carefully you notice that the pickup and the amp contribute about the same amount of noise in the above example—i.e., the amp behaves as though it was also a 50Ω resistor. You can, in fact, characterize the noise behavior of the amplifier with its equivalent noise resistance (Req.). As you will see, this is convenient for calculating the noise of the phono preamp, because the active devices—i.e., the tubes, transistors, and JFETs—can all be described by their equivalent noise resistance. Of course, you (or at least me!) are only interested in how the JFETs behave in this respect, but for the sake of peace with the tube and bipolar fans, I will also mention their devices in comparison!

A SIMPLE MM PHONO PREAMP

The basic amplifier that you can use for amplifying low noise signal is the common source amplifier, which I described in detail in Part 1 of the JFET articles[3]. *Figure 1A* shows the simple JFET stage without the source resistor, because I want to show what the source resistor does in terms of noise. The K170 JFET has an equivalent noise of 0.95nV/√Hz, which over the audio frequency range is 134nV. This is the same as the noise of a 55Ω resistor. For comparison, the discontinued K146 JFET was specified at 0.75nV/√Hz, meaning an equivalent noise resistance of 34Ω. The ROHM bipolar transistor 2SD786 had 0.6nV/√Hz, which is 21.3Ω.

Triode tubes can also be represented by their equivalent noise resistance, which is ≈2.5/g_m, where g_m is the transconductance of the tube in mA/V at the operating point. The ECC88/6DJ8 has a transconductance of 12.5mA/V at 15mA anode current and the Req = 2.5/12.5ma/V = 200Ω. The 5842 and the WE417A tubes can be operated with a g_m = 25mA/V, which will result in Req = 100Ω. The 3A/167M, the WE437A, and the Russian 6C45π can produce, with approximately 40mA anode current, >40mA/V, thus bringing Req down to about 55Ω. This is very close to what the K170 JFET does.

FIGURE 3: The basic four-JFET circuit and its upgraded version for low-noise application.



Coming back to the common source amplifier of *Fig. 1A*, the equivalent short circuit input noise of this stage is 134nV. A normal MM pickup has an equivalent ohmic resistance of about 600Ω, which generates a noise voltage of 445nV. In this case the noise of the pickup dominates and the amp noise has little influence on the overall signal-to-noise ratio.

Of course, the picture is different if you want to use this simple stage with MC pickups. Assuming again an MC with 50Ω resistance, the self-generated noise is 128nV. The amp is generating 134nV. Total noise is $\sqrt{(128^2 + 134^2)}$= 185nV.

The signal-to-noise ratio with a normal MC pickup will now be: 0.5mV/185nV = 2700x, which is only 68dB. The RIAA characteristic also helps a few dB due to its HF rolloff. But being a low-noise freak, I wouldn't consider this adequate for MC preamps.

A customer sent me the schematic of a simple two-stage phono preamp, which he found on the Internet (*Fig. 1B*). It uses two identical stages with K170, both operated at Idss = 5mA, with passive RIAA equalization between the stages. Each stage has a gain of g_m × Rd, where g_m is ≈30mA/V at Idss = 5mA and Rd = 2k4. The gain per stage is about 70×, or 37dB, which means a total gain of 74dB.

The insertion loss of the passive equalization network, due to the 100k gate resistor of the second stage, is about 22dB at 1kHz, so the effective gain is about 52dB. This is a bit too much for an MM, but marginal for an MC. I have also seen a version with Rd = 5k1, which gives you a few dB more gain.

I hooked up the circuits in *Fig. 1B* to check out the performance. First, I measured the noise and distortion in the input stage only. The input noise was 137nV over a 30kHz bandwidth, so the equivalent noise resistance was actually less than the data sheet value of 55Ω. The distortion was dominated by second harmonics: it was 0.88% at 1V RMS/1kHz and 2.5% at 3V/1kHz.

This is to be expected of a single-stage K170 without a source resistor. You could reduce this by using a source resistor as I have shown[3], but since the source resistor is added in series to the equivalent noise resistance, it will increase the noise of the input stage[1]. A 50Ω source resistor would reduce the THD by almost 6dB, but the equivalent noise resistance would increase to 55 + 50 = 105Ω, which would result in an input noise of 186.5nV. The signal-to-noise ratio with a 50Ω MC pickup would be reduced to 66.8dB.

Next I hooked up the whole circuit as shown in *Fig. 1B*. I measured a total gain of 50dB at 1kHz. Overall distortion was reaching 1% at 1.4V RMS/1kHz. RIAA showed a 0.5dB rise at 10kHz, but was flat at the low end.

Thanks to the dominating second harmonics, this phono stage is probably very warm sounding. Indeed, a customer reported that this amp sounded very good with a DENON DL103 MC pickup in spite of the low gain, the second harmonic distortion, and the relatively high equivalent short circuit input noise.

DEVELOPING THE INPUT STAGE FOR THE STARTER KIT

Let's look at the input noise first. Just as paralleling resistors reduces the value of the resistors, the paralleling of JFETs reduces the equivalent noise resistance. Paralleling JFETs with comparable g_m reduces the resistance to half and the noise by approximately 3dB.

I tested the circuit shown in *Fig. 2*, which I published in reference 3. Theoretically, four JFETs in parallel should have 6dB less noise than one JFET. However, you normally use a small source resistor with each to get better matching between the devices and/or to control the drain current, as shown in *Fig. 2*, so the equivalent noise resistance of each is 61.8Ω. The four in parallel are then 15.45Ω, and the input noise drops to about 72nV.

I believe Nelson Pass is using this input stage in his DIY phono stage, albeit with 22Ω source resistors, which increases the equivalent noise resistance and the input noise a bit. Note that the input capacitance also increases 4×, but it is not important here due to the low source impedances in phono preamps.

Paralleling the JFETs solves the noise problem, but does very little for linearity. This four-JFET circuit is a bit better than the single JFET in *Fig. 1A* due to the higher supply voltage: it's 0.32% at 1V RMS/1kHz and reaches 1% at 3V RMS/1kHz.

If you remember my priorities from the first part of this series (March '05 *aX*)4 they are low noise, inherent linearity, and DC coupling. This circuit fails on the second and third of these priorities. I developed a line amp, based on complementary differential circuitry. I mentioned that I use that topology in most, if not all, of my amplifiers.

The "if not all" means that I also use the single-ended complementary topology in some of my amps. In fact, I use both in this two-amp phono preamp, but the single-ended topology offers lower input noise than the complementary differential circuitry, so I use that for the input stage. The complementary differential circuit is used for the second stage.

Figure 3A shows the four-JFET single-ended complementary circuit that was published in reference 5. Its linearity is orders of magnitude better than the single JFET circuit. O.L. THD is 0.032% at 1V RMS/1kHz and reaches only 0.4% at 10V RMS/1kHz!

This topology works fine for an MM-only input stage, but it requires a circuit upgrade for real low noise operation. *Figure 3B* shows the low-noise topology. I have doubled the input stage, making it equivalent to the four-JFET circuit shown in *Fig. 2* in terms of noise resistance.

However, as you will see later, the feedback resistor(s) also contribute to the equivalent noise resistance, so that must be made as small as possible as well. Consequently, I added a complementary source follower to the output to be able to use low value feedback resistors. Granted, this is no longer a one-JFET circuit, and you will find very little of the second harmonic distortion in it. However, you will find low noise, good inherent linearity, and you can also use it without capacitors in the signal path. With the background material presented here, Part 2 will look at the complete phono preamp.

Erno Borbely *received an MSc degree in Electronic Engineering from the Norwegian Institute of Technology, University of Trondheim. In the 60s he worked as a design engineer for the Norwegian Broadcasting Corp., developing studio equipment, both in tube and in semiconductor technology. In 1969 he moved to the US and worked for David Hafler at Dynaco. Mr. Borbely was named Director of Engineering for Dynaco in 1972. He developed the cascode output stage for the Stereo 400 power amplifier, and the Dynatune circuit for the FM5 FM tuner, for which he received a US patent.*

In the (70s he worked at Motorola Semiconductor, developing power amplifiers, low-noise preamplifiers, and FM tuners. Some of this work was put to good use when he joined the David Hafler Company in 1978 as Director of Engineering. Mr. Borbely developed the DH101 preamplifier and the DH200 power amplifier, which was the first to use MOSFETs in the US.

From 1980 until 1997 he worked for National Semiconductor in Germany as its Technical Training Manager. In 1984 Mr. Borbely and his wife Irene started Borbely Audio, which sells high-quality kits to end-users. Since 1982 he has been publishing his designs in Audio Amateur publications. In the last several years he has developed a line of all-FET audio amplifiers. Besides the kit business, he is designing audio products for OEM customers.

Mr. Borbely's web page address is www.borbelyaudio.com and his e-mail address is erno@borbelyaudio.com.

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4. Erno Borbely, "Starter Kits, Part I: EB-604/410 All-JFET Line Amp," *audioXpress*, March 2005, p. 9.

5. Erno Borbely, "The All-FET Line Amp" *audioXpress*, May 2002, p. 28.

Starter Kits Series: All-JFET MM/MC Phono Preamp, Part 2

By Erno Borbely

This flexible phono preamp design benefits from the author's 40 years of experience in the business.

The EB-804/419 is a high quality, low-cost all-JFET DC-coupled MM/MC phono preamp with on-board all-FET regulators. The combination of low input noise and high gain allows you to use it with practically all pickups on the market. The EB-804/419 uses a two-amplifier approach instead of the usual single amplifier. This allows you to optimize it for low noise, RIAA accuracy, high gain, and dynamic range. The block schematic is shown in *Fig. 4*.

You can use the two-amp configuration in three different ways to realize the RIAA function: all active, all passive, or hybrid, meaning combination of passive and active. Two of these are supported in the EB-804/419: the hybrid and the all passive.

In *Fig. 4A* the amp on the left is a high-gain amplifier with extremely low input noise and flat frequency response. You can use it with 20 or 40dB gain by changing the feedback resistor. The R18/C11 network between the two amplifiers provides the 75µs passive rolloff. The second amp provides a further amplification of 24dB at 1kHz and at the same time boosts the bass according to the RIAA characteristics actively through R33/R32/R31/C14.

FIGURE 4: Block schematic for EB-804/419.

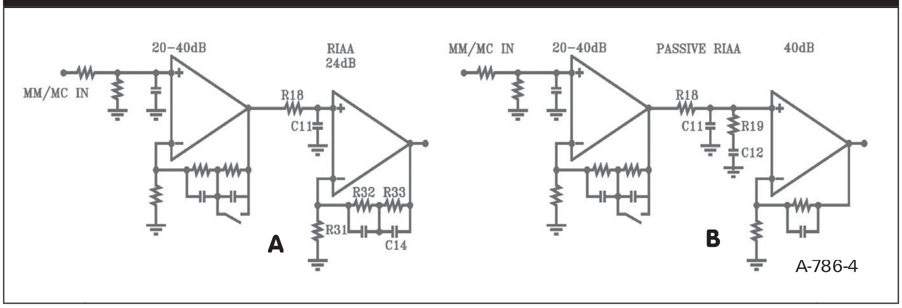
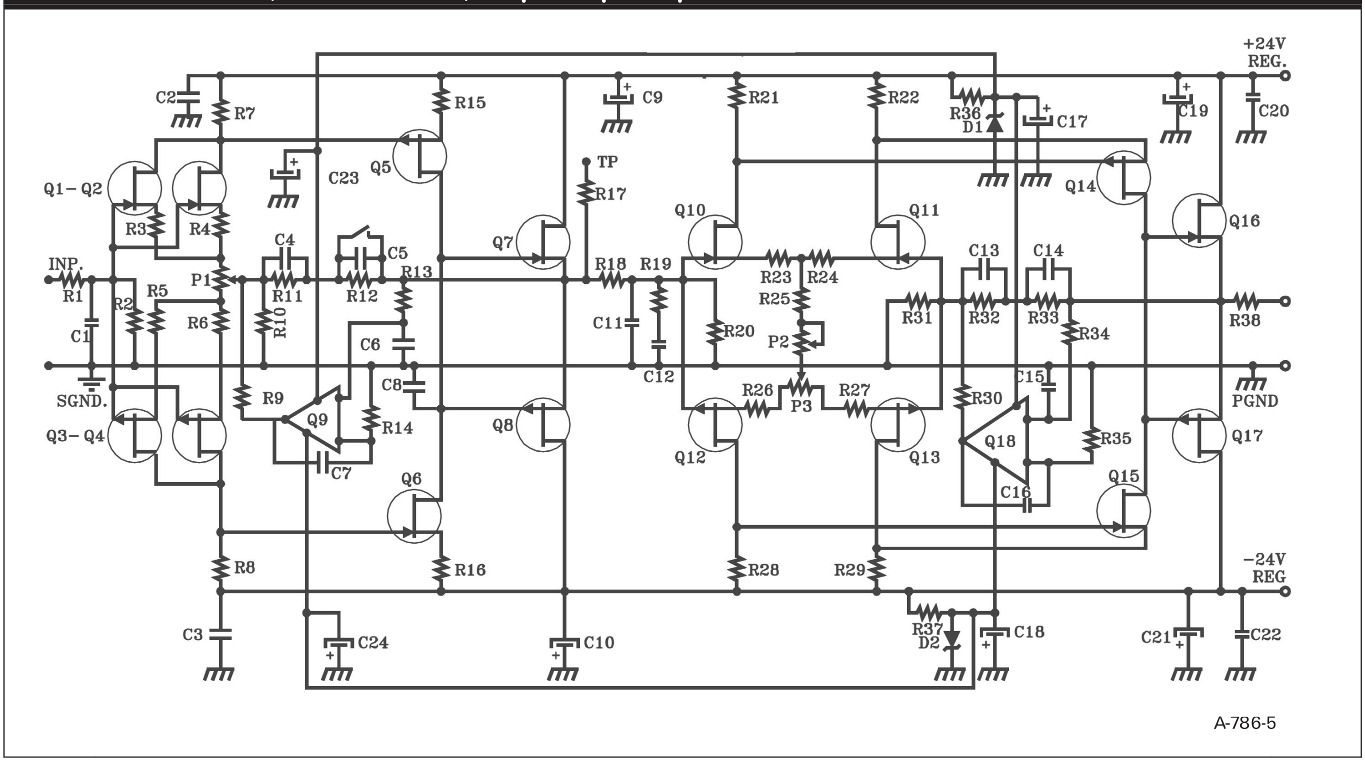


FIGURE 5: The EB-804/419 all-JFET MM/MC phono preamp.

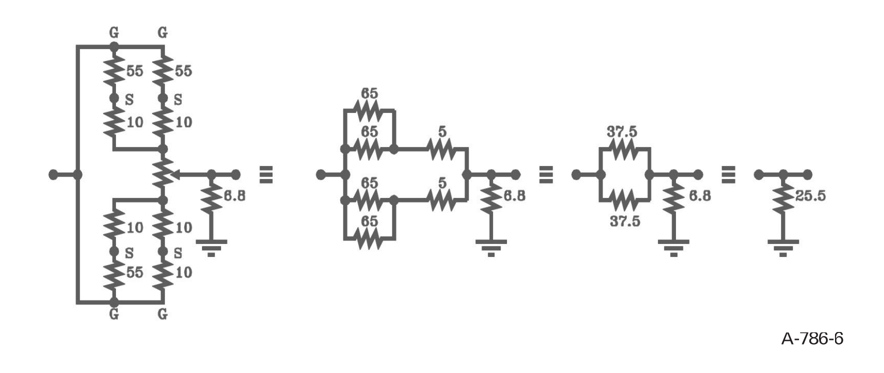


For those of you who want to experiment with all-passive equalization I have included the components for Walt Jung's network N2[6,7] in *Fig. 4B*. Here both amplifiers provide flat frequency response and all RIAA equalization is done passively between the two amplifiers by R18/C11/R19/C12. MM gain is 44dB and MC gain is 64dB for the circuit in *Fig. 4A*. The gain in *Fig. 4B* depends on the all-passive equalization network discussed later. The Starter kit is supplied with components for the circuit in *Fig. 4A*. You can order the all-passive network separately from Borbely Audio.

Figure 5 shows the schematic of the EB-804/419. The input stage is essentially the same as the one in *Fig. 3B*, but I have added source resistors to each of the input JFETs in case these are needed for matching and/or adjusting the drain current. Because the DC gain of the circuit can be very high (MC application and/or all-passive equalization), I use a servo circuit instead of capacitive coupling to keep the offset close to 0V.

I calculated the equivalent noise resistance of the input stage using the drawing in *Fig. 6*. The JFETs are represented by the 55Ω equivalent noise resistors in series with the 10Ω source resistors. These are connected to P1, which is a 10Ω trimpot. Finally, the center point of the trimpot goes to R10 = 6R8 feedback resistor.

FIGURE 6: Calculating the equivalent noise resistance of the input stage.



Because the JFETs are connected in parallel AC-wise, the equivalent noise resistance can be calculated to 25.5Ω. This will produce 92nV over the audio bandwidth. If you use this preamp with a 50Ω MC pickup, you will have a total noise of $\sqrt{128^2 + 92^2} = 157.6$nV, which results in a signal-to-noise ratio of 70dB referred to 0.5mV.

The 50Ω pickup itself contributes 128nV—i.e., if the amp were noiseless, the signal-to-noise ratio would be 71.8dB re 0.5mV. This means that the amp contributes only 1.8dB to the noise. You could decrease this by reducing/eliminating some of the noise-producing resistors in the input circuits.

If you select the input JFETs very closely, you could eliminate the 10Ω source resistors. The close matching could also eliminate the offset trimpot.

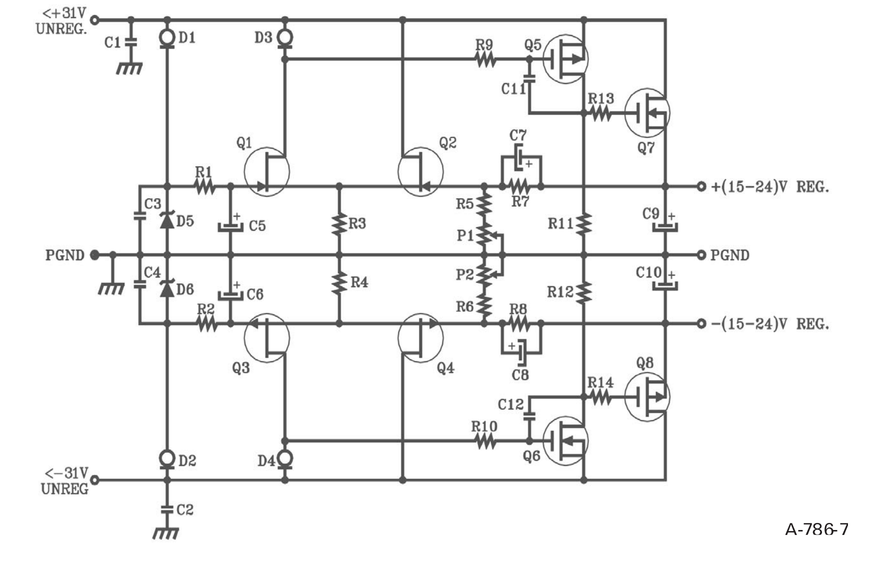
This would result in an equivalent noise resistance of 20.5Ω with a signal-to-noise ratio of 70.4dB, which means that the amp would contribute only about 0.4dB to the overall noise. For a 50Ω MC pickup this is probably the best you can achieve; however, for lower impedance/lower output MC pickups you will need to reduce the amp noise even further. (Calculate the signal-to-noise ratio for an MC pickup with 20Ω resistance!)

For MM the feedback resistor R10 is 75Ω, and the equivalent noise resistance is 93Ω. The equivalent short circuit input noise is 175nV. The total input noise with a 600Ω MM pickup is $\sqrt{(445^2 + 175^2)}$ = 478nV, and the signal-to-noise ratio relative to 5mV is 80dB.

The second stage of the phono preamp uses the same topology as the Starter kit EB-604/410[4]. It provides the active bass boost of the RIAA characteristic through its feedback network. However, due to its high DC gain (in active eq. mode the DC gain is 44dB!) a servo has been added. The servo amps are supplied with ±10V from shunt regulators.

The all-FET phono preamp is a very sensitive amplifier and is capable of picking up small signals not only through its inputs, but also through vibration. It is, therefore, very important to make the whole amp as "dead" as possible for vibration by using rubber stand-offs or mounting the whole board on a Teflon plate, 6-8mm thick. And a mumetal box will certainly put the icing on the cake.

FIGURE 7: EB-704/418 all-FET regulators.



TYPICAL SPECIFICATIONS

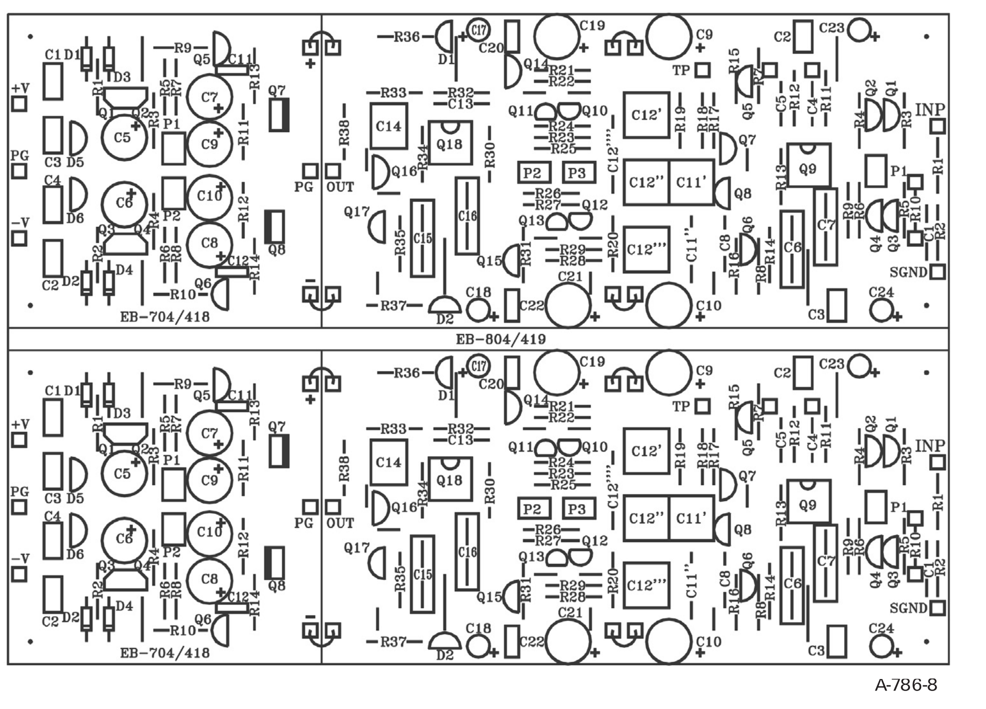
Gain: 1st stage:	20 or 40dB
THD, 20dB gain, 3V/1kHz:	0.0033%
3V/10kHz:	0.005%
Gain: 2nd stage:	24dB
THD, 3V/1kHz:	0.003%
3V/10kHz:	0.007%
Max. output voltage:	approx. 9V RMS
Risetime (without RIAA network):	0.3/0.4μs
Freq. Response:	RIAA ±0.2dB
Equivalent input noise:	<100nV, 20Hz-30kHz

THE EB-704/418 ALL-FET ON-BOARD REGULATOR

The schematic of the on-board EB-704/418 all-FET regulators is shown in *Fig. 7*. The topology is very similar to the EB-604/415 regulators used in the EB-604/410 line amp described in reference 4. However, I added Q7 and Q8 MOSFET source followers to be able to cope with the demand of the phono preamp (70mA). You can change output voltage between 15 and 25V with P1 and P2. If you use them with fixed 24V output, then you can short P1 and P2. Make R5 = R6 = 3k32.

Minimum input/output voltage differential is 3V, but 4V is recommended. Maximum unregulated input voltage is ±31V, and you must heatsink Q7/Q8. The output noise of the regulators is <5μV across the audio bandwidth. Recommended power supply is the EB-604/265. The stuffing guide for the EB-

FIGURE 8: Stuffing guide for MM/MC phono preamp EB-804/419 with EB-704/418 regulator.



804/419 phono preamp is shown in *Fig. 8*.

SETUP PROCEDURE

Start the assembly with the regulators, insert the solder pins and the jumpers, and then install the resistors, FETs, and finally the capacitors. Connect 332R/5W resistors between the +output and ground and the –output and ground of the regulators. Apply ±29V unregulated DC to the regulators. Check the output voltage across the load resistors; it should be 24V, ±5%.

Next, begin the second stage of the

RIAA EQUALIZATION

The RIAA characteristic is defined by three time constants: T1 = 3180µs, T2 = 318µs, and T3 = 75µs, corresponding to the crossover frequencies of 50, 500, and 2122Hz. A fourth time constant of 7950µs has been added by the IEC as an amendment to the RIAA characteristic; however, it has never been officially standardized by the RIAA. This time constant means a 6dB rolloff below 20Hz, and is meant to function as a rumble filter, reducing turntable/record related low-frequency disturbances. This feature is not included in the phono preamp described here.

The RIAA characteristic is shown in *Fig. A* with the asymptoms. It is specified in dB relative to 1kHz (0dB). It appears that the DC (or zero frequency) gain is 20dB (10×) higher than the gain at 1kHz. However, because of the interaction between the time constants, the actual DC gain is 9.898× the gain at 1kHz. This means that in an ideal RIAA preamp the 1kHz gain is always 0.101×DC gain (see reference 6: Walt Jung; *Op Amp Applications*, Chapter 6-1, Analog Devices, 2002, ISBN:0-916550-26-5), so specifying the gain at 1kHz also defines the gain for all other frequencies. ■

phono preamp (R20 and onwards) by installing all solder pins, jumpers, resistors, JFETs, and capacitors. Put in the supply jumpers between the regulators and the second stage and short the input of the second stage (jumper across R20). Do not install Q18 just yet. Connect the ±29V unregulated supply to the regulators. Measure the voltage drop across R22 (or R29) with a DVM and adjust it with P2 to 3V. Then check the offset voltage at the output and adjust it with P3 to 0V. Insert Q18 in the socket and recheck the offset; it should be less than a couple of mV after a few seconds.

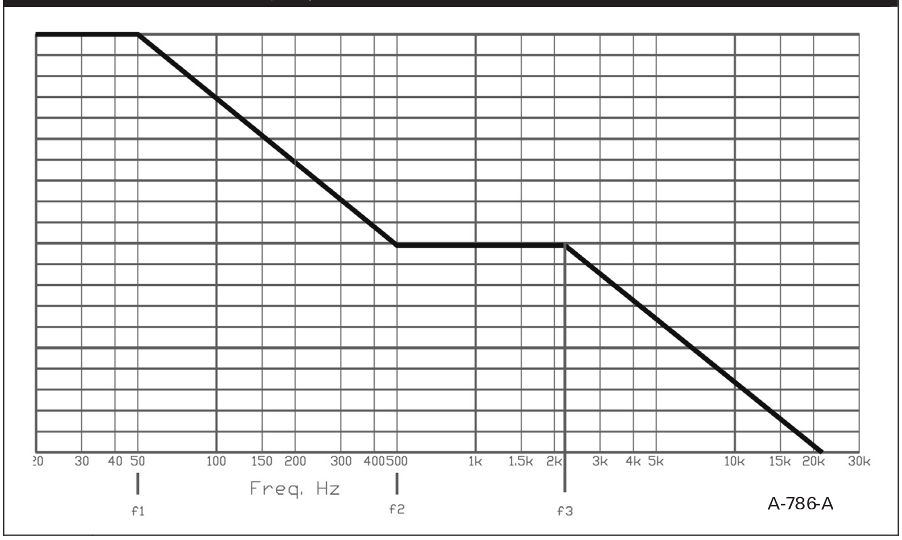
Finally, assemble the input stage, except the RIAA components between the two amps and Q9. Disconnect the supply jumpers between the regulators and the second stage and connect the regulators to the input stage with short wires. Short the +INP-pin of the input stage to SGND and check the voltage across R15 (or R16), which should be between 2.8V and 3.3V. Check the offset voltage of the input stage at the test-point TP and adjust it to zero with P1 (do this very carefully because the DC gain is very high). Install Q9 in the socket and recheck the offset, which should go down to a couple of mV in a few seconds.

Remove the jumper from R20, install the RIAA components, and connect the proper supply jumpers between the regulators and the second stage and between the second stage and the input stage. Leave the short at the +INP and switch on the supply voltage. Check the offset at the output, which, after a couple of minutes, should be a few mV. A certain short-term drift of the offset is normal—remember, you are dealing with a total DC-gain of 84dB!

If you have audio measurement equipment, including an inverse RIAA circuit[8] connect it to the input of the phono preamp and sweep the audio generator from 20Hz to 20kHz. Monitor the output of the phono preamp with an audio mV-meter. The output voltage should be constant with a maximum tolerance of ±0.2dB. If you have a scope, connect it to the output and make sure that the circuit does not overload during the measurement.

The phono preamp is now ready to be connected to your system. Changing between MM and MC means replacing R10. For MM, R10 = 75R and for MC it is 6R8. For MM, R2 is 47k, and for MC you will need to connect the appropriate load resistor at the input. You can solder this resistor either directly across the input terminals of the PCB or across the input connector on the back panel. A shunt capacitor (C1) might also be necessary; consult the datasheet of your pickup. aX

FIGURE A: The RIAA playback characteristic.



DESIGN PROCEDURE

The design procedure for the hybrid circuit in *Fig. 4A* is relatively simple:

1.) 75µs passive network: choose an appropriate value for R18, which should be as low as possible because it influences the noise in the second stage, but high enough so that the input stage can drive it at high frequencies. If you use op amps for the two amps, the value should be between 600Ω and several kΩ, depending on the drive capability of the op amps. I use R18 = 750Ω for the discrete JFET input circuit used in the EB-804/419 because it is easy to make the 75µs time constant with a C11 = 75µ/750 = 0.1µF capacitor. Because resistors are available with many more values than caps, you can also start with the cap and calculate the necessary resistor. If you want to use a 47nF resistor, R18 = 75µs/47nF = 1.6k. Naturally, you can also use two capacitors in parallel to make a convenient value.

2.) 318 and 3180µ feedback network: choose a practical value for R33, which determines the 3180µs time constant. Because I wanted to use the same cap as for the 75µs, I calculated the resistor from R33 = 3180µs/0.1µF = 31k8 (closest practical value is 31k6). Next calculate R31 from the formula that gives the DC (zero frequency) gain. The 1kHz gain is 0.101×DC gain, or the DC gain is 9.898× the 1kHz gain. I want a 1kHz gain of 24dB, which is 15.85×, so the DC gain will be 9.898×15.85 = 156.88×. R1 = (9.898×R3)/(9×DC gain) = 222Ω (closest value is 221R). Finally, calculate R2 from (R3/9)-R1 = 3290 (closest value: 3k32).

For the all-passive equalization in *Fig. 4B* I used Walt Jung's network N2[7] with the following component values:

Theoretical:	Closest fit:
R18 = 7k29	7k32
R19 = 1k06	1k05
C11 = 0.3µF	0.3µF
C12 = 0.1029µF	0.1µF + 3nF

All network components should be precision ones: resistors 1% or better and caps 2% or better. The Starter kit EB-804/419 is delivered with 0.1% Dale CMF-55-143 non-magnetic resistors and 1 and 2% polypropylene and polystyrene capacitors. Using Walt's inverse RIAA network[8] both networks show better than ±0.2dB accuracy across the audio range. ■

Acknowledgements
Many thanks to Walt Jung for letting me use his passive RIAA network in the phono preamp and for his valuable comments and suggestions regarding this article.

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4. Erno Borbely, "Starter Kits, Part I: EB-604/410 All-JFET Line Amp," audioXpress, March 2005, p. 9.

5. Erno Borbely, "The All-FET Line Amp" audioXpress, May 2002, p. 28.

6. Walt Jung, Op Amp Applications, Chapter 6-1, Analog Devices, 2002, ISBN 0-916550-26-5.

7. Walt Jung, "Topology Considerations for RIAA Phono Preamplifiers," AES Preprint 1719, 67th Convention, 1980 Oct. 31/Nov. 3 New York. (http://home.comcast.net/~walt-jung/wsb/PDFs/Topology_Considerations_for_RIAA_Phono_Preamps.pdf)

8. Walt Jung and Stanley Lipshitz, "A High Accuracy Inverse RIAA Network," The Audio Amateur, 1/80, p. 22.

PARTS LIST

PHONO PREAMP EB-804/419

Resistors, Trimpots	
R1	Short for MC 10R for MM
R2	47k5 for MM See spec for MC
R3, R4, R5, R6	10R
R7, R8, R11	681R
R9	2K21
R10	6R8 for MC 75R for MM
R12	Optional for passive eq.
R13, R14, R20, R34, R35	1M
R15, R16, R22, R29	332R
R17, R25, R38	47R5
R18	750R
R19	See passive eq.
R21, R28	1k5
R23, R24	100R
R26, R27	75R
R30	100k
R31	221R
R32	3k32
R33	31k6
R36, R37	1k2
P1	10R COPAL
P2	200R COPAL
P3	50R COPAL
Capacitors	
C1, C8	100pF, 160V, PS
C2, C3, C20, C22	0.1µF, 100V, PP
C4, C13	33pF, 160V, PS
C5	Appl. Dependent
C6, C7, C15, C16	0.22µF, 250V, PP
C9, C10, C19, C21	47µF, 35V
C11, C14	0.1µF, 100V, 1%, ERO PP
C12	See passive eq.
C17, C18, C23, C24	100µF, 25V, ROE EKO
FETs, Diodes	
Q1, Q2	K170GR (Idss: 3.5–4mA)
Q3, Q4	J74GR (Idss: 3.5–4mA)
Q5, Q14	J74V (Idss: >10mA)
Q6, Q15	K170V (Idss: >10mA)
Q7, Q16	K170BL (Idss: 8–10mA)
Q8, Q17	J74 (Idss: 8-10mA)
Q10, Q11	K170BL/V*
Q12, Q13	J74BL/V**
Q9, Q18	LF411CN
D1, D2	LM4040DIZ-10

** Dual JFET K389BL/V can be used for Q10-Q11*
*** Dual JFET J109BL/V can be used for Q12-Q13*

Miscellaneous	
40 × 1mm solder pins	
PCB: EB-804/419	

REGULATORS EB-704/418

Resistors, Trimpots	
R1, R2, R5, R6	3K09
R3, R4	1K82
R7, R8	8K25
R9, R10, R13, R14	100R
R11, R12	10K
P1, P2	5K COPAL
Capacitors	
C1, C2, C3, C4	0.1µF, 100V, PP
C5, C6	220µF, 16V
C7, C8, C9, C10	47µF, 35V
C11, C12	33pF, 400V, MICA
FETs, Diodes	
Q1, Q2	K170BL/V* (Matched, Idss: >9mA)
Q3, Q4	J74BL/V** (Matched, Idss: >9mA)
Q5	J148
Q6	K982
Q7	K2013
Q8	J313
D1, D2	E202 (2mA)
D3, D4	E102 (1mA)
D5, D6	LM329DZ (6.9V ref.)

** Dual JFET K389BL/V can be used for Q1-Q2*
*** Dual JFET J109BL/V can be used for Q3-Q4*