SLAC
TN 65-65
R. S. Larsen
3 August 1965

A LOW-LEVEL ANALYSIS FOR THE DUAL EMITTER CHOPPER TRANSISTOR

R. S. Larsen

SLAC
TN 65-64
R. S. Larsen
3 August 1965

A LOW-LEVEL ANALYSIS FOR THE DUAL EMITTER CHOPPER TRANSISTOR

1. Introduction

The dual or matched pair transistor chopper, consisting of two transistors connected in a series or series-shunt configuration, has become widely used. Furthermore, the inverted connection, wherein the base collector rather than the base-emitter junction is driven, has become popular because of the lower offset voltages obtainable. (The fact that the inverted connection results in higher on-resistances than the normal connection, however, has not been so widely advertised.)

Transistor choppers are commonly specified by their on-resistance as a function of drive current, R_{on}, and by the offset voltage across the switch with zero signal current, Δv_{12}. In the low-level theory, in which bulk resistances, nonlinear recombination, and changes in emitter efficiency with operating point are neglected, the offset voltage of a dual chopper is not a function of drive current but is related only to matching of the two units; whereas the on-resistance varies inversely with drive current.

The dual emitter chopper transistor is a relatively new device which has two emitters fabricated on a single base-collector junction. The obvious advantages to be gained are: better temperature matching of the two base-emitter junctions than would be possible in two separate units; and better matching of the doping characteristics of the junctions. Although one always expects some imperfections, the dual emitter device would appear to have an excellent chance of surpassing in performance matched pair switches, if for no other reason than inherently better temperature matching.

SLAC
TN 65-64
R. S. Larsen
3 August 1965

This report calculates the characteristics of the dual emitter chopper transistor by a straightforward extension of the low-level theory of Ebers and Moll[1]. The low-level theory is expected to apply over only a limited range, for a multitude of reasons. Primarily, as has been emphasized by Gibbons[2], in any practical application of a saturated or super-saturated switch, the conditions for true low-level operation are almost automatically violated.

The low-level theory will, however, give reasonable answers for R_{on} and $\Delta V12$ for drive currents less than 1 mA, but only order of magnitude answers for currents from 1 to 50 mA. The main value of the analysis is to point up some limitations of the low-level theory, and to provide a practical basis for estimating performance.

2. Generalized Ebers-Moll Equations

The dual emitter transistor is shown in Figure 1.

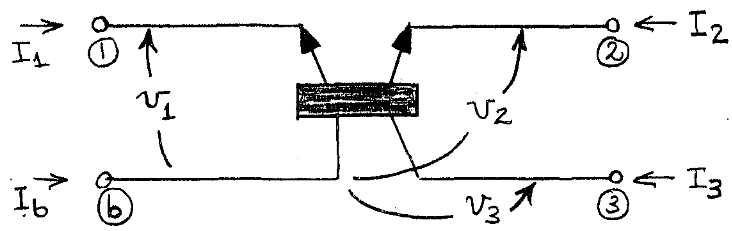


FIGURE 1

The Ebers-Moll equations, which are linear in terms of excess carrier densities in the base region, can be extended to the present case of a 4-terminal device as follows*:

---

*It was pointed out by J. F. Gibbons (private comm.) that this analysis can be easily extended to treat N-dimensional devices, such as multi-port integrated circuits.

$$\begin{pmatrix} I_1 \\ I_2 \\ I_3 \end{pmatrix} = \begin{pmatrix} a_{11} & a_{12} & a_{13} \\ a_{21} & a_{22} & a_{23} \\ a_{31} & a_{32} & a_{33} \end{pmatrix} \begin{pmatrix} \exp qv_1/kT - 1 \\ \exp qv_2/kT - 1 \\ \exp qv_3/kT - 1 \end{pmatrix} \tag{1}$$

where I_1, I_2, I_3 are d. c. junction currents

v_1, v_2, v_3 are junction voltages

$a_{11} \ldots a_{33}$ are generalized circuit parameters relating currents to excess carrier densities.

By alternately short-circuiting terminals 1, 2 and 3 to the base, one can set the excess densities, $Ni/np = (\exp qv_i/kT - 1)$, equal to zero and derive the following relationships between the a's and the d.c. short-circuit gains:

$$\begin{aligned} \alpha_{12} &= -\frac{a_{12}}{a_{22}} & \alpha_{13} &= -\frac{a_{13}}{a_{33}} \\ \alpha_{21} &= -\frac{a_{21}}{a_{11}} & \alpha_{23} &= -\frac{a_{23}}{a_{33}} \\ \alpha_{31} &= -\frac{a_{31}}{a_{11}} & \alpha_{32} &= -\frac{a_{32}}{a_{22}} \end{aligned} \tag{2}$$

where the α's are the d.c. short circuit current gains as defined in Figure 2:

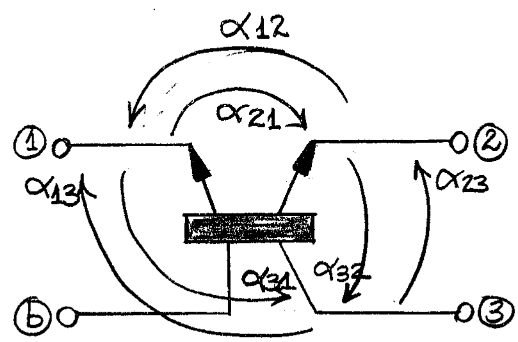


FIGURE 2

Similarly, one can alternately reverse-bias each terminal with the remaining terminals shorted to the base, to obtain relations of the form:

$$\begin{pmatrix} I_1 \\ I_2 \\ I_3 \end{pmatrix} = \begin{pmatrix} a_{11} & a_{12} & a_{13} \\ a_{21} & a_{22} & a_{23} \\ a_{31} & a_{32} & a_{33} \end{pmatrix} \begin{pmatrix} -1 \\ 0 \\ 0 \end{pmatrix} \tag{3}$$

In the reverse-bias condition, the minimum possible density becomes n_p (or p_n) and the current that flows is the reverse saturation current. Thus, one obtains

$$I_1 = a_{11}\,(-1) = -I_{10}$$

which leads to

$$\begin{aligned} a_{11} &= I_{10} \\ a_{22} &= I_{20} \\ a_{33} &= I_{30} \end{aligned} \tag{4}$$

Hence,

$$\begin{pmatrix} a_{11} & a_{12} & a_{13} \\ a_{21} & a_{22} & a_{23} \\ a_{31} & a_{32} & a_{33} \end{pmatrix} = \begin{pmatrix} I_{10} & -\alpha_{12} I_{20} & -\alpha_{13} I_{30} \\ -\alpha_{21} I_{10} & I_{20} & -\alpha_{23} I_{30} \\ -\alpha_{31} I_{10} & -\alpha_{32} I_{20} & I_{30} \end{pmatrix} \tag{5}$$

wherein all the parameters are defined in terms of externally measurable α's and saturation currents.

3. D. C. Circuit Model

The basic lumped diffusion model developed by Linvill[3] for an ordinary transistor can be extended to the dual emitter device as in Figure 3. The model is generally applicable to low-level operation only, defined by the following conditions:[3]

1. Bulk sensitivities are negligible
2. Injected current densities are low
3. Space-charge layer widening is negligible
4. Each junction separately can be described by an equation

$$I = I_s \ (\exp qv/kT - 1)$$

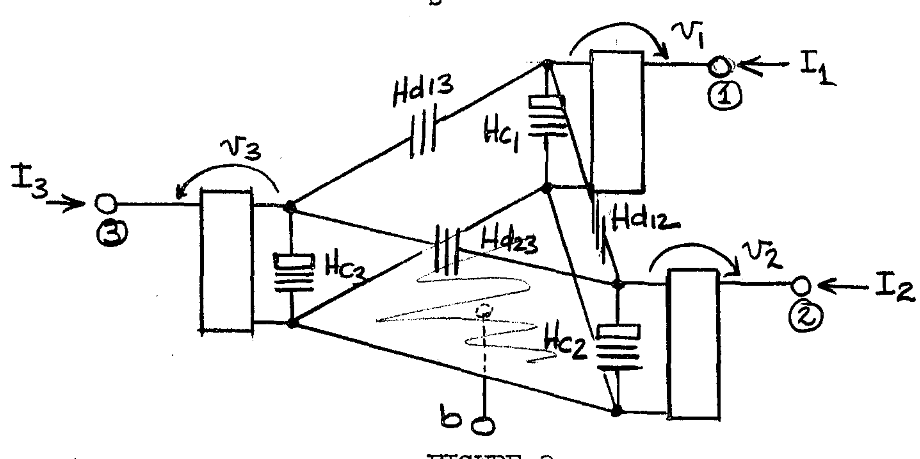


FIGURE 3

This model is useful only for d.c. or low-frequency calculations, since storage effects in the base region and depletion capacitances are not included. The form of the parameters is given by:

$$H_c = \text{(combinance)} \equiv \frac{Aqw}{2\tau_p}$$

$$H_d = \text{(diffusance)} \equiv \frac{AqDp}{w}$$

where A = cross-sectional area of the lump

w = width of the lump

Dp = diffusion constant for holes

τp = minority carrier lifetime

For the particular model at hand, the diffusances and combinances will be related to the externally-measurable parameters derived previously.

The currents are linearly related to the excess densities, N_{E1}, N_{E2}, and N_{E3} as follows:

$$\begin{aligned}
I_1 &= N_{E1}(H_{c1}+H_{d12}+H_{d13}) & -N_{E2}(H_{d12}) -N_{E3}(H_{d13}) \\
I_2 &= -N_{E1}(H_{d12}) & +N_{E2}(H_{c2}+H_{d12}+H_{d23}) -N_{E3}(H_{d23}) \\
I_3 &= -N_{E1}(H_{d13}) & -N_{E2}(H_{d23}) +N_{E3}(H_{c3}+H_{d13}+H_{d23})
\end{aligned} \quad (6)$$

or alternatively,

$$\begin{pmatrix} I_1 \\ I_2 \\ I_3 \end{pmatrix} = \begin{pmatrix} (H_{c1} + H_{d12} + H_{d13}) & - H_{d12} & - H_{d13} \\ -H_{d12} & (H_{c2} + H_{d12} + H_{d23}) & - H_{d23} \\ -H_{d13} & - H_{d23} & (H_{c3} + H_{d13} + H_{d23}) \end{pmatrix} \begin{pmatrix} \exp qv_1/kT - 1 \\ \exp qv_2/kT - 1 \\ \exp qv_3/kT - 1 \end{pmatrix} \qquad (7)$$

Comparing this expression with equation (5), by symmetry of the model we have

$$\begin{aligned} \alpha_{21}I_{10} &= \alpha_{12}I_{20} \\ \alpha_{31}I_{10} &= \alpha_{13}I_{30} \\ \alpha_{32}I_{20} &= \alpha_{23}I_{30} \end{aligned} \qquad (8)$$

These relations will prove useful in calculations to follow.

4. Offset Voltage and On-Resistance

The commonly-specified quantities for a saturated transistor switch are: the offset voltage Δv_{12} appearing across the switch in the on condition, but with zero signal current; and the small-signal resistance R_{on} as a function of drive current, I_3. These quantities will be derived for the model chosen and compared with expressions for conventional single and dual switches.

4.1 Offset Voltage Δv_{12}

Setting $I_1 = I_2 = 0$, the generalized equations become

$$\begin{pmatrix} 0 \\ 0 \\ I_3 \end{pmatrix} = \begin{pmatrix} I_{10} & -\alpha_{12}I_{20} & -\alpha_{13}I_{30} \\ -\alpha_{21}I_{10} & I_{20} & -\alpha_{23}I_{30} \\ -\alpha_{31}I_{10} & -\alpha_{32}I_{20} & I_{30} \end{pmatrix} \begin{pmatrix} \exp qv_1/kT - 1 \\ \exp qv_2/kT - 1 \\ \exp qv_3/kT - 1 \end{pmatrix} \tag{9}$$

Hence: $$\exp qv_1/kT - 1 = \frac{\begin{vmatrix} 0 & a_{12} & a_{13} \\ 0 & a_{22} & a_{23} \\ I & a_{32} & a_{33} \end{vmatrix}}{\Delta}$$

$$= I_3 \frac{(a_{12}a_{23} - a_{13}a_{22})}{\Delta} \tag{10}$$

and $$\exp qv_2/kT - 1 = I_3 \frac{(a_{13}a_{21} - a_{11}a_{23})}{\Delta} \tag{11}$$

The offset voltage is then

$$\Delta v_{12} = (v_1 - v_2) = kT/q \ \ln\left[\frac{a_{12}a_{23} - a_{13}a_{22}}{a_{13}a_{21} - a_{11}a_{23}}\right] \tag{12}$$

Substituting from (2) and (8),

$$\Delta v_{12} = kT/q \ \ln\left[\frac{I_{20}}{I_{30}} \cdot \frac{\alpha_{13}}{\alpha_{23}} \frac{+ \alpha_{12}\alpha_{23}}{\alpha_{21}\alpha_{13}}\right]$$

$$= kT/q \ \ln\left[\frac{\alpha_{32}}{\alpha_{31}} \frac{\alpha_{23}}{\alpha_{13}} \frac{\alpha_{13} + \alpha_{12}\alpha_{23}}{\alpha_{23} + \alpha_{21}\alpha_{13}}\right] \tag{13}$$

4.2 On-Resistance R_{on}

R_{on} is obtained from the current-excess density relations as follows:

$$\frac{N_{E1}}{N_{E2}} = \exp\left(qv_1/kT - qv_2/kT\right);$$

$$kT/q \ln\left(\frac{N_{E1}}{N_{E2}}\right) = v_1 - v_2 \tag{14}$$

$$= \Delta v_{12} + I\, R_{on}$$

where Δv_{12} is the zero-current offset voltage

I is the signal current.

The incremental resistance is then

$$Ron = \frac{d(v_1 - v_2)}{d(I_1 = -I_2)} = \frac{kT}{q} \frac{N_{E2}}{N_{E1}} \left[\frac{N_{E2}\frac{dN_{E1}}{dI_1} - N_{E1}\frac{dN_{E2}}{dI_1}}{N_{E1}^{2}}\right]$$

$$= \frac{kT}{q}\left[\frac{1}{N_{E1}}\frac{dN_{E1}}{dI_1} - \frac{1}{N_{E2}}\frac{dN_{E2}}{dI_1}\right]$$

where $I_1 = -I_2$ is the signal current.

In general, for $I_1 = -I_2 \neq 0$,

$$\begin{pmatrix} I_1 \\ -I_1 \\ I_3 \end{pmatrix} = \begin{pmatrix} a_{11} & & \\ & \ddots & \\ & & a_{33} \end{pmatrix} \begin{pmatrix} N_{E1} \\ N_{E2} \\ N_{E3} \end{pmatrix}$$

from which

$$N_{E1} = I_1 \frac{(a_{22}a_{33}-a_{32}a_{23}+a_{12}a_{33}-a_{13}a_{32})}{\Delta} + I_3 \frac{(a_{12}a_{23}-a_{13}a_{22})}{\Delta}$$

$$= I_1 \cdot \frac{A}{\Delta} + I_3 \cdot \frac{B}{\Delta}$$

$$N_{E2} = I_1 \frac{(-a_{11}a_{33}-a_{21}a_{33} + a_{31}a_{23} + a_{13}a_{31})}{\Delta} + I_3 \frac{(-a_{11}a_{23}+a_{13}a_{21})}{\Delta}$$

$$= I_1 \cdot \frac{C}{\Delta} + I_3 \cdot \frac{D}{\Delta}$$

Therefore

$$\frac{dN_{E1}}{dI_1} = A \ ; \quad \frac{dN_{E2}}{dI_1} = C, \text{ and}$$

$$R_{on} = \frac{kT}{q} \left[\frac{A}{AI_1 + BI_3} - \frac{C}{CI_1 + DI_3} \right] \qquad (16)$$

For zero signal current, $I_1 = -I_2 = 0 = I_s$.

Substituting the α's from (2) and (8),

$$R_{on}\left(I_s = 0\right) = kT/qI_3 \left[\frac{1-\alpha_{12}-\alpha_{32}\alpha_{23}-\alpha_{13}\alpha_{32}}{\alpha_{13}+\alpha_{12}\alpha_{23}} + \frac{1-\alpha_{21}-\alpha_{31}\alpha_{23}-\alpha_{13}\alpha_{31}}{\alpha_{23}+\alpha_{13}\alpha_{21}}\right] \quad (17)$$

This value of R_{on} is the zero or small-signal input resistance as a function of drive current I_3.

For the case where $I_1 = I_2 \neq 0$, we observe that $R_{on} \rightarrow \infty$ where

$$AI_1 = -BI_3 \quad (18)$$

or

$$CI_1 = -DI_3 \quad (19)$$

That is, the switch turns off when either emitter-base junction becomes back-biased, and the maximum current that can be switched is limited by the base-collector drive current, I_3. The equivalent condition is that N_{E1} or $N_{E2} \rightarrow n_p$.

From (18) we find that

$$I_1\left(a_{22}a_{33}-a_{32}a_{23}+a_{12}a_{33}-a_{13}a_{32}\right) = -I_3\left(a_{12}a_{23}-a_{13}a_{22}\right)$$

or

$$\frac{I_1}{I_3} = \frac{a_{13}a_{22} - a_{12}a_{23}}{a_{22}a_{33} - a_{32}a_{21} + a_{12}a_{33} - a_{12}a_{32}} \quad (20)$$

Equation (20) can be evaluated knowing the d.c. α's.

4.3 Special Case of α_{12}, $\alpha_{21} \ll 1$.

In the Texas Instrument devices which were studied, the emitter-emitter short-circuit current gains, α_{12} and α_{21}, proved to be exceedingly small. This is to be expected from the geometry of the device, in which the emitters are diffused on to a thin base region and separated by a relatively large distance (see Figure 4). The resistance to direct current flow across this path is apparently very high, undoubtedly due to both to the small cross-section and the high density of traps near the surface between the two emitters.

As a result of α_{12} and $\alpha_{21} \ll 1$, the foregoing expressions for Δv_{12} and R_{on} can be simplified as follows:

$$\Delta v_{12} \cong kT/q \quad \ln \alpha_{32}/\alpha_{31} \tag{21}$$

$$R_{on}\left(I_s = 0\right) \cong kT/qI_3 \quad \frac{1-\alpha_{32}\alpha_{23}-\alpha_{13}\alpha_{32}}{\alpha_{13}} + \frac{1-\alpha_{31}\alpha_{23}-\alpha_{13}\alpha_{31}}{\alpha_{23}} \tag{22}$$

Also, for finite $I_1 = -I_2$, the ratio of maximum signal to drive current becomes

$$\left.\frac{I_1}{I_3}\right|_{R\to\infty} = \frac{-\alpha_{13}}{1-\alpha_{32}\alpha_{23}-\alpha_{13}\alpha_{32}}, \quad I_1 \text{ negative} \tag{23}$$

$$\left.\frac{I_1}{I_3}\right|_{R\to\infty} = \frac{+\alpha_{23}}{1-\alpha_{31}\alpha_{23}-\alpha_{13}\alpha_{31}}, \quad I_1 \text{ positive} \tag{24}$$

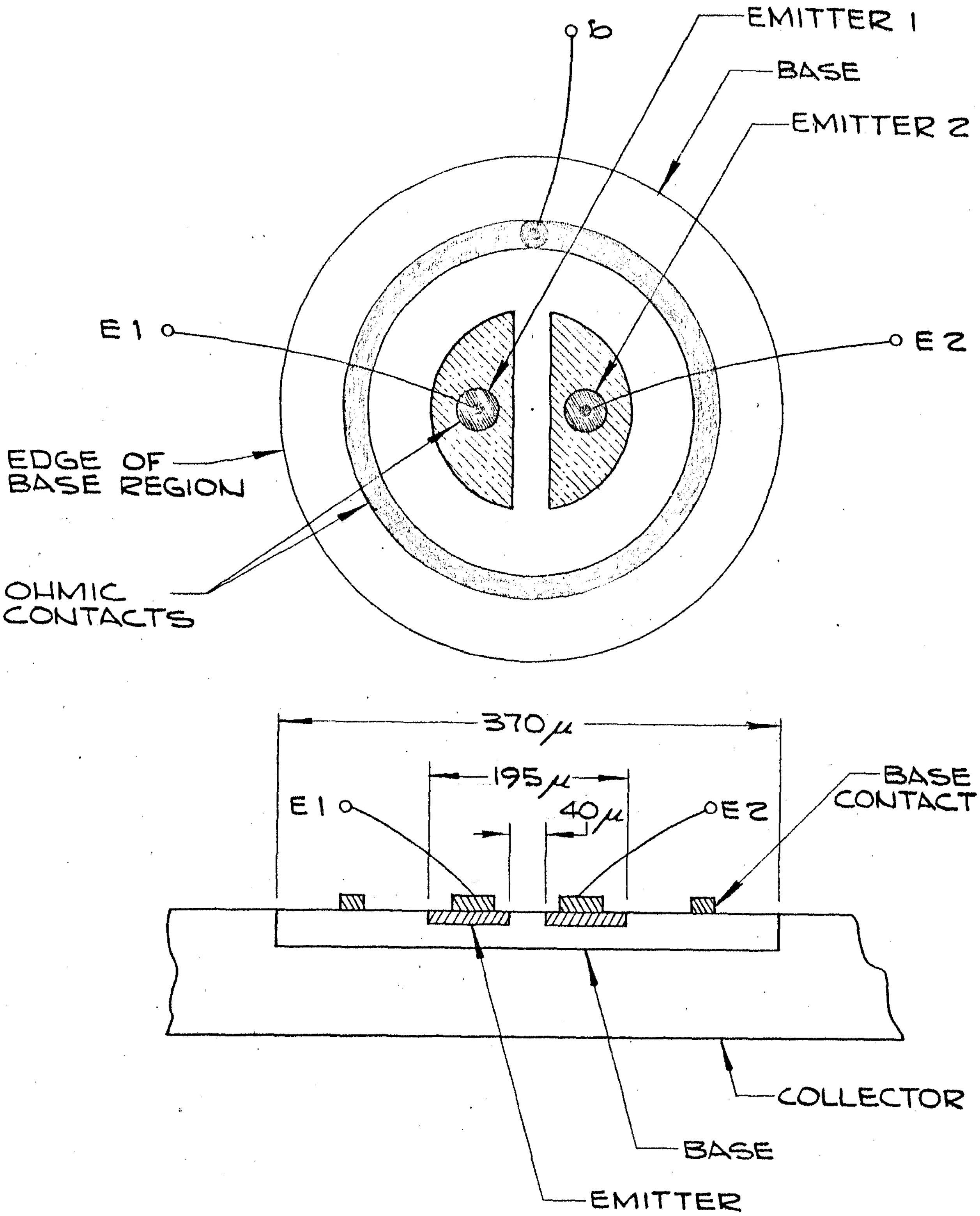

EMITTER 1
b
BASE
EMITTER 2
E1
E2
EDGE OF BASE REGION
OHMIC CONTACTS
370μ
195μ
40μ
E1
E2
BASE CONTACT
COLLECTOR
BASE
EMITTER

FIGURE 4.

4.4 Comparison with Conventional Device Theory

For a single transistor inverted switch, Bright[4] has derived the expression

$$\Delta v_0 = kT/q \cdot \ln (1/\alpha_I) \tag{25}$$

where α_I is the inverted α, collector used as the emitter. For the dual-emitter device, for $\alpha_{12}, \alpha_{21} \ll 1$, the expression

$$\Delta v_{12} = kT/q \cdot \ln (\alpha_{32}/\alpha_{31}) \tag{26}$$

is just what one would expect for two transistors connected back-to back--which is essentially the case where there is very low coupling between emitters.

The low-level theory predicts an on-resistance of

$$R_{on} \cong \frac{\Delta v_0}{I_3} = kT/qI_3 \cdot \ln 1/\alpha_I \cong kT/qI_3 \cdot (1-\alpha_I) \cong kT/qI_3 \cdot 1/\beta_I \tag{27}$$

where Δv_0 is the offset and α_I the inverted α. The corresponding expression for the dual is

$$R_{on} \cong kT/qI_3 \left[\frac{1-\alpha_{32}(\alpha_{23}+\alpha_{13})}{\alpha_{13}} + \frac{1-\alpha_{31}(\alpha_{23}+\alpha_{13})}{\alpha_{23}}\right] \tag{28}$$

For two symmetrical transistors connected as an inverted dual chopper, (27) gives a resistance of

$$R_{on} \cong kT/qI_3 \cdot 2(1-\alpha_I) \tag{29}$$

For comparison, assume in (28) that $\alpha_{31} = \alpha_{32} \cong 1$, a very good approximation. Hence, for the dual,

$$R_{on} \cong kT/qI_3 \left[\frac{1-(\alpha_{13}+\alpha_{23})}{\alpha_{13}} + \frac{1-(\alpha_{13}+\alpha_{23})}{\alpha_{23}}\right] \tag{30}$$

If the two halves are symmetrical, $\alpha_{23} = \alpha_{13}$, and

$$R_{on} \cong kT/qI_3 \, 2\left[\frac{1-2\alpha_{13}}{\alpha_{13}}\right]$$

Since α_{13} has a maximum value of 0.5, assuming symmetry, then we can further approximate

$$R_{on} \cong kT/qI_3 \cdot 4\,(1-2\alpha_{13}) \tag{31}$$

Comparing (31) with (29), if α_{13} (max) $\rightarrow$ 0.5, it appears that the dual unit has approximately twice the on-resistance of two single units, each driven with the same collector current, I_3. However, it seems more proper to assume that the same total drive is available in both cases, in which case each single unit would have a resistance of

$$R_{on} \cong \frac{kT}{qI_3/2} \cdot (1-\alpha_I)$$

or for both units,

$$R_{on} \cong kT/qI_3 \cdot 4\,(1-\alpha_I) \tag{32}$$

Thus, the resistances are nominally the same for a given total drive current.

5. Measurements

The above quantities, Δv_{12}, R_{on}, and I_1/I_3 for $R \rightarrow \infty$, can be calculated from just the d.c. α's of the device. According to our previous definitions, the α's are obtained by a series of short-circuit measurements, each of which assures an excess density of zero at all but one of the terminals.

Over the range of drive currents considered, 0.1 mA to 50 mA, the α's vary considerably; therefore the calculations were made point-by-point to include these effects. This was done mainly because Δv_{12} and R_{on} depend on very small differences between the α's, and no reasonable results can be obtained otherwise. Even when the α-variation is included, the low-level theory will prove inadequate because, as mentioned earlier, low-level operating conditions are being violated in a number of ways.

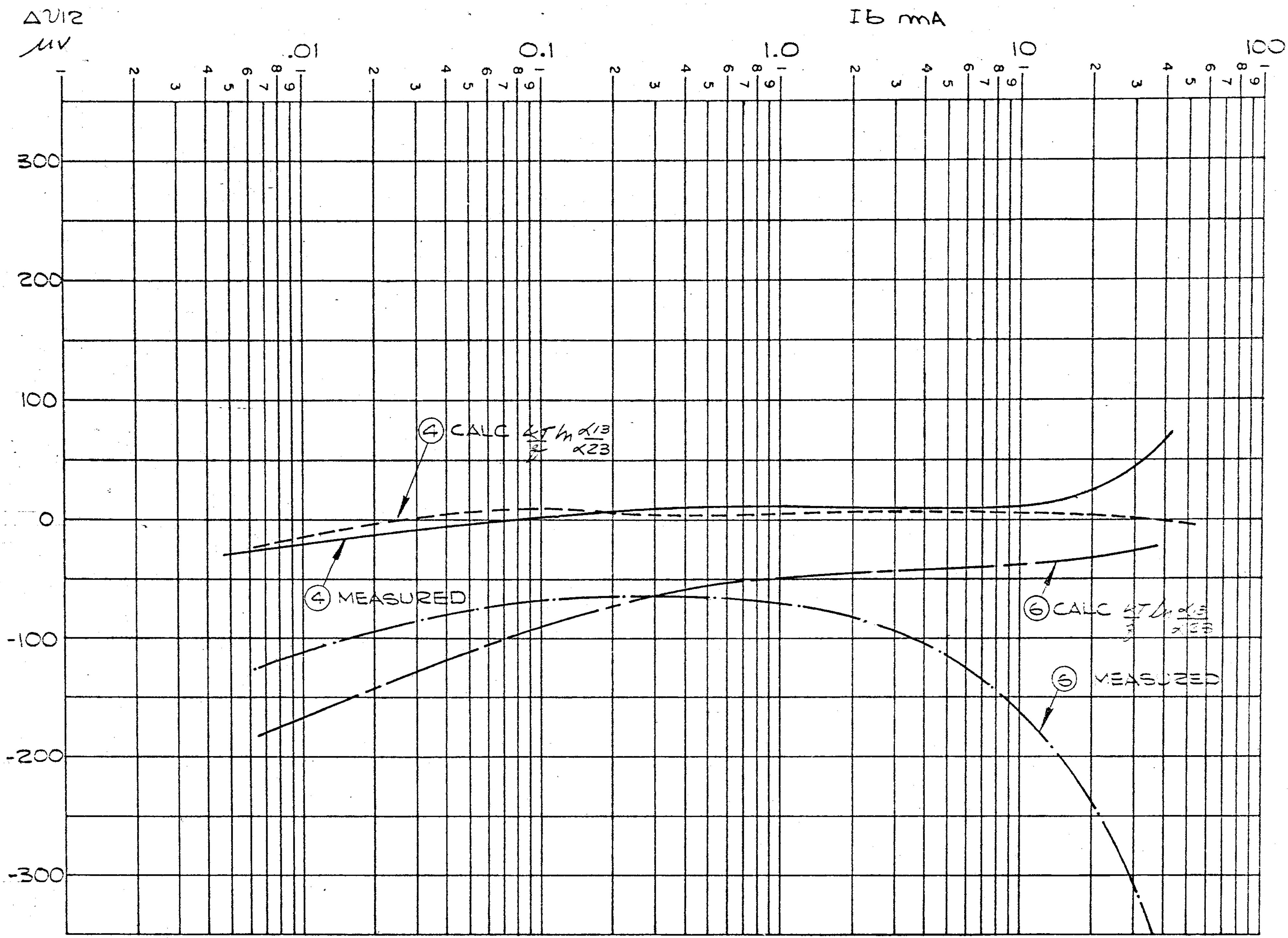
Table I shows the measured α's for 4 different units, 3- 3N76's ($\pm$18 volts Vee max, $\pm$200 μv zero-offset) and 1- 3N79 ($\pm$12 volts Vee max, $\pm$200 μv zero offset). These data were used to calculate Δv_{12} and R_{on}. Measured values of Δv_{12} for the 4 units, and of R_{on} for the 3 units which survived the first measurement (Table II), are compared with the calculated values in Figures 5 through 8.

One additional set of measurements was made on unit No. 6 to determine the limiting signal current as a function of base drive (Figure 9). The slopes of these curves illustrate the variation of R_{on} throughout the range of I_3.

6. Experimental Results

6.1 Offset Voltage Δv_{12}.

The calculated offset voltages in general agree with the measured



ΔV12
μV
Ib mA
.01
0.1
1.0
10
100
300
200
100
0
-100
-200
-300
④ CALC $\frac{kT}{q} \ln \frac{\alpha_{13}}{\alpha_{23}}$
④ MEASURED
⑥ CALC $\frac{kT}{q} \ln \frac{\alpha_{13}}{\alpha_{23}}$
⑥ MEASURED

FIGURE 5

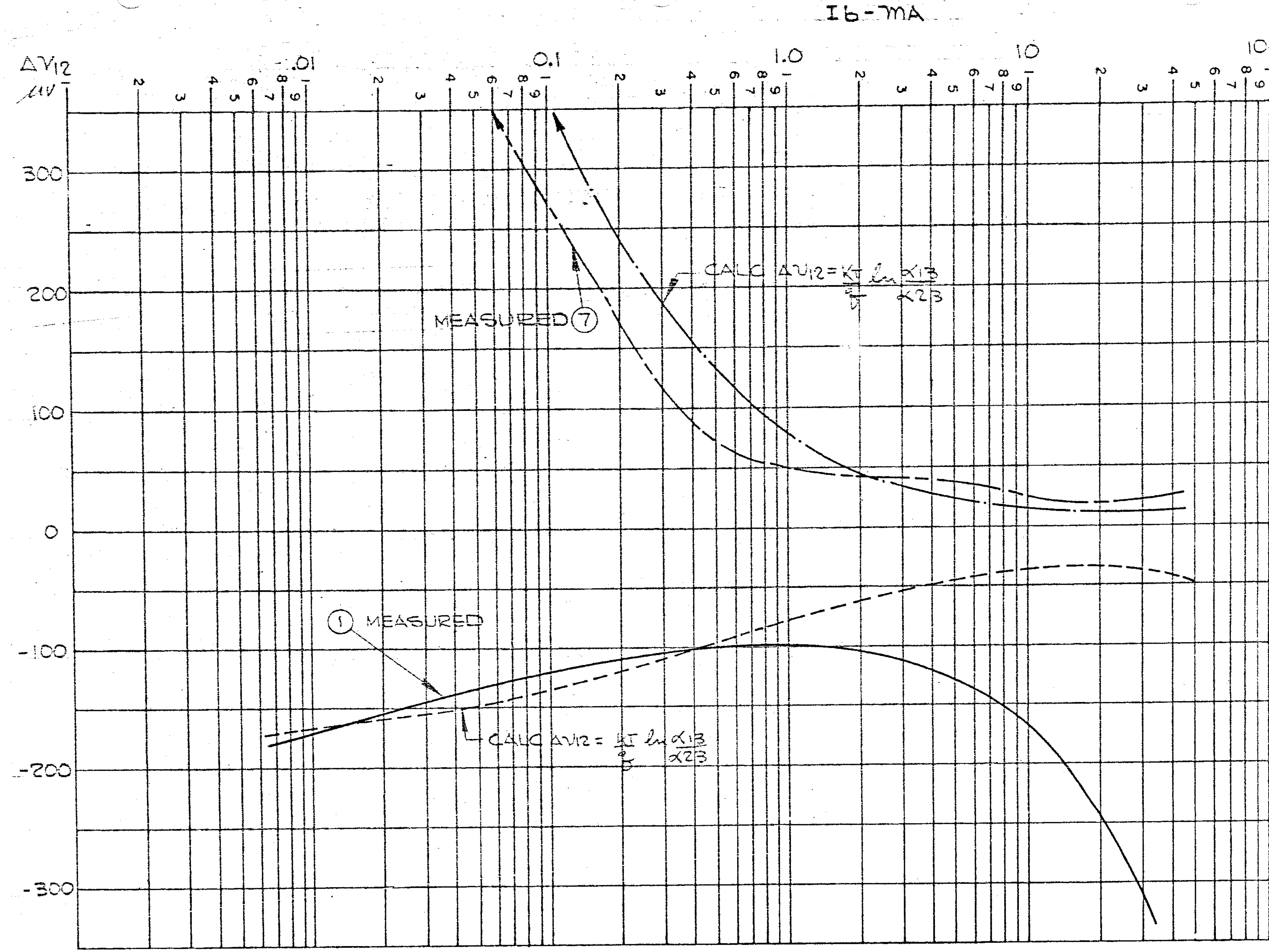
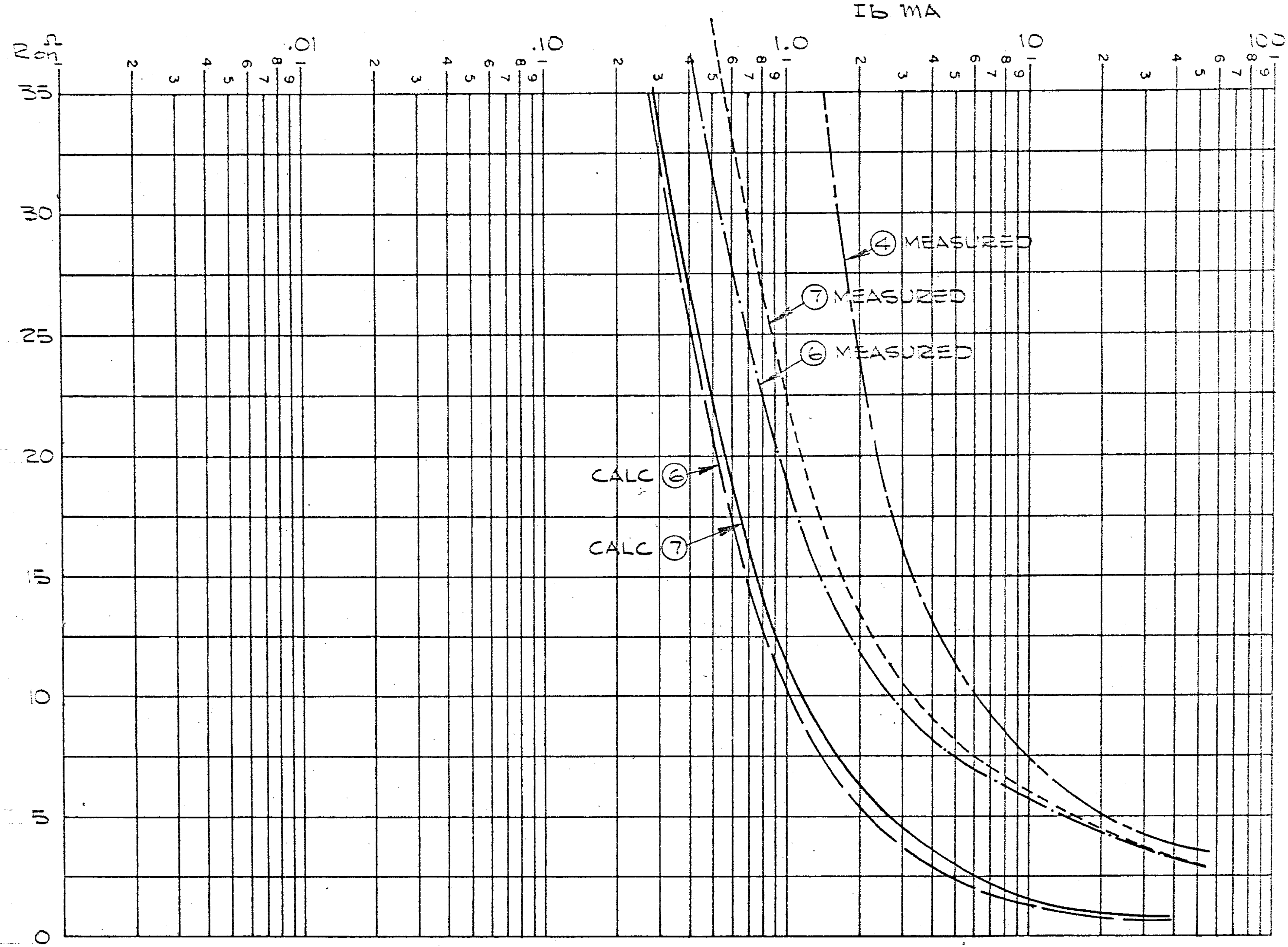
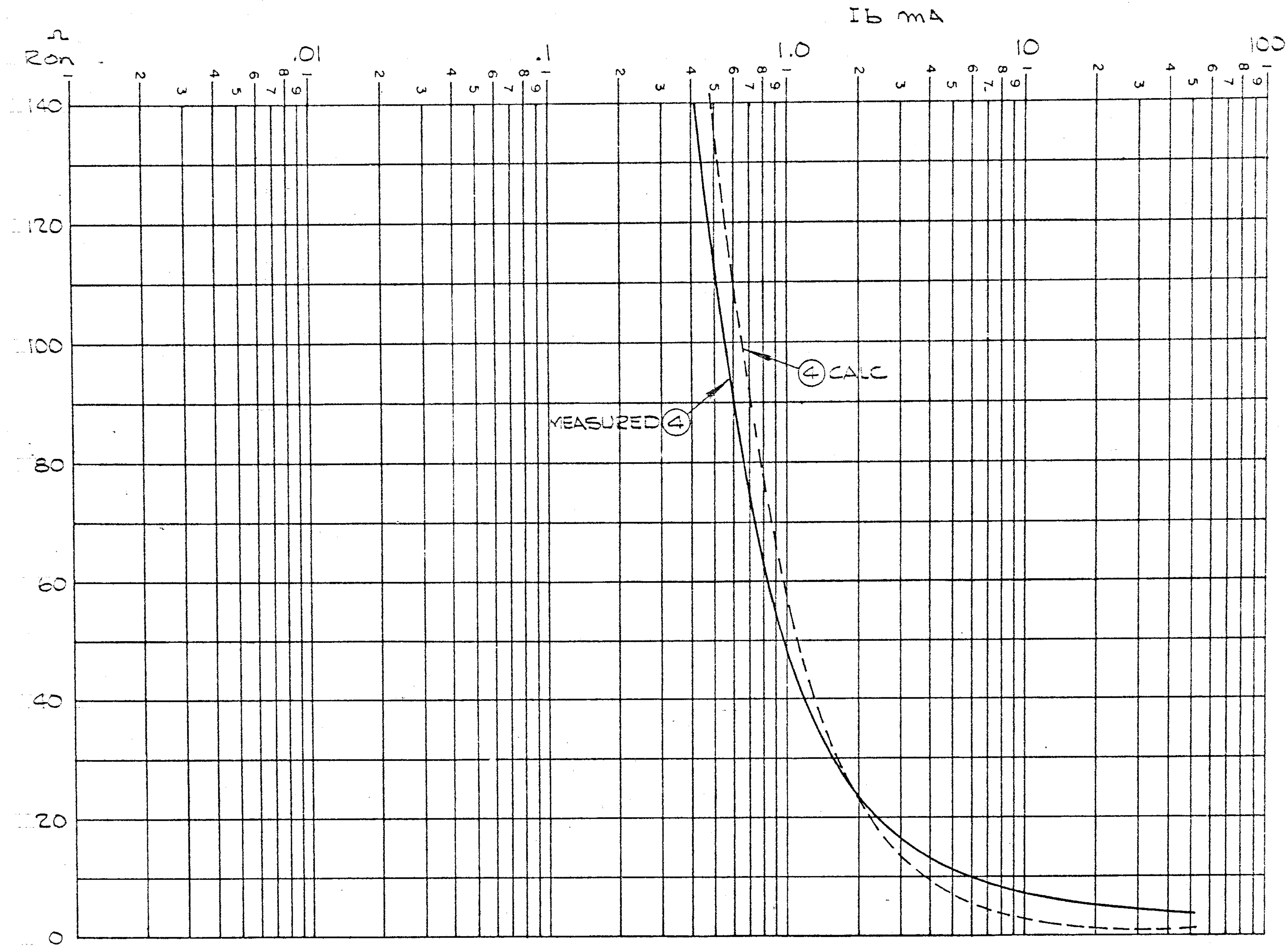

Ib-mA
ΔV12
μV
CALC ΔV12 = kT/q ln α1β/α2β
MEASURED ⑦
① MEASURED
CALC ΔV12 = kT/q ln α1β/α2β

FIGURE 6


Ib MA
R on Ω
4 MEASURED
7 MEASURED
6 MEASURED
CALC 6
CALC 7


Ib mA
.01
.1
1.0
10
100
Ω
Ron
140
120
100
80
60
40
20
0
④ CALC
MEASURED ④

FIGURE

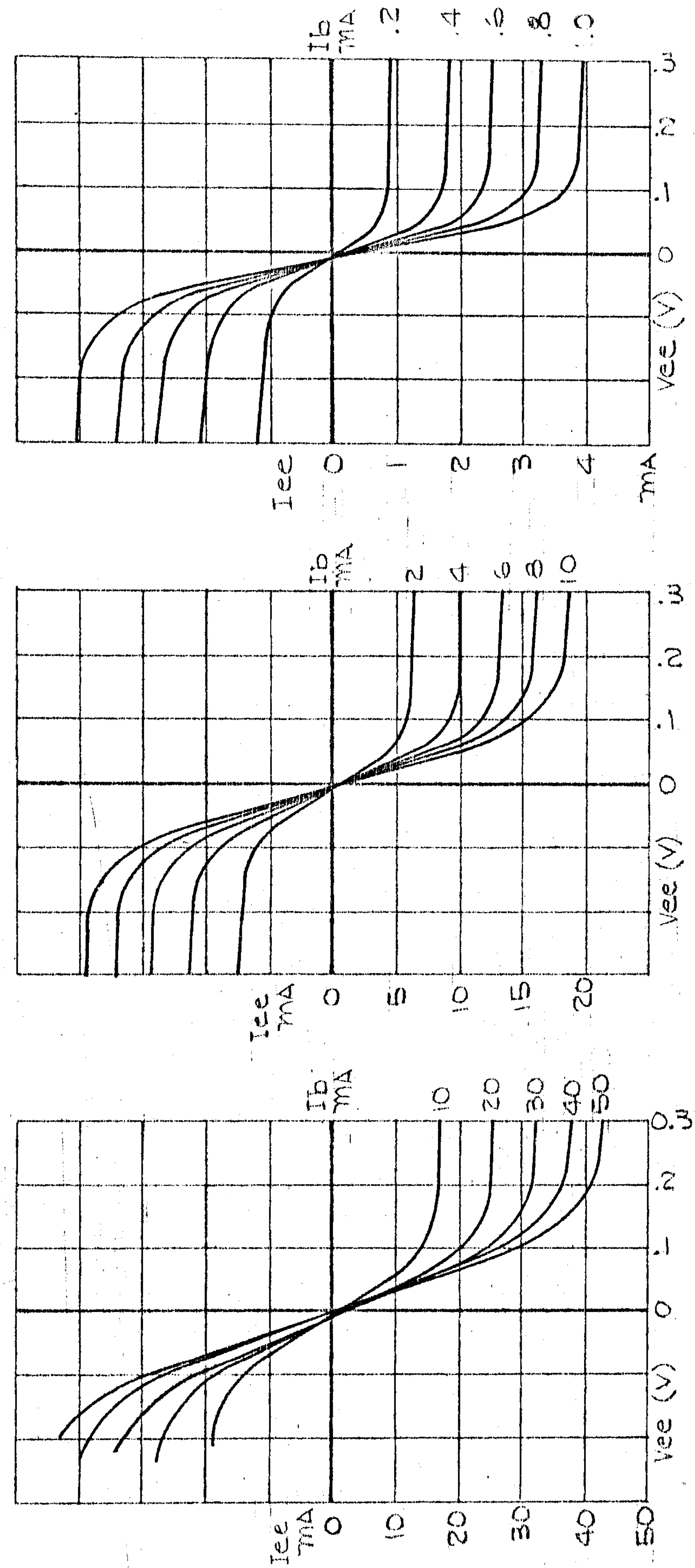


FIGURE 9

values for drive currents between 0.1 mA and 1–2 mA. At higher currents, marked deviations occur. Below 0.1 mA, the calculated and measured values show the same trend but do not agree to better than ± 100 μv.

6.2 On-Resistance R_{on}

Transistors No. 6 and No. 7 give similar results in that their measured resistances agree closely, and their calculated values agree closely. Unit No. 4 exhibits an unusually large resistance, and correspondingly, a phenomenally low offset voltage. (The low offset seems related to unusually high resistance in the base region.)

Units No. 6 and 7 have R_{on}'s which are considerably higher than predicted by the low-level theory. The differences (measured value-calculated value) are as follows:

No. 6		No. 7	
0.5 mA	$\Delta = 10.7\Omega$	0.5 mA	$\Delta = 16.3\Omega$
5 mA	$\Delta = 5.0$	5	$\Delta = 5.3$
1.0	$\Delta = 8.6$	1.0	$\Delta = 11.1$
10	$\Delta = 4.4$	10	$\Delta = 4.4$
3.0	$\Delta = 5.7$	3.0	$\Delta = 6.0$
30	$\Delta = 2.9$	30	$\Delta = 2.9$

For unit No. 6, the difference decreases roughly by a factor of 2 for each decade of drive current. For unit No. 7, the factor is closer to 3.

6.3 Maximum Signal Current $I_{1\,max}$

For operation as a switch, the maximum current which can be switched for a given drive current is often important. For a chopper, the maximum current gives an indication of a "margin of safety;" i.e., the drive should always exceed the maximum signal current by a factor of 2 or 3 to assure a low on-resistance.

Some calculated and measured values for unit No. 6 are compared below, where it is assumed that α_{21}, $\alpha_{12} \ll 1$.

$$\left[\frac{I_1\ \mathrm{max}}{I_3}\right]^- \cong \frac{-\alpha_{13}}{1 - \alpha_{32}\alpha_{23} - \alpha_{13}\alpha_{32}} \tag{33}$$

$$\left[\frac{I_1\ \mathrm{max}}{I_3}\right]^+ \cong \frac{\alpha_{23}}{1 - \alpha_{31}\alpha_{23} - \alpha_{31}\alpha_{13}}$$

Base Drive I_3	$\frac{I_{max}}{I_3}$ (calc)	$\frac{I_{max}}{I_3}$ (meas'd)
0.1 mA	(+) 4.67 (-) 5.43	(+) 4.5 (-) 6.0
1.0 mA	(+) 4.72 (-) 5.30	(+) 4.0 (-) 4.2
10 mA	(+) 3.70 (-) 3.96	(+) 1.7 (-) 1.9
30 mA	(+) 2.51 (-) 2.64	(+) 1.1 (-) 1.2

The measured values were deduced from oscilloscope photos (see Figure 9) and are not reliable to better than $\pm 5\%$. Here again, as in the previous cases, the results are in reasonable agreement only to about $\sim$1 mA of drive current. Also, note that the resistance R_{on}, as given by the slope of the curves in Figure 9, is a minimum at $I_1 = -I_2 = 0$; hence, for any finite signal current I_1, the dynamic resistance will be proportionately greater than the zero value.

6.4 Optimum α_{12}

The complete low-level expression for the on-resistance is

$$R_{on} = \frac{kT}{qI_3}\left[\frac{1 - \alpha_{12} - \alpha_{32}\alpha_{23} - \alpha_{13}\alpha_{32}}{\alpha_{13} + \alpha_{12}\alpha_{23}} + \frac{1 - \alpha_{21} - \alpha_{32}\alpha_{23} - \alpha_{13}\alpha_{31}}{\alpha_{23} + \alpha_{13}\alpha_{21}}\right]$$

First assume symmetry:

$$\alpha_{12} = \alpha_{21}$$

$$\alpha_{23} = \alpha_{13}$$

$$\alpha_{32} = \alpha_{31}$$

Then $$R_{on} = \frac{2kT}{qI_3}\left[\frac{1 - \alpha_{12} - 2\alpha_{23}\alpha_{13}}{\alpha_{13} + \alpha_{12}\alpha_{13}}\right] \qquad (34)$$

We observed experimentally that $\alpha_{12} \ll 1$, due to the geometry wherein the two emitters are separated by a broad, thin base region which forms the main current path. Suppose, however, that α_{12} could become large. For large α_{12}, we can say

$$\alpha_{12} + \alpha_{32} \approx 1$$

or

$$\alpha_{21} + \alpha_{31} \approx 1$$

then

$$\alpha_{32} \approx 1 - \alpha_{12}$$

and

$$R_{on} \cong \frac{2kT}{qI_3}\left[\frac{(1 - \alpha_{12}) - 2(1 - \alpha_{12})\alpha_{13}}{\alpha_{13} + \alpha_{13}\alpha_{12}}\right] \qquad (35)$$

Hence, if $\alpha_{12} \to 1$, $R_{on} \to 0$. At the same time; the forward α's, α_{32} and $\alpha_{31} \to 0$. This implies poor coupling from collector to the emitters through the base region, and at the same time reasonable coupling in the forward directions (α_{13} and α_{23} finite). However, one would expect α_{13} and α_{23} to deteriorate along with α_{31} and α_{32}; so in actual fact, increasing α_{12} and α_{21} would probably result in a very minor decrease in R_{on}.

7. Conclusions

The low-level theory has been shown to give a very rough description of the dual-emitter chopper transistor. It is not too surprising that the offset voltage calculation is not too precise, since we are dealing with minute differences between the two halves of the transistor, and small errors can have a large effect. One would expect to predict the on-resistance with better accuracy, however, at least for currents up to 1 or 2 mA.

For the best-behaved units, 6 and 7, the calculated values of R_{on} at 1 mA are a factor of 2 below the measured values. At 30 mA, the calculated values are $< 1^{\Omega}$, about 5 × less than the measured values of 3.6 and 3.7^{Ω}. Hence, the simple $1/I_3$ dependence of R_{on} predicted by low-level theory is completely unreliable, and at best can be used to give rough answers up to 1 mA. The theory is similarly limited in calculating the maximum signal current for a given drive current.

In comparing with conventional device low-level theory, it was found that the offset voltage is equivalent to that of a matched pair chopper where the two units have different forward α's; and further, that the offset is independent of drive current. This is true only for approximately zero emitter-emitter coupling, which for the transistors studied proved to be a valid assumption.

Similarly, the on-resistance was calculated to be nominally the same as the matched pair chopper if the total drive currents are equal. Although the calculated and measured values do not agree too well, a comparison of our measured values with Gibbons' Table I[2] shows this trend: For a single unit, R_{on} at 1 mA = 6Ω, or 12Ω for a matched pair. For our units 6 and 7, the R_{on} at 2mA were 11.8Ω and 13.3Ω respectively.

In terms of offset voltage and on-resistance, then, there appears to be little difference between the dual emitter device and a matched pair. However, the temperature matching of the former is inherently better, so that its offset remains more stable over wide temperature ranges. Also, experience has shown that the offset for the dual emitter device is better-behaved for high drive currents; this again is due to good thermal matching in the two halves at drive currents large enough to cause considerable internal heating.

One additional measurement was made in an attempt to fit the data to the theory of super-saturated switches developed by Gibbons.[2] Gibbons' theory primarily accounts for a nonlinear recombination law (rather than the linear form used in the present derivation), which for a single unit yielded excellent results up to currents of about 8 mA. Assuming that each half of the unit can separately be considered to have an on-resistance

$$R_{on} = \frac{(1 + Ac/Ae)v_0}{I_b} \tag{36}$$

where

Ac = collector area

Ae = emitter area

v_0 = inverted connection offset voltage

I_b = base drive

then the total resistance would be

$$R_{on} = \frac{(1 + Ac/Ae)(v_{01} + v_{02})}{I_b} \tag{37}$$

Without presenting the details, it was found that the measured values of $(v_{01} + v_{02})$ did not consistently correlate with the measure R_{on} ; and, as in the linear theory, the predicted resistances assuming $Ac/Ae \approx 2$ for each half, were much too low. Hence, for resistance calculations it appears that one cannot make the simplifying assumption as was possible for the offset calculations, that the dual can be treated as two independent units.

8. Acknowledgement

The author is indebted to Professor J. F. Gibbons, Department of Electrical Engineering, Stanford University, for his assistance and many helpful discussions during the course of this work.

9. List of References

1. "Large-Signal Behavior of Junction Transistors," J. J. Ebers and J. L. Moll, Proc. IRE, Dec 1954.
2. "Super-Saturated Transistor Switches," James F. Gibbons, IRE Trans. on Electron Devices, Nov 1965.
3. "Lumped Models for Transistors and Diodes," J. G. Linvill, Proc. IRE, June 1958.
4. "Transistors Used as Switches," R. L. Bright, Trans. AIEE (Commun. and Elect.) March 1955.

SLAC
TN 65-64
R. S. Larsen
3 August 1965

TABLE I

	Forward α's					Reverse α's				
xistor	I_1, I_2 mA	I_{b1} mA	I_{b2} mA	α_{31}	α_{32}	I_3 mA	I_1 μa/mA	I_2 μa/mA	α_{13}	a_{23}
3N76 No. 6	0.01	0.073	0.137	.9927	.9863	0.01	4.95	4.10	.495	.410
	0.100	0.47	0.81	.9953	.9919	0.10	50.	41.6	.500	.416
	0.500	1.92	3.0	.9962	.9940	0.50	245.	201.	.490	.422
	1.0	3.55	5.4	.9965	.9946	1.0	488.	425.	.488	.425
	5.00	14.0	21.3	.9972	.9957	5.0	2.35	2.13	.470	.426
	10.00	26.5	40.0	.9974	.9960	10.0	4.60	4.28	.460	.428
	30.00	99.0	130.	.9967	.9957	30.0	12.9	12.3	.430	.410
3N79 No. 4	0.01	0.210	0.215	.9790	.9785	0.01	1.56	1.60	.156	.160
	0.100	1.42	1.40	.9858	.9860	0.10	24.3	24.9	.243	.249
	0.500	5.35	5.32	.9893	.9893	0.50	151.	155.	.302	.310
	1.00	9.4	9.4	.9906	.9906	1.0	0.325	0.334	.325	.334
	5.00	38.2	37.2	.9924	.9925	5.	1.81	1.84	.362	.368
	10.00	75.	72.5	.9925	.9927	10.	3.7	3.78	.370	.378
	30.00	480.	480.	.984	.984	30.	10.8	10.9	.360	.363
3N76 No. 7	0.01	0.78	0.47	.922	.953	0.01	4.62	4.59	.462	.459
	0.10	4.1	2.72	.959	.9728	0.1	46.7	46.4	.467	.464
	0.50	9.7	7.15	.9806	.9857	0.5	228.	227.	.456	.454
	1.0	13.2	10.2	.9868	.9898	1.0	0.455	0.455	.455	.455
	5.	29.5	24.6	.9941	.9950	5.	2.20	2.19	.440	.438
	10.	45.5	40.	.9954	.9960	10.	4.34	4.35	.434	.435
	30.	153.	139.	.9949	.9954	30.	12.3	12.2	.410	.407
3N76 No. 1	0.01	0.114	0.178	.9886	.9822	0.01	4.5	4.55	.450	.455
	0.10	0.87	1.38	.9913	.9862	0.1	48.4	4.62	.484	.462
	0.50	3.5	5.3	.9930	.9894	0.5	224.	227.	.448	.454
	1.0	6.5	9.5	.9935	.9905	1.0	0.445	0.452	.445	.452
	5.	28.5	37.	.9943	.9926	5.	2.04	2.08	.408	.416
	10.	60.	74.	.9940	.9926	10.	3.7	3.8	.370	.380
	30.	255.	298.	.9915	.9901	30.	8.5	8.76	.283	.292

SLAC
TN 65-64
R. S. Larsen
3 August 1965

TABLE II

OFFSET VOLTAGE CALCULATIONS $\Delta v_{12} = KT/q \ln \frac{\alpha_{32}}{\alpha_{31}}$

I_3	3N76/6	3N79/4	3N76/7	3N76/1
0.01 mA	-166 μv	- 13. μv	+806. μv	-166. μv
0.10	- 88.5	+ 5.2	385.	-135
0.50	- 57.2	+ 1.56	133.	- 93.7
1.00	- 48.1	0	78.	- 78.
5.0	- 38.	+ 5.2	25.4	- 44.2
10.0	- 35.1	+ 6.5	14.3	- 36.4
30.0	- 26.8	0	12.2	- 37.2

OFFSET VOLTAGE MEASUREMENTS $I_1 = I_2 = 0$

I_3				
0.01 mA	-110. μv	- 20. μv	+560. μv	-170. μv
0.10	- 64	0	+270	-120
0.50	- 67	+ 10	70	-100
1.00	- 70	+ 5	50	-100
5.0	- 70	+ 10	40	-130
10.0	-160	+ 10	20	-170
30.0	-300	+ 47	22	-310

ON-RESISTANCE MEASUREMENTS R_{on}

I_3	3N76/6	3N79/4	3N76/7
0.1 mA	122 Ω	950 Ω	145 Ω
0.5	32	115	38
1.0	19	49	22.5
10.	5.8	7.4	6.0
50.	3.0	3.6	3.0