THEORETICAL ANALYSIS OF PREAMPLIFIER
(TRANSISTORIZED) FOR INFRARED TRACKING
STSYEMS WITH PHOTOCONDUCTIVE OR
PHOTOVOLTAIC DETECTORS

By
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SUMMARY.

With detectors having high theoretical performance available, requirements on the preamplifier should be sufficiently stringent that the performance of the photodiode is not degraded by the preamplifier. When the detector is connected to a preamplifier the problem becomes one of determining how the noise figure of the preamplifier affects the detectivity of the detector. Therefore the detector signal noise and detectivity characteristics as well as the preamplifier requirements are discussed. Equations that must be fulfilled by the preamplifier in order to utilize the full detectivity of the detector are derived.

I. INTRODUCTION

The most common form of radiation detectors used in infrared tracking systems is the lead sulfide film and the more recent indium antimonide detectors. Lead sulfide film typically have a surface resistivity of about one million ohms and produce an output rms noise voltage of the order of 1 µV in a 100 cps pass band centered at 1,000 cps. This corresponds to a noise power of $10^{-16}$ watts. The Johnson noise power from a resistor under similar conditions and at room temp. is $4 \times 10^{-13}$ watts. The cooled longer wavelength sensitive detectors such as lead selenide and lead telluride may have resistances 3 to 100 times higher, but they exhibit a similar ratio of noise power to Johnson noise power. The more recent indium antimonide detectors are much lower-impedance devices typically of the order of 200 ohms. The noise output of these cells may be even close to Johnson noise. It is generally difficult to have preamplifiers for use with these detectors in such a way that the combination is limited in detectivity by the noise of the detector. In this paper the detector signal, noise and detectivity characteristics as well as the transistorized preamplifier requirements will be investigated.

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11. GENERAL DISCUSSION ON TRANSISTORIZED PREAMPLIFIER

There are many problems peculiar to low level amplifiers working between input powers of the order of a fraction of a microwatt and output power of the order of a milliwick. Among the problems most generally encountered in these low level applications are the limitation dictated by signal-to-noise ratio; impedance levels from and into which this amplifiers are frequently required to operate; low distortion levels; and occasionally special frequency characteristics. When transistors were first introduced they were unsuitable for operation at power levels much below 1 milliwick because of their high noise level. With the advent of the junction transistor this condition was improved tremendously, since the noise level of the latter devices was three or four orders of the magnitude lower than that of the original point-contact devices. Improvement has continued until now the noise level is of the same order of magnitude as for vacuum tubes. Transistor noise differs radically from thermal or "white" noise in that the noise power per unit bandwidth for the former varies approximately inversely with frequency, as contrasted to the constant noise power in the latter. Fig. 1 shows a typical variation of noise with frequency for a transistor amplifier, indicating the inverse-frequency relationship. It points out one of the major difficulties in the utilization of transistors, in that the noise increases toward the low-frequency end, making the problem of designing d-c amplifier that much more difficult.

![Figure 1](image-url)
THEORETICAL ANALYSIS OF PREAMPLIFIER (TRANSISTORIZED) FOR INFRARED TRACKING SYSTEMS WITH PHOTOCONDUCTIVE OR PHOTOVOLTAIC DETECTORS

1. Noise as a function of operating point.

The noise figure is normally given in terms of a cycle bandwidth at 1000 cycles and room temperature. Current junction transistors have noise figures ranging from an occasional 3-4 db. up to about 30 db, with the average about 15-20 db. This figure is being continually improved, and noise figure differs only slightly between the grounded-base and grounded-emitter arrangements. The transistor noise arises from the emitter and the collector and the total noise is a composite of these two, the relative contributions being a matter of the physics of the transistor and of the operating point. Fig. 2 shows these two components. The contribution due to emitter noise is relatively low and independent of collector voltage. The collector noise increases with collector voltage; thus the overall noise has the rising characteristic shown. Fig. 3 Shows the total noise figure for three junction transistors, indicating the variation with emitter current and with collector voltage. The effect of emitter current is considerably less than that of collector voltage. These curves indicate that best operation from a low-noise standpoint will be obtained if the transistor is operated at collector voltage of about 1 volt, emitter current 1 milliampere or less.

2. Impedance considerations.

The noise figure of a transistor amplifier varies considerably with the source resistance and has a minimum for a peculiar value, depending upon the relative magnitudes of emitter and collector noise. The optimum source resistance will normally lie between 100 and 5000 ohms; however, it does not show very great variation over this range and a source resistance of the order of 500-1000 ohms will usually produce minimum noise for most combinations. Fig. 4 shows the variation of noise figure with source resistance for typical junction units.

In I.R. tracking system, with photoconductive detectors the preamplifier is required to operate from a high-impedance source. We must utilize the best possible
arrangement for operation from this high-impedance source. Under these conditions we must have an amplifier input impedance that is relatively independent of transistor load and of the various transistors themselves.

Three relatively simple methods may be employed to achieve the required high input impedance:

1. use of the grounded-collector configuration;
2. use of the grounded-emitter configuration in series with a large resistor;
3. use of degeneration in conjunction with the grounded-emitter configuration.

The general equations for input impedance and current amplification of a transistor amplifier are as follows:

\[
Z_I = \frac{\Delta h + h_{11}y_e}{h_{22} + y_e} \quad \text{(1)}
\]

\[
A_I = \frac{h_{22}y_e}{h_{22} + y_e} \quad \text{(2)}
\]

where

\[
\Delta h = h_{11}h_{22} - h_{12}h_{21} \quad \text{(3)}
\]

To compare a grounded-collector stage with one employing the grounded-emitter configuration where the input impedance \( h_{11} \) has been supplemented by an additional
resistance $r'$, such that the input resistances of the two amplifiers are equal for the same load admittance $y$. We have, further, the following relation between the corresponding parameters of the two configurations:

$$h_{21}' = h_{21}''$$

$$h_{21}' = h_{21}'' - 1 = \alpha h_{21}''$$

$$h_{22}' = h_{22}''$$

$$\Delta h' = \Delta h'' \Delta h$$

where unprimed quantities refer to the grounded-base parameters, primed refer to grounded-emitter, and double-primed quantities are for the grounded-collector.

From the above it is evident that, since the $h_{22}$ parameters are identical, the denominator in (2) will be the same for the two arrangements. The current amplifications will, therefore, be in the ratio of the $h_{21}$ parameters, or $\alpha$ as above. Since $\alpha$ is very nearly in most junction transistors, the current amplification of the two amplifiers is essentially the same. In compensation for this relatively slight loss in current amplification we have an input resistance largely made up of the series resistor $r'$, hence practically independent of the transistor parameters or loading. This indicates that the grounded-emitter arrangement plus series resistor is superior to the other configuration for this purpose. But do not supply the bias through this resistor $r'$ if good bias stabilization is expected.

One very simple method of obtaining high input resistance with the grounded-emitter arrangement is by means of an unbypped resistor in the emitter lead, as...
shown in fig. 5. This circuit may be analyzed rather easily by adding the resistor \( r_e' \) in series with the parameter \( h_{11} \) for the grounded-base configuration and converting to the corresponding grounded-emitter equation. We now have

\[
\begin{align*}
    h_{11}' &= \frac{h_{11}}{1 + h_{21}} \\
    h_{11}'' &= \frac{h_{11} + r_e'}{1 + h_{21}}
\end{align*}
\]

(the double-prime now refers to the degenerative arrangement)

Again, referring to equation (2) we note that none of the \( h \) parameters in the equation are affected by the addition of the resistor \( r_e' \), hence the current amplification will be unchanged and will also be \( \alpha \) times that of the grounded-collector stage.

The input impedance of the degenerated grounded-collector stage may be found as follows:

From equation (1)

\[
Z = \frac{\Delta^b + h_{11}y_z}{h_{22} + y_z}
\]

For the grounded-emitter stage

\[
\begin{align*}
    \Delta^b &= \frac{\Delta^b}{1 - h_{21}} + \frac{h_{11}h_{22} - h_{12}h_{21}}{1 + h_{21}} \\
    h_{11}' &= \frac{h_{11}}{1 + h_{21}} \\
    h_{22}' &= \frac{h_{22}}{1 + h_{21}}
\end{align*}
\]

Substituting in (1):

\[
Z = \frac{h_{11}h_{22} - h_{12}h_{21} + h_{11}y_z}{1 - h_{21} + h_{21}}
\]

For the degenerated grounded-emitter:

\[
Z = \frac{(h_{11} + r_e')(h_{22} + y_z) - h_{12}h_{21}}{h_{22} + y_z(1 + h_{21})}
\]

We now using the 2N44 transistor and a load admittance of \( 10^{-3} \) mho, to be an example compare the above three arrangements.

For the grounded-collector:

\[
Z_1 = 22,600 \text{ ohms} \\
A_1 = 21.6
\]

For the grounded-emitter:

\[
Z_1 = 865 \text{ ohms} \\
A_1 = 20.6
\]

Series resistor to bring input up to that of grounded-collector: 21,735 ohms. Loss compared to grounded-collector: 0.2 db.

For the degenerated grounded-emitter:

Solving equation (9) gives \( r_e' = 1000 \) ohms.

With this value

\[
Z_1 = 22,000 \text{ ohms} \\
A_1 = 20.6
\]
THEORETICAL ANALYSIS OF PREAMPLIFIER (TRANSISTORIZED) FOR INFRARED TRACKING SYSTEMS WITH PHOTOCONDUCTIVE OR PHOTOVOLTAIC DETECTORS

From this standpoint the two grounded-emitter arrangements are identical. Comparing the three arrangements with respect to effect of load variation by making \( y_e = 2 \times 10^{-4} \), gives the following input resistance for the three,

<table>
<thead>
<tr>
<th>Arrangement</th>
<th>Resistance</th>
</tr>
</thead>
<tbody>
<tr>
<td>Grounded-collector</td>
<td>11,800 ohms</td>
</tr>
<tr>
<td>Grounded-emitter</td>
<td>22,600 ohms</td>
</tr>
<tr>
<td>Degenerated grounded-emitter</td>
<td>22,800 ohms</td>
</tr>
</tbody>
</table>

The above example shows that the degenerated grounded-emitter arrangement is essentially the equal of the undegenerated stage with series resistor. There is, however, one other factor which acts very much in favor of the degenerated arrangement. This is the fact that a series resistor in the emitter lead helps bias stability and also reduces the effects of variation of transistor parameters. The conclusion to draw from the above, therefore, is that where it is desirable to present a high and relatively constant impedance to a source the best configuration is the grounded-emitter with unbypassed emitter resistance.

III. USE INDIUM ANTIMONIDE AS AN EXAMPLE

The indium antimonide single crystal infrared detector shows promise of reaching the ultimate limit of detectivity termed the Blip (Background limited infrared photodetector) condition of the operation. With infrared detectors having such high theoretical performance available, requirements on the preamplifier should be sufficiently stringent that the performance of the photodiode is not degraded by the preamplifier. When the detector is connected to a preamplifier the problem becomes one of determining how the noise figure of the preamplifier affects the detectivity of the detector.

The indium antimonide infrared detector is a simple p-n junction with radiation incident normal to the junction. In thermal equilibrium a p-n junction has voltage-current characteristics given by

\[ I_d = I_s(e^{qV/kT} - 1) \]

where \( I_d \) is the junction current, \( I_s \) is the saturation current, \( q \) is the electronic charge, \( V \) is the drop in potential across the junction of the diode, \( k \) is the Boltzmann's constant, \( T \) is the equilibrium temp. of the diode, and \( \beta \) is a factor which is unity for an ideal diode, but is found to be greater than unity in cooled indium-antimonide junctions.

When radiation from a background with a temperature higher than the equilibrium temperature of the diode falls onto the detector, the voltage-current equation becomes:

\[ I = I_{se} + I_s(e^{qV/kT} - 1) \]

\( I_{se} \) is the current induced by the incident background radiation. \( I_{se} \) can be expressed in terms of the background radiation flux as

\[ I_{se} = q\eta r J_r A \]

where

\[ J_r = \int_{\lambda_1}^{\lambda_2} d\lambda J(\lambda) \]
\( \mathcal{F} \) is the radiation flux expressed in photons per second per square centimeter of area, \( A \) is area in square centimeters, \( \eta_r \) is the quantum efficiency with respect to background radiation and \( \lambda_1 = \frac{hc}{E_1} \) is the long wavelength cut-off of the detector, corresponding to the energy gap \( E_1 \). The radiation flux, \( \mathcal{F} \), and the current density \( j_r \) for an indium antimonide photodiode with an equilibrium temperature of 77°K while looking at 300°K radiation are

\[
\mathcal{F} = 2 \times 10^{16} \, \text{photons/cm}^2 \text{ sec}
\]

and

\[
j_r = qJ_r = 3.2 \times 10^{-3} \, \text{amp/cm}^2.
\]

An additional path for current in diffused alloyed or grown photodiodes is a conductance, \( G_s \), which is in shunt with the diode. When \( G_s \) is incorporated in the voltage current equation (17) becomes

\[
I = -I_{sc} + I_s (e^{qV/\beta kT} - 1) + G_s V
\]

Figure 6 shows the typical voltage-current curves for indium antimonide photodiodes. In curve A, the detector is at thermal equilibrium at 77°K with no excess background radiation; curve B shows the effect of adding 300°K radiation while the diode is held at 77°K and curve C shows the effect of still more background radiation. Note that \( I_{sc} \) can be taken directly from Fig. 6 at the intercept with the current axis. The slope of the curve in the reverse biased condition would approach zero for an ideal photodiode, but in the actual photodiode this slope gives the value for the shunt conductance \( G_s \). By plotting \( \log (1 - I_{sc} - G_s V) \) vs \( V \) for the forward-bias condition the values of \( \beta \) and \( I_s \) can be determined. The value of \( \beta \) has varied from 2 to 3.8 for the photodiodes measured.

1. Small-Signal Properties

Response of the photodiode to a small radiation signal, \( J_s \), in the presence of the background radiation, is obtained from (18) as

\[
\Delta I = \left[ -\frac{dI_{sc}}{dJ_s} + \frac{q}{\beta kT} I_s e^{qV/\beta kT} \frac{dv}{dJ_s} + G_s \frac{dv}{dJ_s} \right] \Delta J_s
\]

The small signal short-circuit current generator for a Norton representation is found by holding \( V \) constant:

\[
i_s = (\Delta I)_{v=\text{const}} = -\frac{dI_{sc}}{dJ_s} \Delta J_s = -q\eta_s A \Delta J_s
\]

where \( \eta_s \) is the quantum efficiency with respect to the signal radiation, The designation \( \eta_s \) is used to distinguish it from \( \eta_r \) because of the possible difference in the spectral character of the signal radiation and background radiation.

Differentiating (18) with respect to \( V \) while holding \( J_s \) constant:

\[
\Delta I = (G_d - G_s) \Delta V
\]

where

\[
G_d = \frac{qI_s}{\beta kT} e^{qV/\beta kT}
\]

and

\[
R_s = \frac{1}{G_s}
\]

At low frequencies (22) and (23) indicate the Norton equivalent circuit that is shown in Fig. 7.
2. Noise Properties

The fundamental noise associated with an ideal narrow-base diode can be expressed in terms of shot noise for each current:

\[ \frac{\bar{\eta}^2}{\Delta f} = 2q \left[ I_s + I_a (e^{qV/\beta kT} + 1) \right] \]

When \( \beta \) is considered, the shot noise equation for the photodiode is

\[ \frac{\bar{\eta}^2}{\Delta f} = 2q \left[ I_{sc} + \frac{I_s}{\beta} (e^{qV/\beta kT} + 1) \right] \]

This equation is correct at thermal equilibrium (\( I_{sc} = 0, V = 0 \)) since

\[ 0 = \frac{4qI_s}{\beta} = 4kT \eta \]

is agreement with the Nyquist result.

The shunt conductance has a Nyquist noise

\[ \frac{\bar{\eta}^2}{\Delta f} = 4kT \eta \]

Experiment has indicated that the main source of 1/f noise is the current which passes through the shunt conductance. An empirical equation for the 1/f noise can be written as

\[ \frac{\bar{\eta}^2}{\Delta f} = \frac{1}{f} \left[ k_1 G_s z^2 + k_2 (1 - G_s V)^2 + k_3 z^2 \right] \]

where \( k_1 > k_2 \) and \( k_3 \)

The values for \( k_1, k_2 \) and \( k_3 \) must be determined experimentally, since no quantitative theory of 1/f noise exists.

The equation for the total short-circuit noise generator is obtained from (27), (31) and (30)

\[ \frac{\bar{\eta}^2}{\Delta f} = 2q \left[ I_{sc} + \frac{I_s}{\beta} (e^{qV/\beta kT} + 1) \right] + 4kT \eta \]

\[ + \left[ \frac{k_1 G_s z^2 + k_2 (1 - G_s V)^2 + k_3 z^2}{f} \right] = N_u \]

3. Signal-To-Noise Ratio and Detectivity

When (22) and (31) are combined, the signal-to-noise ratio the photodiode can be expressed as

\[ \frac{I_s}{\sqrt{\bar{\eta}^2}} = \sqrt{\frac{q \eta A \Delta f}{V}} \]
where $E_a$ is the energy per photon of the signal radiation, and $N_d$ is the detector noise. This is a measure of $S/N$ per unit of incident power, normalized to 1 cm$^2$ area and unit bandwidth.

When $\eta_s$ is substituted into (34) and using (31)

$$D_d^* = \frac{q\eta_s}{E_a \left( j_{se} + \frac{j_s}{\beta} (e^{qV/\beta kT} + 1) \right) + \frac{4kT G_a}{\beta}}$$

and there is no l/f noise, then this equation becomes that of the ideal radiation detector that is the Blip (background-limited infrared photodetector)

$$D^{*}_{Blip} = \frac{1}{E_aV 2J_r}$$

$D^{*}_{Blip}$ for indium antimonide facing hemispherical 300°K background radiation is

$$D^{*}_{Blip}(500°K) = 24 \times 10^9 \text{ watts}^{-1}$$

and $D^{*}_{Blip}(\lambda_f = 5.35\mu) = 13.6 \times 10^{10} \text{ watts}^{-1}$

IV. DETECTOR NOISE FIGURE

A quantitative measure of the performance of a photo-detector is the noise figure, $F_d$ defined by

$$F_d = \frac{(S/N)^2_{Blip}}{D^2_{d}}$$

Substitute (33) and (29) into (37), one finds

$$F_d = \frac{1}{\eta_s^2 \left( \frac{N_n}{2qJ_r} \right)}$$

where $N_n$ is defined in (31) and

$I_r = qJ_r A$

Note that for a detector which is not limited at all by radiation noise, $F_d$ depends on the quantum efficiency as $1/\eta_s^2$, but for detectors which see radiation noise (38) can be written

$$F_d = F_d (\text{quantum eff.}) \cdot F_d (\text{noise})$$

where

$$F_d (\text{quantum eff.}) = \frac{\eta_s}{\eta_s^2 + \frac{2qI_r}{\beta} (e^{qV/\beta kT} + 1) + 4kT G_a + k_1 G_a V^2 + k_2 (1-G_a V) + k_3 I^2}$$

and

$$F_d (\text{noise}) = 1 + \frac{2qI_r}{\beta}$$

For radiation signals of spectra $0 \leq \lambda_i \leq 5.35\mu$, $\eta_s$ can be taken equal to $\eta_r$ so that
THEORETICAL ANALYSIS OF PREAMPLIFIER (TRANSISTORIZED) FOR INFRARED TRACKING SYSTEMS WITH PHOTOCONDUTIVE OR PHOTOVOLTAIC DETECTORS

\[ F_d(\text{quantum eff.}) = \frac{1}{\eta_s} \] (4)

For the example of Fig. 6,

\[ F_d(\text{quantum eff.}) = \frac{1}{0.38} = 2.6 \text{ or } 4.2 \text{ db} \]

\[ \eta_s = \frac{A_{en}}{A_{en}} = 3.16 \times 10^7 \text{ cm}^2 \]

Ambient temp = 300°K, \( V_{oo} = 57 \times 10^{-3} \text{ V} \), \( I_s = 50 \times 10^{-6} \text{ Amp} \), \( R_s = 7140 \Omega \), \( I_s = 2.6 \times 10^{-4} \text{ Amp} \), \( R_a(V=0) = 7700 \Omega \), \( R_a(V=0) = 3700 \Omega \), \( k_e = 0.38 \), \( \beta = 3.0 \)

Fig. 6. Voltage-current characteristics for an InSb photodiode under several conditions of ambient radiation.

To calculate \( F_d(\text{noise}) \) precisely the parameters specified in (4) must be determined. of these, only \( K_1 \), \( K_2 \) and \( K_3 \) require direct noise measurements; all the other parameters can be taken from the voltage-current curves as previously discussed. It is of interest to calculate \( F_d(\text{noise}) \) at \( V=0 \), where the major source \( (K_1 G_s^2 V^2) \) of \( 1/f \) noise is zero. The \( 1/f \) noise associated with the terms containing \( K_2 \) and \( K_3 \) is neglected. Then

\[ [F_d(\text{noise})]_{V=0} = 1 + \frac{1}{M} \]

(42)
where

\[ M = \frac{qI_{S}R_{H}}{2kT} \]

\[ \gamma = \frac{1}{R_{H}} = G_{s} + G_{d} \]

Equation (42) is useful since it gives the lower limit of \( F_{d}(\text{noise}) \) for \( V = 0 \). The presence of \( 1/f \) noise will cause \( F_{a} \) to be larger than this value.

For the example of Fig. 6,

\[ (F_{d}(\text{noise}))_{V = 0} = \frac{15}{14} \text{ or } 0.3 \text{ db} \]

thus the detector should be only 0.3 db from a 100 per cent radiation noiselimited condition if there is no \( 1/f \) noise present.

The total theoretical noise figure for the detector is

\[ F_{a} = 2.6 \left( \frac{15}{14} \right) = 2.8 \text{ or } 4.5 \text{ db} \]

This indium antimonide detector should be only 4.5 db from the fundamental limit of detectivity, that is

\[ D_{d}*(500^\circ K) = \frac{D_{d}*(514^\circ F)}{2.8} = 14 \times 10^{11} \text{ watts}^{-1} \]

V. EFFECTIVE NOISE FIGURE OF PREAMPLIFIER

When the detector is connected to a preamplifier the problem becomes one of determining how the noise figure of the preamplifier affects the detectivity of the detector. The usual amplifier noise figure, \( F_{a} \), is defined relative to the Nyquist noise of a resistor \( R \) at room temperature. The amplifier contribution to the noise can be represented in an equivalent circuit diagram as a series noise generator

\[ \varepsilon_{a}^{2} = 4kT\Delta f \]

as shown in Fig. 8-A.

When the indium antimonide detector is connected in place of the Nyquist noise resistor, a different situation exists: the detector is cooled, and the source noise is not that of a pure resistance but that of the detector. If the equivalent circuit of the detector, Fig. 6 is converted to a voltage equivalent circuit as in Fig. 8-B and added in series with the amplifier noise generator, an effective noise figure \( F_{a}^{*} \), which has meaning relative to the cooled indium antimonide detector can be defined.

The effective noise figure of the amplifier can now be expressed in the same formal manner as the conventional noise figure.

\[ F_{a}^{*} = \left( \frac{S/N}_{a} \right)_{d+} = \left( \frac{S/N}_{a} \right)_{d+} \]

\[ = \left( \frac{D_{d}^{*}}{D_{d}^{*}} \right)^{2} = \left( \frac{N_{d+a}}{N_{d}} \right)^{2} \]

\[ \text{D*}_{d+a} \]

is the detectivity of the detector including the amplifier noise:

\[ \text{D*}_{d+a} = \frac{S\sqrt{\Delta f}}{N_{d+a}E_{a}A} \]

where \( N_{d+a} \) includes both the noise of the detector and preamplifier.

Substituting (32) and (44) into (43) and using the circuit of Fig. 8-B,

\[ F_{a}^{*} = 1 + \frac{4kT_{a}(F_{a} + 1)}{R_{H}N_{a}} \]

is obtained for the effective noise figure of the preamplifier. For indium antimonide detectors at the \( V = 0 \) condition and assuming \( k_{1} = k_{2} = 0 \), (45) becomes
THEORETICAL ANALYSIS OF PREAMPLIFIER (TRANSISTORIZED) FOR INFRARED TRACKING SYSTEMS WITH PHOTOCONDUCTIVE OR PHOTOVOLTAIC DETECTORS

**Fig. 7.** Norton's Equivalent Circuit for Indium-antimonide photodiode.

\[ \overline{e_a}^2 = 4kTR(F_a - 1) \Delta t \]
\[ \overline{e_n}^2 = 4kR \Delta t \]

**Fig. 8-A.** Equivalent Circuit for Preampifier Input

\[ \overline{e_a}^2 = 4kTR(F_a - 1) \Delta t \]
\[ \overline{e_n}^2 = 4kR \Delta t \]

**Fig. 8-B.** Equivalent Circuit for Detector-Preamplifier Combination.

\[ \overline{e_a}^2 = 4kTR_{11}(F_a - 1) \Delta t \]
\[ R_{11} = R_2 R_s / R_s R_a \]
\[ \overline{e_n}^2 = \overline{is}^2 R_{11}^2 \]

\[ c_s = is R_{11} \]

where \( T_d = 77^\circ K \) and \( T_a = 300^\circ K \). If the data from Fig. 6 are used along with \( F_a = 2 \) (3 db) as an example, \( F_a^* \) is 1.0 db, or 1.27 db. This means that the detectivity as measured through this amplifier \( (D_{d+a}^*) \) is a factor of 1.27 less than the actual detectivity of the detector \( (D_a^*) \) as can be seen from the definition of \( F_a^* \) in (30):

\[ D_{d+a}^*(500^\circ K) = \frac{D_a^*}{1.27} = 12.5 \times 19^y \text{ watts}^{-1} \]

To minimize the noise introduced by the amplifier \( Isc \) and \( R_{11} \) need to be as large as possible, and \( F_a \) should be as nearly unity as possible. A recent increase in
detector impedance as a result of improved detector technology has helped in its
direct effect in [4] as well as in its effect on transistor preamplifiers, which have
better noise figures with source impedances of a few thousand ohms rather than a
few hundred ohms.

VI. NOISE FIGURE OF DETECTOR-AMPLIFIER COMBINATION

The over-all noise figure of the detector with the amplifier can be defined as

\[ F_{d+a} = \frac{(S/N)_{d+a}}{(S/N)_{d}} \times \frac{(D_{d+a}^* D_{d}^*)^2}{(D_{d}^*)^2} \]  

or

\[ F_{d+a} = F_{d} F_{a}^* \]  

where \( F_{d} \) is given by (33) and \( F_{a}^* \) by (47). Using the example from Fig. 6,

\[ F_{d+a} = 4.5 \text{ db} + 1.0 \text{ db} = 5.5 \text{ db} \]  

and

\[ D_{d+a}^*(500^\circ \text{ K}) = \frac{D_{d}^* D_{11p}^*}{3.54} = 12.5 \times 10^9 \text{ watts}^{-1} \]  

in agreement with the results of the previous section.

This theoretical value for \( D_{d+a}^* \) is very satisfactory and indicates that contemporary indium antimonide photodetectors constitute very sensitive infrared detectors which can be connected directly into transistor preamplifiers without the use of a coupling transformer.

VII. EXPERIMENTAL RESULTS

A simple I.R. tracking system model has been constructed (Fig. 10). Various
transistorized preamplifiers have been used in it and measurements have been taken
to investigate the above conclusions. A simple preamplifier circuit using a selected
germanium transistor (2N185) is shown in Fig. 9. The characteristics of the amplifier

![Circuit Diagram for Transistor Preamplifier With InSb Detector](image)

\( F_{d+a} \) is given by (33) and \( F_{a}^* \) by (47). Using the example from Fig. 6,

\[ F_{d+a} = 4.5 \text{ db} + 1.0 \text{ db} = 5.5 \text{ db} \]  

and

\[ D_{d+a}^*(500^\circ \text{ K}) = \frac{D_{d}^* D_{11p}^*}{3.54} = 12.5 \times 10^9 \text{ watts}^{-1} \]  

in agreement with the results of the previous section.

This theoretical value for \( D_{d+a}^* \) is very satisfactory and indicates that contemporary indium antimonide photodetectors constitute very sensitive infrared detectors which can be connected directly into transistor preamplifiers without the use of a coupling transformer.

VII. EXPERIMENTAL RESULTS

A simple I.R. tracking system model has been constructed (Fig. 10). Various
transistorized preamplifiers have been used in it and measurements have been taken
to investigate the above conclusions. A simple preamplifier circuit using a selected
germanium transistor (2N185) is shown in Fig. 9. The characteristics of the amplifier

![Circuit Diagram for Transistor Preamplifier With InSb Detector](image)

\( F_{d+a} \) is given by (33) and \( F_{a}^* \) by (47). Using the example from Fig. 6,

\[ F_{d+a} = 4.5 \text{ db} + 1.0 \text{ db} = 5.5 \text{ db} \]  

and

\[ D_{d+a}^*(500^\circ \text{ K}) = \frac{D_{d}^* D_{11p}^*}{3.54} = 12.5 \times 10^9 \text{ watts}^{-1} \]  

in agreement with the results of the previous section.

This theoretical value for \( D_{d+a}^* \) is very satisfactory and indicates that contemporary indium antimonide photodetectors constitute very sensitive infrared detectors which can be connected directly into transistor preamplifiers without the use of a coupling transformer.
as a function of frequency are shown in Fig. 11 and the dependance of $F_a$ on source resistance is shown in Fig. 12. At present the noise figure for alloyed germanium transistors is superior to that of other transistors when operated in the "starved" condition. Silicon transistors can be obtained with noise figure between 3 and 6 db.

Two methods are used be biasing to the zero-voltage condition. A choke can be placed across the photodiode to provide a dc short while maintaining a high impedance for ac signals or the detector can be back-biased to the condition of zero voltage across the diode as shown in Fig. 10. The latter method has been used for making the experimental measurements. $R_b$ should be large compared to $R_s$ so that no
Nyquist noise is introduced by the biasing resistor. If $R_b$ is not large compared to $R_s$, the above formulas are still applicable provided $R_s$ is considered to be the parallel combination of $R_b$ and $R_s$.

Fig. 12. Preamplifier Noise Figure As a Function of Source Resistance

The experimental values for $D*_{d+s}$ (500°K) are measured with the transistor preamplifier shown in Fig. 10 coupled directly into a standard amplifier. Both noise voltage and signal voltage are measured with the detector biased to the zero dc voltage condition. Thus $D*_{d+s}$ as defined in (46) is measured experimentally without transformers.

The value of $D*_{d+s}$ for the detector of Fig. 6 as measured above was

$$ D*_{d+s}(500°K) \quad v=0 \quad f=1000cps \quad -8.7 \times 10^9 \, \text{watts}^{-1} \quad (53) $$

and

$$ F*_{d+s}(500°K) = \frac{24}{8.7} = 2.76 \quad \text{or} \quad 8.8 \, \text{db} \quad (54) $$

When the value from (54) or (53) is compared to the theoretical value from (51) or (53) it is found that the experimental value of $D*_{d+s}$ is 3.3 db below the theoretical...
value. The difference is due to the neglect of $1/f$ noise in calculating $D*_{d+a}$ (theoretical).

Summarizing the values obtained for the example of Fig. 6 it is found that

$$D*_{d+a} \text{ (exp)} = 8.7 \times 10^9 \text{ watts}^{-1}$$

and

$$F*_{d+a} \text{ (exp)} = 8.8 \text{ db}$$

of which

$$F_d \text{ (quantum eff.)} = 4.2 \text{ db}$$

$$F_d \text{ (noise other than } 1/f = 0.3 \text{ db}$$

$$F_d \text{ (i/f noise)} = 3.3 \text{ db}$$

$$F*_{d+a} \text{ (exp)} = 8.8 \text{ db}$$

VIII. CONCLUSION

The informations which have been discussed and formulas which have been derived for the treatment and calculation of the noise figure of the detector and the preamplifier separately should be useful in the design of preamplifiers for infrared tracking system with photoconductive or photovoltaic detectors.

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裝用光電導型或光電壓型感觸器之紅外線追綜系內
電晶體先期放大器靈敏性理論研討

郭銳冰

摘要

在紅外線追綜系統內，假如裝用靈敏度極高或近乎理論靈敏度的紅外線感觸器時，對先期放大器的要求必須十分嚴格。當紅外線感觸體接於先期放大器後，所要考量的問題將是究竟先期放大器的雜音度對感觸體的性能發生了多大的消弱影響。所以在本文詳述討論有關紅外線感觸器靈敏度的問題，信號雜音的問題，牠們之間的關係以及先期放大器雜音量與感觸體靈敏度的關係。根據理論及實驗獲得若干有關公式及曲線。籍這些公式及曲線將判定設計紅外線追綜系中電晶體先期放大器時所應遵守的的條件。

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