Part I

A MOVING COIL PREAMP

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

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.

MC-preamp considerations

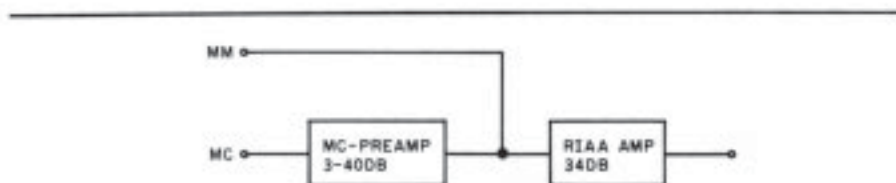If you look at the annual directory published by *Audio* (October 1985), you will find the lowest output MC is the Ortofon MC2000, which has an output of 0.05mV at 5cm/S RMS lateral velocity. The one with the highest output appears to be the Dynavector DV-20B2, with 3.6mV at the same groove velocity. 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-
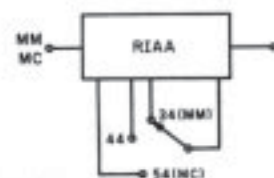

FIGURE 2: Incorporating 20dB extra gain in the RIAA stage will allow you to use medium-to-high output MCs.

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 agitation 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 $(1.38 \times 10^{-23}$ 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:
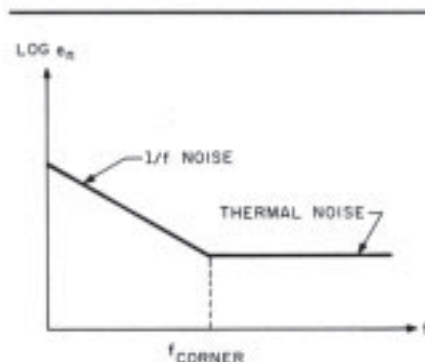i_n is the RMS noise current in Amp.

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

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

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
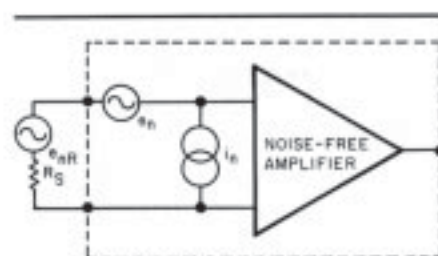

FIGURE 4: Amplifier noise model.

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 $1/f\infty$, with ∞ 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.

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:

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

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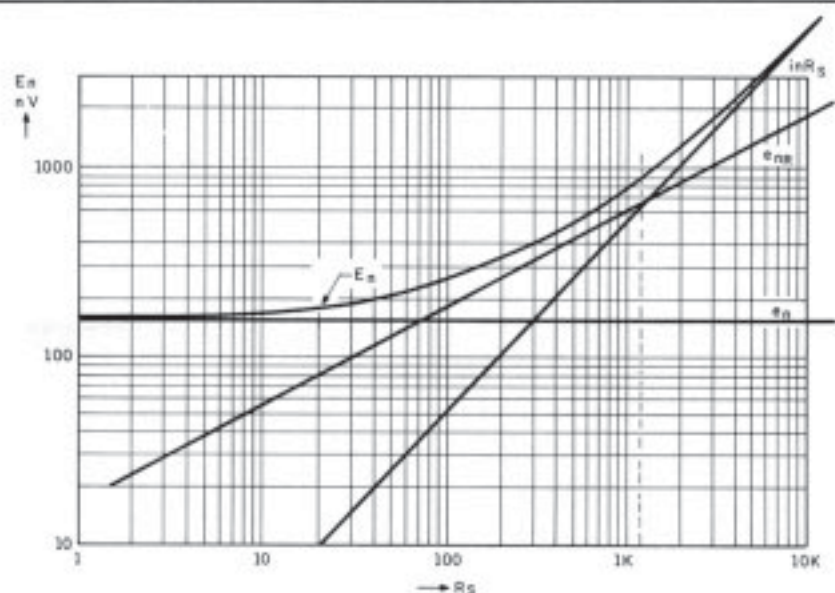
FIGURE 5: Equivalent input noise (E_n) versus R_s for a hypothetical amplifier (20-20kHz, flat, RMS).

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:

$$e_nR = \sqrt{4 \times 1.38 \times 10^{-23} \times 300 \times 2000} = 0.575\mu V \quad (4)$$

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 $0.128\mu V$ of noise in the audio frequency range. More about this later.
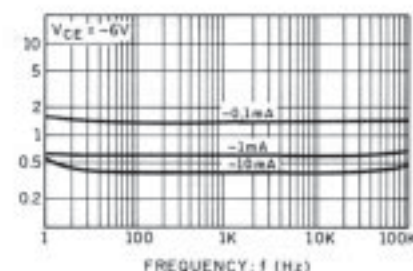
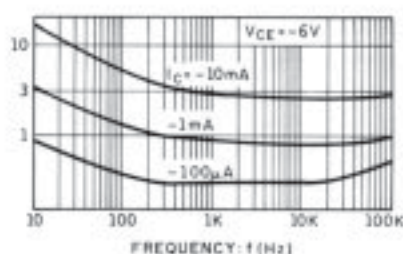FIGURE 6a: Noise voltage versus frequency for 2SB737. h_{FE} = 270-560 (S-Group).

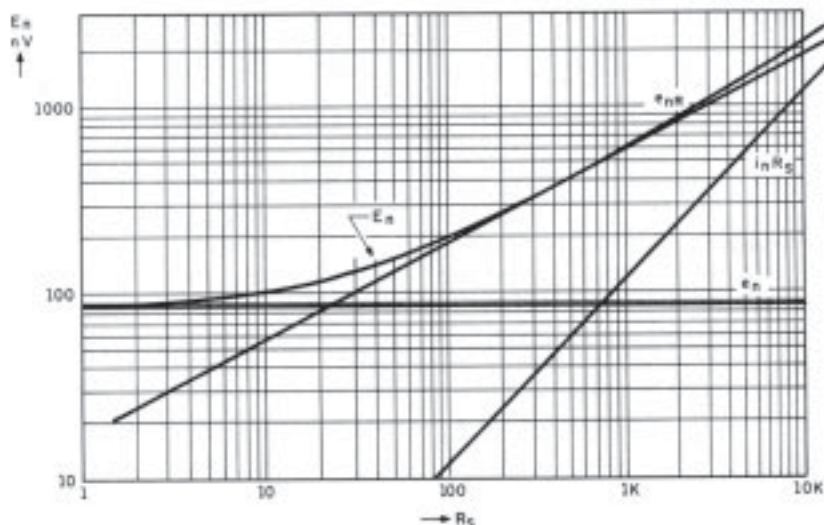FIGURE 6b: Noise current versus frequency for 2SB737. h_{FE} = 270-560 (S-Group).

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:

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

and

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

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

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/$\sqrt{Hz}$ and 0.85pA/$\sqrt{Hz}$. 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
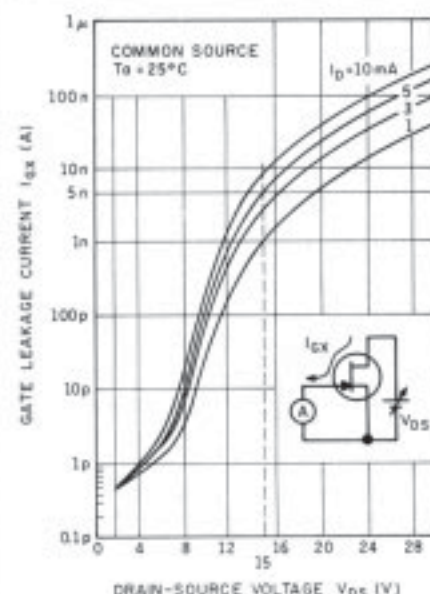

FIGURE 8: Gate leakage current for 2SK147.

[R₄] 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 0.75nV/√Hz. For 20kHz, this results in $e_n = 106$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, $I_G = 5$nA, which produces an $i_n = 5.6 \times 10^{-12}$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
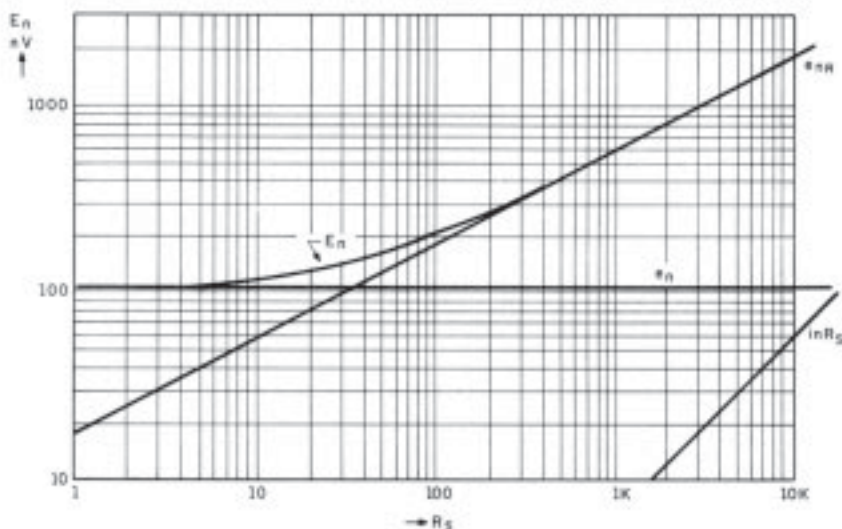

FIGURE 9: *Equivalent input noise (E_n) versus R_s for an amplifier using 2SK147 FET.*

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
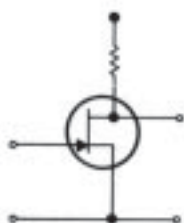

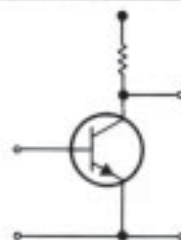FIGURE 10a: *Basic common emitter configuration.*
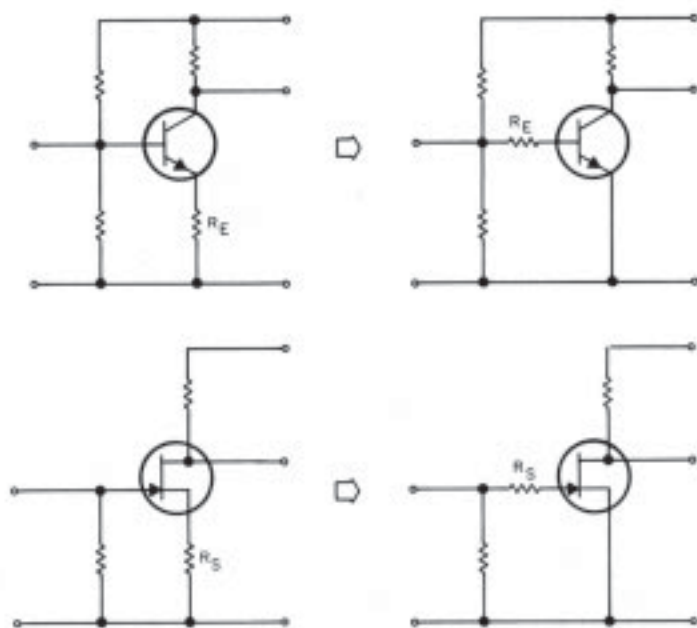

FIGURE 10b: *Basic common source configuration.*

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,
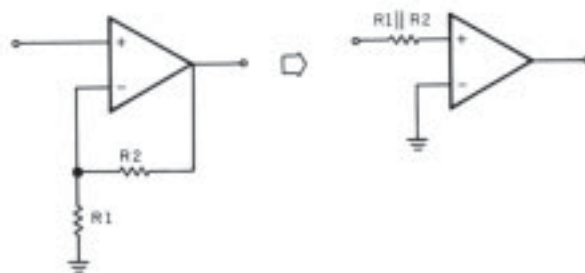

FIGURE 12: The overall feedback network also contributes to the noise.

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
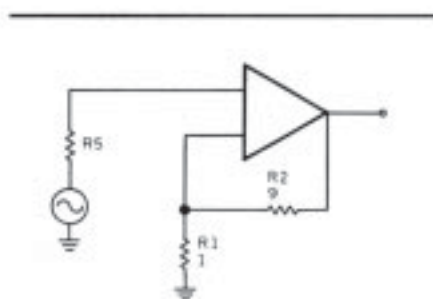
FIGURE 13: *For low noise, R2||R1 should be lower than R₅.*

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.

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.

REFERENCES

1. Netzer, Y., "The Design of Low-Noise Amplifiers," *Proceedings of the IEEE* (Volume 69, Number 6), June 1981.
2. Maxwell, J., "Noise of Sources," *National Semiconductor*, February 1977.
3 Motchenbacher and Fitchen, *Low-Noise Electronic Design*, John Wiley and Sons, Inc., 1973.
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*.