DEVELOPMENT OF A COMPUTER PROGRAM FOR TRANSISTOR NOISE ANALYSIS

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DEVELOPMENT OF A COMPUTER PROGRAM

FOR TRANSISTOR NOISE ANALYSIS

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DEVELOPMENT OF A COMPUTER PROGRAM FOR TRANSISTOR NOISE ANALYSIS

ABSTRACT

A computer program is developed for designing low-noise systems and calculating the spot and broadband values of input and output noise, the noise bandwidth and the noise figure.

The transistor base-spreading resistance, an parameter of the noise model, is determined experimentally from 1/f noise measurements.

BİLGİSAYAR YARDIMLI TRANZİSTÖR GÜRÜLTÜ ANALİZİ

ÖZETÇE

Elektrik cihaz tasarımı, endüstri ve askeri uygulamalar gibi pek çok konuda gürültü önemli bir sorundur. Çünkü gürültü herhangi bir ölçümün doğruluğunu ve elektronik olarak işleme tabi olan sinyalin en düşük değerini belirler.

Düşük gürültü sistem tasarımı için tek bir frekansta veya bir frekans bandında giriş ve çıkış gürültüsünün hesaplanmasında, yükselteç ve sezici devrelerin tepki eğrilerinin bulunmasında ve gürültü faktörünün belirlenmesinde bilgisayar yardımlı gürültü analizi etkili bir çözümdür.

Bu tez'de ilk olarak sezici devrelerdeki ve tranzistörlü yükselteçlerdeki gürültü kaynakları incelenmiştir. Devre elemanları ve alt devreler için gürültü modelleri geliştirilmiş ve bu modellerdeki gürültünün belirlenmesindeki hesaplama zorluklarını yenmek için bir bilgisayar programı geliştirilmiştir.

Bazı tranzistör parametreleri gürültünün ölçülmesi yardımıyla bulunabilir. Tez'in ikinci kısmında da iki kutuplu eklem tranzistörü-

nün parametrilerinden olan baz-gövde direnci, alçak frekans gürültü ölçme yöntemi kullanılarak belirlenmiştir. Bu amaç için bir ölçme cihazı kurulmuş ve kullanılmıştır.

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$A_v(f)$: Voltage gain as a function of frequency

A_p : Power gain

e_n : r.m.s. equivalent noise voltage generator

e_t : r.m.s equivalent thermal noise

F . Noise figure

$G(f)$: Power gain as a function of frequency

g_m : Transconductance

i_d : Drain-noise current of FET

i_g : Gate-noise current of FET

i_f : 1/f noise current

i_n : r.m.s. equivalent noise current generator

$K_g(\eta)$, $K_d(\eta)$, $K_{gd}(\eta)$: Working point-dependent factors, characterizing the channel noise and gate noise of MIS transistor and their correlation

k : Boltzman constant

P_{no} : Two-port output noise power

P_{ni} : Two-port input noise power

R_n : Equivalent noise resistance

r_e : Emitter resistance

$S(f)$: Power density

q : The charge of the electron

T : Temperature in degrees Kelvin

V_{ni} : Equivalent input noise

Y_{21} : Complex g_m of transistor

Δf : Noise bandwidth in Hertz

I. INTRODUCTION

Electrical noise is a problem in industrial, military, and consumer equipment design. The designer must be cognizant not only of the sources of noise, but also of the methods of noise reductions that one available to him. He will strive toward an optimum design. In this quest he must use all available and applicable tools. One such tool is the digital computer.

The main purpose of calculating the noise in an electronic system is to discover what is the least signal it will detect or what is the accuracy with which a large signal can be measured, and to show how the result depends on the parameters of the system.

In this thesis,first the sources of noise are examined; these sources include sensors and other devices, amplifier, and associated circuitry. Noise models are developed for circuit components and for subsystems. In the design of low noise systems,that information is used to predict performance. However, it should be noted that associated calculations can be very time consuming. Network and sensor impedances can be complex. To predict the total noise in a usable bandwidth it is necessary to calculate the noise at many frequencies and integrate the mean square noise voltage over the bandwidth of interest. To simplify the design process a computer program is developed for noise analysis.

In the second part, the low-frequency noise measurement is used to determine the base-spreading resistance of a bipolar junction transistor. For this purpose a laboratory calibre equipment is built and used.

II. NOISE MECHANISMS

2.1 INTRODUCTION

Noise is undesired random signals that obscure or interfere with a desired signal. These extraneous signals may be derived from sources external to the system such as the electrostatic or electromagnetic coupling from the 50 Hz power lines. Another source of noise is the physics of the devices and materials that make up the electrical system. We will be concerned with this type noise.

The fluctuating quantities such as noise voltages or currents are called random variables. Random variables X(t) can be characterized on a statictial basis. One important way is to characterize them by their statistical averages. The most widely used averages are the mean value $\overline{X}$ and the mean square value $\overline{X^2}$. Often $\overline{X}$ is rigorously zero, and then the most meaningful quantity is $\overline{X^2}$.

The most significant noise sources give rise to fluctuating quantities that have averages and mean square averages that are independent of time. Such random variables are called stationary random variables. Stationary random variables have probability density functions that not explicity depend on time.

A fluctuating quantity X(t) can be described by its spectral density function Sx(f). By introducing this quantity, a fluctuating voltage V(t) in a small frequency interval Δf can be represented by a noise voltage $\sqrt{Sv(f)\ \Delta f}$, where Sv(f) is the spectral intensity of V(t) at frequency f.

The main types of noise are thermal noise, low frequency noise, shot noise, and burst noise.

2.2 THERMAL NOISE

Thermal noise, also known as white noise or Johnson noise, is caused by the thermally excited random vibration of the charge carriers in a conductor. The expression for the r.m.s. noise voltage in the conductor over a frequency interval Δf is,

$$e_t = (\overline{e_t^2})^{\frac{1}{2}} = \sqrt{4\ k\ T\ R\ \Delta f} \qquad (2.1)$$

where

k = Boltzman constant = 1,38 x 10^{-23} W - sec/K^{o}

T = Temperature of the conductor in degrees Kelvin

Δf = Noise bandwidth in Hertz

R = Resistance of real part of the conductor impedance

For the purpose of analysis, a noisy resistor R is represented by an equal noiseless resistor in series with a noise generator of r.m.s.

value e_t equal to $(4\ k\ T\ R\ \Delta f)^{\frac{1}{2}}$ as shown in Fig. 2.1. This series arrangement can be replaced by an equivalent constant-current generator in parallel with the resistor.

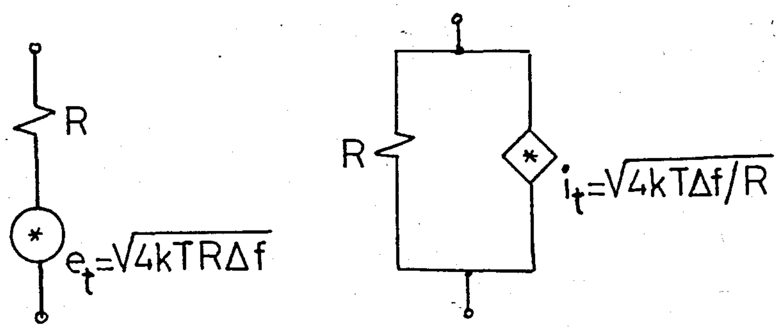


Figure 2.1 Equivalent circuits for thermal noise

2.3 SHOT NOISE

Shot noise is generated when the charge carriers supplying the current, pass through an energy gap in tubes, transistors, and diodes. Shot noise is associated with current flow accross a potential barrier. Such a barrier exists in every p-n junction semiconductor device and at the cathode surface in vacuum tubes. The mean square current fluctuation is a function of the bandwith Δf:

$$\overline{i^2} = 2qI\ \Delta f \qquad (2.2.)$$

No barrier is present in a simple conductor, therefore no shot noise is generated. Thermal noise and shot noise both have a flat frequency spectrum.

2.4 1/f NOISE

1/f noise is also referred to as flicker noise or low-frequency noise. The thin metallic or semiconductor layers, carbon resistors, carbon microphones, vacuum tubes, solid state devices, etc, are sources of another type of noise which is important at low frequencies. The power spectrum of this noise source may be expressed in the general form:

$$S(f) = K \frac{I^{\beta}}{f^{\alpha}} \qquad (2.3)$$

where

I = DC current flowing through the device

f = frequency

$\beta \cong 2$

$\alpha \cong 1$

K = constant

The major cause of 1/f noise in semiconductor devices is traceable to properties of the materials. The generation and recombination of carriers in surface energy states and the density of surface states are important factors.

2.5 BURST NOISE

A pulse-type noise, called burst noise or "popcorn" noise, has been observed in carbon or carbon-film resistors, at reverse or forward biased p-n junctions, and discrete or integrated circuit transistors. This type of noise appears as a square wave width random fluctuations.

The power spectral density is a $1/f^{\alpha}$ function with $1<\alpha<2$. It is often found to vary as $1/f^{2}$. This noise is masked by other mechanisms such as shot noise and 1/f noise.

2.6 NOISE BANDWIDTH

The effective noise bandwidth Δf represents the bandwidth of the equivalent rectangular filter which will pass the same rms noise as the actual filter. Noise bandwidth is the area under the power curve; the integral of power gain versus frequency, divided by the peak amplitude of the curve. This can be stated in an equation form:

$$\Delta f = \frac{1}{G_o} \int_o^\infty G(f)\, df \qquad (2.4)$$

where

Δf = noise bandwith in Hertz

$G(f)$ = power gain as a function of frequency

G_o = peak power gain

The equivalent noise bandwidth can also be written as

$$\Delta f = \frac{1}{A_{vo}^2} \int_o^\infty \left[A_V^2(f)\right] df \qquad (2.5)$$

where

$Av(f)$ = voltage gain as a function of frequency

A_{vo} = midband voltage gain

Let us assume the bandwidth of the overall frequency range is

$f_2 - f_1$. The transfer function of the two-port is H(f) and the one-sided power density of the noise source connected to the input is S(f). Then the squared output noise voltage is

$$\overline{A_v^2} = \int_{f_1}^{f_2} S(f) \, | \, H(f) \, |^2 \, df \qquad (2.6)$$

Let H(f) be the transfer function of a band-pass filter. For a sinusoidal signal, the bandwidth is defined between the 3-dB points where the output power is decreased to half the maximum value. However for a stochastic driving signal, the output power is a function of both H(f) and S(f), the latter being the spectrum of the driving signal the definition of the bandwidth is

$$\Delta f = \frac{\int_0^\infty | \, H(f) \, |^2 \, S(f) \, df}{| \, H(f_o) |^2 \, S(f_o)} \qquad (2.7)$$

where f_o is a reference frequency suitably chosen within the pass-band.

2.7 ADDITION OF NOISE VOLTAGES AND CORRELATION

Equivalent noise generators represent a very large number of component frequencies with a random distribution of amplitudes and phases. The output power is the sum of these separate output powers and consequently it is valid to combine such sources so that the resultant mean square voltage is the sum of the mean square voltages of the individual generators. This statement can be extended to noise current sources in parallel.

If e_1 and e_2 represent uncorrelated noise sources in series, the mean square of the sum $\overline{e^2}$ is given by,

$$\overline{e^2} = \overline{e_1^2} + \overline{e_2^2} \qquad (2.8)$$

In order to sum correlated waves, the general expression is

$$\overline{e^2} = \overline{e_1^2} + \overline{e_2^2} + 2C\, \overline{e_1^2}\, \overline{e_2^2} \qquad (2.9)$$

where C is called the correlation coefficient and can have any value between -1 and $+1$. When $C=0$ the voltages are uncorrelated. When $C = 1$ the signals are totally correlated.

III. NOISE PARAMETERS OF LINEAR NETWORKS

3.1 ONE-PORT

The noise equivalent circuit of a resistor R consists of a noiseless resistor R in series with a noise voltage source. The mean square value of the thermal noise component of the voltage is given by

$$\overline{e^2} = \int_{f}^{f+\Delta f} S_V(f)df = 4\ k\ T\ R\,\Delta f \qquad (3\text{-}1)$$

The source voltage of the voltage generator in Fig. 2.1 is

$$e = (\overline{e^2})^{1/2} = \sqrt{4\ k\ T\ R\,\Delta f} \qquad (3.2)$$

After a Norton - Thevenin transformation, we have the elements of the current generator equivalent circuit shown in Fig. 2.1

$$i = (\overline{i^2}\ )^{1/2} = \sqrt{4\ k\ T\ G\ \Delta\ f} \qquad (3.3)$$

If S(f) is assumed to be constant within a narrow frequency band Δf, the following expressions are frequently applied :

$$i = \sqrt{Si(f)} \quad \sqrt{\Delta f} \quad \text{where} \quad \sqrt{Si(f)} \quad = A/Hz^{\frac{1}{2}} \qquad (3\text{-}4)$$

$$e = \sqrt{S_V(f)} \quad \sqrt{\Delta f} \quad \text{where} \quad \sqrt{S_V(f)} \quad = V/Hz^{\frac{1}{2}} \qquad (3\text{-}5)$$

The noise equivalent circuits may be easily be applied to networks comprising reactive elements. The appearence of a susceptance B in parallel with conductance G will not change the source voltage in Eq. (3.1); an ideal susceptance does not generate noise. However, the terminal voltage of an admittance $Y = G + j\,w\,C$ wil be frequency dependent :

$$e_o = \frac{i}{Y} = \frac{i}{G + j\,w\,C} \qquad (3\text{-}6)$$

3.2 TWO-PORTS

For a two-port network, the admittance equation is

$$\begin{bmatrix} i_1 \\ i_2 \end{bmatrix} = Y \begin{bmatrix} e_1 \\ e_2 \end{bmatrix} + \begin{bmatrix} i^{I} \\ i^{II} \end{bmatrix} \qquad (3\text{-}7)$$

which may be expressed in the form:

$$i_1 = Y_{11}\; e_1 + Y_{12}\, e_2 + i^{I} \qquad (3\text{-}8)$$

$$i_2 = Y_{21}\; e_1 + Y_{22}\, e_2 + i^{II} \qquad (3\text{-}9)$$

Figure 3-1 shows the definitions of these quatities. Superscripts I and II refer to the input and output respectively.

There is a difference between one-ports and two-ports. A single parameter, which may be either voltage or current, is sufficient to characterize a one-port. However, two uncontrolled generators (e.g. i^{I} and i^{II}) are required to characterize a two-port. These generators are not necessarily independent. The noise equivalent circuit of the network is shown in Fig. 3.2

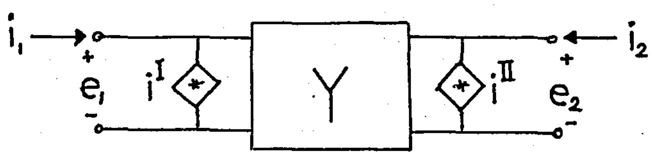


Figure 3.1 Two-port with noise sources

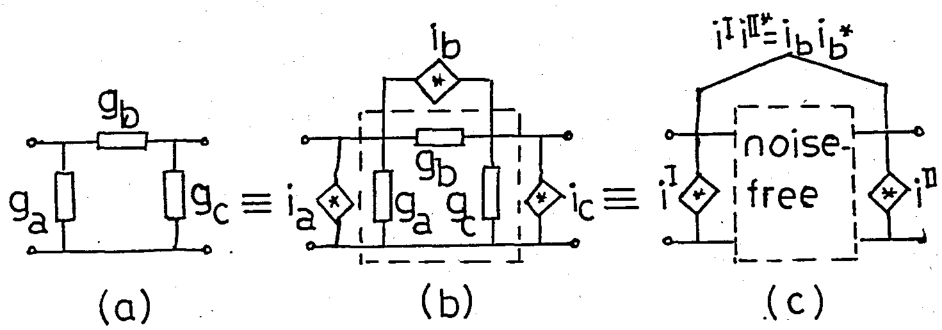


Figure 3.2 (a) Noisy two-port (b) Exclusion of noise sources (c) Contraction of noise sources

where

$$i_a = \sqrt{4\ k\ T\ g_a\ \Delta f}$$

$$i_b = \sqrt{4\ k\ T\ g_b\ \Delta f}$$

$$i_c = \sqrt{4\ k\ T\ g_c\ \Delta f}$$

The source current of the input equivalent noise generator is obtained by shorting the output and connecting an ideal current meter of zero resistance accross the input. In this case, we have:

$$\overline{i^{I^2}} = (i_a + i_b) (i_a^* + i_b^*) \qquad (3\text{-}10)$$

$$\overline{i^{II^2}} = (i_b + i_c) (i_b^* + i_c^*) \qquad (3.11)$$

(Assuming i_a, i_b and i_c are independent)

The correlation between i^I and i^{II} is characterized by the product

$$\overline{i^I \, i^{II*}} = (i_a + i_b) (i_b^* + i_c^*) = i_b \, i_b^* \qquad (3.12)$$

The two-port block will be regarded as noise-free, as the excluded internal noise sources have been substituted by i^I and i^{II} and by their correlation.

The equivalent circuit of Fig 3.3 may be utilized even more conveniently.

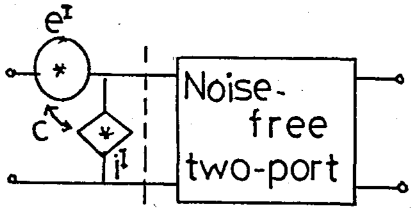


Figure 3.3 Two-generator equivalent circuit.

The two-port, regarded as noise free is characterized by its chain matrix L and the noise sources are reduced to the input.

$$\begin{bmatrix} e_1 \\ i_1 \end{bmatrix} = L \begin{bmatrix} e_2 \\ i_2 \end{bmatrix} + \begin{bmatrix} e^I \\ i^I \end{bmatrix} \qquad (3.13)$$

and the correlating factor can be defined as, |1|

$$C = \frac{e^{I*} \; i^{I}}{\sqrt{|e^{I}|^{2} \; |i^{I}|^{2}}} \tag{3-14}$$

3.3 AMPLIFIER NOISE

In preceeding section a universal noise model for any two-port network is described. The network is considered as a noise-free box, and the internal sources of noise are represented by a pair of noise generators located at one port only. Figure 3.4 represents a noisy amplifier including the signal source V_S and the source resistance R_S.

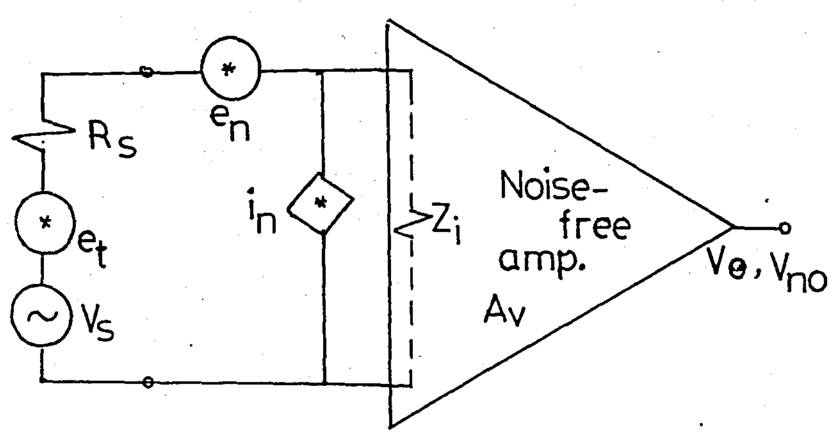


Figure 3.4 Amplifier noise model with the signal source

Amplifier noise is represented completely by a zero impedance voltage generator e_n in series with the input port, an infinite impedance current generator i_n in parallel with the input and by a complex correlation coefficient C (omitted). The thermal noise of the signal source is represented by a noise generator e_t.

The equivalent input noise is given by,

$$\overline{V_{ni}^2} = \frac{\overline{V_{no}^2}}{K^2} \tag{3-15}$$

Where $K = A_V Z_i \,/\, (R_S + Z_i)$

and $$\overline{V_{ni}^2} = \overline{e_t^2} + \overline{e_n^2} + \overline{i_n^2}\, R_s^2 \tag{3-16}$$

The equation can be applied to a system using any type of active device. The noise voltage and current generators are not necessarily independent. We must introduce the correlation coefficient C. And Eq. (3-16) results

$$\overline{V_{ni}^2} = \overline{e_t^2} + \overline{e_n^2} + \overline{i_n^2}\, R_S^2 + 2C\, \overline{e_n^2}\, \overline{i_n^2}. \tag{3-17}$$

In Figure 3-4, the correlation term can be represented as a voltage generator with r.m.s. value $(2C\, \overline{e_n^2}\, \overline{i_n^2})^{\frac{1}{2}}$ in series with e_n or an appropriate current generator in parallel with i_n.

3.4 NOISE FIGURE

The noise figure, also called the noise factor, of a two-port device is the ratio of the available output noise power per unit bandwidth to the portion of that noise caused by the actual source connected to the input terminals of the device, measured at the standard temperature of 290 °K.

$$F = \frac{P_{no}}{A_p\, P_{ni}} = 1 + \frac{P_{no}'}{A_p\, P_{ni}} \tag{3-18}$$

Where P_{no} = the output noise power

P_{ni} = the signal source noise power

A_p = power gain

P'_{no} = Fraction of the output noise power generated within the two-port

An equivalent definition of noise figure is:

$$F = \frac{\text{Input signal-to-noise ratio}}{\text{Output signal-to-noise ratio}}$$

The noise figure F can be defined in terms of e_n and i_n

$$F = \frac{\overline{e_t^2} + \overline{e_n^2} + \overline{i_n^2}\, R_s^2}{\overline{e_t^2}} = 1 + \frac{\overline{e_n^2} + \overline{i^2}\, R_s^2}{\overline{e_t^2}} \qquad (3\text{-}19)$$

The equation shows that the noise figure can be expressed as the ratio of the total mean square equivalent input noise to the mean square thermal noise of the source. A minimum noise figure may be obtained by differentiating Eq. (3.19) with respect to Rs. Then,

$$F(\min) = 1 + \frac{e_n\,(\text{r.m.s.})\; i_n\,(\text{r.m.s.})}{2\,k\,T\;\Delta f} \qquad (3\text{-}20)$$

and

$$R_s(\text{opt}) = \frac{e_n\,(\text{r.m.s.})}{i_n\,(\text{r.m.s.})} \qquad (3\text{-}21)$$

3.5 NOISE IN CASCADED STAGES

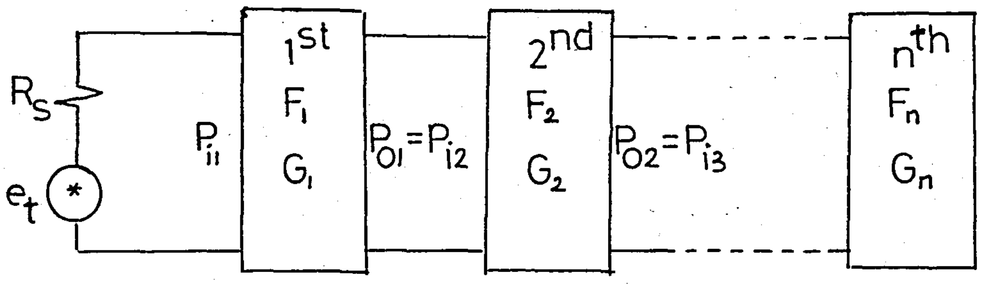


Figure 3-5 Cascaded networks.

The system shown in Fig. 3-5 is a cascaded n-stage network; where G_n is the available power gain and P_n is the available power. According to Eq. 3-18 noise figure is:

$$F = \frac{1}{G} \frac{P_O}{k\ T\ \Delta f} \qquad (3\text{-}22)$$

The available noise power at the input to network 2 P_{i2} is

$$P_{i2} = P_{O1} = F_1\ G_1\ k\ T\ \Delta f \qquad (3\text{-}23)$$

By considering the second stage separately Eq. 3-22 becomes:

$$F = \frac{P_{O2}}{G_2\ k\ T} \qquad (3\text{-}24)$$

Noise generated in stage-2 is;

$$F_2G_2 k\ T\ \Delta f\ -\ G_2 k\ t\ \Delta f\ =\ (F_2-1)\ G_2 k\ T\ \Delta f \qquad (3\text{-}25)$$

The total output noise P_t is given by the sum of term E_q 3-23 and 3-25 :

$$P_t = G_2\ (F_1 G_1\ k\ T\ \Delta f\ ) + (F_2-1)\ G_2 k\ T\ \Delta f$$

$$= (F_1G_1G_2 + F_2G_2 - G_2)\ k\ T\ \Delta f \qquad (3\text{-}26)$$

The noise figure of the cascaded pair is:

$$F = \frac{P_t}{G_1G_2 k\ T\ \Delta f} \qquad (3\text{-}27)$$

By substitution of Eq. 3-26 into 3-27 we obtain:

$$F_{12} = F_1 + \frac{F_2 - 1}{G_1}$$

The general expression for the noise figure of an n-stage amplifier is [4] :

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1G_2} + \ldots\ldots\ldots\ldots + \frac{F_n - 1}{G_1G_2\ldots\ldots G_n}$$

It can be concluded that for an n-stage amplifier the overall noise figure is nearly the noise figure of the first stage if the gain of the first stage is sufficiently large.

IV. NOISE IN BIPOLAR TRANSISTORS

4.1 BIPOLAR TRANSISTOR NOISE MECHANISMS

The common bipolar transistor contains sources of thermal noise, 1/f noise, and shot noise. The widely used hybrid $-\pi$ small signal equivalent circuit is modified to include noise sources. The equivalent input noise parameter V_{ni} is derived for the bipolar transistor by making use of the noise circuit model.

The base-spreading resistance $r_{bb'}$, the resistance of the lightly doped base region between the external base contact and active base region, exhibits thermal noise. Fluctuating base current I_B and collector current I_C are responsible for shot noise at the respective junction. The flow of base current I_B through the base-emitter depletion region gives rise to 1/f noise. These noise generators are shown in the hybrid-π noise model in Fig. (4.2) Feedback elements $C_{b'e}$ and $r_{b'e}$ have been omitted for simplicity.

The noise voltage generator e_b represents the thermal noise of the base-spreading resistance. The noise current generator i_1 is the shot noise of the total base current and i_2 is the shot noise of the collector current. These generators are:

$$\overline{e_b^2} = 4\ k\ T\ r_{bb'}\ \Delta f \qquad (4\text{-}1)$$

$$\overline{i_1^2} = 2q\ I_B\ \Delta f \qquad (4\text{-}2)$$

$$\overline{i_2^2} = 2q\ I_C\ \Delta f \qquad (4\text{-}3)$$

The source and load resistance thermal noise generators are respectively,

$$\overline{e_t^2} = 4\ k\ T\ R_S\ \Delta f \qquad (4\text{-}4)$$

$$\overline{e_L^2} = 4\ k\ T\ R_L\ \Delta f \qquad (4\text{-}5)$$

The correlation between i_1 and i_2 is | 1 |

$$i_1^* i_2 = 2\ k\ T\ (\ Y_{21}\ \ - 9m\ ) \qquad (4\text{-}6)$$

At low frequencies, $gm \cong R_e\ (Y_{21}) \cong Y_{21}$, so Eq. (4.6) is simplified to

$$i_1^*\ i_2 = 0$$

According to the discussion in Section 2-4, the spectral density of 1/f noise current can be written as

$$\overline{i_f^2} = \frac{K\ I_B^{\gamma}}{f} \qquad (4\text{-}7)$$

γ ranges between 1 and 2, but often can be taken as unity. It has been found that the constant K can be replaced by $2qf_L$ where q is the electron charge and f_L may have values from 3.7 kHz to 7 kHz.

The value for f_L is a representation of the noise corner frequency. The equation (4-7) can be rearranged in the form :

$$\overline{i_f^2} = \frac{2q\, f_L\, I_B^{\gamma}}{f} \qquad (4\text{-}8)$$

To derive an equivalent input noise V_{ni}, first we need to calculate the total noise at the transistor output, the gain from source to output and then divide the output noise by the gain.

The final expression for the equivalent input noise V_{ni} is given by |4| :

$$\overline{V_{ni}^2} = 4k\,T\,(r_{bb'} + R_S) + 2qI_B\,(r_{bb'} + R_S)^2 + \frac{2qI_C}{\beta_o^2}\,(r_{bb'} + R_S + r_{b'e})^2$$

$$+ \frac{2qf_L\, I_B^{\gamma}\,(R_S + r_{bb'})^2}{f} + 2qI_C\,\left(r_{bb'} + \frac{1}{w\,C_{b'e}} + R_S\right)^2 \left(\frac{f}{f_T}\right)^2 \qquad (4\text{-}9)$$

Eq. (4-9) is a good engineering approximation to actual behavior. The errors are at high frequencies beyond the transistor cut-off.

The $\overline{e_n^2}$ of the transistor can be written in the absence of the source noise as :

$$\overline{e_n^2} = 4\,kT\,r_{bb'} + 2qI_B\,(r_{bb'})^2 + \frac{2qI_C}{\beta_o^2}\,(r_{bb'} + r_{b'e})^2$$

$$+ \frac{2q\,f_L\,I_B^{\gamma}\,r_{bb'}^2}{f} + 2qI_C\,\left(r_{bb'} + \frac{1}{w\,C_{b'e}}\right)^2 \left(\frac{f}{f_T}\right)^2 \qquad (4\text{-}10)$$

And by assuming that R_s is very large, $R_s \gg r_{b'e}$, i_n^2 is obtained from Eq. (4.9). After dividing each term by R_s, we obtain the $\overline{i_n^2}$ parameter :

$$\overline{i_n^2} = 2qI_B + \frac{2qf_L\, I_B^{\gamma}}{f} + 2qI_c \left(\frac{f}{f_T}\right)^2 + \frac{2qI_c}{\beta_o^2} \qquad (4\text{-}11)$$

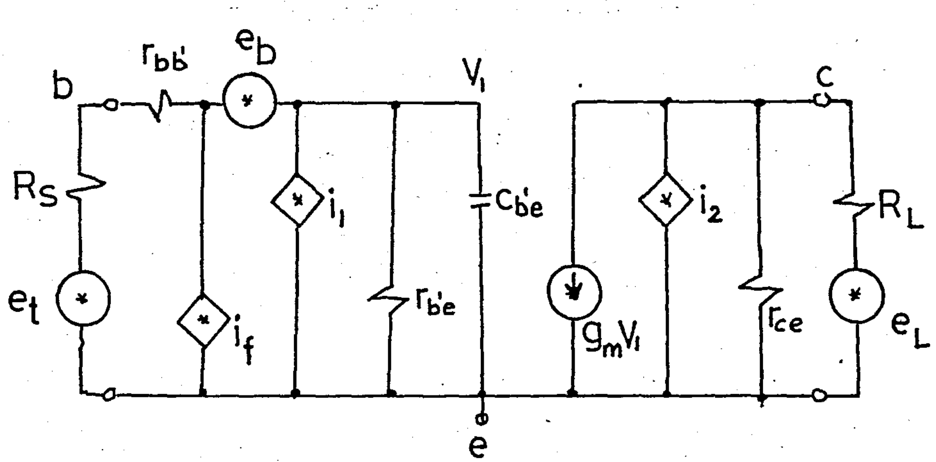


Figure 4.1 Hybrid - π noise model with source and load resistances

V. NOISE IN FET's

5.1 NOISE IN FET's

In a p-n junction device, the current flows inside the semiconductor material, while in the Metal-Insulator-Semiconductor or Metal-Oxide Semiconductor the current flow is confined to an extremely thin layer on the surface. Both JFET and the MOSFET can be assembled in either conductivity type, depending on whether the channel is p or n-type silicon. These differences appear naturally in the noise parameters too.

The noise equivalent circuit for common-source operation is shown in Fig 5.1. This circuit applies to both JFET or the MOSFET [4].

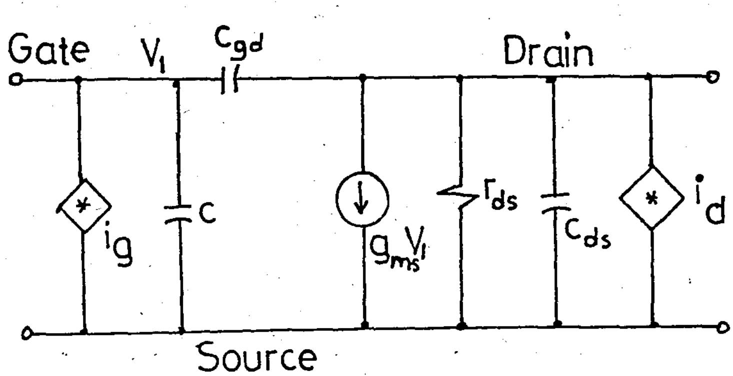


Figure 5.1 Noise equivalent circuit for JFET and MOSFET

The noise current generator i_g is the result of three physical processes : Shot noise of the current flowing through the gate, 1/f noise, and thermal fluctuations in the drain circuit. Noise generator i_d is the result of the thermal excitation of carriers in the channel of the device. Following the relations in the literature [1] for the channel noise, the expression for i_d is :

$$\overline{i_d^2} = 4\ k\ T\ \Delta f\ \ g_{ms}\ K_d\ (\eta) \qquad (5\text{-}1)$$

where $1 > K_d\ (\eta) > 2/3$ if $0 < \eta < 1$; g_{ms} is the mutual conductance.

And the gate noise is [1] :

$$\overline{i_g^2} = 4\ k\ T\ \Delta f\ \frac{w^2\ C_{gs}^2}{g_{ms}}\ K_g\ (\eta) \qquad (5\text{-}2)$$

For $0 < \eta < 1$ $1/12 < K_g\ (\eta) < 16/135$

Both the channel noise and the gate noise are brought about by random fluctuation of the channel potential, so there is correlation between i_g and i_d especially at high frequencies.

$$i_g^*\ i_d = 4\ k\ T\ \Delta f\ \ j\, w\, C_{gs}\ K_{dg}\ (\eta) \qquad (5\text{-}3)$$

if $0 < \eta < 1$, $0 < K_{dg}(\eta) < 1/9$

For practical purposes for which the value pertaining to $\eta = 1$ is extrapolated, is of importance. Then Eqs. (5-1), (5-2), and (5-3) will take the forms :

$$\overline{i_d^2} = \frac{2}{3} \; 4 \; k \; T \; g_{ms} \; \Delta f \tag{5-4}$$

$$\overline{i_g^2} = \frac{16}{135} \; 4 \; k \; T \; \frac{w^2 \; C_{gs}^2}{g_{ms}} \quad \Delta f \tag{5-5}$$

$$i_g^* \; i_d = \frac{1}{9} \; 4 \; k \; T \; jw \; C_{gs} \; \Delta f \tag{5-6}$$

and the complex correlation factor is:

$$C = \frac{i_g^* \; i_d}{\sqrt{|i_g|^2 |i_d|^2}} = 0{,}39 \; j \tag{5-7}$$

This factor is independent of the frequency, the mutual conductance, and the capacity C_{gs}. We see that C is purely imaginary and its magnitude is smaller than 0,5 which means that the correlation has little effect on either the design or the performance of the low-noise amplifier [2] . For most purposes we can ignore it. In the noise equivalent circuit of the field effect transistor shown in Fig. (5-2), the noise sources are reduced to the input port.

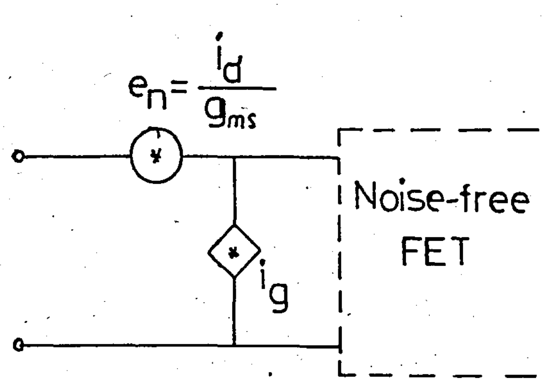


Figure 5-2 Noise sources reduced to the input in FET

In junction field effect transistors the gate leakage current I_g adds a further uncorrelated component to the gate-current fluctuations and the Eq. (5-5) becomes as:

$$\overline{i_g^2} = \frac{16}{135} 4 k T \frac{w^2 C_{gs}^2}{g_{ms}} \Delta f + 2qI_g \Delta f \qquad (5\text{-}8)$$

The MOSFET has no shot-noise component.

5-2 FLICKER NOISE IN FET's

The devices designed for low-noise low-frequency operation have a considerable increase of noise output at low frequencies. This is associated mainly with surface phenomena. The surface effects can be eliminated, however, in junction - gate devices, and the residual noise is then due to fluctuations in the recombination rate of carriers in the gate-channel depletion layer. These fluctuations modulate the channel current and therefore are equivalent to an additional noise voltage generator in the gate circuit. There is a little effect on the gate current noise. It is convenient to express the flicker noise power spectrum as:

$$\overline{e_f^2} = \frac{2}{3} \frac{4 k T}{g_{ms}} \frac{f_o}{f} \Delta f \qquad (5\text{-}9)$$

and the frequency f_o determines the frequency at which flicker noise first becomes noticable. The input noise voltage generator is then expressed as ;

$$\overline{e_n^2} = \frac{2}{3} \frac{4\,k\,T}{g_{ms}} \left(1 + \frac{f_o}{f} \right) \tag{5-10}$$

VI. NOISE IN SENSORS

6.1 NOISE MODELS

To develop the noise model of a sensor one can start with the equivalent circuit diagram. To each resistance and current generator we add the appropriate noise generators to obtain a noise equivalent circuit. The resistances have thermal noise and possibly excess noise. The current generators may have shot noise, 1/f noise, and excess noise. An expression for equivalent input noise can be derived using the noise equivalent circuit. The derivation of the equivalent input noise for the system follows three steps:

i. Determine the total output noise

ii. Calculate the system gain

iii. Divide the total output noise by the system gain to get equivalent input noise

For six classes of sensors, noise models and equivalent input noise expressions are derived[4].These sensors are :

1. Resistive sensor
2. Biased resistive source
3. RLC source

4. Biased diode sensor

5. Transformer model

6. Piezoelectric sensor

6.1.1 Resistive Sensor

Resistive detectors include the thermocouple, thermopile, and PEM infrared cell. In Fig. 6-1a the sensor is symbolized by signal source V_S and series resistance R_s. A coupling capacitor C_C can be used if we are interested exclusively in the time-varying output of the sensor. Element R_L may be useful for impedance matching. A noise model of the sensor-amplifier system is shown in Fig 6.1.b. Shunt capacitance C_p can be the result of the sensor assembly or it may represent the capacitance of the connecting wires.

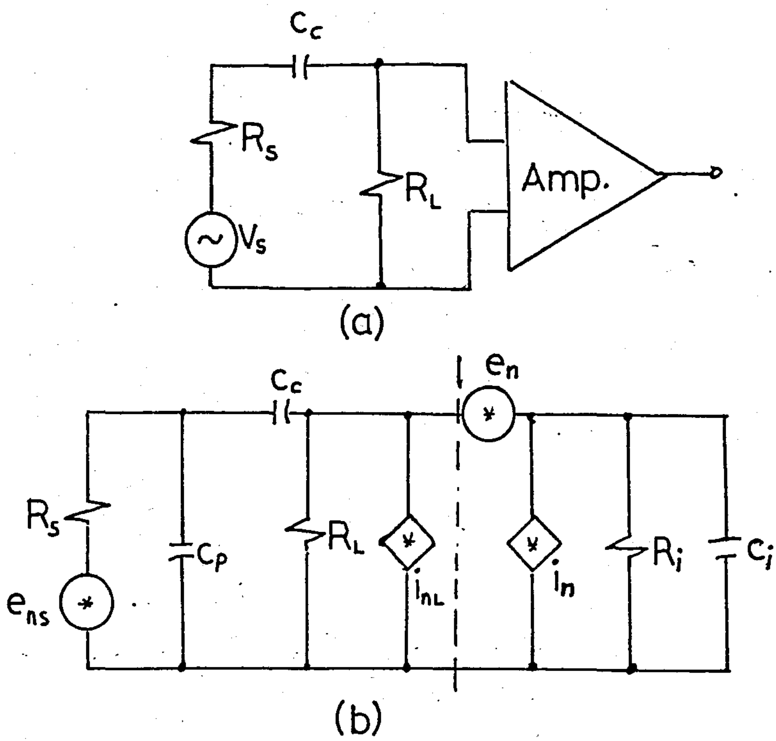


Figure 6.1 (a) General system diagram for resistive sensor

(b) Noise model for computer

The amplifier is represented by the noise parameters e_n and i_n, and by its input resistance R_i and input capacitance C_i. Following the procedure, we obtain :

$$\overline{V_{ni}^2} = 4\ k\ T\ R_s + \overline{e_n^2}\left[1 + \frac{R_s}{R_L}\left(1 + \frac{c_p}{c_c}\right) + j\left(wR_s\ C_p - \frac{1}{w\ R_L c_c}\right)\right]^2$$

$$+ \left(\overline{i_n^2} + \frac{4\ k\ T}{R_L}\right)\left[\frac{1 + j\ w\ R_s\ (c_p + c_c)}{j\ w\ c_c}\right]^2 \qquad (6\text{-}1)$$

and the gain is

$$K_t = \frac{j\ w\ R_L\ c_c\ R_i}{R_i + R_L - w^2 R_i R_L R_s\ (c_i c_p + c_i c_c + c_c c_p) + jw\left[(c_i + c_c)\ R_i R_L + (c_p + c_c)\ R_s\ (R_i + R_L)\right]} \qquad (6\text{-}2)$$

6.1.2 Biased Resistive Source

Biased resistive sources include photoconductive cells, piezoresistive strain gauges, and other elements which resistance changes with the sensed parameter. The noise model is shown in Fig. 6-2b. Following the method, the expression for input noise is:

$$\overline{V_{ni}^2} = 4\ k\ T\ R_s\ \left(\frac{f_L}{f}\right) + \left(\overline{i_n^2} + \frac{4\ k\ T}{R_L}\right)\left[\frac{R_s\ (c_p + c_c)}{c_c} - j\ \frac{R_s + R_b}{w\ R_b\ c_c}\right]^2$$

$$+ \left[\frac{c_c R_L (R_b + R_s) + R_b R_s (c_p + c_c) + j(w\ R_s R_b R_L c_p c_p - (R_s + R_b)/w}{R_b\ R_L\ c_c}\right]^2 \overline{e_n^2}$$

$$+ \frac{4\ k\ T}{R_b} R_s^2 \tag{6-3}$$

and the system gain is

$$K_t = \frac{R_b\ R_p\ c_c}{R_p\ (R_b+R_s)\ (c_c+c_i) + R_sR_b\ (c_p+c_c) + j\ \left(w\ R_sR_b\ ci\ (c_p+c_c) + w\ R_sR_bc_cc_pR_p - \frac{R_s + R_b}{w}\right)} \tag{6-4}$$

where $R_p = R_iR_L\ /\ (R_i + R_L)$

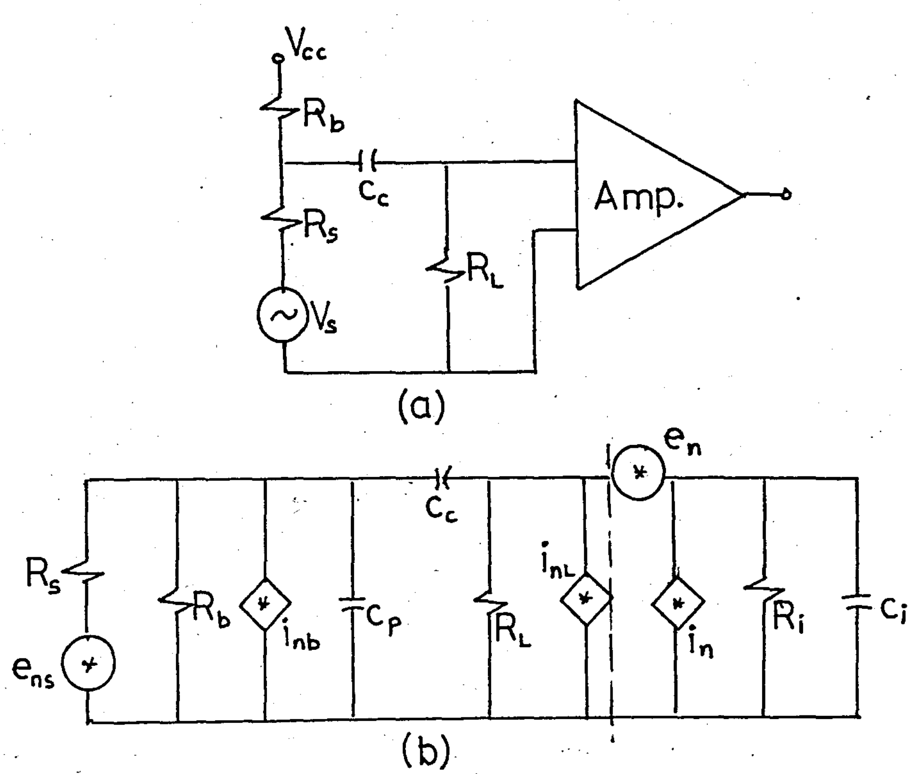


Figure 6-2 (a) Biased resistive source system.

(b) Noise model for computer.

6.1.3 R L C Source

R L C source models coils, inductive pickups, dynamic microphones, and various other inductive sensors. A general system diagram is shown in Fig. 6.3.

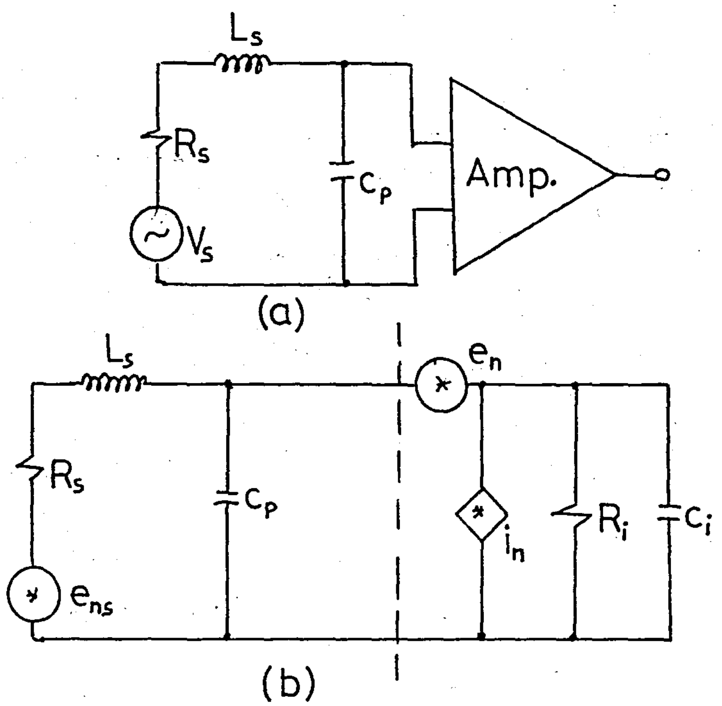


Figure 6-3) (a) System diagram for R L C sensor

(b) Noise model for Computer

The equivalent input noise is found as :

$$\overline{V_{ni}^2} = 4\ k\ T\ R_s + \left[(1-w^2 c_p L_s)^2 + (w\ R_s C_p)^2 \right] \overline{e_n^2} + (R_s^2 + w^2 L_s^2)\ \overline{i_n^2} \quad (6\text{-}5)$$

and the system gain is:

$$K_t = \frac{R_i}{R_i+R_s-w^2 R_iL_s\ (c_p+c_i) + jw\left[R_sR_i\ (c_p+c_i) + L_s\right]} \tag{6-6}$$

6.1.4 Biased Diode Sensor

An example of the reverse-biased diode sensor is the photodiode. A dc voltage reverse biases the diode and direct current flows through a bias resistor. Incoming radiation causes the current flow to increase. The shot noise of the dc bias current is an extra noise mechanism. A system diagram is shown in Fig. 6-4a and the noise model is shown in Fig. 6-4b. Where c_s is cell capacitance and c_w is wiring capacitance. For this circuit, it is most convenient to derive the expression for equivalent input noise current I_{ni}. That equivalent is the sum of the noise currents entering the amplifier.

$$\overline{I_{ni}^2} = \frac{4\ k\ T}{R_L} + 2q\ I_{DC}\ \left(\frac{f_L}{f} + 1 \right) + \frac{4\ k\ T}{R_b} + \overline{i_n^2} + \frac{\overline{e_n^2}}{Z^2} \tag{6-7}$$

where

$$Z = \frac{j\ w\ L_p\ R_p}{R_p - w^2L_pc_pR_p + j\ w\ L_p},\ R_p = R_b\ //\ R_L$$

and the system gain is:

$$K_t = \frac{j\ w\ R_i R_b R_L L_p}{R_i R_b R_L \left[1 - w^2\ L_p\ (c_p + c_i) \right] - j\ w\ L_p\ (R_i R_L + R_i R_b + R_b R_L)} \qquad (6\text{-}8)$$

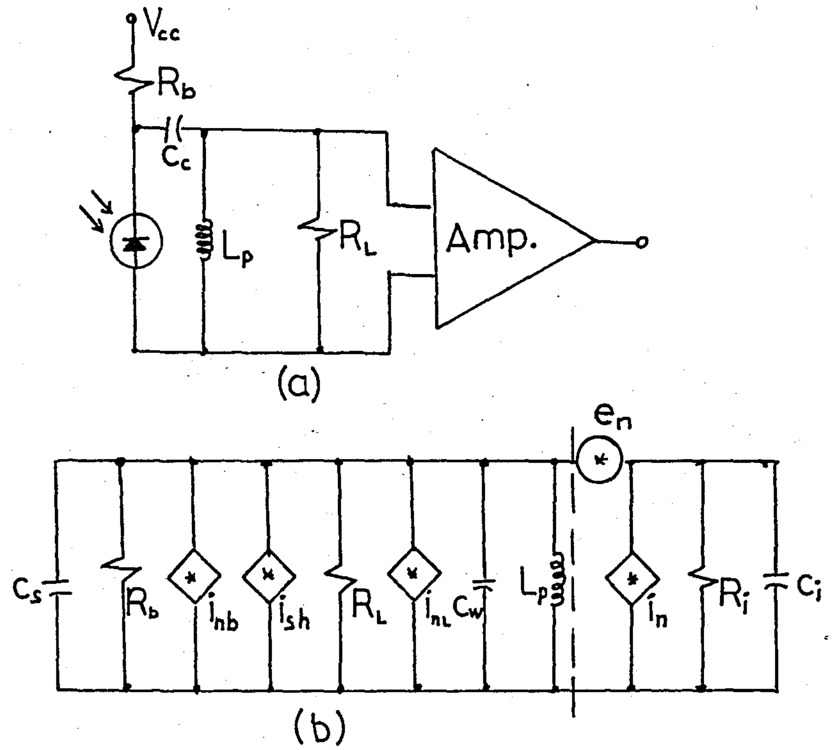


Figure 6-4 (a) System diagram for diode sensor

(b) Noise model for Computer

6.1.5 Transformer Model

There are three main reasons for using an input transformer to couple the signal source to the amplifier. The first is to transform the impedance of the source to match the noise resistance of the amplifier and therefore minimize the system noise figure. The second is to provide isolation between the source and amplifier. A third reason is for impedance matching to obtain maximum signal power transfer.

Altough the transformer can reduce the equivalent input noise of the amplifier, its own noise mechanism can contribute to the overall system noise. A system diagram and the small signal ac equivalent of a transformer-coupled input stage are shown in Fig.6-5 The transformer is represented by primary winding resistance r_p and primary inductance and reflected secondary series resistance r_s'. The prime designation is to indicate the reflected values.

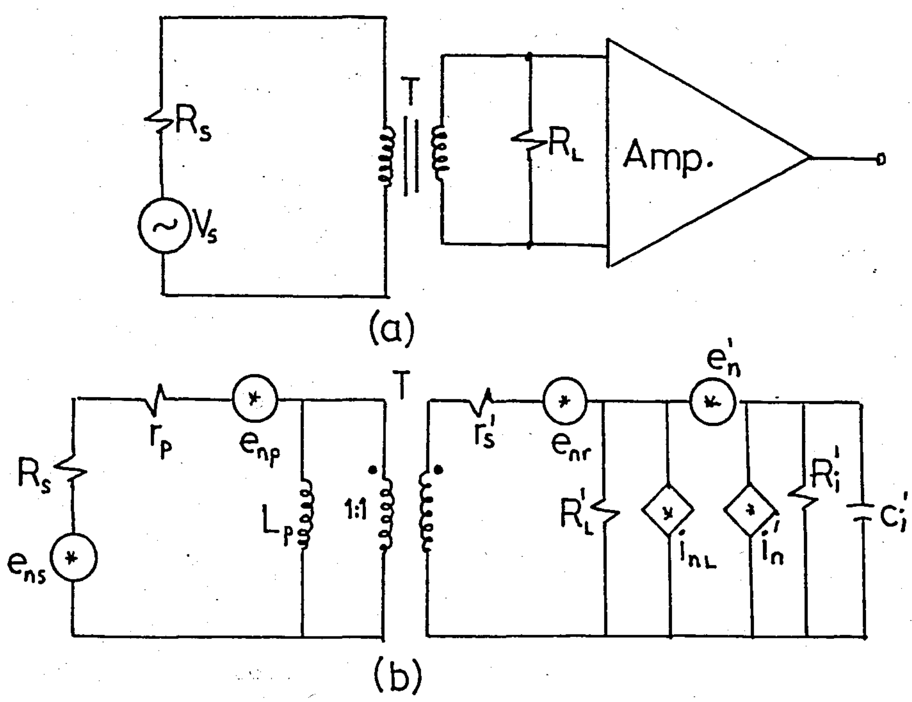


Figure 6-5 (a) System diagram for transformer coupled source.

(b) Noise model for computer

The results of circuit analysis are :

$$\overline{V_{ni}^2} = 4\ k\ T\ R_s + 4\ k\ T\ r_p + 4\ k\ T\ r_s' \left[1 + \frac{R_s + r_p}{jw\ L_p} \right]^2 +$$

$$\overline{e_n^2} \left[\frac{R_s + r_p + r_s' + R_L'}{R_L'} + \frac{(R_s + r_p)\ (r_s' + R_L')}{j\ w\ R_L'\ L_p} \right]^2 +$$

$$\left[\overline{i_n^2} + \frac{4\ k\ T}{R_L'} \right] \left[(R_s + r_p + r_s') + \frac{r_s'\ (R_s + r_p)}{j\ w\ L_p} \right]^2 \qquad (6\text{-}9)$$

and the system gain is:

$$K_t = \frac{j\ w\ L_p\ R_p}{R_A(r_s' + R_p) - w^2 L_p\ R_p\ C_i'\ (R_A + r_s') + j\ w \left[(R_A + r_s' + R_p + r_s'\ R_p\ C_i'\ R_A \right]} \qquad (6\text{-}10)$$

where $R_p = R_i' // R_L'$ and $R_A = R_s + r_s'$

6.1.6 Piezoelectric Sensor

Ferroelectric ceramic elements and quartz crystals are examples of piezoelectric transducer. These elements are used in microphones, sismic detectors, vibration sensors, and other devices where there is mechanical to electrical energy conversion. A system diagram and the noise model for the system are shown in Fig. 6-6. Where L_m and Cm are mechanical inductance and capacitance respectively, C_b is bulk capacitance.

The system equivalent input noise is:

$$\overline{V_{ni}^2} = 4\,k\,T\,R_s + \overline{e_n^2}\left[\frac{Z_s+Z_L}{Z_L}\right]^2 + \left[\overline{i_n^2} + \frac{4\,k\,T}{R_L}\right] Z_L^2 \qquad (6\text{-}11)$$

and the system gain is:

$$K_t = \frac{Z_L\,Z_i}{Z_s\,(Z_L+Z_i) + Z_i\,Z_L} \qquad (6\text{-}12)$$

where

$$Z_L = \frac{j\,w\,R_L\,L_x}{R_L - w^2 R_L\,L_x\,(c_b+c_p) + j\,w\,L_x}$$

$$Z_s = R_s - j\left(\frac{1 - w^2 L_m\,c_m}{w\,c_m}\right)$$

$$Z_i = \frac{R_i}{1 + j\,w\,R_i\,c_i}$$

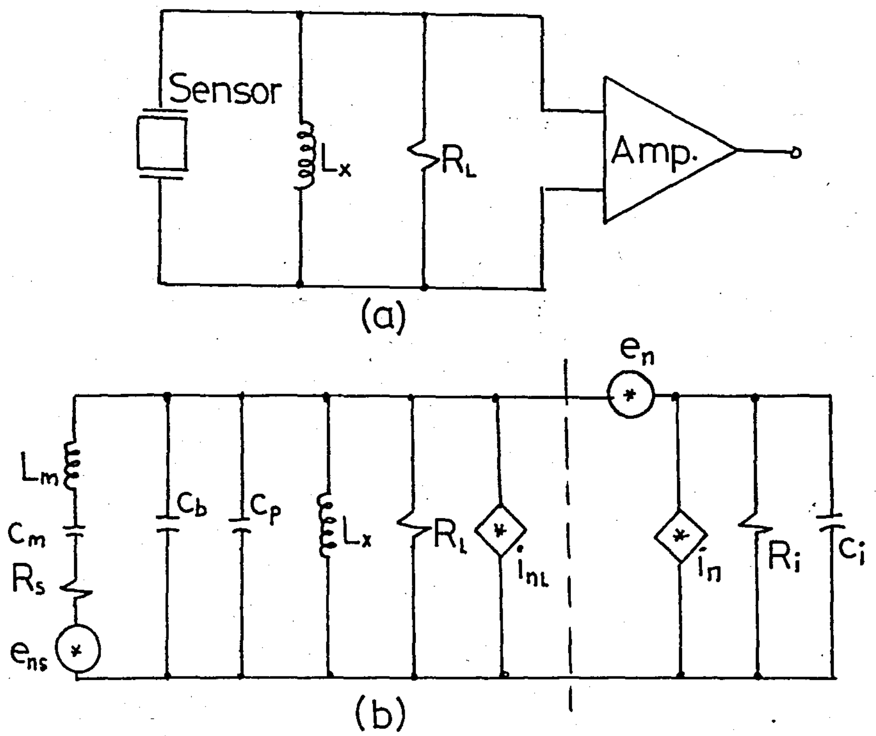


Figure 6-6 (a) System diagram for piezoelectric sensor

(b) Noise model for computer

VII DEVELOPMENT OF THE COMPUTER PROGRAM

7.1 INTRODUCTION

This program provides a noise analysis of an electronic system. The system is considered to be composed of three subsystems.

1. The sensor and its circuitry.
2. The amplifier or the first transistor stage.
3. The gain and frequency response of the system.

The program must be provided with data concerning sensor characteristics, response information, and frequency ranges. The program performs the following calculations [4].

1. Total equivalent input noise over a band of frequencies.
2. Input network frequency response.
3. Input noise at one frequency.
4. Input and output noise versus frequency.
5. Total noise at the output.
6. Total system gain.
7. Noise bandwidth.
8. Noise figure versus source resistance.

Calculation 4,6, and 8 are coupled in a plot routine.

Calculation 1 integrates and prints equivalent input noise V_{ni} for each frequency interval selected. This is the total noise at the input independent of amplifier gain and input impedance.

Calculation 2 determines the transfer function from the sensor to the amplifier input impedance. The frequency range must be selected. The program calculates at 10 frequency/decade stepping by $\sqrt[10]{10}$

Calculation 3 determines the equivalent input noise V_{ni} in a 1-Hz bandwidth at any frequency specified.

Calculation 4 determines the noise voltage spectral density at the input as a function of frequency and plots the noise versus frequency. The frequency range must be specified. The computer steps through the band, printing the noise at 10 frequency/decade as in calculation 2. Also the noise at the output of the amplifier is plotted as a function of frequency. The equivalent input noise is calculated and multiplied by the input network transfer function and the amplifier transfer function. This gives the total noise at the output after it has been amplified and equalized.

Calculation 5 integrates the total noise at the output of the amplifier over the intervals selected.

Calculation 6 prints and plots the total system gain versus frequency. For the selected frequency range, it prints the results at 10 frequency/decade. The total gain is the product of the input transfer function as determined in calculation 2 and the amplifier response as supplied in the data

Calculation 7 determines the noise bandwidth as discussed in chapter 2.

Calculation 8 determines and plots the noise figure versus source resistance at the any frequency selected.

7.2 SUBROUTINES

The system to be analyzed is shown in Fig. 7.1. The three major subdivisions are sensors, amplifier noise, and gain and response.

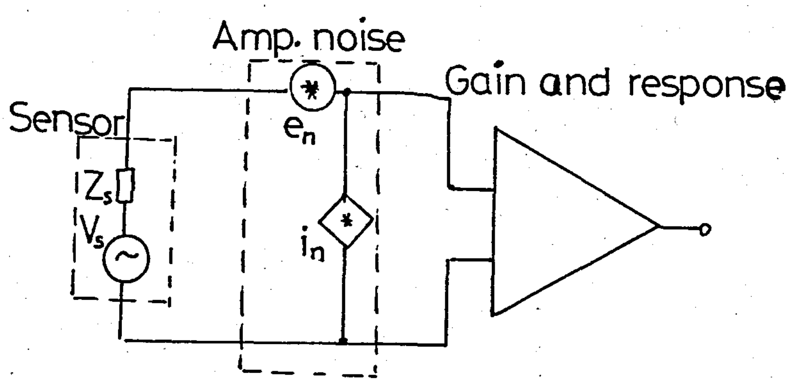


Figure 7.1 Subsystems

The user selects a sensor from the six available subroutines to describe a specific source. The data describing the source has to be entered. The subroutines are representative of sensors according to the following table :

Subroutine	Sensor model
SOURCE 1	Resistive source
SOURCE 2	Biased resistive source
SOURCE 3	Coil-RLC source
SOURCE 4	Biased diode source
SOURCE 5	Transformer model

SOURCE 6 Piezoelectric sensor

To represent the amplifier or first transistor, the user selects either the TMFET soubroutine or the BJT subroutine.

The third major section of the system is referred to as AMPLG. The AMPLG subroutine is concerned with the voltage transfer function of the post amplifier and equalizer.

7.2.1 "BJT" and "TMFET" Model

It is shown in chapter 4 that the noise of a bipolar transistor can be calculated from its hybrid- π values. The equivalent noise voltage and current generators e_n and i_n are represented in Eqs. 4-10 and 4-11. These equations are valid for a 1-Hz frequency band.

The equivalent noise voltage and current generators e_n and i_n for FET's are represented in Eqs. 5-8 and 5-10. Symbol equivalents used in BJT and TMFET are listed in Appendix B.

7.2.2 Amplifier and Equalizer Response "AMPLG"

Noise at the load depends on the system transfer function. The transfer function for each frequency-shaping network is stored in subprogram AMPLG. The Bode plot of each network is illustrated in Fig. 7.2.

By providing the proper values for the frequency corners, these curves can be combined to match most amplifiers and equalizers. The resulting transfer function is then the pdoduct of them all. The midband gain is referred to as K2 and must also form part of the data.

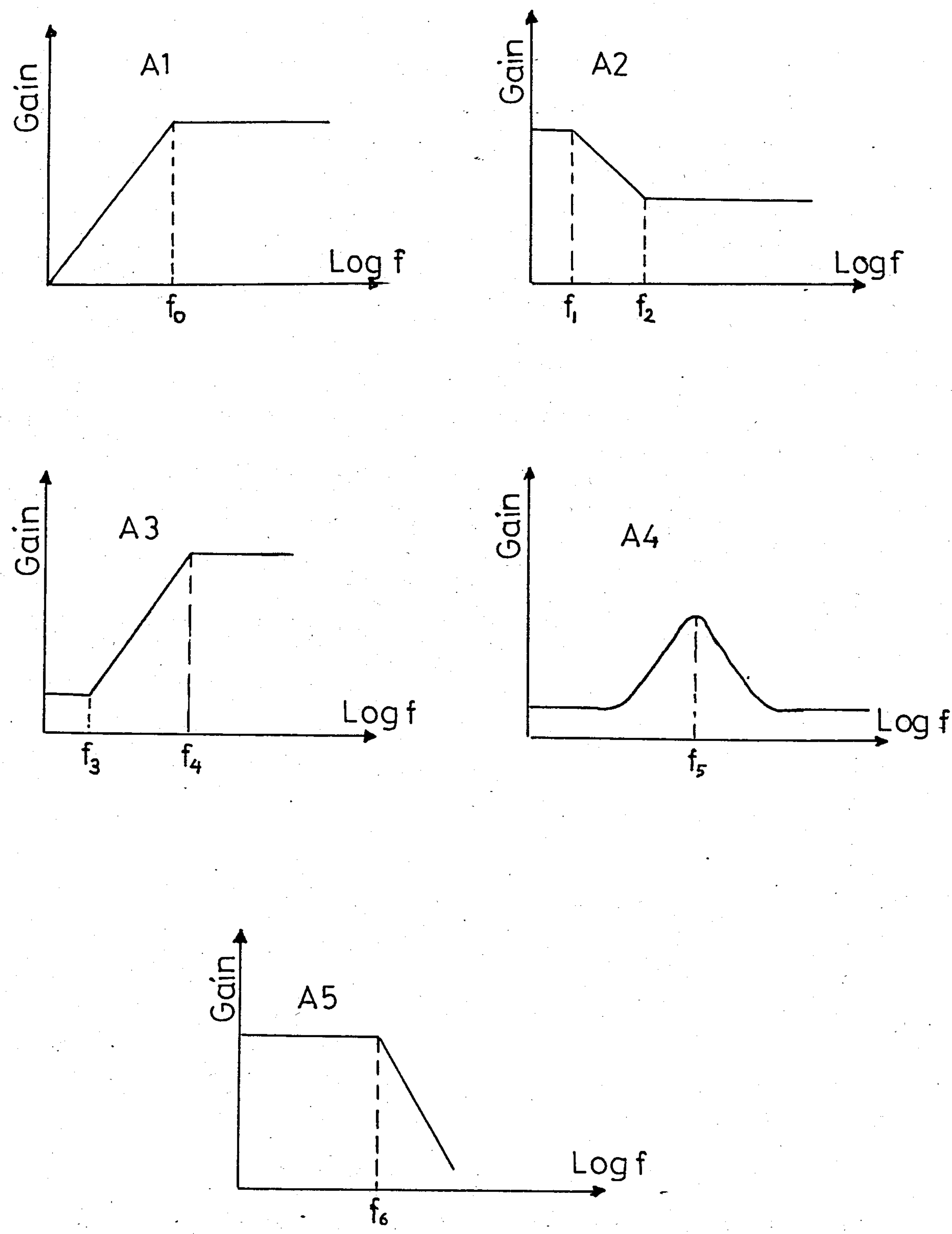


Figure 7-2 The Bode plots of the frequency-shaping networks

For the first-order high pass characteristic, the transfer function A1 is :

$$A1 = \left[\frac{f/f_o}{\sqrt{1 + (f/f_o)^2}} \right]^{N8} \qquad (7\text{-}1)$$

N8 = 1 corresponds to a slope of 6 dB/octave and N8 = 2 produces a 12 dB/octave roll-off.

The transfer function A5 is given by,

$$A2 = \left[\frac{(f/f_2)^2 + 1}{(f/f_1)^2 + 1} \right]^{1/2} \qquad (7\text{-}2)$$

The gain decreases linearly at 20 dB/decade from f_1 to f_2. It remains constant for frequencies higher than f_2.

The transfer function A3 is :

$$A3 = \left[\frac{(f/f_3)^2 + 1}{(f/f_4)^2 + 1} \right]^{1/2} \qquad (7\text{-}3)$$

This is a type of equalization curve used to linearize a sensor high-frequency roll-off.

The transfer function A4 for the resonant peak is :

$$A4 = \frac{f/f_5}{\left[\left((1-(f/f_5))^2 \right)^2 + \left(2f/f_5 \left(\sqrt{(1 + 1/Q)} - 1\right) \right)^2 \right]^{1/2}} \qquad (7\text{-}4)$$

This curve can describe an RLC resonant circuit or peaking as can be found in a feedback amplifier. The circuit Q is entered as QO.

For the first-order low pass characteristic, the transfer function A5 is described by,

$$A5 = \left[\frac{1}{\sqrt{1 + (f/f_6)^2}} \right]^{N9} \qquad (7\text{-}5)$$

The total amplifier and equalization gain is the product of these five transfer functions and gain constant K2, the reference gain.

7.2.3 Sensor Subroutines

The equivalent input noise for the six sources is given in chapter 6. Symbol equivalents used in SOURCE subroutines are summerized in the Appendix B.

In the calculations 1,3,4, and 5 the computer prints the four output separately. The first output called SENSOR EN is the thermal noise of the sensor. The second output called LOAD EN is the thermal noise of the load resistance R_L and the other coupling network noise. The third output called AMPL EN is the contribution of the noise voltage of the amplifier. Then the fourth output called AMPL IN; ZS is the contribution of the noise current of the amplifier. Thus, by looking at the computer printout we can see the noise contribution of each section; sensor noise, coupling network noise, and amplifier noise. The total noise is also printed out.

VIII MEASUREMENT OF BASE-SPREADING RESISTANCE

8.1 EQUIVALENT NOISE CIRCUIT OF A SINGLE STAGE TRANSISTOR AMPLIFIER [24]

The low-frequency hybrid-π noise model for a common-emitter amplifier stage is shown in Fig. 8.1 at low frequencies [8]. Where e_n and i_n represent the equivalent noise voltage and equivalent noise current sources respectively.

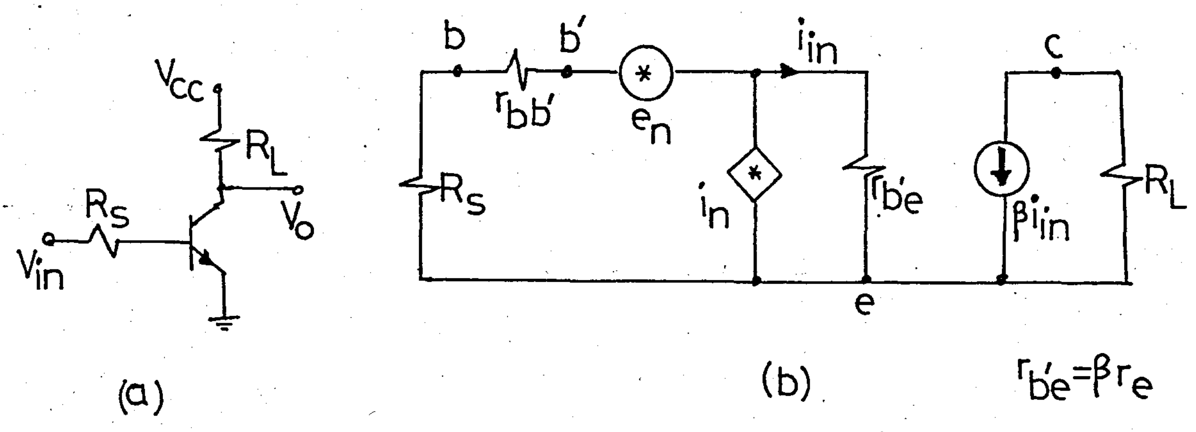


Figure 8.1 (a) A Common-emitter transistor circuit.
(b) The low-frequency noise equivalent circuit

$$e_n = (\overline{e_n^2})^{\frac{1}{2}} = \left[4\ k\ t\ \Delta f\ (r_{bb'} + \frac{r_e}{2}) \right]^{\frac{1}{2}} \tag{8-1}$$

$$i_n = (\overline{i_n^2})^{\frac{1}{2}} = \left[4\ k\ T\ \Delta f\ (\frac{1}{2\beta\ r_e}) \right]^{\frac{1}{2}} \tag{8-2}$$

The noise factor F can be written as follows :

$$F = 1 + \frac{(R_s + r_{bb'} + r_e)^2}{2\beta\ r_e\ R_s} + \frac{(r_{bb'} + r_e/2)}{R_s} \qquad (8\text{-}3)$$

The noise factor is independent of transistor configuration, but depends on the d.c collector current and the transistor type. For the optimal value of R_s, noise factor given in Eq. 8-3 will be minimized.

$$R_s(\text{opt}) = (2\beta\ r_e\ (r_{bb'} + \frac{r_e}{2}))^{\frac{1}{2}} \qquad (8\text{-}4)$$

$$F(\text{min}) \simeq 1 + (\frac{2r_{bb'}}{\beta\ r_e} + \frac{1}{\beta})^{\frac{1}{2}} \qquad (8\text{-}5)$$

For small values of collector current the minimum value of noise factor also becomes smaller. When the condition $I_C << V_T/r_{bb'}$ is satisfied one can define F, R_s(opt), and F(min) as follows :

$$F \cong 1 + \frac{(R_s + r_e)^2}{2\beta\ r_e\ R_s} + \frac{r_e}{2R_s} \qquad (8\text{-}6)$$

$$R_s(\text{opt}) = \sqrt{\beta}\ r_e \qquad (8\text{-}7)$$

$$F(\text{min}) = 1 + \beta^{-\frac{1}{2}} \qquad (8\text{-}8)$$

It is clearly seen that the equations above are base-spreading resistance independent. On the other hand, the collector current which minimizes the noise factor is:

$$I_C(opt) = \frac{V_T\ (1+\beta)^{\frac{1}{2}}}{(R_s + r_{bb'})} \tag{8-9}$$

It can be simplified as :

$$I_c(opt) \simeq \sqrt{\beta}\ \frac{V_T}{R_s} \tag{8-10}$$

When a low noise amplifier is designed for a known source resistance, the first stage transistor operating point collector current can be found from the above equation.

For the small source resistance, p-n-p transistor is preferred to n-p-n type, because p-n-p's have lower base-spreading resistance $r_{bb'}$.

When the base-spreading resistance is not negligible, in order to calculate the values of $R_S(opt)$ and $F(min)$ $r_{bb'}$ must be defined accurately. The base-spreading resistance can be best determined by noise measurement methods [9]. Gibbons and Chenette proposed a method for determining $r_{bb'}$ from the 1/f noise data ; Hsu proposed the method of measurement from thermal noise data [10,11,12,13].

8.2 THE 1/F NOISE REGION

For the frequencies at which 1/f noise is effective (f < 1 k Hz), the noise current source i_{f_1} representing the base emitter junction surface 1/f noise, and the noise current source i_{f_2} representing active base region 1/f noise should be added (Fig.8-2). The power density spectra of i_{f_1} and i_{f_2} are proportional to the bias current of the transistor and inversely proportional to the frequency [16].

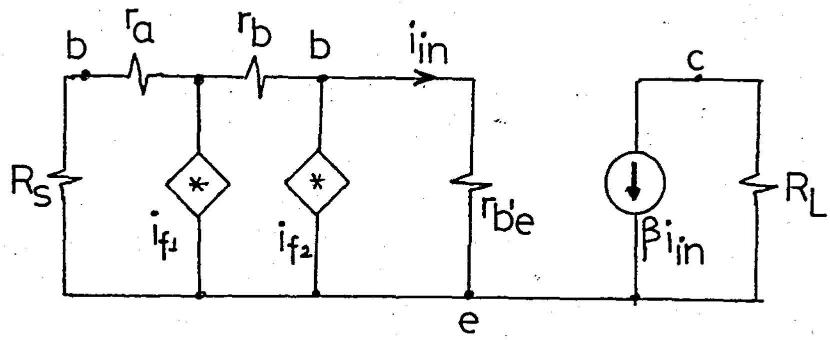


Figure 8 2 The low-frequency noise equivalent circuit of a common-emitter transistor $(r_a + r_b = r_{bb'})$.

$$\overline{i^2_{f1}} = K_1 \, I_B \, \frac{\Delta f}{f} \tag{8-11}$$

$$\overline{i^2_{f2}} = K_2 \, I_B \, \frac{\Delta f}{f} \tag{8-12}$$

The 1/f noise equivalent voltage sources can be obtained as the product of the resistors parallel to i_{f_1} and i_{f_2} noise current sources by these two current sources correspondingly.

The experiments show that the resistance r_a parallel to i_{f1} noise current source should be smaller than the base-spreading resistance $r_{bb'}$ [13]. The total base-spreading resistance can be considered to be composed of two sections: the relatively small resistance from the base contact to the edge of the emitter junction (r_a) and a larger resistance lying beneath the emitter (r_b).

For the frequencies at which the other low-frequency noise sources are negligible, the noise factor of a common-emitter transistor amplifier is:

$$F = 1 + \frac{\overline{i_{f1}^2}\,(R_s + r_a)^2}{4\,k\,T\,R_s\,\Delta f} + \frac{\overline{i_{f2}^2}\,(R_s + r_{bb'})^2}{4\,k\,T\,R_s\,\Delta f} \tag{8-13}$$

The value of R_s which minimizes the noise factor is:

$$R_s(\min) = \left[\left(r_a^2 + \frac{\overline{i_{f2}^2}\, r_{bb'}^2}{\overline{i_{f1}^2}}\right) \Big/ \left(1 + \frac{\overline{i_{f2}^2}}{\overline{i_{f1}^2}}\right) \right]^{\frac{1}{2}} \tag{8-14}$$

In the equivalent noise circuit, the current sources i_{f1} and i_{f2} can be replaced by a single noise current source i_f to be connected after $r_{bb'}$.

Accordingly the power spectrum of noise current source i_f can be written as [15]:

$$\overline{i_f^2} = 4\,k\,T\,\frac{\rho_o}{r_{b'e}^2}\,\frac{\Delta f}{f} \tag{8-15}$$

ρ_o = the equivalent noise resistance which equals to 1/f noise of transistor with the base open circuit, at 1-Hz.

$$r_{b'e} = \beta\, r_e = \beta\, V_T/I_C$$

The value of ρ_o can be considered to be constant when the collector biasing current is between 10 μA and 1 mA. The dependence of 1/f noise on the operating point of the transistor can be obtained from the dependence of input resistance $r_{b'e}$ on the collector current.

For the values of collector currents I_C in the order of microamperes, the current gain β is changed as $\beta \propto I_C^{1/3}$. That is why the effective value of $1/f$ noise is changed as $I_C^{2/3}$.

In $1/f$ noise region, the noise factor is as :

$$F = 1 + \overline{i_f^2}\,\frac{(R_s + r_{bb'})^2}{4\,k\,T\,R_s\,\Delta f} \tag{8-16}$$

from Egs. 8-15 and 8-16

$$F = 1 + \frac{1}{f}\,\frac{\rho_o}{r_{b'e}^2}\,\frac{(R_s + r_{bb'})^2}{R_s} \tag{8-17}$$

For $R_S(min) = r_{bb'}$ the noise factor will be minimum :

$$F(min) = 1 + \frac{4}{f}\,\frac{\rho_o}{r_{b'e}^2}\,r_{bb'} \tag{8-18}$$

8.3 MEASUREMENT OF BASE SPREADING-RESISTANCE [24]

The noise factor of a low noise amplifier which operates in $1/f$ noise region will be minimum, if R_S equals to $r_{bb'}$. For determining optimum source resistance and calculating the noise factor, the exact value of the base-spreading resistance $r_{bb'}$ should be known.

In the method proposed by Gibbons and Chenette, for a suitably selected source resistor, emitter current, and collector d.c. voltage, the noise factor F is measured in the 1/f noise region [10,11].

When the value of the collector d.c. voltage is sufficiently low and the value of the emitter current is sufficiently high, 1/f noise factor is changed proportionally with respect to $(R_s + r_{bb'}) / R_s$ [Eq. (8-16)]. Then the value of R_S which minimizes the noise factor will be equal to the base-spreading resistance $r_{bb'}$.

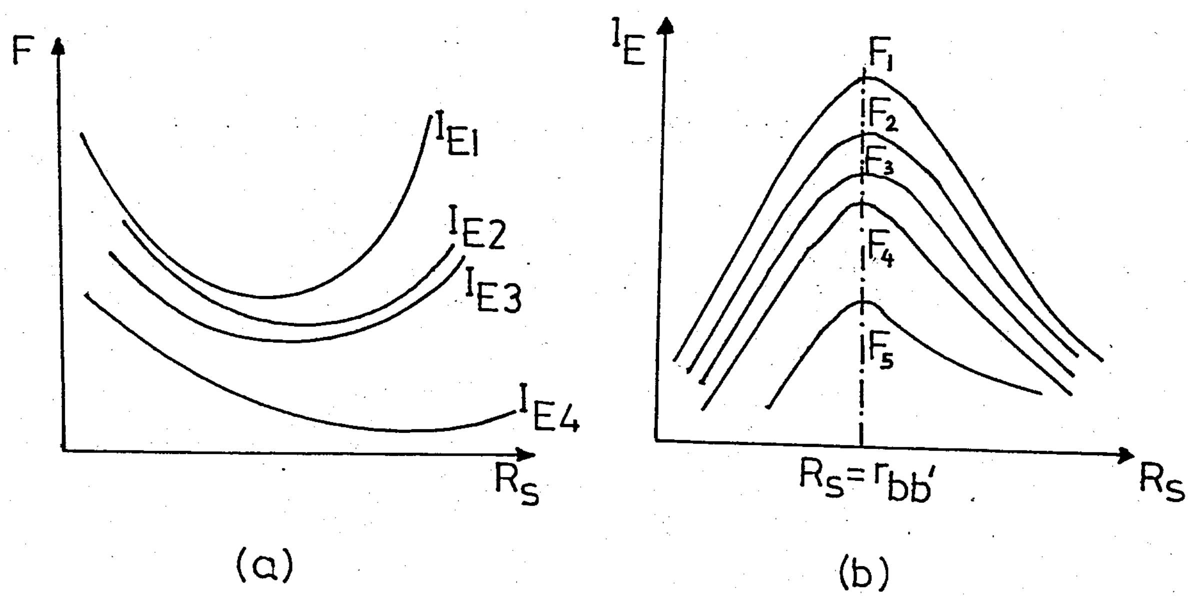


Figure 8-3 (a) The noise figure versus R_S for different values of I_E.

(b) The noise figure contours.

Since the minimum points are not changed for different emitter currents, as it is shown in Fig.8-3, the minimum source resistance value is obtained from the curves. When compared with the others, it is not a very accurate method because one reads a minimum region rather than a minimum point for R_S [11,12].

The method proposed by Chenette has been adopted for $r_{bb'}$

measurements. Transistor noise is measured for the different values of source resistor R_s in the 1/f noise region, the value of Rs(opt) which minimizes the noise is found [13]. By repeating the experiments for different collector operating currents, the values of R_s(opt) are plotted as a function of V_T/I_E. The point at which the line intercepts the R_s(opt) axis, R_s(opt) is equal to $-r_{bb'}$ (Fig. 8-4).

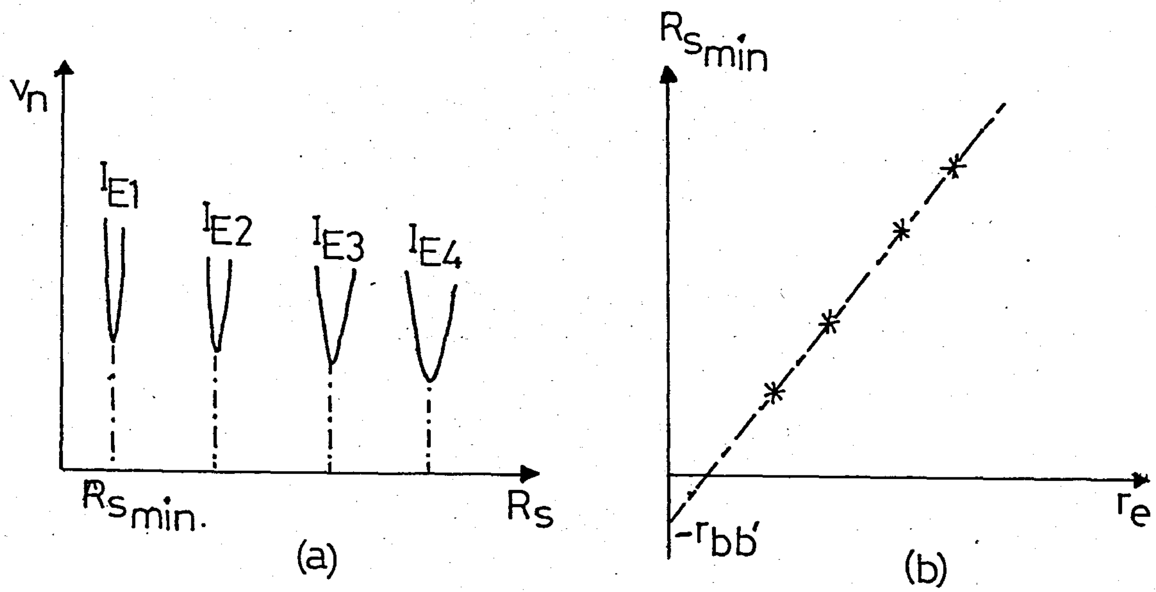


Figure 8-4 (a) Relative noise power output versus R_s for different values of $I_E (I_{E1} > I_{E2} > I_{E3} > I_{E4})$
(b) R_s(opt) versus $r_e = (V_T/I_C)$

8.4 MEASUREMENT OF $r_{bb'}$ IN HIGH-GAIN TRANSISTORS

In the high-gain transistors, the flicker noise is found to be much smaller than that of a low-gain transistor. The reason is that in order to achieve large current gain the generation-recombination center and the surface state density in the device have to be greatly reduced. Therefore thermal noise measurement methods are used in place of methods of 1/f noise region measurement.

When the thermal noise of the base-spreading resistance suppresses all other noise sources, the measured noise power is the thermal noise of $r_{bb'}$.

By selecting the operating point such that $r_{bb'} > r_{b'e}/\beta$ is satisfied for a common-emitter transistor, it can be shown that the output short-circuit current, when the input terminal is a.c. shorted, is given by

$$\overline{i_{no}^2} = \frac{4\ k\ T\ r_{bb'}\ \beta^2\ \Delta f}{(r_{bb'} + r_{b'e})^2} \tag{8-19}$$

If the transistor is biased in the small collector currents region ($I_C < 10\ \mu A$ and $r_{b'e} \gg r_{bb'}$), shot noise is the dominant noise. Then $\overline{i_{no}^2}$ is changed in proportion to the squared value of collector current;

$$\overline{i_{no}^2} = \frac{4}{V_T}\ r_{bb'}\ I_C^2\ \Delta f \tag{8-20}$$

The base-spreading resistance $r_{bb'}$ is found easily from the Eq.8-19 and Eq. 8-20 .

As an alternative method, the output short-current is measured when a R_S resistance which satisfies the condition $(R_S + r_{bb'}) \ll r_{b'e}$ is connected to the input terminal.

$$i_{no}^2 = \frac{4}{V_T}\ (r_{bb'} + R_s)\ I_C^2\ \Delta f \tag{8-21}$$

When the output noise current is plotted as a function of source resistor R_s, the point at which the line intercepts the R_s axis will give the base-spreading resistance $r_{bb'}$ directly.

8.5 MEASURING EQUIPMENT

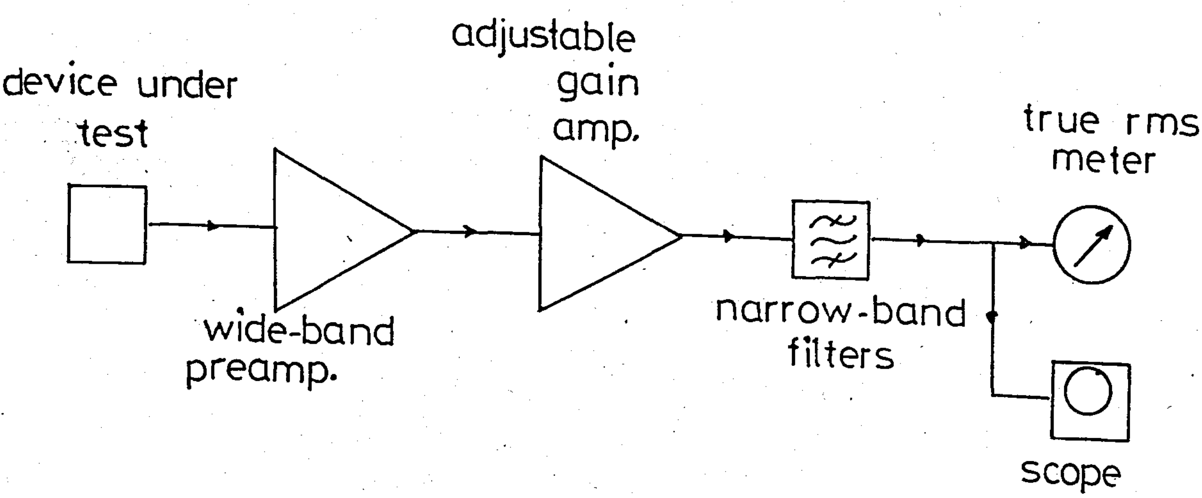


Figure 8.5 Block diagram of the measuring system

A block of the noise measurement set-up is shown in Fig 8.5. The equipment used the base-spreading resistance measurement comprises the following devices.

8.5.1 Transistor Test Circuit

Altough the noise of a transistor is about the same for all three configurations, it is normally measured in the common-emitter configuration. The biasing network must not contribute additional noise. The low noise biasing method shown in Fig. 8.6 has been adopted [4].

The network contains a resistor R_D connected between the junction of R_A and R_B and the transistor base. The capacitor C_B is chosen so that its reactance at the lowest operating frequency is small compared to the resistor R_B. This C_B provides an a.c. ground that eliminates any noise generated in R_A and R_B from getting to the transistor. The only biasing resistance that contributes noise is R_D. The d.c. drop accorss R_D need not be large. Because R_D parallels the source resistor, its thermal noise is important. However attenuation of that noise is present: R_S/R_D. It can be shown that the maximum thermal noise contribution exists when R_D is low valued.

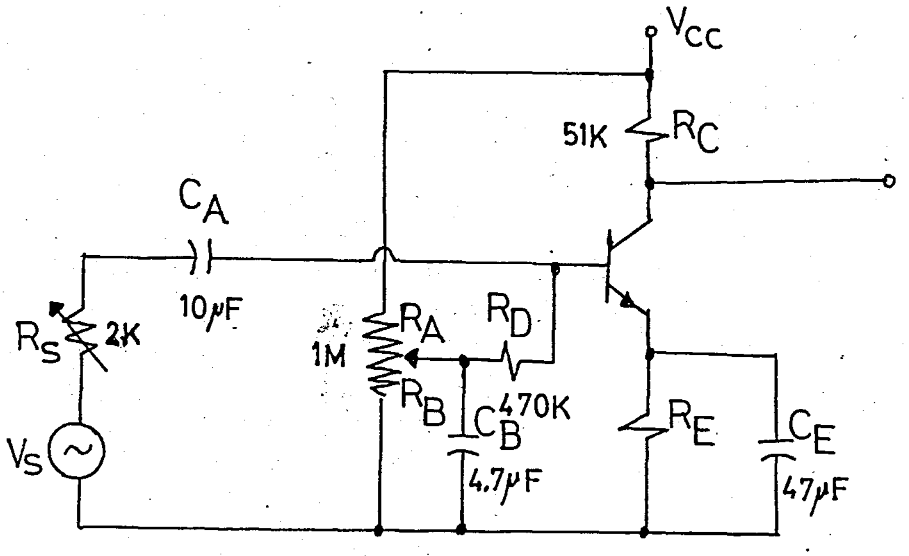


Figure 8.6 The Transistor stage with noiseless biasing

The reactance of C_A X_{CA} must be much less than R_s at the lowest frequency of interest. We would select the reactance of C_B to be no greater than 0.1 R_D at the lowest frequency of interest. To meet the noise specification C_E must effectively bypass the noise of R_E. The impedance of C_E-R_E network must be low compared to $(R_T + R_i + R_e)/\beta$, where R_T is the Thevenin equivalent of all resistance to the left of the base terminal, and the R_i is the input resistance of the transistor alone. The resistor R_D is of low noise type.

8.5.2 The Wide Band- FET Input Stage Preamplifier [24]

(Fig.8-7) offers a nominal gain of 1000 in a 50 kHz bandwidth. The common-drain FET configuration is considered for the input device. As a further precaution the preamplifier is battery driven (2x9 Volt batteries). [25]

If the amplifier is represented by an equivalent noise resistance R_n in series with the input of noiseless amplifier then the noise output power with white noise input from a known resistance R will be proportional to $(R_n + R)$, Fig. (8-8) [25]. For the conditions $R = 0$ and $R >> R_n$ the measured noise powers are V_1^2 and V_2^2 respectively, and the ratio $(V_1^2 R/V_2^2)$ yields the noise resistance R_n. The experimental results are plotted in Fig.(8-9) for the range of frequncies of interest. All measurements will be relative to the reference resistor R.

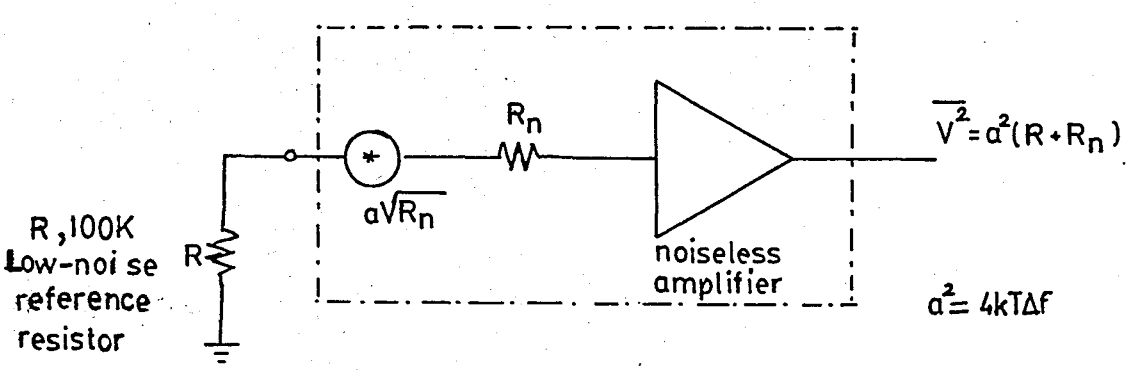


Figure 8.8 Equivalent representation of the amplifier by its noise resistance in series with the input of an identical but noise free amplifier

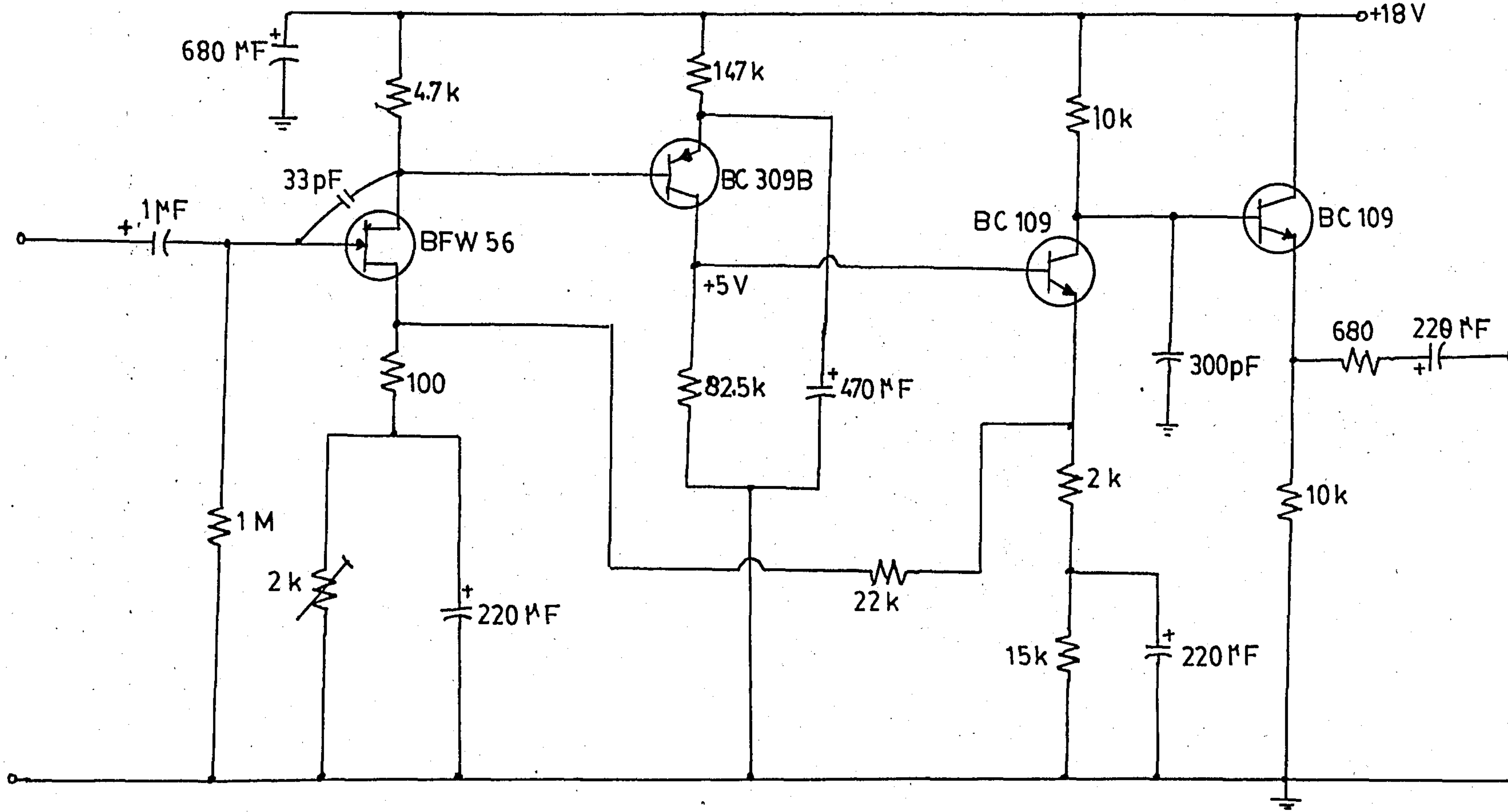


Figure 8.7 The low-noise FET input stage preamplifier

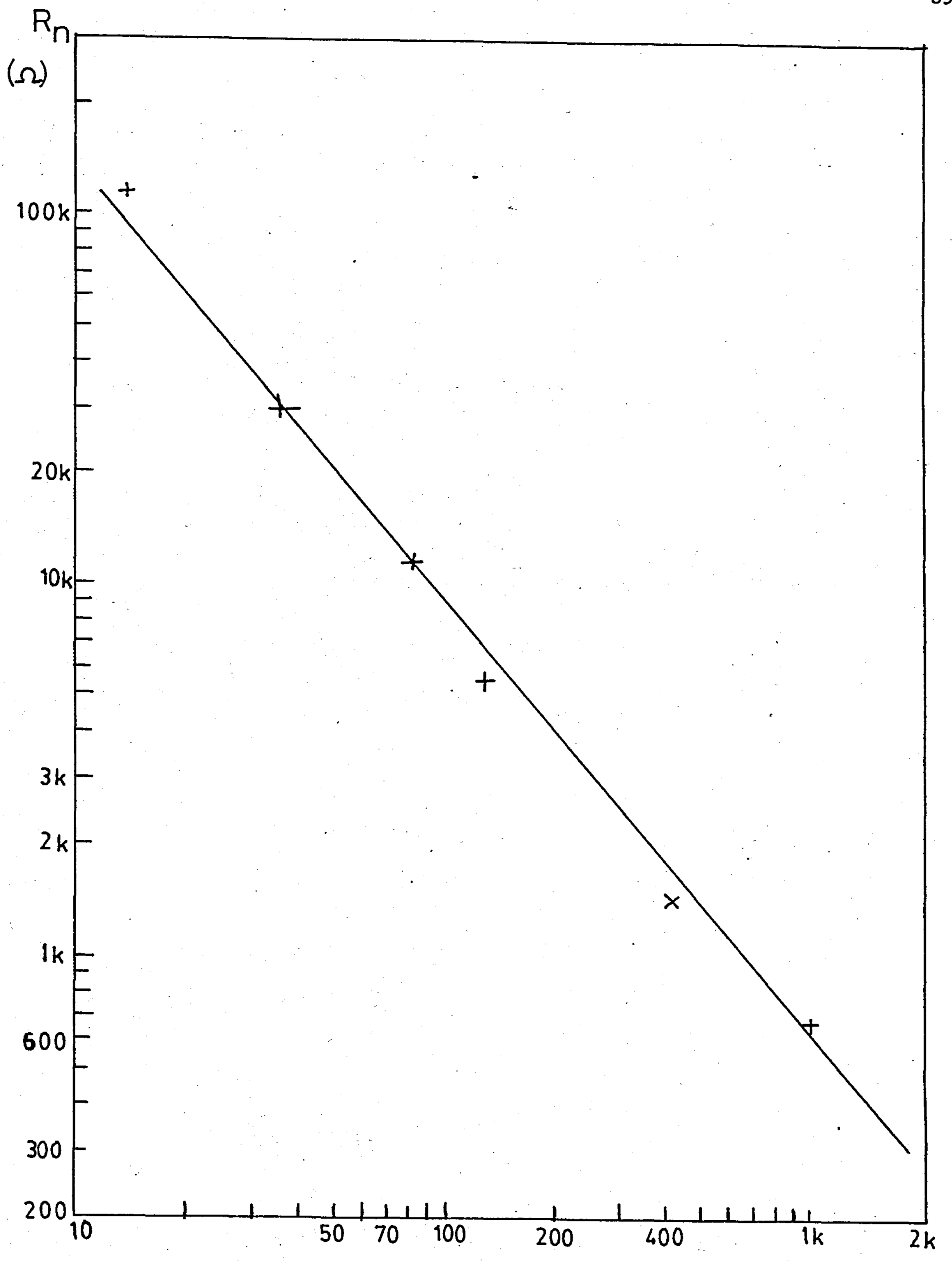


Figure 8.9 Equivalent series noise resistance of the preamplifier at room temperature.

8.5.3 Adjustable-Gain Amplifier

Fig. (8-10) shows the adjustable-gain amplifier and the buffer stage used in the measuring instrumentation. The gain can be adjusted between -10 dB and 10 dB in 10 dB steps.

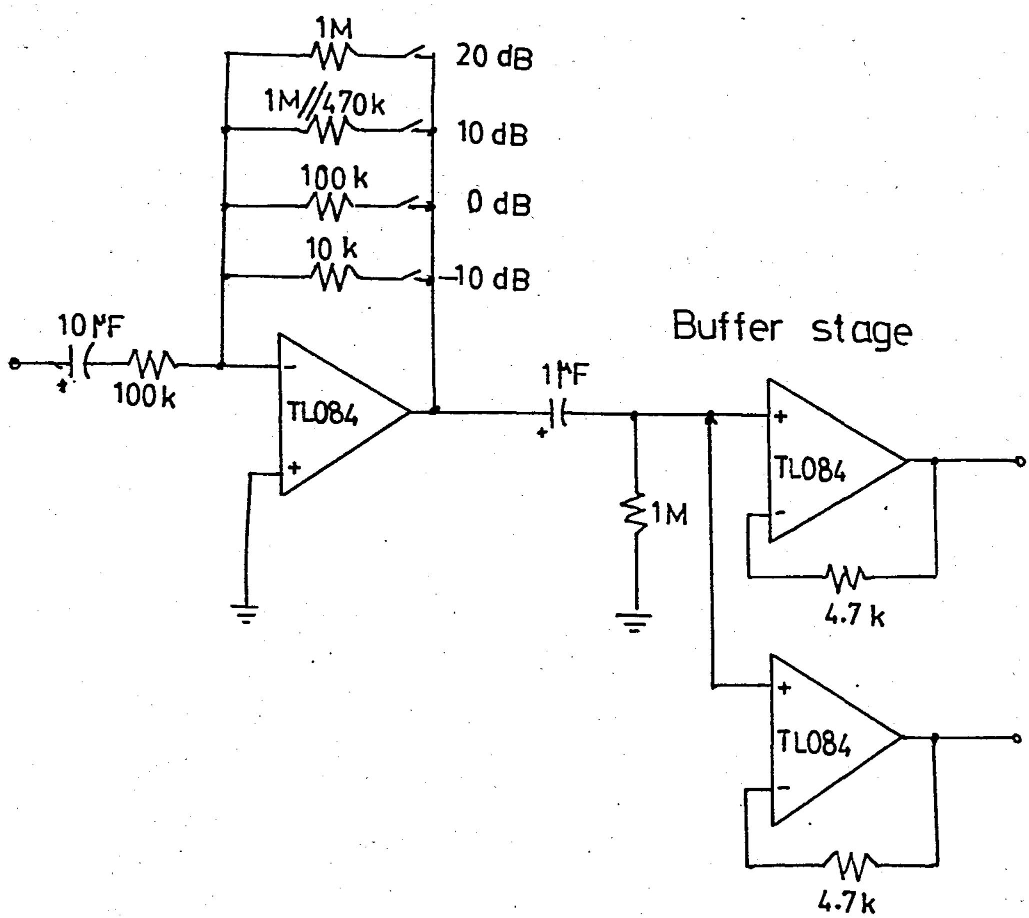


Figure 8.10 Adjustable-gain amplifier and buffer stage

8.5.4 Band-Pass Filter

Seven BIQUAD band- pass filters centered at different frequencies are built and used in the measurement (14 Hz, 35 Hz, 84 Hz, 424 Hz, 1 KHz, and 10 kHz). A BIQUAD band-pass filter is shown in Fig. (8.11) and the design equations are as follows :

The center frequenciy f_o and the Q factor are :

$$f_o = \frac{1}{2\pi RC} \qquad Q = \frac{R_Q}{R}$$

The bandwidth : $BW = \dfrac{f_o}{Q}$

The gain at resonance : $G(f_o) = \dfrac{R_Q}{R_{in}}$

R_{in} is adjusted to make the gain and the square root of bandwidth product constant ($G(f_o)\ \sqrt{BW} = 100$) for each filter. The component values, the center frequency, and the bandwidth for seven filters are listed in Table 8.1

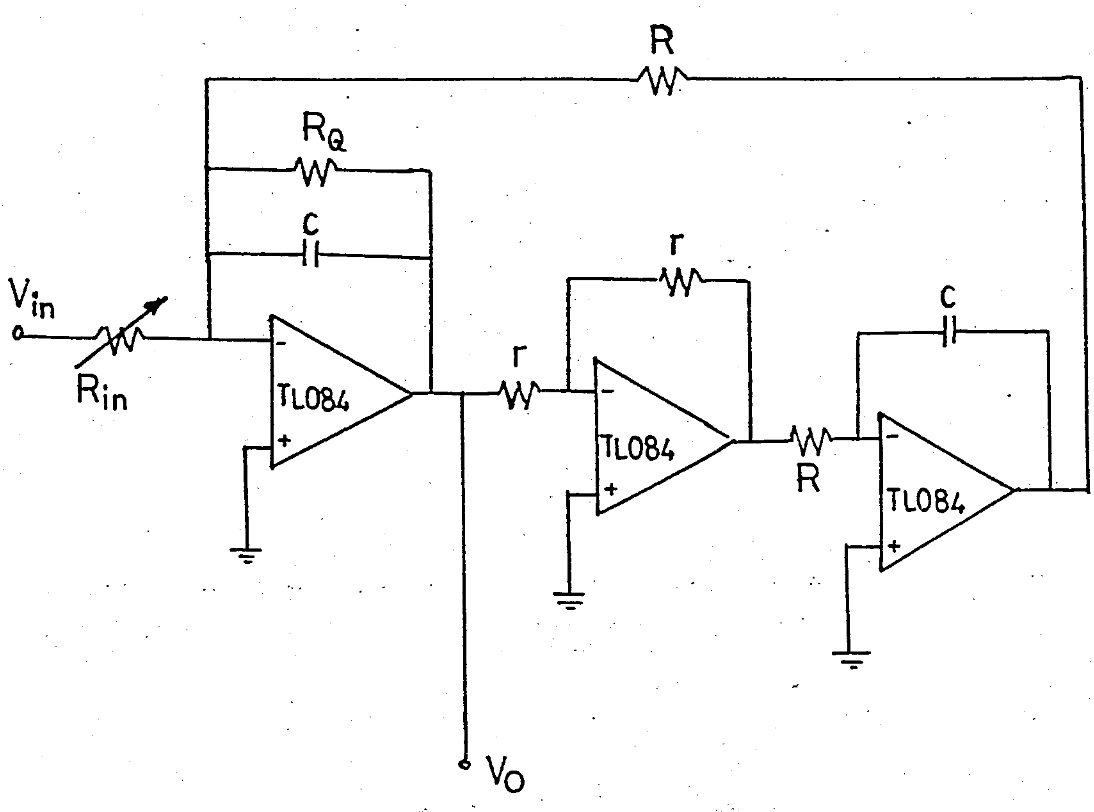


Figure 8.11 BIQUAD filter

f_o	R_{in} (Ω)	R_Q (Ω)	R (Ω)	C	G	BW
10 kHz	243 k	1 M	56.6 k	270 pF	4	590 Hz
1 kHz	77 k	1 M	56.6 k	2.7 nF	13	59 Hz
424 Hz	47 k	1 M	51 k	7.35 nF	21.5	22 Hz
128 Hz	27 k	1 M	56.6 k	22 nF	37	7.25 Hz
84 Hz	22 k	1 M	56.6 k	33.4 nF	45	4.75 Hz
35 Hz	13.33 k	1 M	51 k	90 nF	75	1.80 Hz
14 Hz	5.8 k	470 k	51 k	220 nF	81	1.60 Hz

Table 8.1 The component values and the design parameters for the BIQUAD filters with $G \times \sqrt{BW} = 100$.

IX. SAMPLE PROGRAM AND EXPERIMENTAL RESULTS

9.1 A SAMPLE PROGRAM

The program is used to analyze the noise in a magnetic cartridge circuit shown in Fig. 9.1.

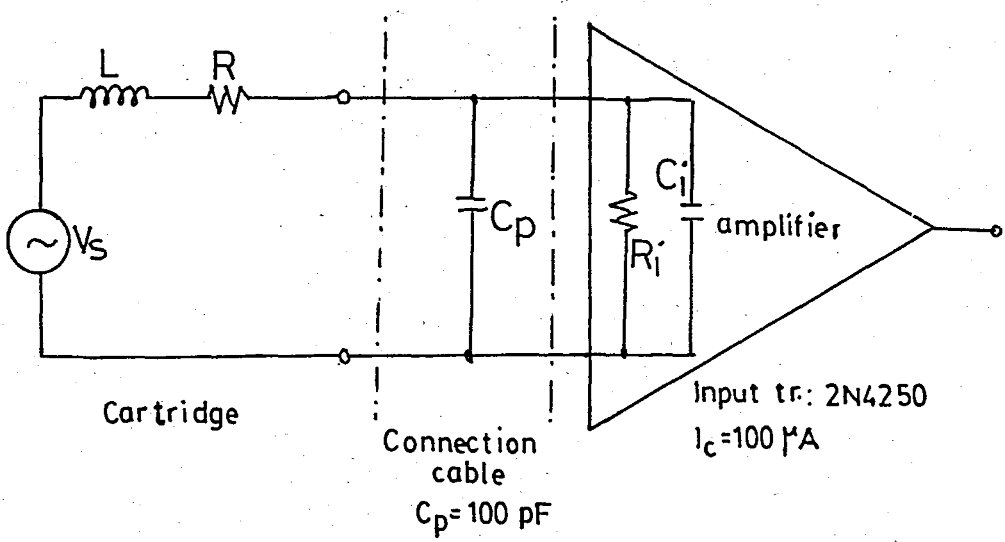


Figure 9.1 The equivalent circuit diagram for a magnetic cartridge.

Three cartridges from different manufactures are chosen as an illustration to the RLC sensor circuit. The characteristics for each type are listed in Table 9.1

Cartridge types	L	R	R_i	C_i
(1) SHURE V15 - III	500 mH	1350 Ω	47 kΩ	400 pF
(2) STANTON 881 S	510 mH	900 Ω	47 kΩ	400 pF
(3) ORTOFON SME 30 H	500 mH	600 Ω	47 kΩ	275 pF

Table 9-1

The frequency response of the magnetic cartridge preamplifier is as shown in Fig 9-2.

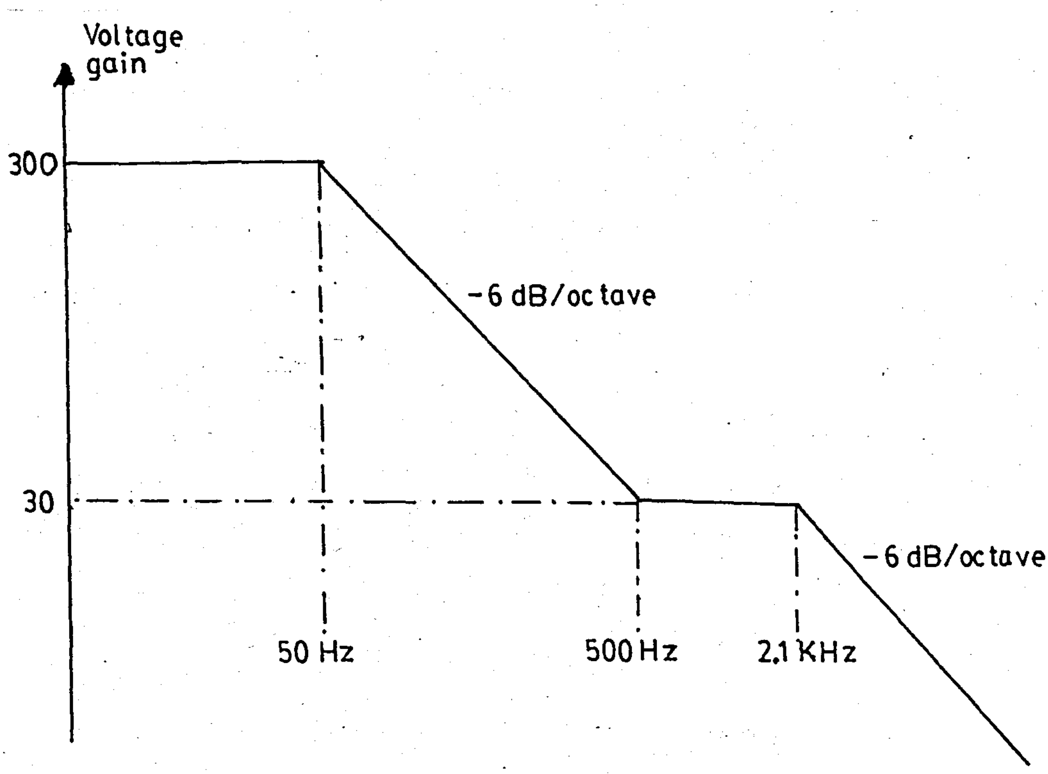


Figure 9.2 The frequency response of cartridge preamplifier according to RIAA Replay characteristic.

The program requires that the user assembles13 data cards.

The first card contains the following

1. Coloumn 1 : the sensor model (a number from 1 to 6)
2. Coloumn 2 : the input device (1 for BJT, 2 for TMFET)
3. Coloumn 3 : number of calculations

Card 2 is used for BJT or TMFET model. If the user selects BJT, he must supply meaningful values for all nine locations. Symbol equivalents are given Appendix B. When the TMFET is used the user must supply meaningful values for four locations. Source data are contained in cards 3 and 4. A value for each element must be provided even if that element is not present in the noise model of the system. For example, a resistive sensor as discussed in Section 6.1.1 does not have elements C2, R6, L1, and L2. Harmless constants must be provided for these locations on the card. A set of such constants for the cartridge (SHURE V15 III) is shown in Appendix C. Card 5 is used to enter data for AMPLG. For this data card, Harmless constants are given by Appendix C. This example requires that the values for the break frequencies F0, F1, F2, F6, and the midband gain K2 must be supplied. A2 and A5 representations discussed in Section 7.2.2 are used with N9 = 1. All values in card 5 must be provided as shown in Appendix C.

Data cards from 6 to 13 contain information regarding the frequency or frequencies to be used in each calculation. These data are listed in Appendix D. for the example being considered.

9.2 SAMPLE RESULTS

The results of calculation 1 for the cartridge "SHURE V15 III" are listed in Table 9.2.

TOTAL EQUIVALENT INPUT NOISE OVER A BAND

VOLTAGE SOURCE MODEL

FREQUENCY	SENSOR EN	LOAD EN	AMPL EN	AMP IN;ZS	SUM NOISE
1. TO 50.	32.87	0.00	21.27	12.83	41.20
50. TO 500.	100.30	0.00	64.77	15.26	120.37
500. TO 2100.	189.13	0.00	121.68	59.75	232.70

TOTAL NOISE VOLTAGE= 265.203 NANOVOLTS

Table 9.2 Total equivalent input noise over a band

The computer calculates and integrates the noise in each frequency band. It prints the r.m.s. value of the noise contribution of each noise generator as well as the sum referred to the input side.

For the input network transfer function, that is the gain of the input network including the amplifier input impedance, the output of calculation 2 is shown in Table 9.3 (for "SHURE V15-III" Cartridge)

Calculation 3 determines the equivalent input noise at the selected frequencies. The computer output for the cartridge "SHURE V15-III" is shown in Table 9.4

INPUT NETWORK FREQUENCY RESPONSE

FREQUENCY	GAIN	GAIN DB
30	.972	-.246
37	.972	-.246
47	.972	-.246
59	.972	-.246
75	.972	-.246
94	.972	-.245
119	.972	-.245
150	.972	-.245
189	.972	-.244
238	.972	-.242
299	.973	-.240
377	.973	-.237
475	.974	-.232
598	.975	-.224
753	.976	-.210
948	.978	-.190
1194	.982	-.157
1503	.988	-.104
1892	.998	-.021

Table 9.3 Input network frequency response.

The equivalent input noise versus frequency is determined from calculation 4. The program calculates and prints the noise at each of 10 frequency/decade over the range requested. As it is shown in Table 9.5, at about 20 kHz which is the resonance frequency ($w^2 R_S L_S = 1$), the contribution of the noise voltage generator of the transistor (e_n) is minimum.

FREQUENCY	SENSOR EN	LOAD EN	AMPL EN	AMP IN:ZS	SUM NOISE
40	4.73	0.00	3.06	1.08	5.73
600	4.73	0.00	3.05	.83	5.69
15000	4.73	0.00	1.70	14.49	15.33
20000	4.73	0.00	.64	19.27	19.86

Table 9.4 Input noise at selected frequencies.

INPUT NOISE VERSUS FREQUENCY

VOLTAGE SOURCE MODEL

FREQUENCY	SENSOR EN	LOAD EN	AMPL EN	AMP IN;ZS	SUM NOISE
10	4.73	0.00	3.06	2.04	5.99
12	4.73	0.00	3.06	1.83	5.92
15	4.73	0.00	3.06	1.64	5.87
19	4.73	0.00	3.06	1.47	5.82
25	4.73	0.00	3.06	1.33	5.78
31	4.73	0.00	3.06	1.20	5.76
39	4.73	0.00	3.06	1.09	5.73
50	4.73	0.00	3.06	.99	5.72
63	4.73	0.00	3.05	.91	5.70
79	4.73	0.00	3.05	.83	5.69
99	4.73	0.00	3.05	.77	5.68
125	4.73	0.00	3.05	.73	5.68
158	4.73	0.00	3.05	.69	5.67
199	4.73	0.00	3.05	.67	5.67
251	4.73	0.00	3.05	.66	5.67
316	4.73	0.00	3.05	.67	5.67
398	4.73	0.00	3.05	.71	5.67
501	4.73	0.00	3.05	.77	5.68
630	4.73	0.00	3.05	.86	5.69
794	4.73	0.00	3.05	.98	5.71
999	4.73	0.00	3.05	1.16	5.74
1258	4.73	0.00	3.04	1.39	5.79
1584	4.73	0.00	3.04	1.69	5.87
1995	4.73	0.00	3.03	2.07	5.98
2511	4.73	0.00	3.02	2.55	6.16
3162	4.73	0.00	2.99	3.17	6.43
3981	4.73	0.00	2.96	3.95	6.83
5011	4.73	0.00	2.90	4.93	7.42
6309	4.73	0.00	2.81	6.17	8.27
7943	4.73	0.00	2.67	7.73	9.45
9999	4.73	0.00	2.45	9.70	11.06
12589	4.73	0.00	2.10	12.18	13.23
15848	4.73	0.00	1.54	15.30	16.09
19952	4.73	0.00	.66	19.23	19.81
25118	4.73	0.00	.75	24.18	24.65

Table 9.5 Input noise versus frequency.

The program plots the contribution of noise generators and total noise at the input and output. The plot of the total output noise of the cartridge 1,2, and 3 are shown in Figure 9-3 a,b,c respectively.

The total noise at the output of the amplifier shown in Table 9.6 is obtained by using calculation 5. (for "SHURE V15 III")

TOTAL NOISE AT THE OUTPUT

VOLTAGE SOURCE MODEL

FREQUENCY	SENSOR EN	LOAD EN	AMPL EN	AMP IN;ZS	SUM NOISE nV
1. TO 50.	8344.44	0.00	5402.43	3534.08	10550.15
50. TO 500.	7877.90	0.00	5088.42	1309.89	9469.38
500. TO 2100.	4859.30	0.00	3128.70	1405.56	5947.87
2100. TO 3000.	2565.02	0.00	1635.53	1396.09	3347.14

TOTAL NOISE VOLTAGE= 15733.890 NANOVOLTS

Table 9.6 Total noise at the output

The total equalized system gain is determined by using calculation 6 (in Table 9.7). Calculation 6 plots the gain versus frequency too.

In the calculation 8 the program calculates the noise figure versus source resistor at a selected frequency. The plot of the noise figure versus source resistor at 100 Hz is shown in Fig. 9.4 (for "SHURE V15-III" Cartridge)

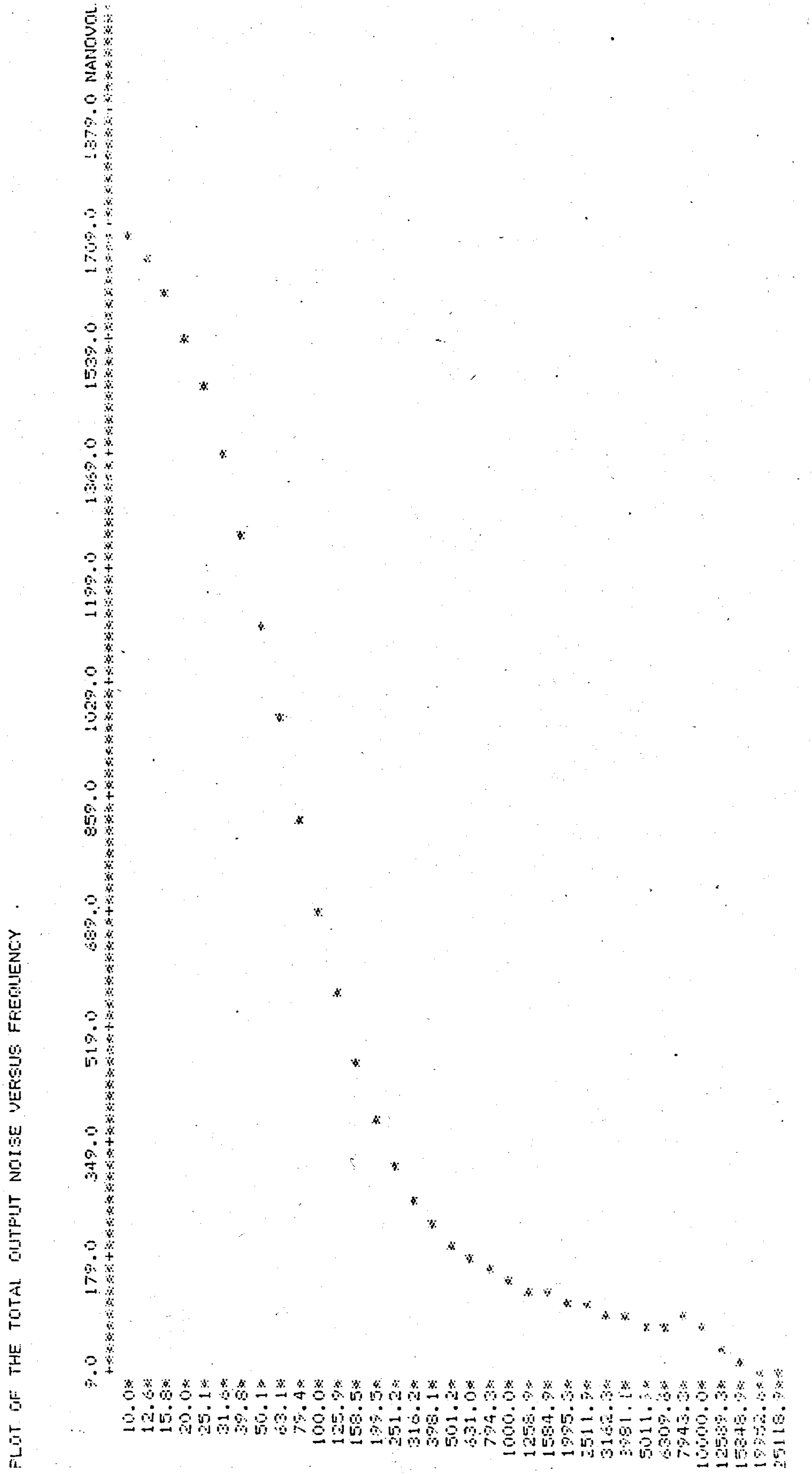


Figure 9.3a The plot of the total output noise of the cartridge "SHURE V15-III".

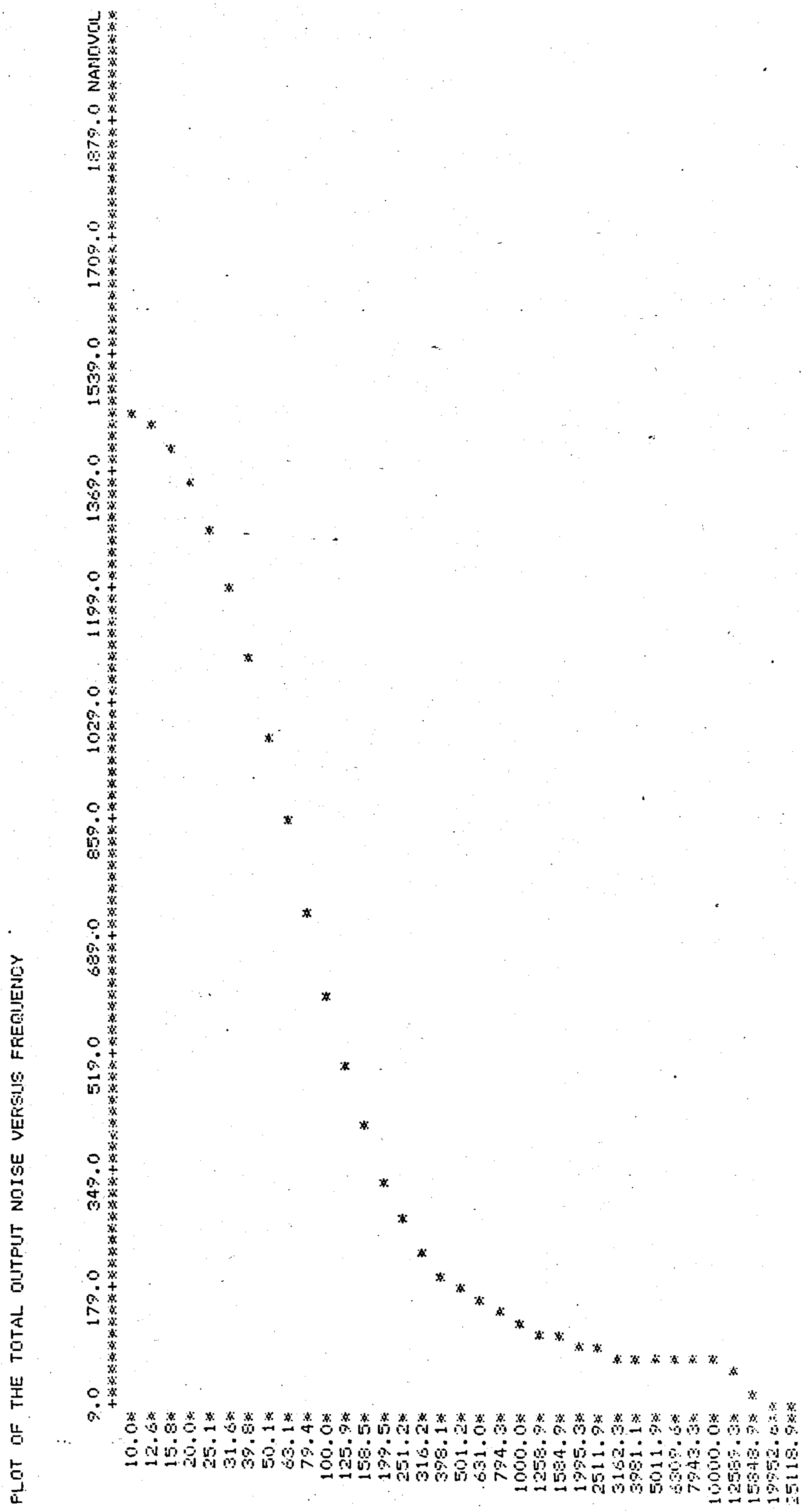


Figure 9.3b The plot of the total output noise of the cartridge" STANTON 881S".

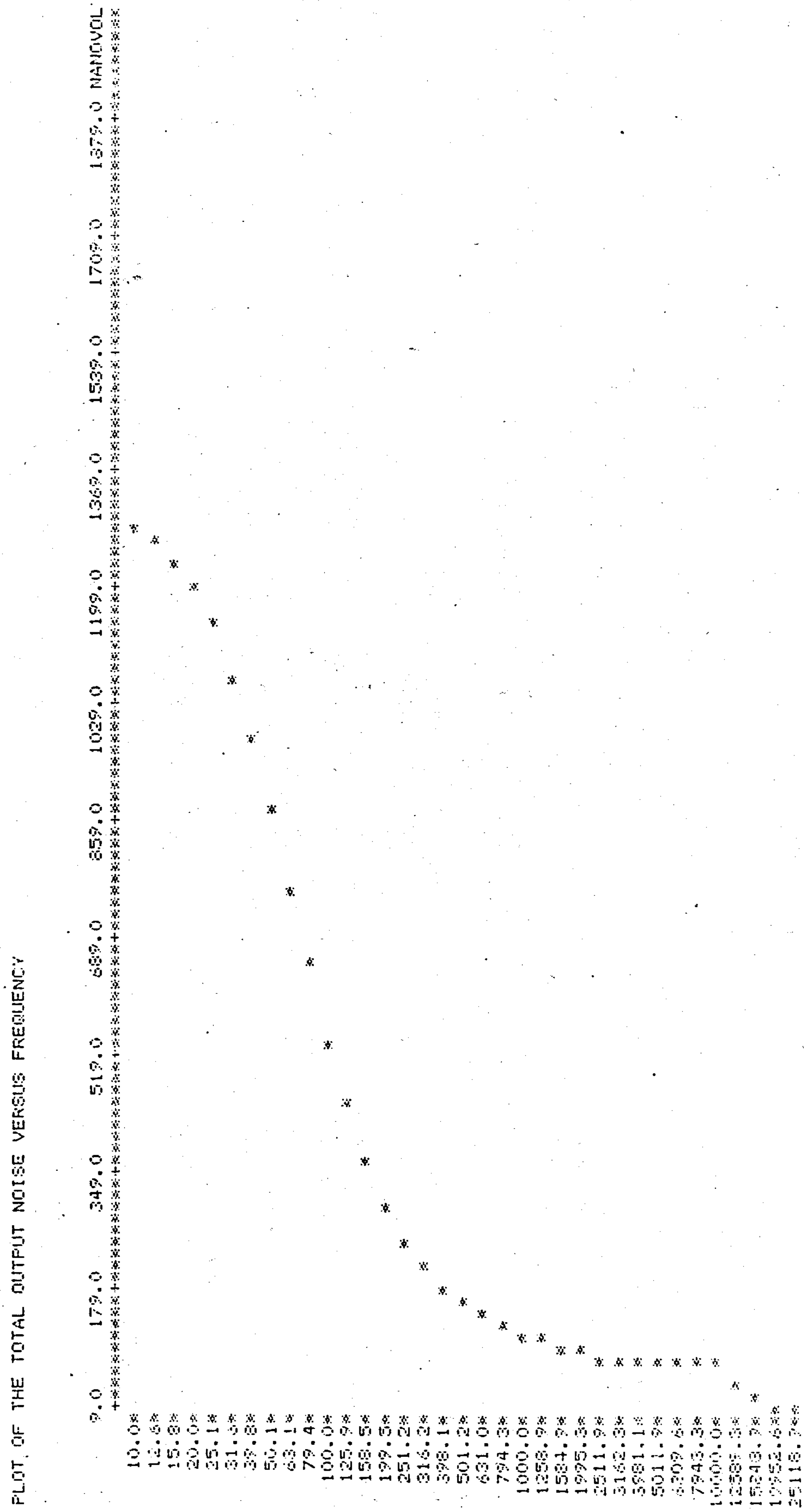


Figure 9.3c The plot of the total output noise of the cartridge "ORTOFON SME 30 H".

TOTAL SYSTEM GAIN

FREQUENCY	GAIN	GAIN DB
5	289.831	49.242
6	288.852	49.213
7	287.302	49.166
9	284.883	49.093
12	281.166	48.979
15	275.560	48.804
19	267.329	48.540
25	255.686	48.154
31	240.025	47.605
39	220.258	46.858
49	197.093	45.893
62	172.001	44.710
79	146.812	43.335
99	123.165	41.809
125	102.156	40.185
158	84.282	38.514
199	69.586	36.850
250	57.835	35.243
315	48.664	33.744
397	41.657	32.393
499	36.389	31.219
629	32.440	30.221
792	29.413	29.371
997	26.951	28.611
1255	24.753	27.872
1581	22.608	27.085
1990	20.410	26.196
2505	18.168	25.186
3154	15.984	24.074
3971	14.002	22.924
4999	12.356	21.837
6294	11.114	20.917
7924	10.056	20.048
9976	7.915	17.969
12559	4.340	12.749
15811	1.968	5.880
19905	.892	-.997
25059	.417	-7.605
31547	.200	-13.996
39716	.097	-20.243

Table 9.7 Total system gain

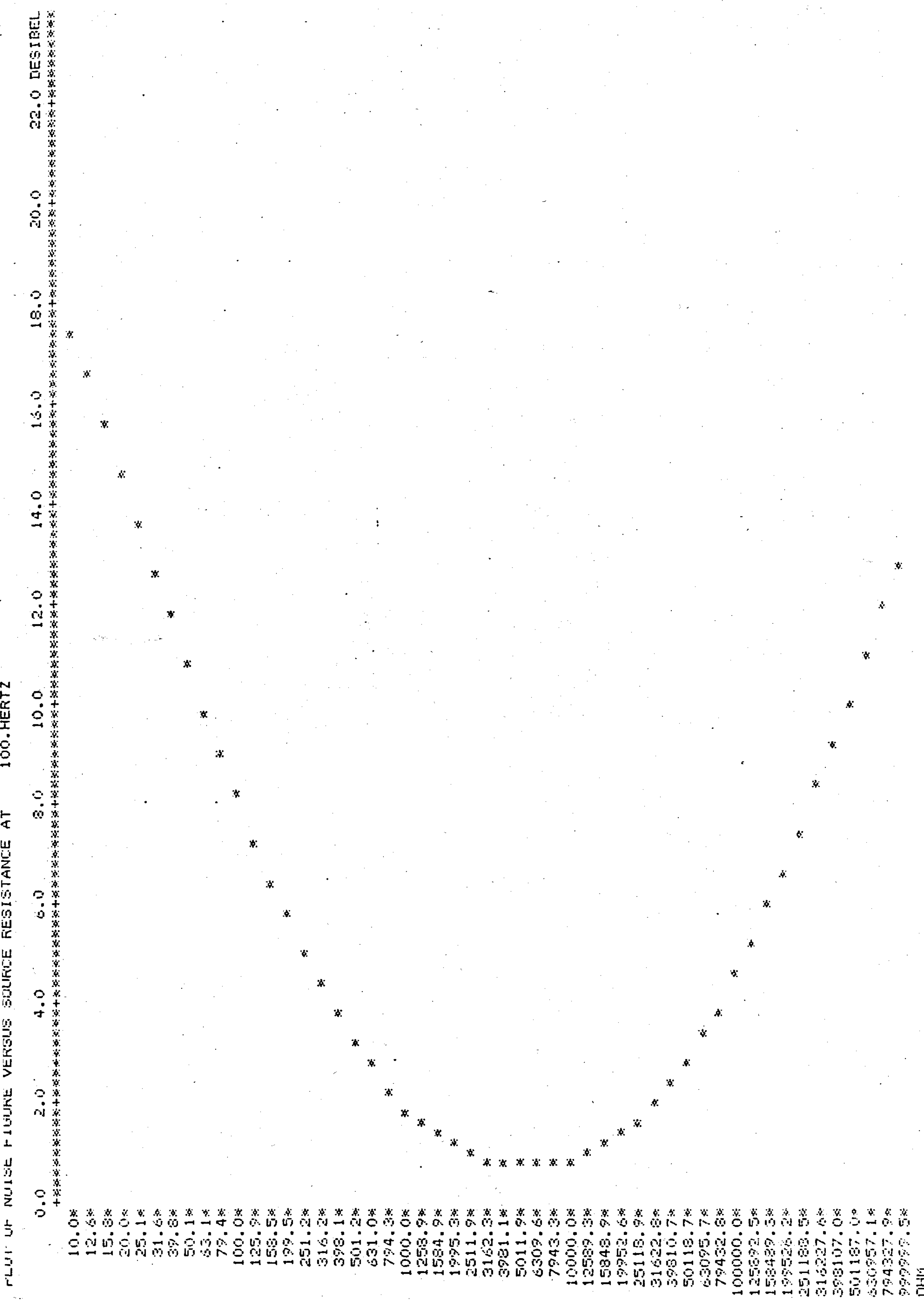


Figure 9.4 The plot of the noise figure versus source resistor for the cartridge "SHURE V15-III" at 100 Hz.

9.3 EXPERIMENTAL RESULTS

The BC 109- Silicon planar epitaxial transistor is selected for measuring the base-spreading resistance beacuse of its popularity and low $r_{bb'}$ resistance. Figure 9.5 shows the plot of the relative noise output voltage as a function of the source resistor R_S with the emitter bias current as parameter.

Noise is measured at four different values of emitter currents. For each emitter current value, the source resistor R_S is varied to determine $R_S(min)$. The r.m.s. readings of the noise voltages are averaged by eye.

The values of $R_S(min)$ could be determined from Figure 9.5. The approximate values of $R_S(min)$ are 180 Ω, 290 Ω, 640 Ω, and 1310 Ω.

Figure 9.6 is a plot of the $R_S(min)$ as a function of $kT/q\,I_E$. The best fit to the data is a straight line intercepting the R_S axis at -220 Ω. This yields the base-spreading resistance $r_{bb'}$ of the test transistor.

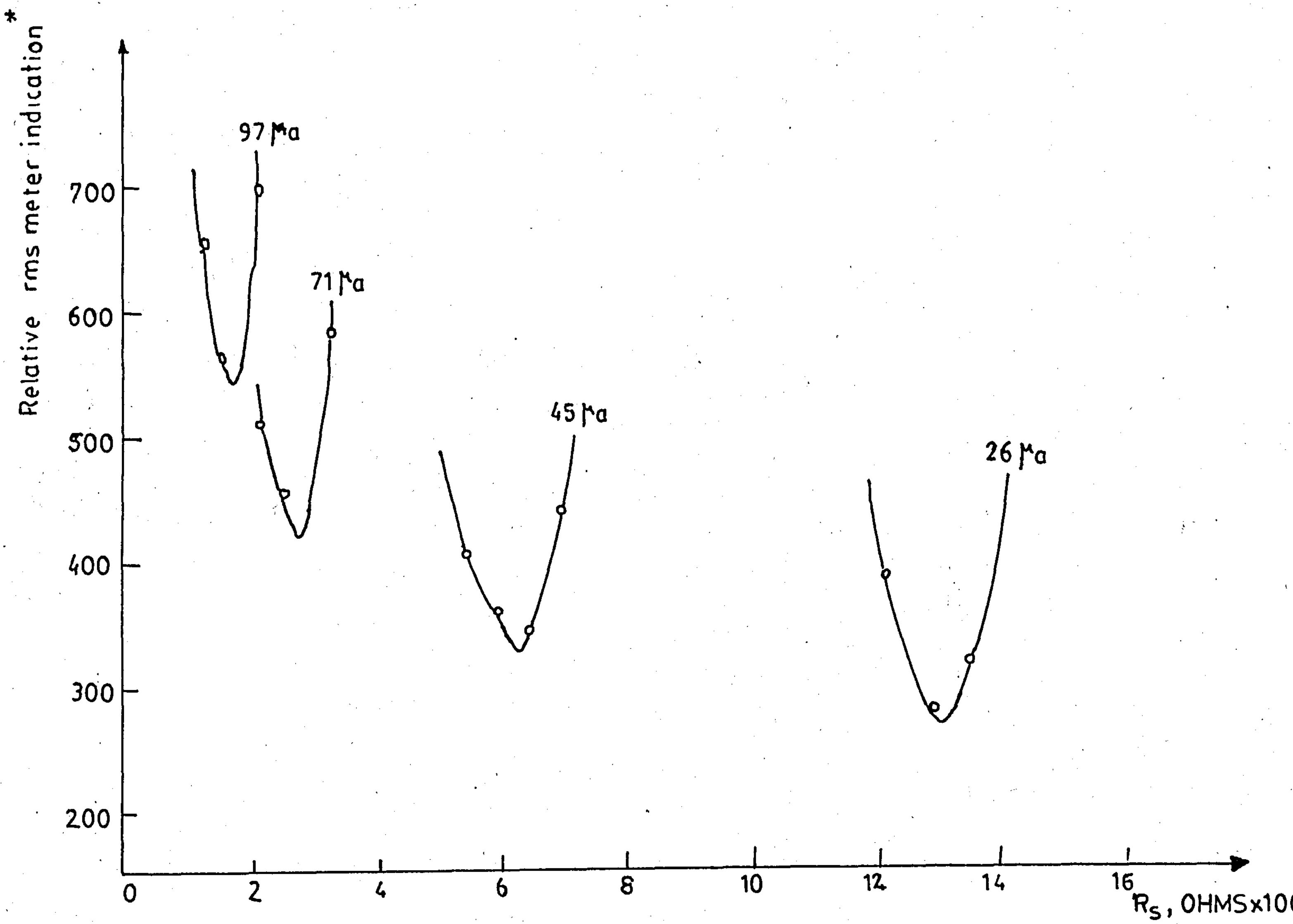


Figure 9.5 The plot of the relative output noise voltage as a function of R_S for different values of I_E (*The thermal noise of 100 K Resistor = 43 nV.)

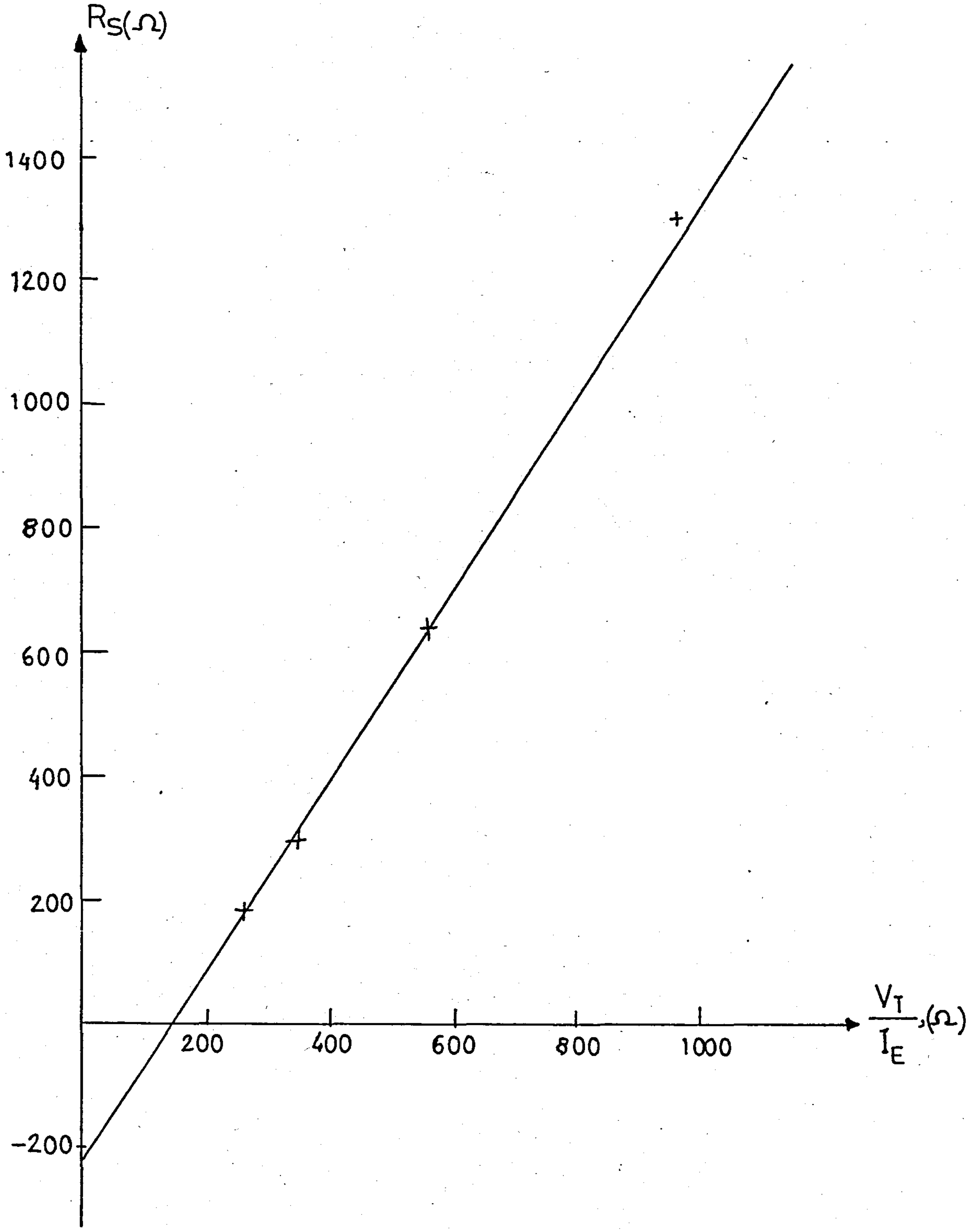


Figure 9.6 The plot of R_s(min) as a function kT/qI_E for BC 109 transistor.

CONCLUSION

Various circuit analysis programs are available in the literature for computer-aided design. Programs for a.c. analysis, such as ECAP and HICAP, can be used in noise problems. One has to introduce the noise mechanisms in the computer model in the form of noise voltage and current generators. The computer calculates the output voltage for each generator acting alone. ECAP is capable of analyzing the noise of cascaded stages.This universal circuit program calculates the noise as a function of frequency only. Although it is possible to use a general circuit analysis program in noise problems, it seems to be much faster and convenient to use the program given in this thesis.

Noise is essentially unaffected by the circuit configuration and the overall negative feedback; therefore the transistor and its operating point can be selected to meet the circuit noise requirements, and then the configuration or feedback determined to meet the gain, bandwidth, and impedance requirements. This approach allows the circuit designer to optimize for the noise and for other circuit parameters independently.

The ultimate limit on the equivalent input noise is determined by the sensor imdedance and the first stage of the amplifier. Initial steps in the design procedure are the choice of the type of the input device such as bipolar transistor or field effect transistor, and the associated operating point to minimize the generated noise. If the amplifier is operated over a band of frequencies, the noise must be integrated over this interval. Since the noise mechanisms and sensor impedances are frequency dependent, the computer program must also perform an integration. The computations may indicate that there is too much noise to meet the requirements or it may suggest a different operating point. By changing devices and/or operating points, the performance can approach the optimum point. The program can be used in computing the design equations.

Low-frequency noise measurements are shown to provide a convenient and reasonably accurate means of measuring the base-spreading resistance $r_{bb'}$. This method requires a large amount of 1/f noise and relatively small value of $r_{bb'}$ and therefore is not applicable to high-gain transistors. In the measuring instrumentation a low noise FET input preamplifier is used. Since a wave analyzer was not available, BIQUAD type band-pass filters operating selected frequencies are designed. For the instrumentation and test circuitry, the screenning has been provided because interference may be effective on the noise measurements.

APPENDIX A

```
      PROGRAM CNX(XNPUT,KUTPUT,TAPE5=XNPUT,TAPE6=KUTPUT)
      DIMENSION FR(10),FG(200),EG(200),GI(200),SUMG(200),GK(200)
      DIMENSION FQ(10),FACTOR(200),RES(200),EOG(200),GOI(200)
      DIMENSION SUMOG(200)
      COMMON B,B0,B1,BK0,C1,C2,C3,C4,C8,CK0,D,D0,D1,D10,E,E0,F,F0,F1,F2
      COMMON F4,F5,F6,F7,F8,F9,I,I0,I2,IB,IC,ICB,K,K0,K1,K2,K9,KT,L1,L2
      COMMON M2,M4,N2,N8,N9,NM,P,R1,R2,R3,R4,R6,R7,R10,R11,R12,Q,Q0,Q1
      COMMON SUM,T,T0,W,A1,A2,A3,A4,A5,A6,A7,A8,A9,A0,R13,F3,M1,Q2,LX
      REAL I,I0,I2,IB,IC,ICB,K,K0,K2,K4,KT,L1,L2,N2,N8,N9,K9
      DATA JP,JR,JS,JT,JU,JV,JY,JZ,JW/9*0/
      DATA IZ,IW,IQ,IV,IU,IM,IP,IR,IS/9*0/
C     READ IN VALUES FOR M4(SOURCE USED),NM(NOISE MODEL),
C     NN(NUMBER OF CALCULATION)
      READ(5,*) M4,NM,NN
      DO 100 LP=1,NN
      IF(NM-1)101,102,103
  101 WRITE(6,42)
   42 FORMAT(1H1,//,'NOISE MODEL IMPROPERLY SELECTED')
      GO TO 1506
C     READ IN VALUES FOR BJT
  102 READ(5,*) R10,B,F7,F8,Q1,Q2,IC,ICB,C8
      GO TO 104
C     READ IN VALUES FOR TMFET
  103 READ(5,*) F8,C8,ICB,B
C     READ IN VALUES FOR SENSOR MODEL
  104 READ(5,*) R1,R2,R3,R4,R6,C1,C2,C3,C4
      READ(5,*) L1,L2,I2,T,T0,F9
C     READ IN VALUES FOR AMPLG
      READ(5,*) F0,F1,F2,F3,F4,F5,F6,Q0,N8,N9,K2
      P=3.14159
      Q=1.602E-19
      K=1.38E-23
      DO 99 M2=1,8
      GO TO (505,505,505,515,505,505),M4
  505 GO TO (710,720,730,740,750,760,770,780),M2
  515 GO TO (810,720,830,840,850,760,770,780),M2
C     TOTAL EQUIVALENT INPUT NOISE OVER A BAND
  710 WRITE(6,21)
   21 FORMAT(//,2X,'TOTAL EQUIVALENT INPUT NOISE OVER A BAND',/)
      K1=1
      CALL PRINT1
      CALL PRINT2
      CALL INTEGR
  703 N2=SQRT(N2)
      WRITE(6,81) N2
   81 FORMAT(//,2X,'TOTAL NOISE VOLTAGE=',F12.3,3X,'NANOVOLTS',//)
      GO TO 99
C     INPUT NETWORK FREQUENCY RESPONCE
  720 WRITE(6,22)
   22 FORMAT(//,2X,'INPUT NETWORK FREQUENCY RESPONSE',/)
      CALL READ1
      CALL PRINT5
  719 M1=2
      CALL SOURCE
      CALL PRINT4
      GO TO (719,99),LX
C     INPUT NOISE AT ONE FREQUENCY
  730 WRITE(6,23)
   23 FORMAT(//,2X,'INPUT NOISE AT ONE FREQUENCY',/)
      READ(5,*) (FR(IK),IK=1,10)
      WRITE(6,82)
   82 FORMAT(/,2X,'FREQUENCY')
      CALL PRINT2
  731 IL=1
   31 F=FR(IL)
      IF(F)99,99,1501
 1501 K1=1
      CALL KIT
```

```
      SUM=SQRT(E+I+D+D1)
      E=SQRT(E)
      I=SQRT(I)
      D=SQRT(D)
      D1=SQRT(D1)
      KF=F
      WRITE(6,84) KF,D,D1,E,I,SUM
   84 FORMAT(I7,5F12.2)
      IL=IL+1
      GO TO 31
C     INPUT AND OUTPUT NOISE VERSUS FREQUENCY
  740 WRITE(6,24)
   24 FORMAT(//,2X,'INPUT NOISE VERSUS FREQUENCY',/)
      K1=1
      CALL READ1
      CALL PRINT1
      CALL PRINT2
      LK=1
 1498 CALL KIT
      JF=F
      SUM=SQRT(E+I+D+D1)
      E=SQRT(E)
      I=SQRT(I)
      D=SQRT(D)
      D1=SQRT(D1)
      WRITE(6,84) JF,D,D1,E,I,SUM
      EG(LK)=E
      GI(LK)=I
      SUMG(LK)=SUM
      M1=2
      CALL SOURCE
      CALL AMPLG
      KO=KO*KT
      KO=SQRT(KO)
      EOG(LK)=KO*E
      GOI(LK)=KO*I
      SUMOG(LK)=KO*SUM
      FG(LK)=F
      IF(F-B1)1499,1499,91
 1499 F=F*1.2589254
      LK=LK+1
      GO TO 1498
   91 WRITE(6,71)
   71 FORMAT(///,2X,'PLOT OF THE  CONTRIBUTION OF THE NOISE '
     + ,'VOLTAGE  OF THE AMPLIFIER AT INPUT',///)
      CALL PLOT1(LK,EG,FG,1,JP,IZ,LP)
      WRITE(6,72)
   72 FORMAT(///,2X,'PLOT OF THE CONTRIBUTION OF THE NOISE ',
     + 'CURRENT OF THE AMPLIFIER AT INPUT',///)
      CALL PLOT1(LK,GI,FG,1,JR,IW,LP)
      WRITE(6,73)
   73 FORMAT(///,2X,'PLOT OF THE TOTAL INPUT NOISE VERSUS FREQ',///)
      CALL PLOT1(LK,SUMG,FG,1,JS,IQ,LP)
      WRITE(6,745)
  745 FORMAT(///,2X,'PLOT OF THE CONTRIBUTION OF THE NOISE ',
     + 'VOLTAGE OF THE AMPLIFIER AT THE OUTPUT',///)
      CALL PLOT1(LK,EOG,FG,1,JT,IV,LP)
      WRITE(6,746)
  746 FORMAT(///,2X,'PLOT OF THE CONTRIBUTION OF THE NOISE ',
     + 'CURRENT OF THE AMPLIFIER AT THE OUTPUT',///)
      CALL PLOT1(LK,GOI,FG,1,JU,IU,LP)
      WRITE (6,747)
  747 FORMAT(///,2X,'PLOT OF THE TOTAL OUTPUT NOISE VERSUS ',
     + 'FREQUENCY',///)
      CALL PLOT1(LK,SUMOG,FG,1,JV,IM,LP)
      GO TO 99
C     TOTAL NOISE AT THE OUTPUT
  750 WRITE(6,25)
   25 FORMAT(//,2X,'TOTAL NOISE AT THE OUTPUT',/)
      K1=2
      CALL PRINT1
      CALL PRINT2
      CALL INTEGR
      GO TO 703
C     TOTAL SYSTEM GAIN
  760 WRITE(6,26)
   26 FORMAT(//,2X,'TOTAL SYSTEM GAIN',/)
```

```
      CALL READ1
      CALL PRINT5
      LK=1
 761  M1=2
      CALL SOURCE
      CALL AMPLG
      KO=KO*KT
      CALL PRINT4
      GK(LK)=KO
      FG(LK)=F/1.2589254
      LK=LK+1
      GO TO (761.92),LX
  92  WRITE(6,94)
  94  FORMAT(//,2X,'PLOT OF TOTAL SYSTEM GAIN',//)
      LK=LK-1
      CALL PLOT2(LK,GK,FG)
      GO TO 99
 810  WRITE(6,21)
      K1=1
      CALL PRINT3
      CALL INTEGR
 811  N2=SQRT(N2)
      WRITE(6,87) N2
  87  FORMAT(//,2X,'TOTAL NOISE CURRUNT=',F12.4,3X,'PICOAMPERS',/)
      GO TO 99
 830  WRITE(6,23)
      READ(5,*) (FR(IK),IK=1,10)
      WRITE(6,88)
  88  FORMAT(//,2X,'FREQUENCY SENSOR IN  LOAD IN')
      GO TO 731
 840  WRITE(6,24)
      K1=1
      CALL READ1
      CALL PRINT3
      LK=1
 841  CALL KIT
      JF=F
      SUM=SQRT(E+I+D+D1)
      E=SQRT(E)
      I=SQRT(I)
      D=SQRT(D)
      D1=SQRT(D1)
      WRITE(6,84) JF,D,D1,E,I,SUM
      EG(LK)=E
      GI(LK)=I
      SUMG(LK)=SUM
      M1=2
      CALL SOURCE
      CALL AMPLG
      KO=KO*KT
      KO=SQRT(KO)
      EOG(LK)=KO*E
      GOI(LK)=KO*I
      SUMOG(LK)=KO*SUM
      FG(LK)=F
      IF(F-B1)1491,1491,93
1491  F=F*1.2589254
      LK=LK+1
      GO TO 841
  93  WRITE(6,74)
  74  FORMAT(///,2X,'PLOT OF THE CONTRIBUTION OF NOISE ',
     + 'VOLTAGE OF THE AMPLIFIER',///)
      CALL PLOT1(LK,EG,FG,1,JP,IZ,LP)
      WRITE(6,75)
  75  FORMAT(///,2X,'PLOT OF THE CONTRIBUTION OF NOISE ',
     + 'CURRENT OF THE AMPLIFIER',///)
      CALL PLOT1(LK,GI,FG,1,JR,IW,LP)
      WRITE(6,76)
  76  FORMAT(///,2X,'PLOT OF THE TOTAL INPUT NOISE VERSUS FREQ',///)
      CALL PLOT1(LK,SUMG,FG,1,JS,IQ,LP)
      WRITE(6.845)
 845  FORMAT(///,2X,'PLOT OF THE CONTRIBUTION OF THE NOISE ',
     + 'VOLTAGE OF THE AMPLIFIER AT THE OUTPUT',///)
      CALL PLOT1(LK,EOG,FG,1,JT,IV,LP)
      WRITE(6,846)
 846  FORMAT(///,2X.'PLOT OF THE CONTRIBUTION OF THE NOISE ',
     + 'CURRENT OF THE AMPLIFIER AT THE OUTPUT',///)
```

```
      CALL PLOT1(LK,GOI,FG,1,JU,IU,LP)
      WRITE(6,847)
 847  FORMAT(///,2X,'PLOT OF THE TOTAL OUTPUT NOISE VERSUS ',
     + 'FREQUENCY ',///)
      CALL PLOT1(LK,SUMOG,FG,1,JV,IT,LP)
      GO TO 99
 850  WRITE(6,25)
      K1=2
      CALL PRINT3
      CALL INTEGR
      GO TO 811
      NOISE BANDWIDTH
 770  TOP=0.
      READ(5,*) (FQ(I),I=1,10)
      K1=2
      M=1
      N=2
   1  IF(FQ(N)) 16,17,17
  17  S=(FQ(N)-FQ(M))/24.
      R=0.
      F=FQ(M)
  18  F=F+S
      CALL KIT
      R=R+K9*2.
      IF(F-FQ(N)+2.*S) 18,18,19
  19  F=FQ(M)
      F=F+S/2.
  20  CALL KIT
      R=R+K9*4.
      IF(F-FQ(N)+S) 11,11,12
  11  F=F+S
      GO TO 20
  12  F=FQ(M)
      SX=FQ(N)-FQ(M)
  13  CALL KIT
      R=R+K9
      IF(F-FQ(N)+2.*S) 14,15,15
  14  F=F+SX
      GO TO 13
  15  TOP=TOP+R*S/6.
      M=M+1
      N=N+1
      IF(N-10) 1,1,16
  16  TOP=TOP/(K2*K2)
      WRITE(6,32) TOP
  32  FORMAT(////,2X,'NOISE BANDWIDTH=',F12.0,///)
      GO TO 99
      NOISE FIGURE VERSUS SOURCE RESISTANCE
 780  READ(5,*) FF
      F=FF
      R1=10.
      K1=1
      LK=1
 781  CALL KIT
      SUM=(E+I+D+D1)
      FACTOR(LK)=10.*(ALOG10(SUM/D))
      RES(LK)=R1
      R1=R1*1.2589254
      LK=LK+1
      IF(R1.LT.1000000.) GO TO 781
      LK=LK-1
      WRITE(6,782) FF
 782  FORMAT(///,2X,'NOISE FIGURE VERSUS SOURCE RESISTANCE',//,2X,
     + 'FREQUENCY=',F7.0,'HERTZ',//,2X,'SOURCE RESISTANCE NOISE FIGURE'
     + ,' DESIBEL',/)
      DO 783 J=1,LK
 783  WRITE(6,784) RES(J),FACTOR(J)
 784  FORMAT(2X,F12.0,5X,F7.3)
      WRITE(6,785) FF
 785  FORMAT(///,2X,'PLOT OF NOISE FIGURE VERSUS SOURCE RESISTANCE',
     + ' AT ',F7.0,'HERTZ',/)
      CALL PLOT1(LK,FACTOR,RES,3,JW,IS,LP)
      WRITE(6,786)
 786  FORMAT(1X,'OHM')
  99  CONTINUE
 100  CONTINUE
1506  STOP
      END
```

```
      SUBROUTINE READ1
      COMMON B,BO,B1,BKO,C1,C2,C3,C4,C8,CKO,D,DO,D1,D10,E,EO,F,FO,F1,F2
      COMMON F4,F5,F6,F7,F8,F9,I,IO,I2,IB,IC,ICB,K,KO,K1,K2,K9,KT,L1,L2
      COMMON M2,M4,N2,N8,N9,NM,P,R1,R2,R3,R4,R6,R7,R10,R11,R12,Q,QO,Q1
      COMMON SUM,T,TO,W,A1,A2,A3,A4,A5,A6,A7,A8,A9,AO,R13,F3,M1,Q2,LX
      READ(5,*) BO,B1
      F=BO
      RETURN
      END
      SUBROUTINE MODEL
      COMMON B,BO,B1,BKO,C1,C2,C3,C4,C8,CKO,D,DO,D1,D10,E,EO,F,FO,F1,F2
      COMMON F4,F5,F6,F7,F8,F9,I,IO,I2,IB,IC,ICB,K,KO,K1,K2,K9,KT,L1,L2
      COMMON M2,M4,N2,N8,N9,NM,P,R1,R2,R3,R4,R6,R7,R10,R11,R12,Q,QO,Q1
      COMMON SUM,T,TO,W,A1,A2,A3,A4,A5,A6,A7,A8,A9,AO,R13,F3,M1,Q2,LX
      REAL I
      IF(NM-1)11,12,13
11    STOP
12    CALL BJT
      RETURN
13    CALL TMFET
      RETURN
      END
      SUBROUTINE BJT
      COMMON B,BO,B1,BKO,C1,C2,C3,C4,C8,CKO,D,DO,D1,D10,E,EO,F,FO,F1,F2
      COMMON F4,F5,F6,F7,F8,F9,I,IO,I2,IB,IC,ICB,K,KO,K1,K2,K9,KT,L1,L2
      COMMON M2,M4,N2,N8,N9,NM,P,R1,R2,R3,R4,R6,R7,R10,R11,R12,Q,QO,Q1
      COMMON SUM,T,TO,W,A1,A2,A3,A4,A5,A6,A7,A8,A9,AO,R13,F3,M1,Q2,LX
      REAL I,IB,IC,ICB,K
      W=2.*P*F
      IB=IC/B+ICB
      X1=4.0*K*T*R10
      X2=(2.0*Q*IC*(R10+B/(IC*40.))**2.)/(B*B)
      X3=2.0*Q*IC*(R10+1./(W*C8))**2.*(F*F)/(F8*F8)
      X4=2.0*Q*IB*(R10)**2.
      X5=2.0*Q*F7*(IB**Q1)*R10*R10/(F**Q2)
      E=X1+X2+X3+X4+X5
      X6=2.0*Q*IB+(2.0*Q*F7*(IB**Q1))/(F**Q2)
      X7=2.0*Q*IC/(B*B)+2.0*Q*IC*F*F/(F8*F8)
      I=X6+X7
      RETURN
      END
      SUBROUTINE TMFET
      COMMON B,BO,B1,BKO,C1,C2,C3,C4,C8,CKO,D,DO,D1,D10,E,EO,F,FO,F1,F2
      COMMON F4,F5,F6,F7,F8,F9,I,IO,I2,IB,IC,ICB,K,KO,K1,K2,K9,KT,L1,L2
      COMMON M2,M4,N2,N8,N9,NM,P,R1,R2,R3,R4,R6,R7,R10,R11,R12,Q,QO,Q1
      COMMON SUM,T,TO,W,A1,A2,A3,A4,A5,A6,A7,A8,A9,AO,R13,F3,M1,Q2,LX
      REAL I,K
      W=2.*P*F
      X1=4.*K*T
      E=((2./3.)*X1/B)*(1.+F8/F)
      I=(16./135.)*W*W*C8*C8*X1/B+2.*Q*ICB
      RETURN
      END
      SUBROUTINE AMPLG
      COMMON B,BO,B1,BKO,C1,C2,C3,C4,C8,CKO,D,DO,D1,D10,E,EO,F,FO,F1,F2
      COMMON F4,F5,F6,F7,F8,F9,I,IO,I2,IB,IC,ICB,K,KO,K1,K2,K9,KT,L1,L2
      COMMON M2,M4,N2,N8,N9,NM,P,R1,R2,R3,R4,R6,R7,R10,R11,R12,Q,QO,Q1
      COMMON SUM,T,TO,W,A1,A2,A3,A4,A5,A6,A7,A8,A9,AO,R13,F3,M1,Q2,LX
      REAL N8,N9,K2,KT
      A1=((F/FO)/(SQRT(1.+(F/FO)**2.)))**N8
      A2=((F/F2)**2.+1.)/((F/F2+F/F1)**2.+1.)
      A3=((F/F3+F/F4)**2.+1.)/((F/F4)**2.+1.)
      A6=F/F5
      A7=1.-A6*A6
      A8=2.*A6*(SQRT(1.+1./QO)-1)
      A4=1.+(A6*A6)/((A7*A7)+(A8*A8))
      A5=(1./(SQRT(1.+(F/F6)**2.)))**N9
      KT=K2*K2*A1*A1*A2*A3*A4*A5*A5
      RETURN
      END
      SUBROUTINE INTEGR
      COMMON B,BO,B1,BKO,C1,C2,C3,C4,C8,CKO,D,DO,D1,D10,E,EO,F,FO,F1,F2
      COMMON F4,F5,F6,F7,F8,F9,I,IO,I2,IB,IC,ICB,K,KO,K1,K2,K9,KT,L1,L2
      COMMON M2,M4,N2,N8,N9,NM,P,R1,R2,R3,R4,R6,R7,R10,R11,R12,Q,QO,Q1
      COMMON SUM,T,TO,W,A1,A2,A3,A4,A5,A6,A7,A8,A9,AO,R13,F3,M1,Q2,LX
      DIMENSION FR(10)
      REAL I,IO,N2
      N2=0.0
```

```
      READ(5,*) (FR(IK),IK=1,10)
      L=1
      J=2
 133  IF(FR(J))131,131,132
 132  WRITE(6,93) FR(L),FR(J)
  93  FORMAT(/,F7.0,' TO',F7.0)
      H7=(FR(J)-FR(L))/24.
      D0=0.
      D10=0.
      E0=0.
      I0=0.
      F=FR(L)
      H9=FR(J)-FR(L)
 144  CALL KIT
      E0=E0+E
      I0=I0+I
      D0=D0+D
      D10=D10+D1
      IF(F-FR(J)+2.*H7)141,142,142
 141  F=F+H9
      GO TO 144
 142  F=FR(L)
      F=F+H7/2.
 154  CALL KIT
      E0=E0+E*4.
      I0=I0+I*4.
      D0=D0+D*4.
      D10=D10+D1*4.
      IF(F-FR(J)+H7)151,151,152
 151  F=F+H7
      GO TO 154
 152  F=FR(L)
 161  F=F+H7
      CALL KIT
      E0=E0+E*2.
      I0=I0+I*2.
      D0=D0+D*2.
      D10=D10+D1*2.
      IF(F-FR(J)+2.*H7)161,161,162
 162  N2=N2+(E0+I0+D0+D10)*H7/6.
      H8=H7/6.
      SUMD=SQRT(D0*H8)
      SUMD1=SQRT(D10*H8)
      SUME=SQRT(E0*H8)
      SUMI=SQRT(I0*H8)
      SUMSUM=SQRT((E0+I0+D0+D10)*H8)
      WRITE(6,94) SUMD,SUMD1,SUME,SUMI,SUMSUM
  94  FORMAT(/,6X,5F12.2)
      L=L+1
      J=J+1
      IF(J-10)133,133,131
 131  RETURN
      END
      SUBROUTINE KIT
      COMMON B,B0,B1,BK0,C1,C2,C3,C4,C8,CK0,D,D0,D1,D10,E,E0,F,F0,F1,F2
      COMMON F4,F5,F6,F7,F8,F9,I,I0,I2,IB,IC,ICB,K,K0,K1,K2,K9,KT,L1,L2
      COMMON M2,M4,N2,N8,N9,NM,P,R1,R2,R3,R4,R6,R7,R10,R11,R12,Q,Q0,Q1
      COMMON SUM,T,T0,W,A1,A2,A3,A4,A5,A6,A7,A8,A9,A0,R13,F3,M1,Q2,LX
      REAL I,K0,K2,K9
      IF(K1-1)1500,1510,1520
1510  CALL MODEL
      M1=1
      CALL SOURCE
      RETURN
1500  STOP
1520  CALL KIT2
      RETURN
      END
      SUBROUTINE KIT2
      COMMON B,B0,B1,BK0,C1,C2,C3,C4,C8,CK0,D,D0,D1,D10,E,E0,F,F0,F1,F2
      COMMON F4,F5,F6,F7,F8,F9,I,I0,I2,IB,IC,ICB,K,K0,K1,K2,K9,KT,L1,L2
      COMMON M2,M4,N2,N8,N9,NM,P,R1,R2,R3,R4,R6,R7,R10,R11,R12,Q,Q0,Q1
      COMMON SUM,T,T0,W,A1,A2,A3,A4,A5,A6,A7,A8,A9,A0,R13,F3,M1,Q2,LX
      REAL I,K0,K9,KT
      CALL MODEL
      M1=1
      CALL SOURCE
```

```
      M1=2
      CALL SOURCE
      CALL AMPLG
      K9=K0*KT
      E=K9*E
      I=K9*I
      D=K9*D
      D1=K9*D1
      RETURN
      END
      SUBROUTINE SOURCE
      COMMON B,B0,B1,BK0,C1,C2,C3,C4,C8,CK0,D,D0,D1.D10,E,E0.F,F0.F1,F2
      COMMON F4,F5,F6,F7,F8,F9,I,I0,I2,IB,IC,ICB,K,K0,K1,K2,K9,KT,L1,L2
      COMMON M2,M4.N2,N8,N9,NM,P,R1,R2,R3,R4,R6,R7,R10,R11,R12.Q,Q0,Q1
      COMMON SUM,T,T0,W,A1,A2,A3,A4,A5,A6,A7,A8,A9,A0,R13,F3,M1,Q2,LX
      REAL I
      GO TO (1,2,3,4,5,6),M4
 1    CALL SORCE1
      GO TO 9
 2    CALL SORCE2
      GO TO 9
 3    CALL SORCE3
      GO TO 9
 4    CALL SORCE4
      GO TO 9
 5    CALL SORCE5
      GO TO 9
 6    CALL SORCE6
 9    RETURN
      END
      SUBROUTINE SORCE1
      COMMON B,B0,B1,BK0,C1,C2,C3,C4,C8,CK0,D,D0,D1,D10.E,E0,F,F0,F1,F2
      COMMON F4,F5,F6,F7,F8,F9,I,I0,I2,IB,IC,ICB,K,K0,K1,K2,K9,KT,L1,L2
      COMMON M2,M4,N2,N8,N9,NM,P,R1,R2,R3,R4,R6,R7,R10,R11,R12,Q,Q0,Q1
      COMMON SUM,T,T0,W,A1,A2,A3,A4,A5,A6,A7,A8,A9,A0,R13,F3,M1,Q2,LX
      REAL I,K,K0
      W=2.*P*F
      GO TO (10,110),M1
 10   A0=1.+(R1/R3)*(1.+C1/C3)
      A1=W*R1*C1-(1./(W*C3*R3))
      A2=W*R1*(C1+C3)
      A3=(1.+A2*A2)/(W*W*C3*C3)
      E=E*(A0*A0+A1*A1)*1.E18
      I=I*A3*1.E18
      D=(4.*K*T*R1)*1.E18
      D1=(4.*K*T*A3/R3)*1.E18
      RETURN
 110  R7=R3*R4/(R3+R4)
      A4=W*R7*R1*(C4*(C1+C3)+C1*C3)-(1./W)
      A5=R1*(C1+C3)+R7*(C3+C4)
      K0=(R7*R7*C3*C3)/(A4*A4+A5*A5)
      RETURN
      END
      SUBROUTINE SORCE2
      COMMON B,B0,B1,BK0,C1,C2,C3,C4,C8,CK0,D,D0,D1,D10,E,E0,F,F0,F1,F2
      COMMON F4,F5,F6,F7,F8,F9,I,I0,I2,IB,IC,ICB,K,K0,K1,K2,K9,KT,L1,L2
      COMMON M2,M4,N2,N8,N9,NM,P,R1,R2,R3,R4,R6,R7,R10.R11,R12,Q,Q0.Q1
      COMMON SUM,T,T0,W,A1,A2,A3,A4,A5,A6,A7,A8,A9,A0,R13,F3,M1,Q2,LX
      REAL I,K,K0
      W=2.0*P*F
      GO TO (20,120),M1
 20   A0=R3*C3*(R1+R2)+R1*R2*(C1+C2)
      A1=W*C1*C3*R1*R2*R3-(R1+R2)/W
      A2=(R1*(C1+C3))/C3
      A3=(R1+R2)/(W*R2*R3)
      D=4.0*K*T*R1*(F9/F+1.)*1.E18
      A4=(A0*A0+A1*A1)/R2*R2*R3*R3*C3*C3
      I=I*(A2*A2+A3*A3)*1.E18
      E=E*A4*1.E18
      D1=((4.*K*T/R3)*(A2*A2+A3*A3)+4.*K*T*R1*R1/R2)*1.E18
      RETURN
 120  R7=R3*R4/(R3+R4)
      A5=(R1+R2)*R7*(C3+C4)+R1*R2*(C1+C3)
      A6=W*R1**C1*C3*R7-(R1+R2)/W+W*R1*R2*C4*R7*(C1+C3)
      K0=R2*R2*R7*R7*C3*C3/(A5*A5+A6*A6)
      RETURN
      END
      SUBROUTINE SORCE3
```

```
      COMMON B,B0,B1,BK0,C1,C2,C3,C4,C8,CK0,D,D0,D1,D10,E,E0,F,F0,F1,F2
      COMMON F4,F5,F6,F7,F8,F9,I,I0,I2,IB,IC,ICB,K,K0,K1,K2,K9,KT,L1,L2
      COMMON M2,M4,N2,N8,N9,NM,P,R1,R2,R3,R4,R6,R7,R10,R11,R12,Q,Q0,Q1
      COMMON SUM,T,T0,W,A1,A2,A3,A4,A5,A6,A7,A8,A9,A0,R13,F3,M1,Q2,LX
      REAL I,K,K0,L2
      W=2.0*P*F
      GO TO (30,130),M1
 30   D=4.0*K*T*R1*1.E18
      IF(R3-1.)37,129,37
 37   A0=1.-W*W*L2*(C1+C3)-W*W*R1*R3*C1*C3
      A1=W*(C3*(R1+R3)+C1*R1-W*W*R3*L2*C1*C3)
      A2=1.-W*W*L2*(C1+C3)
      A3=W*R1*(C1+C3)
      E=E*((A0*A0+A1*A1)/(R3*R3*C3*C3*W*W))*1.E18
      I=I*((A2*A2+A3*A3)/(W*W*C3*C3))*1.E18
      D1=(4.*K*T/R3)*((A2*A2+A3*A3)/(W*W*C3*C3))*1.E18
 129  I=I*(R1*R1+W*W*L2*L2)*1.E18
      A0=1.-W*W*L2*C1
      A1=W*R1*C1
      E=E*(A0*A0+A1*A1)*1.E18
      D1=0.0
      RETURN
 130  IF(R3-1.)128,127,128
 128  R7=R3*R4/(R3+R4)
      A0=W*R7*C3
      A1=1.-W*W*L2*(C1+C3)-W*W*R1*R7*(C1*C4+C1*C3+C3*C4)
      A2=W*(R7*(C3+C4)+R1*(C1+C3)-W*W*R7*L2*(C1*C3+C3*C4+C1*C4))
      K0=A0*A0/(A1*A1+A2*A2)
      RETURN
 127  A0=R1+R4-W*W*R4*L2*(C1+C4)
      A1=W*(R1*R4*(C1+C4)+L2)
      K0=R4*R4/(A0*A0+A1*A1)
      RETURN
      END
      SUBROUTINE SORCE4
      COMMON B,B0,B1,BK0,C1,C2,C3,C4,C8,CK0,D,D0,D1,D10,E,E0,F,F0,F1,F2
      COMMON F4,F5,F6,F7,F8,F9,I,I0,I2,IB,IC,ICB,K,K0,K1,K2,K9,KT,L1,L2
      COMMON M2,M4,N2,N8,N9,NM,P,R1,R2,R3,R4,R6,R7,R10,R11,R12,Q,Q0,Q1
      COMMON SUM,T,T0,W,A1,A2,A3,A4,A5,A6,A7,A8,A9,A0,R13,F3,M1,Q2,LX
      REAL I,I2,K0,K9,L1,K
      W=2.*P*F
      GO TO (40,140),M1
 40   C5=C1+C2
      R5=R2*R3/(R2+R3)
      A1=1./R5*R5+((1.-W*WC5*L1)/(W*L1))**2.
      E=E*A1*1.E24
      I=I*1.E24
      D=(((4.*K*T)/R3)+(2.*Q*I2*(F9/F+1.)))*1.E24
      D1=(4.*K*T/R3)*1.E24
      RETURN
 140  C6=C1+C4+C2
      R7=(R2*R3*R4)/(R2*R4+R2*R3+R3*R4)
      A0=(1.-W*W*L1*C6)/W*L1
      K0=(1./((1./(R7*R7))+(A0*A0)))
      RETURN
      END
      SUBROUTINE SORCE5
      COMMON B,B0,B1,BK0,C1,C2,C3,C4,C8,CK0,D,D0,D1,D10,E,E0,F,F0,F1,F2
      COMMON F4,F5,F6,F7,F8,F9,I,I0,I2,IB,IC,ICB,K,K0,K1,K2,K9,KT,L1,L2
      COMMON M2,M4,N2,N8,N9,NM,P,R1,R2,R3,R4,R6,R7,R10,R11,R12,Q,Q0,Q1
      COMMON SUM,T,T0,W,A1,A2,A3,A4,A5,A6,A7,A8,A9,A0,R13,F3,M1,Q2,LX
      REAL I,K,K0,L1
      W=2.*P*F
      GO TO (50,150),M1
 50   R8=R3/(T0*T0)
      A0=((R6+R8)*(R1+R6))/(W*L1*R8)
      A1=(R1+R6+R6+R8)/R8
      A2=(R6*(R1+R6))/(W*L1)
      A3=R1+R6+R6
      E=(E*(A0*A0+A1*A1)/(T0*T0))*1.E18
      I=I*(A2*A2+A3*A3)*T0*T0*1.E18
      A4=(4.*K*T/R8)*(A2*A2+A3*A3)
      A5=1.+(R1+R6)*(R1+R6)/(W*W*L1*L1)
      D=4.*K*T*R1*1.E18
      D1=((4.*K*T*R6)+(4.*K*T*R6)*A5+A4)*1.E18
      RETURN
 150  R7=(R3+R4)/((R3+R4)*T0*T0)
```

```
      A0=W*L1*R7
      A1=(R1+R6)*(R6+R7)-W*W*L1*R7*C4*(R1+R6)
      A2=W*(R1+R6+R6)+R7+R6*R7*C4*(R1+R6)
      K0=(A0*A0*T0*T0)/(A1*A1+A2*A2)
      RETURN
      END
      SUBROUTINE SORCE6
      COMMON B,B0,B1.BK0,C1,C2.C3,C4,C8,CK0,D,D0,D1.D10.E,E0,F,F0.F1,F2
      COMMON F4,F5,F6,F7,F8,F9,I,I0,I2,IB,IC,ICB,K,K0,K1,K2,K9,KT.L1.L2
      COMMON M2,M4,N2,N8,N9,NM,P,R1,R2,R3,R4,R6.R7,R10,R11,R12,Q,Q0,Q1
      COMMON SUM,T,T0,W,A1,A2,A3,A4,A5,A6,A7,A8,A9,A0,R13,F3,M1,Q2,LX
      REAL I,K,K0,L1,L2
      W=2.*P*F
      A0=W*L2-(1./(W*C1))
      Z0=SQRT(R1*R1+A0*A0)
      B5=ATAN(A0/R1)
      A1=W*L1*R3
      A2=R3*(1.-W*W*L1*C2)
      B6=ATAN(W*L1/A2)
      B6=90.-B6
      Z2=SQRT(1.+W*W*C4*C4*R4*R4)
      B7=ATAN(W*R4*C4)
      Z2=R1/Z2
      B7=B7*(-1)
      GO TO (60.160),M1
 60   Z3=(R1+Z1*COS(B6))**2.+(A0+Z1*SIN(B6))**2.
      E=(E*Z3/(Z1*Z1))*1.E18
      I=I*Z1*Z1*1.E18
      D=4.*K*T*R1*1.E18
      D1=(4.*K*T*Z1*Z1/R3)*1.E18
 160  Z4=Z0*Z1*COS(B5+B6)+Z1*Z2*COS(B6+B7)+Z0*Z2*COS(B5+B7)
      Z5=Z0*Z1*SIN(B5+B6)+Z1*Z2*SIN(B6+B7)+Z0*Z2*SIN(B5+B7)
      K0=Z2*Z2*Z1*Z1/(Z4*Z4*Z5*Z5)
      RETURN
      END
      SUBROUTINE PRINT1
      WRITE(6.71)
 71   FORMAT(/,8X,'VOLTAGE SOURCE MODEL',/,2X,'FREQUENCY')
      RETURN
      END
      SUBROUTINE PRINT2
      WRITE(6.72)
 72   FORMAT(11X.'SENSOR EN     LOAD EN     AMPL EN     AMP IN',
     + ':ZS    SUM NOISE NANOVOLTS')
      RETURN
      END
      SUBROUTINE PRINT3
      WRITE(6,76)
 76   FORMAT(1H1.//,'FREQUENCY')
      WRITE(6,77)
 77   FORMAT(12X,'SENSOR IN     LOAD IN      AMP IN     AMP EN;ZS',
     +'   SUM NOISE  PICOAMPERS')
      RETURN
      END
      SUBROUTINE PRINT4
      COMMON B,B0,B1,BK0,C1,C2,C3,C4,C8,CK0.D,D0,D1,D10,E,E0,F,F0.F1,F2
      COMMON F4.F5,F6,F7,F8,F9,I,I0,I2,IB,IC,ICB,K,K0,K1,K2,K9,KT,L1.L2
      COMMON M2.M4,N2,N8,N9,NM,P,R1,R2,R3,R4,R6,R7,R10,R11,R12,Q,Q0,Q1
      COMMON SUM,T,T0,W,A1,A2,A3,A4,A5,A6,A7,A8,A9,A0,R13,F3,M1,Q2,LX
      REAL K0
      BK0=SQRT(K0)
      CK0=4.3429*ALOG(K0)
      MF=F
      WRITE(6,74) MF.BK0,CK0
 74   FORMAT(I8,2F10.3)
      IF(F-B1+0.1*F) 99,99,98
 99   F=F*1.2589254
      LX=1
      RETURN
 98   LX=2
      RETURN
      END
      SUBROUTINE PRINT5
      WRITE(6,73)
 73   FORMAT(/,2X,'FREQUENCY    GAIN     GAIN DB')
      RETURN
      END
```

```
      SUBROUTINE PLOT1(NSPC,SPLOT,FPLOT,JB,KP,IT,JT)
      DIMENSION FPLOT(200),SPLOT(200)
      DIMENSION ISARET(4),IBUF(300),IS(300),IY(300),YI(300)
      DATA ISARET/1H ,1H*,1H+,1H./
      DATA AUNIT/10HNANOVOLTS /
      DATA BUNIT/10HPICOAMPERS/
      DATA CUNIT/10HDESIBEL    /
      GO TO (19,20,16),JB
   19 UNIT=AUNIT
      GO TO 21
   16 UNIT=CUNIT
      GO TO 21
   20 UNIT=BUNIT
   21 PLOTX=SPLOT(1)
      K=1
   11 K=K+1
      IF(K.GT.NSPC) GO TO 12
      IF(SPLOT(K).GT.PLOTX) GO TO 11
      PLOTX=SPLOT(K)
      GO TO 11
   12 DO 13 K=1,NSPC
   13 SPLOT(K)=SPLOT(K)-PLOTX
      TA=SPLOT(1)
      DO 14 L=2,NSPC
   14 IF(SPLOT(L).GT.TA) TA=SPLOT(L)
      IF(JT-1) 25,25.24
   24 KT=KP
      GO TO 26
   25 KT=IFIX(TA/10.+1.)
   26 TX=TA-PLOTX
      IF(TA.GT.1.) GO TO 17
      TC=10.
      TD=100.
      IF(JT-1) 28,28,27
   27 IY(1)=IT
      GO TO 18
   28 IY(1)=IFIX(PLOTX)
      GO TO 18
   17 TC=1.
      TD=10.
      IF(JT-1) 30,30,29
   29 IY(1)=IT
      GO TO 31
   30 IY(1)=PLOTX-1.
   31 IF(PLOTX.LT.1.) IY(1)=0.
   18 IT=IY(1)
      YI(1)=FLOAT(IY(1))
      DO 2 I=1,11
      IY(I+1)=IY(I)+KT
    2 YI(I+1)=YI(I)+(FLOAT(KT))/TC
      WRITE(6,100) (YI(JK),JK=1,12),UNIT
      DO 3 KI=1,125
    3 IBUF(KI)=ISARET(2)
      DO 4 I1=1,121,10
    4 IBUF(I1)=ISARET(3)
      WRITE(6,111) (IBUF(KL),KL=1,125)
  111 FORMAT(10X,125A1)
      DO 5 I=1,NSPC
      IS(I)=IFIX((SPLOT(I)+(PLOTX-(FLOAT(IY(1)))))*TD/(FLOAT(KT)))+2
      KP=KT
      DO 6 J=2,124
      IBUF(J)=ISARET(1)
      IF(J.EQ.IS(I)) IBUF(J)=ISARET(2)
    6 CONTINUE
      IBUF(1)=ISARET(2)
      IBUF(123)=ISARET(2)
      IBUF(125)=ISARET(2)
    5 WRITE(6,110) FPLOT(I),(IBUF(JK),JK=1,125)
  100 FORMAT(6X,F6.1,11F10.1,1X,A10)
  110 FORMAT(F9.1,125A1)
      RETURN
      END
      SUBROUTINE PLOT2(NSPC,SPLOT,FPLOT)
      DIMENSION ISARET(4),IBUF(300),IS(300),IY(300)
      DIMENSION FPLOT(200),SPLOT(200)
      DATA ISARET/1H ,1H*,1H+,1H./
      WRITE(6,90)
      PLOTX=900.
```

```
      K=1
  11  K=K+1
      IF(K.GT.NSPC)  GO TO 12
      IF(SPLOT(K).GT.PLOTX) GO TO 11
      PLOTX=SPLOT(K)
      GO TO 11
  12  DO 13 K=1,NSPC
  13  SPLOT(K)=SPLOT(K)-PLOTX
      DO 10 I=1,NSPC
  10  IS(I)=IFIX(10.*(ALOG10(SPLOT(I)+0.01)))
      IBIG=IS(1)
      DO 20 I=2,NSPC
  20  IF(IS(I).GT.IBIG)   IBIG=IS(I)
      NORM=IFIX(FLOAT(IBIG-100)/10.)*10
  30  DO 40 I=1,12
  40  IY(I)=NORM+(I-1)*10
      WRITE(6,100) (IY(JK),JK=1,12)
      DO 7 KI=1,125
   7  IBUF(KI)=ISARET(2)
      DO 50 I1=1,121,10
  50  IBUF(I1)=ISARET(3)
      WRITE(6,111) (IBUF(KL),KL=1,125)
 111  FORMAT(8X,125A1)
      DO 1 I=1,NSPC
      IS(I)=IS(I)-NORM
      DO 2 J=2,124
      IBUF(J)=ISARET(1)
      IF(J.EQ.IS(I).AND.J.LT.111) IBUF(J)=ISARET(2)
   2  CONTINUE
   5  IBUF(1)=ISARET(2)
      IBUF(111)=ISARET(2)
      IBUF(125)=ISARET(2)
   1  WRITE(6,110) FPLOT(I),(IBUF(JK),JK=1,125)
  90  FORMAT(5X,'GAIN')
 100  FORMAT(5X,I5,11I10,'DESIBEL')
 110  FORMAT(1X,F8.0,125A1)
      RETURN
      END

READY.
```

APPENDIX B

Symbol equivalents used in "BJT" and "TMFET" Subroutines :

R10 = $r_{bb'}$ = Base-spreading resistance

C8 = $C_{b'e}$ = Base-emitter capacitance

IC = I_C = Collector current

ICB= I_{CBO} = Collector-to-base leakage

F7 = f_L = 1/f noise corner

F8 = f_T = frequency at which $\beta = 1$

Q1 = γ

Q2 = α

B = β_o = Short-circuit current gain

C8 = C_{gs} = Gate-channel capacity of FET's

B = g_{ms} = Mutual conductance of FET's

ICB= I_g = Gate Leakage Current

F8 = f_o = 1/f noise corner

Symbol equivalent used in "SOURCE" subroutines :

R1 = R_s = Sensor resistance

R2 = R_b = Bias resistance

R3 = R_L = Load resistance

R4 = R_i = Amplifier input resistance

R6 = r_p = Resistance of the transformer primary

C1 = c_p = Shunt capacitance of sensor or wiring

C2 = C_W = Wiring capacitance

C3 = c_c = Coupling capacitance

C4 = C_i = Amplifier input capacitance

C1 = C_S = Mechanical capacitance

L1 = L_p = Sensor series inductance

L1 = L_p = Transformer secondary inductance

L2 = L_M = Mechanical inductance

F9 = f_L = 1/f noise corner

I2 = I_{DC}= Sensor Leakage current

T0 = T = Transformer secondary to primary turn ratio

L1 = L_x = External inductance

APPENDIX C

CONTENTS OF DATA CARDS 3.4 AND 5

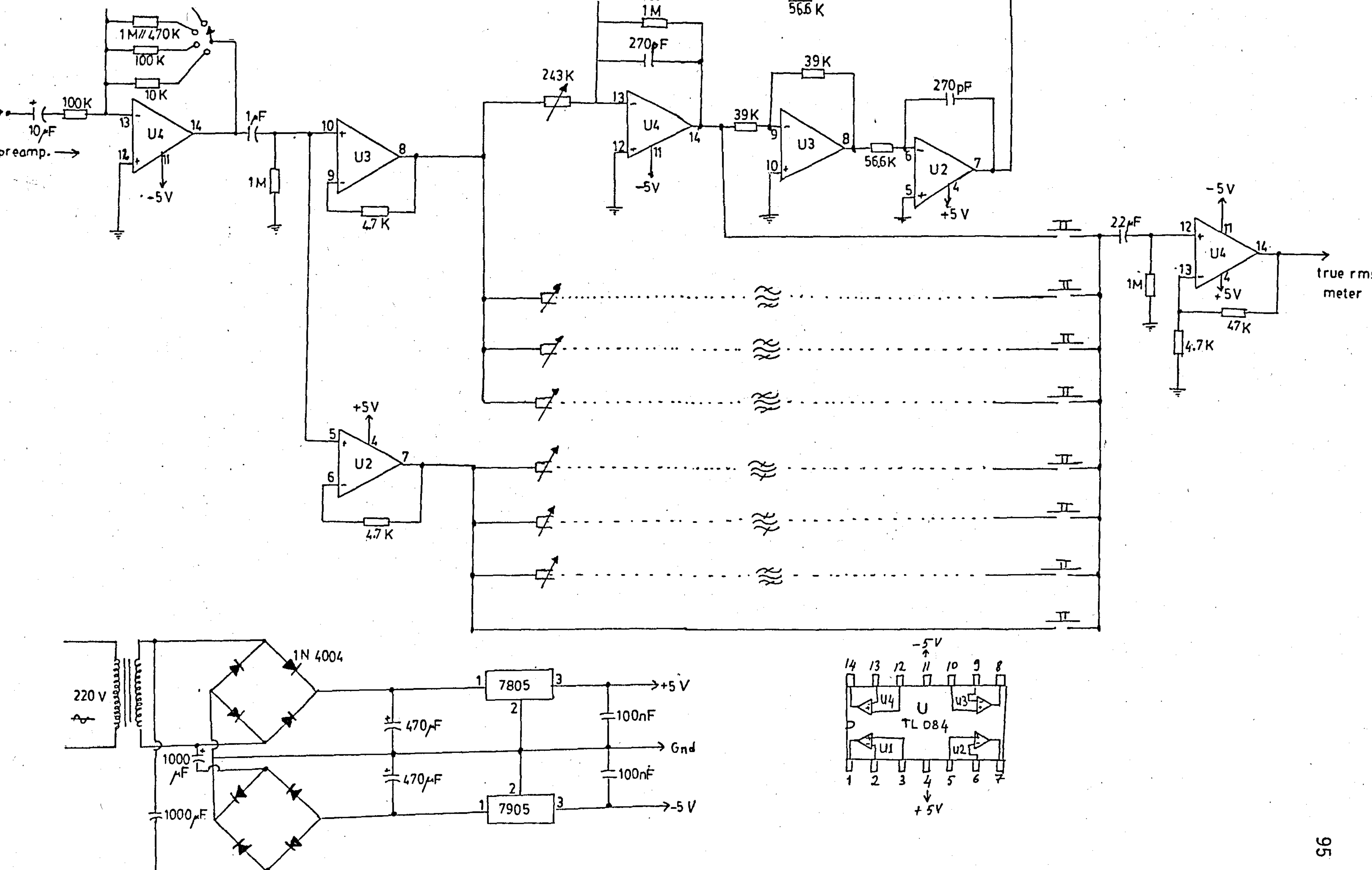
SOURCE DATA				AMPLG DATA	
Card 3		Card 4		Card 5	
Qnty.	Value	Qunty.	Value	Qunty.	Value
R1	1300.	L1	1.E4	F0	0.1
R2	1.E8	L2	500.E-3	F1	50.
		I2	I.E-12	F2	500.
R3	1.E0	T	3.E2	F3	1.E9
R4	47.E3	T0	1.	F4	1.E9
R6	1.E0	F9	1.E-3	F5	1.E10
C1	1.E-10			F6	2.1E3
C2	2.E-12			Q0	1.E-3
C4	400. E-12			N8	1.
				N9	1.
				K2	300.

APPENDIX D

CONTENTS OF DATA CARDS 6,7,8,9,10,11,12, AND 13

Calculation	Card Number	frequencies (Hz)
1. Equivalent input noise over a band	6	1.50.500.2100.
2. Input network frequency response	7	30.2000.
3. Equivalent input noise at one frequency	8	40.500.15000.20000.
4. Equivalent input noise versus frequency	9	10.25000.
5. Total noise at the output	10	1.50.500.2100.3000.
6. Total system gain	11	5.40000.
7. Noise bandwidth	12	1.50.1000.2100.3000.5000. 1.E4 15.E3 2.E4
8. Noise figure	13	100.

APPENDIX E



preamp.
10μF
100K
1M// 470K
100K
10K
U4
-5V
1μF
1M
U3
4.7K
U2
+5V
243K
270pF
39K
56.6K
22μF
47K
4.7K
true rms meter
220 V
1N 4004
7805
7905
1000 μF
470μF
100nF
+5 V
Gnd
-5 V
TL 084
U1

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