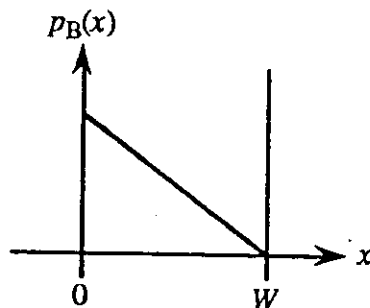


CHAPTER 11

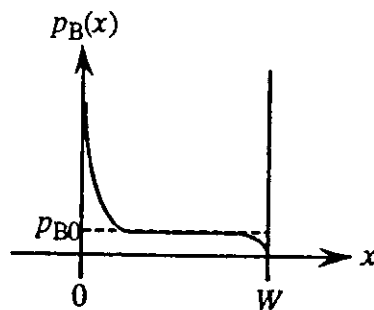
11.1

(a) The E, B, and C dopings are assumed to be nondegenerate and constant throughout a given region. The quasineutral widths of the emitter and collector are assumed to be much greater than the minority carrier diffusion lengths in the respective regions.

(b)

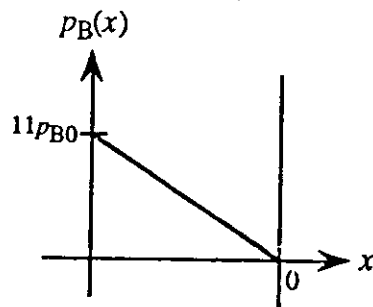
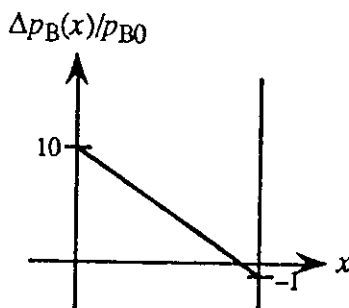
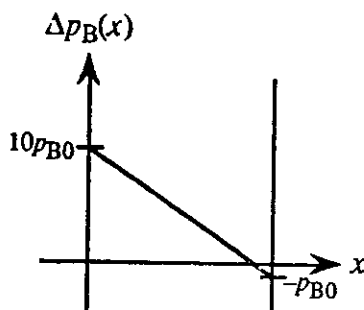


$W/L_B \ll 1$



$W/L_B \gg 1$

(c)



(d) γ , α_T , and α_{dc} are all typically slightly less than unity. $\beta_{dc} \sim 100$ to 1000 .

(e) The desired equation is simply obtained by subtracting Eq. (11.47b) from Eq. (11.47a).

$$I_B = I_E - I_C = (1 - \alpha_F) I_{F0} (e^{qV_{EB}/kT} - 1) + (1 - \alpha_R) I_{R0} (e^{qV_{CB}/kT} - 1)$$

(f) As noted in Eq. (11.46)

$$\alpha_F I_{F0} = \alpha_R I_{R0}$$

Thus

$$I_{R0} = \alpha_F I_{F0} / \alpha_R = (0.9944)(4.749 \times 10^{-15}) / (0.4286) = 1.102 \times 10^{-14} \text{ A}$$

(g) The quasi-linear slope at low V_{EC} is caused by base width modulation. The sharp upward curvature when $V_{EC} \rightarrow V_{CE0}$ is caused by carrier multiplication and feedback.

(h) Yes, it is possible. Even though V_{CB0} may be punch-through limited, the voltage where $M \rightarrow 1/\alpha_{dc}$ may be less than the V_{CB0} due to punch-through.

(i) "Current crowding" is caused by the voltage drop associated with the base current flowing laterally across the face of the emitter.

(j) ...The drift-enhanced transport of carriers across the base decreases the transit time, thereby reduces recombination in the base, which ultimately increases α_T and the current gains.

...The built-in \mathcal{E} -field and attendant decrease in the carrier transit times across the base lead to an improved high-frequency response.

(k) A Gummel plot is a simultaneous semilog plot of I_B and I_C as a function of the input voltage V_{EB} .

(l) The fall-off in β_{dc} at low I_C is due to an increasingly important I_{R-G} component in I_B as $I_C(V_{EB})$ is reduced. The fall-off in β_{dc} at high I_C levels is caused by high-level injection in the collector, current crowding, and/or series resistance.

(m) [No]...A totally poly-Si emitter would be essentially the same (function the same) as a totally crystalline-Si emitter.—The minority carrier distribution in a totally poly-Si emitter would be similar to Fig. 11.18(a). (Actually, a totally poly-Si emitter would undoubtedly lead to a poor E-B interface and a poorly operating BJT.)

(n) In general, "heterojunction" refers to a junction between two dissimilar materials. With reference to the HBT, however, a "heterojunction" is understood to be a junction between two dissimilar semiconductors.

(o) In a standard BJT the device is fabricated in a single semiconductor and $N_E \gg N_B$. In HBTs the emitter is a wider band gap semiconductor and the base is the heaviest doped of the transistor regions ($N_E \ll N_B \gg N_C$).

11.2

$$(a) \quad \gamma = \frac{I_{Ep}}{I_{Ep} + I_{En}} = \frac{I_I}{1.001I_I} = 0.9990$$

$$(b) \quad \alpha_T = \frac{I_{Cp}}{I_{Ep}} = \frac{0.999I_I}{I_I} = 0.9990$$

$$(c) \quad \alpha_{dc} = \gamma\alpha_T = 0.9980$$

$$\beta_{dc} = \frac{\alpha_{dc}}{1 - \alpha_{dc}} = 499$$

(d) There are two possible solution approaches:
Approach #1...

$$I_E = I_{Ep} + I_{En} = 1.001I_I$$

$$I_C = I_{Cp} + I_{Cn} = 0.999I_I + 10^{-6}I_I$$

$$I_B = I_E - I_C = (1.999 \times 10^{-3})I_I$$

Approach #2...

$$I_{B1} = I_{En} = 0.001I_I$$

$$I_{B2} = I_{Ep} - I_{Cn} = 0.001I_I$$

$$I_{B3} = -I_{Cn} = -(10^{-6})I_I$$

$$I_B = I_{B1} + I_{B2} + I_{B3} = (1.999 \times 10^{-3})I_I$$

Note: I_{Cn} was incorrectly shown on Fig. P11.2 as a negative current in the first printing.

(e) **Yes**. All currents on the figure are constant across the depletion regions. This implies (see the Depletion Region Considerations in Subsection 6.1.2) that recombination-generation is negligible in these regions.

11.3

Provided the direction of positive current flow for I_E , I_C , and I_B are as defined in Fig. 10.2(b), the equations in Subsection 11.1.1 and 11.1.2 can be appropriately modified for an *npn* BJT by simply making the following substitutions:

$$nE \rightarrow pE$$

$$pB \rightarrow nB$$

$$nC \rightarrow pC$$

$$I_{Ep} \rightarrow I_{En}; I_{En} \rightarrow I_{Ep}$$

$$I_{Cp} \rightarrow I_{Cn}; I_{Cn} \rightarrow I_{Cp}$$

$$V_{EB} \rightarrow V_{BE}; V_{CB} \rightarrow V_{BC}$$

11.4

The desired sample computational results along with the generation program (m-file P_11_04.m), are reproduced below. *Note:* The assumed device area, $A = 10^{-4} \text{ cm}^2$, was omitted in the first printing.

$$\begin{aligned}\text{(a)} \quad \cosh(W/L_P) &= 1.0018 \\ \sinh(W/L_P) &= 5.9559 \times 10^{-2} \\ W/L_P &= 5.9524 \times 10^{-2}\end{aligned}$$

$$\begin{aligned}\text{(b)} \quad I_{Ep} &= 4.9473 \times 10^{-3} \text{ A} \\ I_{En} &= 7.2194 \times 10^{-6} \text{ A} \\ I_{Cp} &= 4.9386 \times 10^{-3} \text{ A} \\ I_{Cn} &= 9.4373 \times 10^{-15} \text{ A}\end{aligned}$$

$$\begin{aligned}\text{(c)} \quad I_E &= 4.9545 \times 10^{-3} \text{ A} \\ I_C &= 4.9386 \times 10^{-3} \text{ A} \\ I_B &= 1.5971 \times 10^{-5} \text{ A}\end{aligned}$$

$$\begin{aligned}\text{(d)} \quad I_{B1} &= 7.2194 \times 10^{-6} \text{ A} \\ I_{B2} &= 8.7515 \times 10^{-6} \text{ A} \\ I_{B3} &= -9.4373 \times 10^{-15} \text{ A}\end{aligned}$$

$$\begin{aligned}\text{(e)} \quad \gamma &= 0.99854 \\ \alpha_T &= 0.99823 \\ \alpha_{dc} &= 0.99678 \\ \beta_{dc} &= 309.22\end{aligned}$$

MATLAB program script...

```
%Sample Ideal-Transistor Computations...Problem 11.4
```

```
%Initialization
```

```
clear; format compact  
format short e
```

```
%Parameters and constants
```

NE=1.0e18;	NB=1.0e16;	NC=1.0e15;
μE=263;	μB=437;	μC=1345;
DE=6.81;	DB=11.3;	DC=34.8;
tauE=1.0e-7;	tauB=1.0e-6;	tauC=1.0e-6;
LE=8.25e-4;	LB=3.36e-3;	LC=5.90e-3;
	WB=2.0e-4;	

```

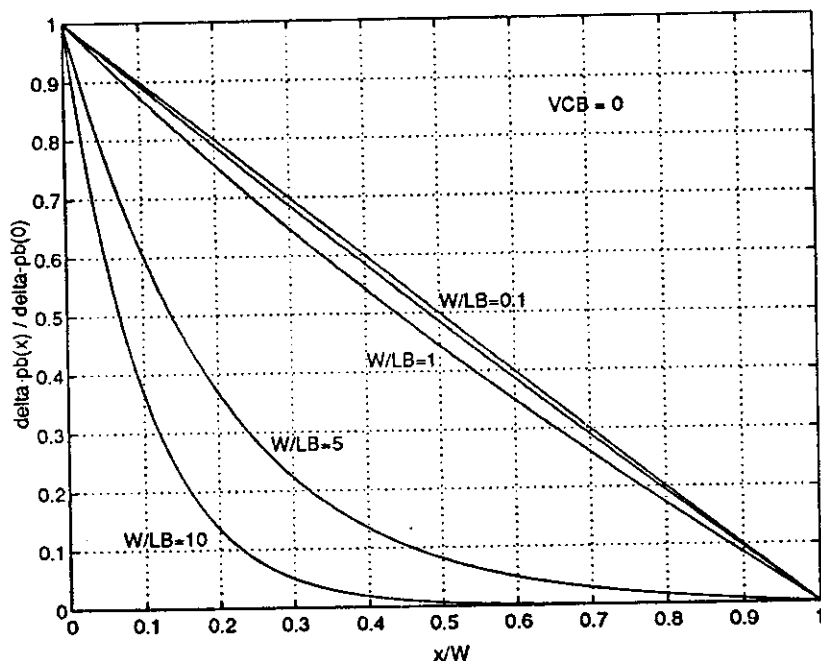
VEB=0.7;
VCB=-5;
W=WB;
A=1.0e-4;
q=1.6e-19;
ni=1.0e10;
kT=0.0259;

%Computations
%Part (a)
c=cosh(W/LB)
s=sinh(W/LB)
r=W/LB
%Part (b)
nE0=ni^2/NE;
pB0=ni^2/NB;
nC0=ni^2/NC;
IEp=q*A*DB*pB0/LB*(c/s*(exp(VEB/kT)-1)-1/s*(exp(VCB/kT)-1))
IEn=q*A*DE/LE*nE0*(exp(VEB/kT)-1)
ICp=q*A*DB*pB0/LB*(1/s*(exp(VEB/kT)-1)-c/s*(exp(VCB/kT)-1))
ICn=-q*A*DC/LC*nC0*(exp(VCB/kT)-1)
%Part (c)
IE=IEp+IEn
IC=ICp+ICn
IB=IE-IC
%Part (d)
IB1=IEn
IB2=IEp-ICp
IB3=-ICn
%Part (e)
gamma=IEp/(IEp+IEn)
aT=ICp/IEp
adc=gamma*aT
Bdc=adc/(1-adc)

```

11.5

The plot displayed on the next page illustrates how the minority carrier distribution in the base of a BJT varies with W/L_B . The plot is consistent with the fact that we expect a linear distribution when $W/L_B \ll 1$ and an exponentially decaying distribution when $W/L_B \gg 1$. It is perhaps somewhat surprising, however, that W/L_B values as large as unity yield a nearly linear distribution.



MATLAB program script...

%Minority Carrier Distribution in the base of a BJT
 %Variation with W/LB

%Initialization

clear; close

% $\Delta p_b(x)/\Delta p_b(0)$ vs. x/W ($V_{CB}=0$) calculation

%Let $r=W/LB$, $z=x/W$, $\Delta p_{br}=\Delta p_b(x)/\Delta p_b(0)$

$r=[10 \ 5 \ 1 \ 0.5 \ 0.1]$;

$z=\text{linspace}(0,1)$;

for $i=1:5$,

$\Delta p_{br}(i,:)=\sinh(r(i).*(1-z))./\sinh(r(i));$

end

%Plotting result

plot($z,\Delta p_{br}$);

axis([0 1 0 1]); grid

xlabel('x/W'); ylabel('delta-pb(x) / delta-pb(0)');

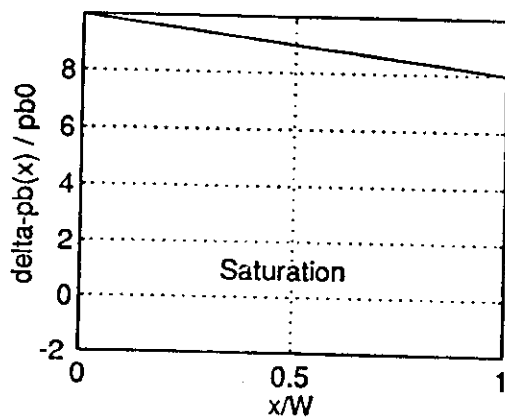
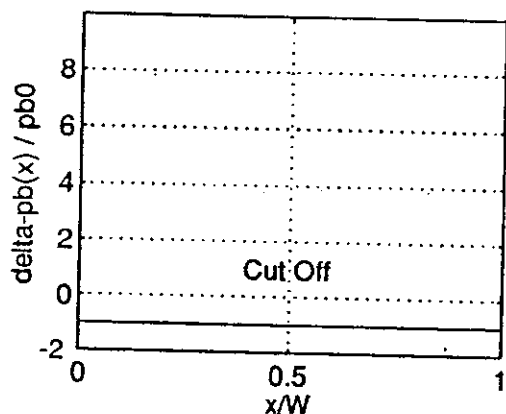
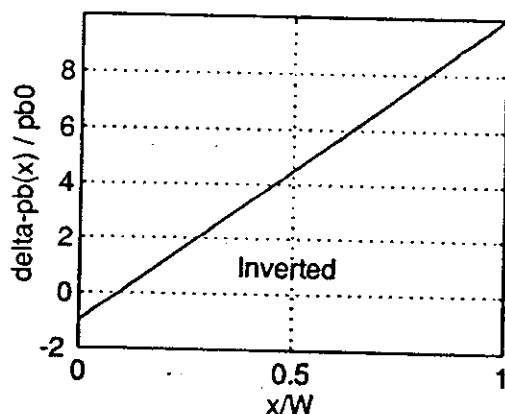
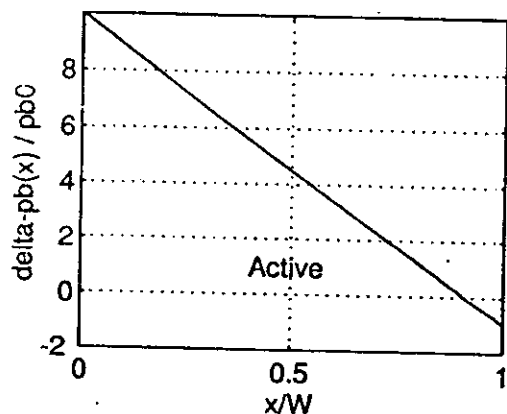
text(0.65,0.85,'VCB = 0');

text(0.5,0.52,'W/LB=0.1'); text(0.4,0.42,'W/LB=1');

text(0.27,0.28,'W/LB=5'); text(0.07,0.12,'W/LB=10');

11.6

"Demonstration plots" illustrating the general nature of the minority carrier distribution in a base of a BJT corresponding to the four biasing modes are displayed below. There also follows a listing of the MATLAB m-file used to produce the plots.



MATLAB program script...

%BJT Biasing Modes Illustration

%Initialization

clear; close

% $\Delta p_b(x)/p_{b0}$ vs. x/W calculation

%Let $r=W/LB$, $z=x/W$, and $\Delta p_{br}=\Delta p_b(x)/p_{b0}$

$r=0.1$;

$z=\text{linspace}(0,1)$;

$A=[10 \ -1 \ -1 \ 10]$;

$B=[-1 \ 10 \ -1 \ 8]$;

$\Delta p_{br}=A*\sinh(r*(1-z))/\sinh(r) + B*\sinh(r*z)/\sinh(r)$;

%Constructing Plots

subplot(2,2,1); plot(z, $\Delta p_{br}(1,:)$);

axis([0 1 -2 10]); grid

xlabel('x/W'); ylabel('delta-pb(x) / pb0');

text(0.4,1,'Active');

subplot(2,2,2); plot(z, $\Delta p_{br}(2,:)$);

axis([0 1 -2 10]); grid

xlabel('x/W'); ylabel('delta-pb(x) / pb0');

text(0.37,1,'Inverted');

subplot(2,2,3); plot(z, $\Delta p_{br}(3,:)$);

axis([0 1 -2 10]); grid

xlabel('x/W'); ylabel('delta-pb(x) / pb0');

text(0.39,1,'Cut Off');

subplot(2,2,4); plot(z, $\Delta p_{br}(4,:)$);

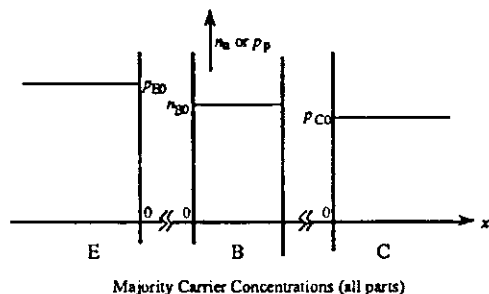
axis([0 1 -2 10]); grid

xlabel('x/W'); ylabel('delta-pb(x) / pb0');

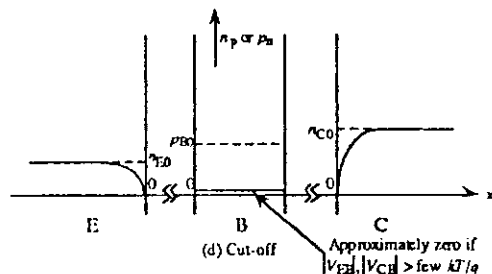
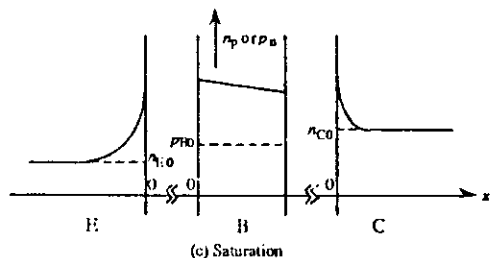
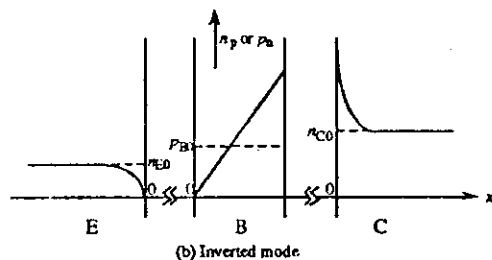
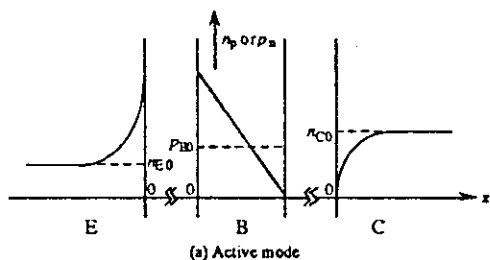
text(0.33,1,'Saturation');

11.7

Majority carriers... Assuming low-level injection, the majority carrier concentrations will be essentially unperturbed from their equilibrium values in all cases. Thus the majority carrier sketches will all be the same, with the solid carrier-distribution lines lying on top of the dashed equilibrium values in the three device regions.



Minority carriers... With $W \ll L_B$, the minority carrier concentration will vary linearly with position in the base under all biasing conditions. In the emitter and collector regions the carrier concentrations will decay exponentially with distance from the edges of the respective E-B and C-B depletion regions. The carrier concentrations will be greater than the equilibrium values when the applied biases are positive and less than the equilibrium values when the applied biases are negative. The distributions are concluded to be of the general form pictured below. Note that $p_B(x)$ in the base is approximately zero everywhere under cutoff biasing if $|V_{EB}|$ and $|V_{CB}|$ are greater than a few kT/q volts.



11.8

(a) $V_{EB} < 0$. The E-B junction is concluded to be reverse biased because both $\Delta n_E/n_{E0}$ and $\Delta p_B/p_{B0}$ are less than zero at the edges of the E-B depletion region.

(b) $V_{CB} > 0$. The C-B junction is concluded to be forward biased because both $\Delta p_B/p_{B0}$ and $\Delta n_C/n_{C0}$ are greater than zero at the edges of the C-B depletion region.

(c) The boundary condition at the base edge of the C-B depletion region (Eq. 11.4b) requires

$$\Delta p_B(W) = p_{B0}(e^{qV_{CB}/kT} - 1)$$

Thus

$$\Delta p_B(W)/p_{B0} = e^{qV_{CB}/kT} - 1 = 10$$

and assuming room temperature operation

$$V_{CB} = (kT/q) \ln(11) = (0.0259) \ln(11) = 0.062 \text{ V}$$

(d) **Inverted**

(e) For an *npn* BJT, $n_E \rightarrow p_E$, $p_B \rightarrow n_B$, and $n_C \rightarrow p_C$. However, little else changes. The E-B junction would still be reverse biased, the C-B junction forward biased, and the BJT operated in the inverted mode with $V_{BC} = 0.062 \text{ V}$. (Note that V_{BC} replaces V_{CB} .)

11.9

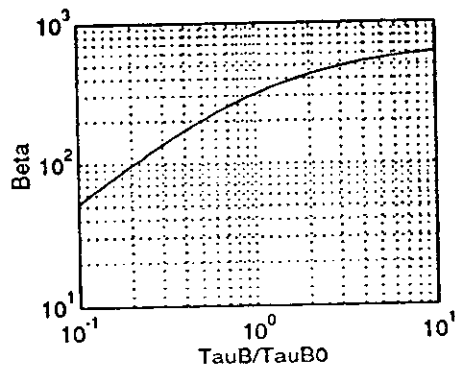
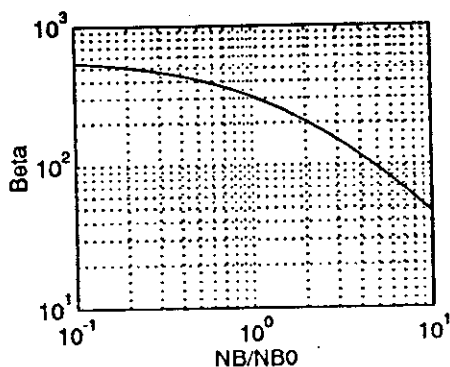
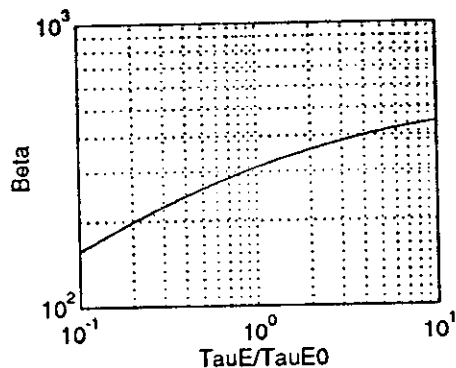
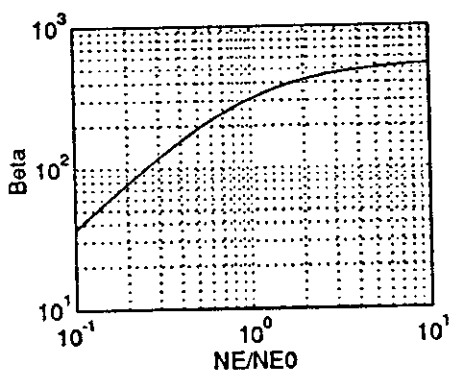
Change	Effect on γ	Effect on α_T	Effect on β_{dc}
Increase W_B	Decreases	Decreases	Decreases
Increase τ_B	No Effect (if $W \ll L_B$)	Increases	Increases
Increase N_B	Decreases	Decreases	Decreases
Increase τ_E	Increases	No Effect	Increases
Increase N_E	Increases	Essentially No Effect	Increases

Explanations:

- (1) In all cases, the effect on β_{dc} is determined by noting the effect on γ and α_T . Specifically, β_{dc} exhibits the same modification as α_{dc} , and $\alpha_{dc} = \gamma\alpha_T$. Thus, if the $\gamma\alpha_T$ product decreases, then β_{dc} decreases; if the $\gamma\alpha_T$ product increases, β_{dc} increases.
- (2) Assuming increasing W_B likewise increases W , the effect of increasing W_B on γ and α_T is obvious by inspecting Eqs. (11.41) and (11.42), respectively.
- (3) Since $L_B = (D_B \tau_B)^{1/2}$, increasing τ_B increases L_B , which in turn increases α_T . In a typical BJT, $W \ll L_B$ and from Eq. (11.41) we conclude increasing τ_B has no effect on γ . However, if W is allowed to be arbitrary, then per Eq. (11.31) increasing τ_B increases L_B , which decreases the sinh/cosh term and thereby increases γ .
- (4) Inspecting Eq. (11.41) it is obvious γ decreases if N_B increases. Noting that the mobility decreases if N_B increases and given $L_B = (D_B \tau_B)^{1/2} = [(kT/q)\mu_B \tau_B]^{1/2}$, it follows that α_T also decreases if N_B increases.
- (5) Since $L_E = (D_E \tau_E)^{1/2}$, an increase in τ_E leads to an increase in L_E , which in turn leads to an increase in γ as is obvious from Eq. (11.41). The τ_E parameter has no effect on transport through the base and α_T .
- (6) It is obvious from an inspection of Eqs. (11.41) and (11.42) that γ increases with increasing N_E and that N_E does not explicitly affect α_T . Changing N_E will slightly change the V_{bi} across the emitter-base junction and thereby affect the emitter-base depletion width. The net effect on α_T , however, should typically be small.

11.10

The requested plots and the generating program are reproduced below. The graphical results here are indeed consistent with the answers to Problem 11.9.— β_{dc} increases with increasing N_E , τ_E , and τ_B . β_{dc} decreases with increasing N_B .



MATLAB program script...

```
%Problem 11.10...Beta Calculation
%Initialization
clear; close
%Reference Values and Constants
NE0=1.0e18;
TauE0=1.0e-7;
NB0=1.0e16;
TauB0=1.0e-6;
WB=2.0e-4;
kT=0.0259;
```

```

W=WB;

%Mobility Fit Parameters
NDref=1.3e17; NAref=2.35e17;
μnmin=92; μpmin=54.3;
μn0=1268; μp0=406.9;
an=0.91; ap=0.88;

%Beta vs. NE/NE0
%Base Parameters
NB=NB0;
μB=μpmin+μp0./(1+(NB./NAref).^ap);
DB=kT.*μB;
TauB=TauB0;
LB=sqrt(DB.*TauB);
%Emitter Parameters
NE=logspace(log10(0.1*NE0),log10(10*NE0));
μE=μnmin+μn0./(1+(NE./NDref).^an);
DE=kT.*μE;
TauE=TauE0;
LE=sqrt(DE.*TauE);
%Beta Calculation
Y=1./((DE./DB).*(NB./NE).*(W./LE)+(0.5).*(W./LB).^2);
X=NE./NE0;
%Plotting
subplot(2,2,1); loglog(X,Y); grid
xlabel('NE/NE0'); ylabel('Beta')

%Beta vs. TauE/TauE0
%Revised Emitter Parameters
NE=NE0;
μE=μnmin+μn0./(1+(NE./NDref).^an);
DE=kT.*μE;
TauE=logspace(log10(0.1*TauE0),log10(10*TauE0));
LE=sqrt(DE.*TauE);
%Beta Calculation
Y=1./((DE./DB).*(NB./NE).*(W./LE)+(0.5).*(W./LB).^2);
X=TauE/TauE0;
%Plotting
subplot(2,2,2); loglog(X,Y); grid
xlabel('TauE/TauE0'); ylabel('Beta')

%Beta vs. NB/NB0
%Emitter Parameters
%NE, μE, and DE same as prior calculation
TauE=TauE0;
LE=sqrt(DE.*TauE);

```

```

%Base Parameters
NB=logspace(log10(0.1*NB0),log10(10*NB0));
μB=μpmin+μp0./(1+(NB./Naref).^ap);
DB=kT.*μB;
TauB=TauB0;
LB=sqrt(DB.*TauB);
%Beta Calculation
Y=1./((DE./DB).*(NB./NE).*(W./LE)+(0.5).*(W./LB).^2);
X=NB/NB0;
%Plotting
subplot(2,2,3); loglog(X,Y); grid
xlabel('NB/NB0'); ylabel('Beta')
%Beta vs. TauB/TauB0
%Revised Base Parameters
NB=NB0;
μB=μpmin+μp0./(1+(NB./Naref).^ap);
DB=kT.*μB;
TauB=logspace(log10(0.1*TauB0),log10(10*TauB0));
LB=sqrt(DB.*TauB);
%Beta Calculation
Y=1./((DE./DB).*(NB./NE).*(W./LE)+(0.5).*(W./LB).^2);
X=TauB/TauB0;
%Plotting
subplot(2,2,4); loglog(X,Y); grid
xlabel('TauB/TauB0'); ylabel('Beta')

```

11.11

(a) Like in the standard ideal-transistor analysis, the general solution for $\Delta n_E(x'')$ is

$$\Delta n_E(x'') = A_1 e^{-x''/L_E} + A_2 e^{x''/L_E}$$

However, because of the finite width of the emitter, $A_2 \neq 0$. Rather

$$\Delta n_E(x''=0) = n_{E0}(e^{qV_{EB}/kT} - 1) = A_1 + A_2$$

$$\Delta n_E(x''=W_E) = 0 = A_1 e^{-W_E/L_E} + A_2 e^{W_E/L_E}$$

Solving for A_1 and A_2 yields

$$A_1 = \frac{\Delta n_E(0) e^{W_E/L_E}}{e^{W_E/L_E} - e^{-W_E/L_E}} \quad \text{and} \quad A_2 = -\frac{\Delta n_E(0) e^{-W_E/L_E}}{e^{W_E/L_E} - e^{-W_E/L_E}}$$

Substituting back into the general solution then gives

$$\Delta n_E(x'') = \Delta n_E(0) \left[\frac{e^{(W_E - x'')/L_E} - e^{-(W_E - x'')/L_E}}{e^{W_E/L_E} - e^{-W_E/L_E}} \right]$$

or

$$\Delta n_E(x'') = n_{E0} \frac{\sinh[(W_E - x'')/L_E]}{\sinh(W_E/L_E)} (e^{qV_{EB}/kT} - 1)$$

Applying Eq. (11.7), we obtain the revised I_{En} expression

$$I_{En} = qA \frac{D_E}{L_E} n_{E0} \frac{\cosh(W_E/L_E)}{\sinh(W_E/L_E)} (e^{qV_{EB}/kT} - 1)$$

(b) Relative to the standard result (Eq. 11.20), the finite width of the emitter leads to the modification

$$n_{E0} \rightarrow n_{E0} \frac{\cosh(W_E/L_E)}{\sinh(W_E/L_E)}$$

or equivalently, since $n_{E0} = n_i^2/N_E$,

$$\frac{1}{N_E} \rightarrow \frac{1}{N_E} \frac{\cosh(W_E/L_E)}{\sinh(W_E/L_E)}$$

Revised expressions for the performance parameters analogous to Eqs. (11.31)–(11.34) are therefore obtained by simply making the above noted substitution. Specifically,

$$\left(\frac{D_E L_B N_B}{D_B L_E N_E} \right) \rightarrow \left(\frac{D_E L_B N_B}{D_B L_E N_E} \right) \frac{\cosh(W_E/L_E)}{\sinh(W_E/L_E)}$$

in the γ , α_{dc} , and β_{dc} expressions. α_T remains unchanged.

(c) With $W_E/L_E \ll 1$,

$$\frac{\cosh(W_E/L_E)}{\sinh(W_E/L_E)} \cong \frac{L_E}{W_E}$$

and L_E in the Eq. (11.41)–(11.44) expressions is to first order simply replaced by W_E . This yields, for example,

$$\gamma = \frac{1}{1 + \frac{D_E N_B W}{D_B N_E W_E}}$$

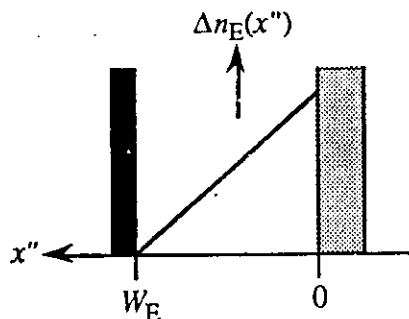
A similar modification applies to α_{dc} and β_{dc} . Being independent of L_E , the α_T expression remains unchanged.

(d) A systematic decrease in W_E obviously causes a monotonic decrease in γ and a corresponding decrease in β_{dc} . Reducing the width of the emitter has the negative effect of degrading the gain of the BJT (which is the point of the problem).

(e) If $W_E \ll L_E$ the part (a) result simplifies to

$$\Delta n_E(x'') = \Delta n_E(0) \left(1 - \frac{x''}{W_E}\right)$$

or one obtains a linear distribution analogous to the situation in the base. Moreover, since the emitter is forward biased under active mode biasing, $\Delta n_E(0) > 0$. Thus we conclude



11.12

"Diode" Configuration (a)...

(a) Note that $I = I_E$, $V_{EB} = V_A$, and $I_C = 0$ for the given circuit configuration. Since $I_C = 0$, it follows from Eq. (11.47b) that

$$\alpha_F I_{F0} (e^{qV_{EB}/kT} - 1) = I_{R0} (e^{qV_{CB}/kT} - 1)$$

or

$$(e^{qV_{CB}/kT} - 1) = (\alpha_F I_{F0} / I_{R0}) (e^{qV_{EB}/kT} - 1)$$

and

$$I = I_E = I_{F0} (e^{qV_{EB}/kT} - 1) - (\alpha_R I_{R0}) (\alpha_F I_{F0} / I_{R0}) (e^{qV_{EB}/kT} - 1)$$

Therefore

$$I = (1 - \alpha_R \alpha_F) I_{F0} (e^{qV_A/kT} - 1)$$

(b) $\Delta p_B(0)/p_{B0} = (e^{qV_{EB}/kT} - 1) = (e^{qV_A/kT} - 1)$

$$\Delta p_B(W)/p_{B0} = (e^{qV_{CB}/kT} - 1) = (\alpha_F I_{F0} / I_{R0}) (e^{qV_A/kT} - 1)$$

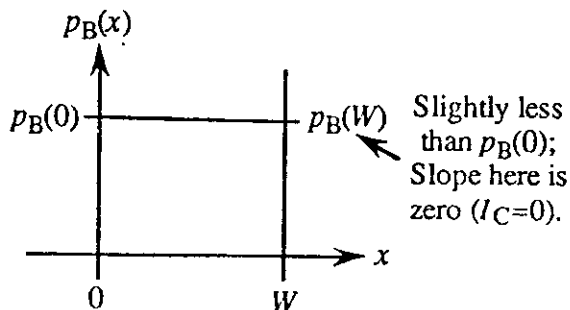
(c) Simplifying

$$\Delta p_B(0)/p_{B0} = (e^{qV_A/kT} - 1)$$

$$\Delta p_B(W)/p_{B0} = \alpha (e^{qV_A/kT} - 1)$$

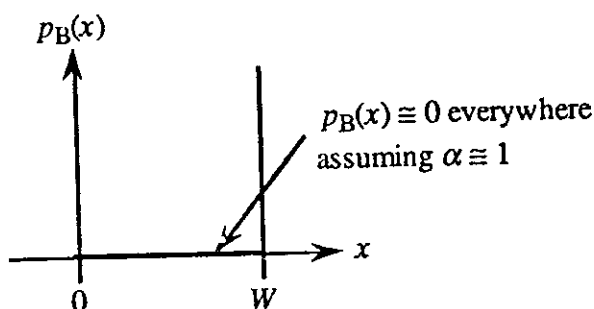
(d) If $V_A \gg kT/q$,

$$p_B(0) = p_{B0} e^{qV_A/kT} \quad \text{and} \quad p_B(W) \equiv \alpha p_{B0} e^{qV_A/kT}$$



(e) If $-V_A \gg kT/q$,

$$p_B(0) \equiv 0 \quad \text{and} \quad p_B(W) \equiv (1-\alpha)p_{B0} \equiv 0$$



"Diode" Configuration (b)...

(a) For the given circuit configuration, $V_{EB} = V_A$, $V_{CB} = 0$, and $I = I_E$. It therefore follows from Eq. (11.47a) that

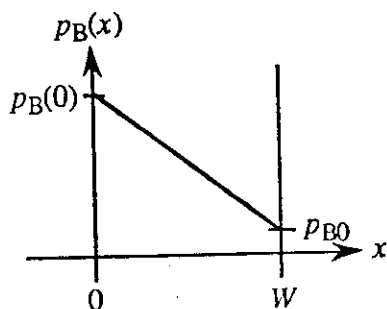
$$I = I_{F0} (e^{qV_A/kT} - 1)$$

(b) $\Delta p_B(0)/p_{B0} = (e^{qV_{EB}/kT} - 1) = (e^{qV_A/kT} - 1)$

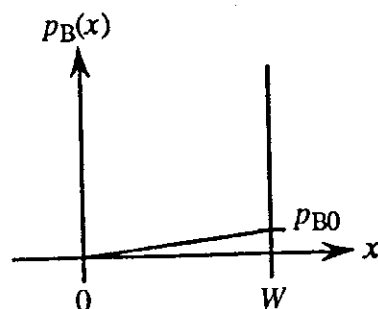
$$\Delta p_B(W)/p_{B0} = (e^{qV_{CB}/kT} - 1) = 0$$

(c) There is no further simplification of the part (b) result.

(d)



(e)



"Diode" Configuration (c)...

(a)/(b) For the given configuration, $V_{EB} - V_{CB} = V_A$ and $I = I_E = I_C$. Equating the Ebers-Moll relationships for I_E and I_C yields

$$(1-\alpha_F) I_{F0} (e^{qV_{EB}/kT} - 1) = -(1-\alpha_R) I_{R0} (e^{qV_{CB}/kT} - 1)$$

or

$$(e^{qV_{EB}/kT} - 1) = -\xi (e^{qV_{CB}/kT} - 1) \quad \dots \xi \equiv \frac{(1-\alpha_R)I_{R0}}{(1-\alpha_F)I_{F0}}$$

Next noting

$$(e^{qV_{EB}/kT} - 1) = (e^{qV_A/kT} e^{qV_{CB}/kT} - 1) = e^{qV_A/kT} (e^{qV_{CB}/kT} - 1) + (e^{qV_A/kT} - 1)$$

Thus

$$(e^{qV_A/kT} + \xi)(e^{qV_{CB}/kT} - 1) = -(e^{qV_A/kT} - 1)$$

or

$$(e^{qV_{CB}/kT} - 1) = -\frac{(e^{qV_A/kT} - 1)}{(e^{qV_A/kT} + \xi)} = \frac{\Delta p_B(W)}{p_{B0}}$$

and

$$(e^{qV_{EB}/kT} - 1) = \frac{\xi(e^{qV_A/kT} - 1)}{(e^{qV_A/kT} + \xi)} = \frac{\Delta p_B(0)}{p_{B0}} \quad \text{-----Part (b) Answer}$$

Finally, substituting into Eq. (11.47a) and simplifying

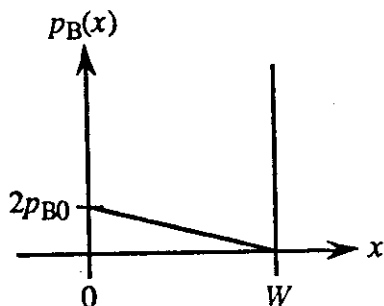
$$I = \frac{(1-\alpha_R\alpha_F)I_{R0}I_{F0}(e^{qV_A/kT} - 1)}{(1-\alpha_F)I_{F0}e^{qV_A/kT} + (1-\alpha_R)I_{R0}}$$

(c) Under the given simplifications, $\xi \rightarrow 1$ and

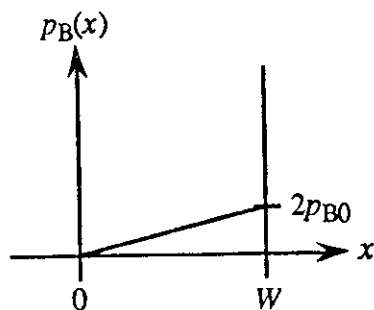
$$\frac{\Delta p_B(0)}{p_{B0}} = -\frac{\Delta p_B(W)}{p_{B0}} = \frac{e^{qV_A/kT} - 1}{e^{qV_A/kT} + 1}$$

(d)/(e) If $V_A \gg kT/q$, $\Delta p_B(0)/p_{B0} = -\Delta p_B(W)/p_{B0} = 1$ or $p_B(0) = 2p_{B0}$ and $p_B(W) = 0$.
If $-V_A \gg kT/q$, $\Delta p_B(0)/p_{B0} = -\Delta p_B(W)/p_{B0} = -1$ or $p_B(0) = 0$ and $p_B(W) = 2p_{B0}$.

(d)



(e)



"Diode" Configuration (d)...

(a) For the given connection, $I = I_B$ and $V_{EB} = V_{CB} = V_A$. In general, employing Eqs. (11.47a) and (11.47b),

$$I_B = I_E - I_C = (1 - \alpha_F) I_{F0} (e^{qV_{EB}/kT} - 1) + (1 - \alpha_R) I_{R0} (e^{qV_{CB}/kT} - 1)$$

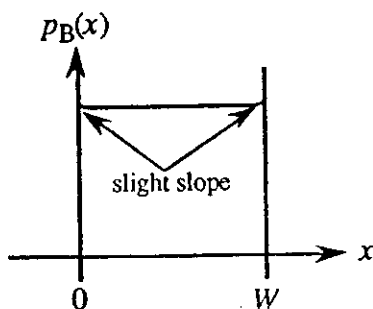
Thus

$$I = [(1 - \alpha_F) I_{F0} + (1 - \alpha_R) I_{R0}] (e^{qV_A/kT} - 1)$$

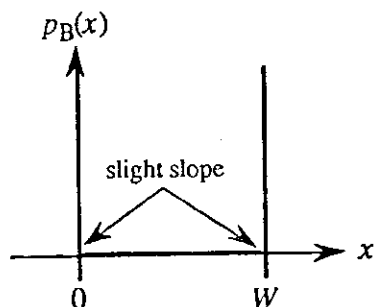
(b) $\Delta p_B(0)/p_{B0} = \Delta p_B(W)/p_{B0} = (e^{qV_A/kT} - 1)$

(c) There is no further simplification of the part (b) result.

(d)



(e)



"Diode" Configuration (e)...

(a) Here $I = -I_C$, $V_{CB} = V_A$, and $I_E = 0$. Since $I_E = 0$, it follows from Eq. (11.47a) that

$$I_{F0} (e^{qV_{EB}/kT} - 1) = \alpha_R I_{R0} (e^{qV_{CB}/kT} - 1)$$

or

$$(e^{qV_{EB}/kT} - 1) = (\alpha_R I_{R0} / I_{F0}) (e^{qV_{CB}/kT} - 1)$$

and

$$I = -I_C = -(\alpha_F I_{F0}) (\alpha_R I_{R0} / I_{F0}) (e^{qV_{CB}/kT} - 1) + I_{R0} (e^{qV_{CB}/kT} - 1)$$

Therefore

$$I = (1 - \alpha_R \alpha_F) I_{R0} (e^{qV_A/kT} - 1)$$

(b) $\Delta p_B(0)/p_{B0} = (e^{qV_{EB}/kT} - 1) = (\alpha_R I_{R0} / I_{F0}) (e^{qV_A/kT} - 1)$

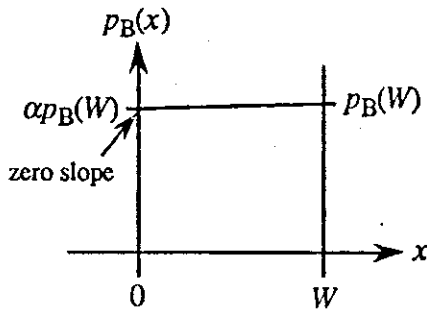
$$\Delta p_B(W)/p_{B0} = (e^{qV_{CB}/kT} - 1) = (e^{qV_A/kT} - 1)$$

(c) Simplifying

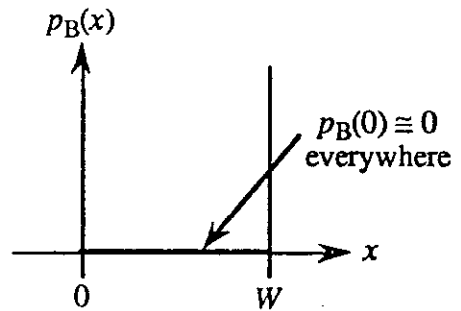
$$\Delta p_B(0)/p_{B0} = \alpha (e^{qV_A/kT} - 1)$$

$$\Delta p_B(W)/p_{B0} = (e^{qV_A/kT} - 1)$$

(d)



(e)



“Diode” Configuration (f)...

(a) Here $V_{EB} = 0$, $V_{CB} = V_A$, and $I = -I_C$. It therefore follows from Eq. (11.47b) that

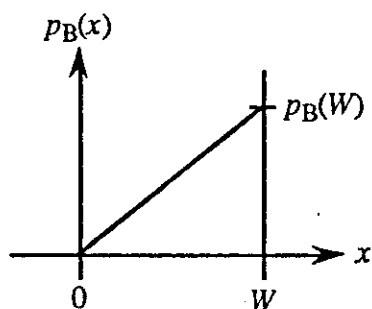
$$I = I_{R0} (e^{qV_A/kT} - 1)$$

(b) $\Delta p_B(0)/p_{B0} = (e^{qV_{EB}/kT} - 1) = 0$

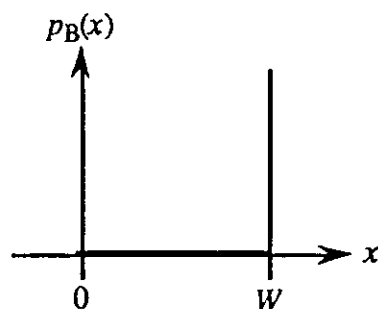
$$\Delta p_B(W)/p_{B0} = (e^{qV_{CB}/kT} - 1) = (e^{qV_A/kT} - 1)$$

(c) There is no further simplification of the part (b) result.

(d)

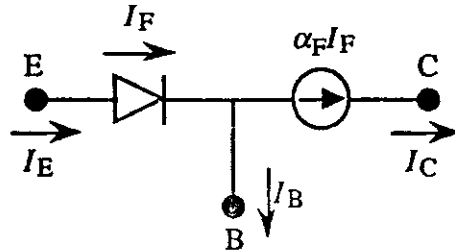


(e)



11.13

(a) Since $V_{CB} = 0$ at the active-mode/saturation-mode boundary, it follows that $I_R = 0$. The two Ebers-Moll circuit elements involving I_R therefore vanish yielding the following simplified circuit.



(b) As deduced from the part (a) circuit,

$$I_B = I_E - I_C = I_F - \alpha_F I_F = (1 - \alpha_F) I_{F0} (e^{qV_{EB}/kT} - 1)$$

but

$$V_{EC} = V_{EB} - V_{CB} = V_{EB} \quad (V_{CB} = 0)$$

Thus

$$I_B = (1 - \alpha_F) I_{F0} (e^{qV_{EC}/kT} - 1)$$

and

$$V_{EC} = \frac{kT}{q} \ln \left[1 + \frac{I_B}{(1 - \alpha_F) I_{F0}} \right]$$

11.14

(a) Solving Eq. (11.47a) for $\exp(qV_{EB}/kT) - 1$, one obtains

$$e^{qV_{EB}/kT} - 1 = \frac{I_E + \alpha_R I_{R0}(e^{qV_{CB}/kT} - 1)}{I_{F0}}$$

Substituting the $\exp(qV_{EB}/kT) - 1$ expression into Eq. (11.47b) then yields

$$I_C = \alpha_F I_E + \alpha_F \alpha_R I_{R0}(e^{qV_{CB}/kT} - 1) - I_{R0}(e^{qV_{CB}/kT} - 1)$$

or

$$I_C = \alpha_F I_E - (1 - \alpha_F \alpha_R) I_{R0}(e^{qV_{CB}/kT} - 1) \quad \Leftarrow \text{desired equation}$$

(b) Making use of Eqs. (11.47a) and (11.47b), we can write

$$I_B = I_E - I_C = (1 - \alpha_F) I_{F0}(e^{qV_{EB}/kT} - 1) + (1 - \alpha_R) I_{R0}(e^{qV_{CB}/kT} - 1)$$

Now

$$V_{CB} = V_{CE} + V_{EB} = V_{EB} - V_{EC}$$

and therefore

$$I_B = (1 - \alpha_F) I_{F0} e^{qV_{EB}/kT} - (1 - \alpha_F) I_{F0} + (1 - \alpha_R) I_{R0} e^{qV_{EB}/kT} e^{-qV_{EC}/kT} - (1 - \alpha_R) I_{R0}$$

or rearranging to obtain the desired equation

$$I_B = [(1 - \alpha_F) I_{F0} + (1 - \alpha_R) I_{R0} e^{-qV_{EC}/kT}] e^{qV_{EB}/kT} - [(1 - \alpha_F) I_{F0} + (1 - \alpha_R) I_{R0}]$$

(c) Eq. (11.47b) can be rewritten

$$\begin{aligned} I_C &= \alpha_F I_{F0} e^{qV_{EB}/kT} - \alpha_F I_{F0} - I_{R0} e^{qV_{EB}/kT} e^{-qV_{EC}/kT} + I_{R0} \\ &= (\alpha_F I_{F0} - I_{R0} e^{-qV_{EC}/kT}) e^{qV_{EB}/kT} + I_{R0} - \alpha_F I_{F0} \end{aligned}$$

Referring to the I_B solution in part (b), we note

$$e^{qV_{EB}/kT} = \frac{I_B + (1 - \alpha_F) I_{F0} + (1 - \alpha_R) I_{R0}}{(1 - \alpha_F) I_{F0} + (1 - \alpha_R) I_{R0} e^{-qV_{EC}/kT}}$$

which when substituted into the I_C expression yields

$$I_C = \frac{(\alpha_F I_{F0} - I_{R0} e^{-qV_{EC}/kT}) [I_B + (1 - \alpha_F) I_{F0} + (1 - \alpha_R) I_{R0}]}{(1 - \alpha_F) I_{F0} + (1 - \alpha_R) I_{R0} e^{-qV_{EC}/kT}} + I_{R0} - \alpha_F I_{F0}$$

11.15

(a) As recorded in Exercise 11.7, the general relationship for the common emitter input characteristic is

$$I_B = [(1-\alpha_F)I_{F0} + (1-\alpha_R)I_{R0}e^{-qV_{EC}/kT}]e^{qV_{EB}/kT} - [(1-\alpha_F)I_{F0} + (1-\alpha_R)I_{R0}]$$

For a routine bias of $V_{EC} > \text{few } kT/q$ volts, the term involving $\exp(-qV_{EC})$ will be negligible. Moreover, the second group of terms is typically small and may be completely ignored for V_{EB} biases normally encountered in practice. Thus the I_B expression reduces to

$$I_B \cong (1-\alpha_F)I_{F0}e^{qV_{EB}/kT}$$

Using Eqs. (11.45a) and (11.46) to display the explicit parametric dependencies, one obtains

$$I_B \cong qA \left[\frac{D_E}{L_E} n_{E0} + \frac{D_B}{L_B} p_{B0} \frac{\cosh(W/L_B) - 1}{\sinh(W/L_B)} \right] e^{qV_{EB}/kT}$$

In a standard transistor $W \ll L_B$ and

$$\frac{\cosh(W/L_B) - 1}{\sinh(W/L_B)} \cong \frac{(1/2)(W/L_B)^2}{W/L_B} = W/2L_B$$

and

$$I_B = qA \left(\frac{D_E}{L_E} n_{E0} + \frac{1}{2} \frac{D_B W}{L_B L_B} p_{B0} \right) e^{qV_{EB}/kT}$$

or

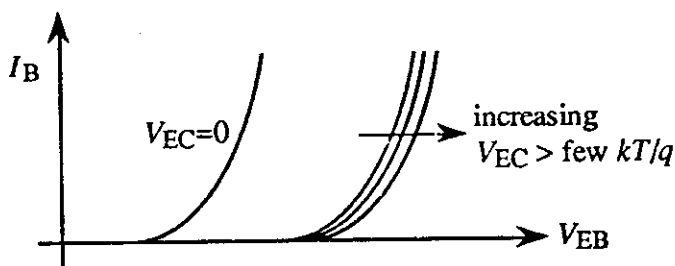
$$I_B = qA \left(\frac{D_E}{L_E} \frac{n_i^2}{N_E} + \frac{1}{2} \frac{W D_B}{L_B L_B} \frac{n_i^2}{N_B} \right) e^{qV_{EB}/kT}$$

If $D_E N_B W / D_B N_E L_E \gg (1/2)(W/L_B)^2$, then the first term inside the parentheses in the above expression will be much greater than the second term. We conclude for the specific situation

$$I_B \cong qA \left(\frac{D_E}{L_E} \frac{n_i^2}{N_E} \right) e^{qV_{EB}/kT}$$

There is no W -dependence in the above expression. Therefore the common emitter input characteristics are expected to be relatively insensitive to base width modulation if the gain is limited by the emitter efficiency.

(b) If the gain is not limited by the emitter efficiency, then the second term in the simplified I_B expression developed in part (a) must be retained. This second term decreases with decreasing W , and I_B is therefore predicted to decrease with increasing V_{EC} . For the specified situation, the expected form of the common emitter input characteristics are as sketched below.



11.16

If $V_{EB} = 0.7$ V instead of zero as assumed in Exercise 11.8, then x_{nEB} will be reduced and V_{CB} is expected to increase. Specifically, recalculating x_{nEB} and V_{CB} ,

$$x_{nEB} = \left[\frac{2K_S \epsilon_0}{q} \frac{N_E}{N_B(N_E + N_B)} (V_{bi(EB)} - V_{EB}) \right]^{1/2}$$

$$= \left[\frac{(2)(11.8)(8.85 \times 10^{-14})(10^{18})(0.835 - 0.7)}{(1.6 \times 10^{-19})(10^{16})(1.01 \times 10^{18})} \right]^{1/2} = 1.32 \times 10^{-5} \text{ cm}$$

$$V_{CB} = V_{bi(CB)} - \frac{(W_B - x_{nEB})^2}{\frac{2K_S \epsilon_0}{q} \frac{N_C}{N_B(N_C + N_B)}} \quad \dots \text{at punch-through}$$

$$= 0.656 - \frac{(2 \times 10^{-4} - 1.32 \times 10^{-5})^2}{\left[\frac{(2)(11.8)(8.85 \times 10^{-14})(10^{15})}{(1.6 \times 10^{-19})(10^{16})(1.1 \times 10^{16})} \right]} = -293 \text{ V}$$

11.17

(a) **Transistor A**. Referring to Eq. (11.41),

$$\gamma = \frac{1}{1 + \frac{D_E N_B W}{D_B N_E L_E}}$$

In Transistor A, $N_E \gg N_B$ and $\gamma \rightarrow 1$. In Transistor B, $N_E < N_B$ and γ is expected to be considerably less than unity.

One might alternatively argue that $I_{Ep} \gg I_{En}$ in Transistor A, while $I_{En} > I_{Ep}$ in Transistor B. Since $\gamma = I_{Ep}/(I_{Ep} + I_{En})$ in a *pnp* transistor, Transistor A clearly has the greater emitter injection efficiency.

(b) Under active mode biasing $V_{EB} > 0$ and $V_{CB} < 0$. Considering the more important reverse-bias collector-base junction, there is very little incursion of the depletion region into the base in Transistor A. For Transistor B, however, most of the depletion region lies in the base because $N_C \gg N_B$. Thus **Transistor B** will be more sensitive to base width modulation.

(c) **Transistor A**. V_{CB0} is approximately equal to V_{BR} of the C-B junction if the BJT is limited by avalanche breakdown. V_{BR} in turn is roughly inversely proportional to the doping on the lightly-doped side of the *pn* junction. In Transistor A, the collector is the lower doped with $N_C = 10^{14}/\text{cm}^3$; in Transistor B, the base has the lighter doping, $N_B = 10^{15}/\text{cm}^3$. Since N_C of Transistor A is less than N_B of Transistor B, Transistor A will exhibit the larger V_{CB0} .

11.18

The following "data" was obtained by running the BJTplus program (E_11_10.m) with base width modulation included (but excluding carrier multiplication).

$I_B(\mu A)$	$V_{EC}(V)$	$I_C(mA)$
2.5	40.9	0.807
2.5	101.2	1.089
5.0	40.9	1.614
5.0	101.2	2.177
7.5	40.9	2.421
7.5	101.2	3.266
10.0	40.9	3.228
10.0	101.2	4.354

Seeking to fit the data with the relationship

$$I_C = B(V_{EC} - V_E)I_B$$

one computes

$$B = \frac{I_{C2} - I_{C1}}{I_B(V_{EC2} - V_{EC1})}$$

and

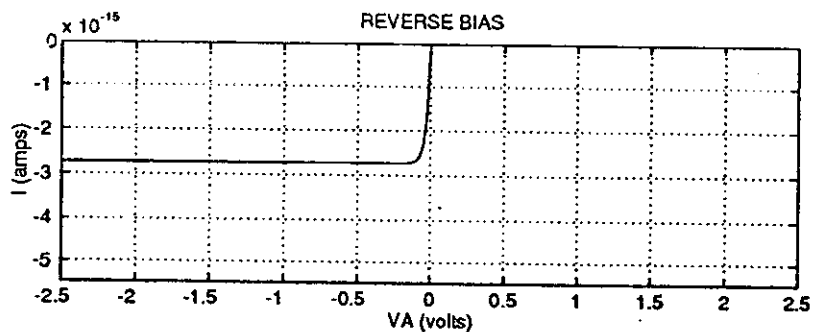
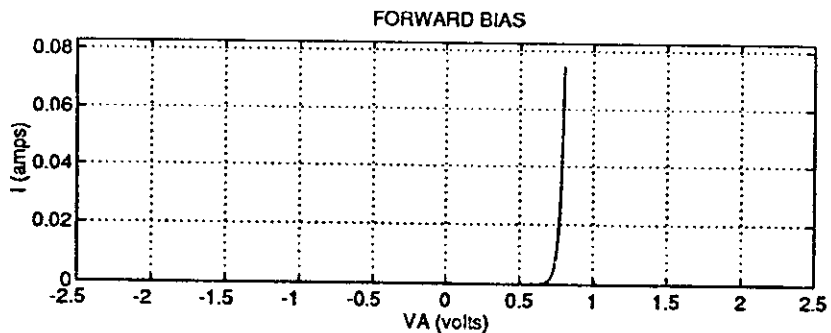
$$V_E = V_{EC2} - \frac{I_{C2}}{B I_B}$$

The same answer, $B = 1.87/V$ and $V_E = -132V$ is obtained employing all four data sets; i.e., for the given device

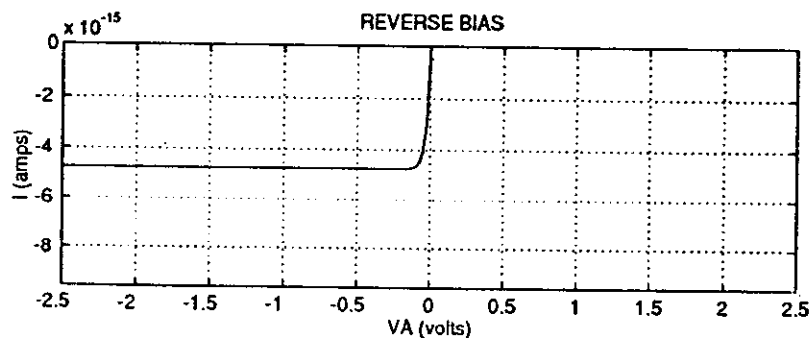
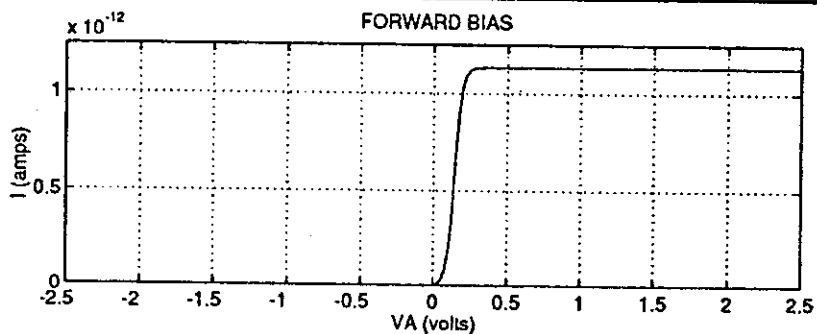
$$I_C = (1.87 I_B)(V_{EC} + 132) \quad \dots V_{EC} \text{ in volts; } I_B, I_C \text{ in amps}$$

11.19

(a) Basically, one obtains the same "ideal diode" type characteristic for all of the "diode" connections EXCEPT for the floating-base (c)-connection. The (c)-connection exhibits a reverse-bias type characteristic under both reverse and forward biasing. Representative print-outs of the I - V characteristics generated by the P_11_19a.m program (included on the instructor's disk) are reproduced on the next page.



Configuration (a)



Configuration (c)

(b) Generation of the I - V characteristics with base-width modulation and carrier multiplication included is considerably more challenging than initially envisioned by the author. This is especially true of the floating terminal connections (connections a, c, e). For one, it is necessary to associate separate carrier multiplication factors with the E-B and C-B junctions. If V_{EB0} and V_{CB0} are the avalanche breakdown voltages of the E-B and C-B junctions respectively, then $MF = 1/[1-(I_{VA}/V_{EB0})^m]$ is associated with carrier multiplication in the E-B junction under reverse bias conditions and $MR = 1/[1-(I_{VA}/V_{CB0})^m]$ is associated with carrier multiplication in the C-B junction. In the computations, carrier multiplication is included by replacing I_{F0} by $MF \cdot I_{F0}$ and I_{R0} by $MR \cdot I_{R0}$. (Note, however, that $\alpha_F I_{F0}$ and $\alpha_R I_{R0}$ are NOT multiplied by the M -factor.) Since base width modulation and carrier multiplication has little effect on the $V_A > 0$ characteristics (except for the (c)-connection), only the reverse bias characteristics are included in the revised computations (except for the (c)-connection).

The voltages required for the computations are straightforward in cases where two of the terminals are shorted together. Specifically, in these cases one has

- case (b) ... $V_{EB} = V_A$; $V_{CB} = 0$
- case (d) ... $V_{EB} = V_A$; $V_{CB} = V_A$
- case (f) ... $V_{EB} = 0$; $V_{CB} = V_A$

In connections (a) and (e) the voltage across the open junction floats up or down to assure that no current flows across this junction. The voltage across the open junction can be determined from the part (b) answers to Problem 11.12. Specifically, we find

- case (a) ... $V_{EB} = V_A$; $V_{CB} = (kT/q) \ln[1+(\alpha_F I_{F0}/I_{R0})(e^{qV_A/kT}-1)]$
- case (e) ... $V_{EB} = (kT/q) \ln[1+(\alpha_R I_{R0}/I_{F0})(e^{qV_A/kT}-1)]$; $V_{CB} = V_A$

The floating voltages pose somewhat of a computational problem. The Ebers-Moll parameters are needed to compute the floating voltage, and the computation of the Ebers-Moll parameters in turn depend on the floating voltage when base-width modulation is included. This is handled in the calculations by first computing the Ebers-Moll parameters ignoring base-width modulation, computing the floating voltage, redoing the Ebers-Moll parameters computation with base-width modulation included and employing the first-guess floating voltage, and then repeating the iteration procedure if necessary. It was found that only two complete iterations (two floating voltage and two base-width included computations of the Ebers-Moll parameters) was necessary.

As a final point relative to the applied voltages, please note that the avalanche breakdown in the E-B junction limits the maximum applied voltage in cases (a), (b) and (d). Avalanche breakdown in the C-B junction limits the maximum applied voltage in cases (e) and (f). With $N_B > N_C$, the breakdown voltage of the E-B junction is of course lower than the breakdown voltage of the C-B junction.

The (c)-connection poses the most serious computational problems. It is the only connection where both negative and positive values of V_A must be considered. Also, with the base floating, there is a built-in feedback effect as described in Subsection 11.2.4. One

therefore expects the reverse and forward limiting voltages to be somewhat less than the respective avalanche breakdown voltages. Finally, the junction voltages are coupled, with $V_{EB} - V_{CB} = V_A$. Note from the (c)-connection analysis in Problem 11.12 that

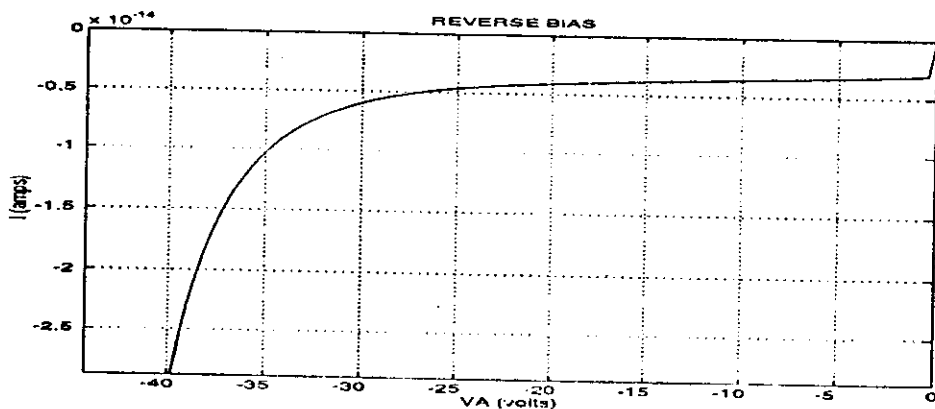
$$(e^{qV_{EB}/kT} - 1) = (e^{qV_{CB}/kT} e^{qV_A/kT} - 1) = -\frac{(1-\alpha_R)I_{R0}}{(1-\alpha_F)I_{F0}} (e^{qV_{CB}/kT} - 1)$$

and upon solving for V_{CB} ,

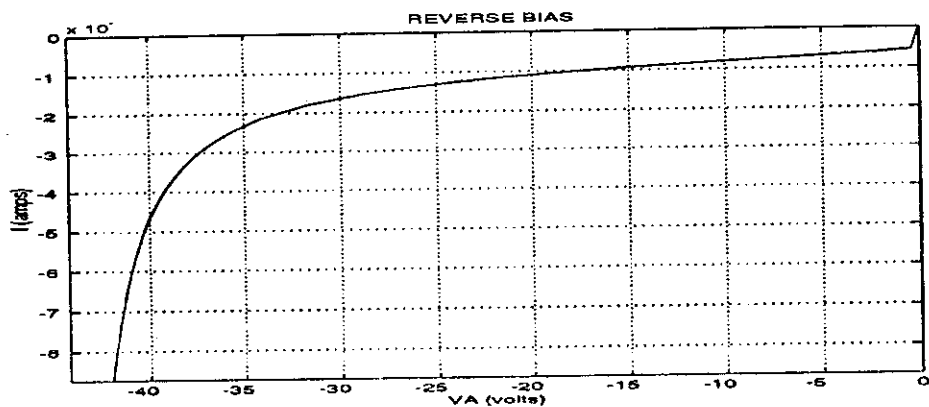
$$V_{CB} = \frac{kT}{q} \ln \left[\frac{1 + \frac{(1-\alpha_R)I_{R0}}{(1-\alpha_F)I_{F0}}}{e^{qV_A/kT} + \frac{(1-\alpha_R)I_{R0}}{(1-\alpha_F)I_{F0}}} \right]$$

Again, it is necessary to perform an iteration to determine the junction voltages and the Ebers-Moll parameters. Some additional care must be taken in handling $\exp(qV_A/kT)$ for forward biases. Unless the V_{CB} expression is modified, the exponential will exceed the largest value handled by the MATLAB program. When this happens, the exponential is artificially set equal to ∞ , leading to erroneous results.

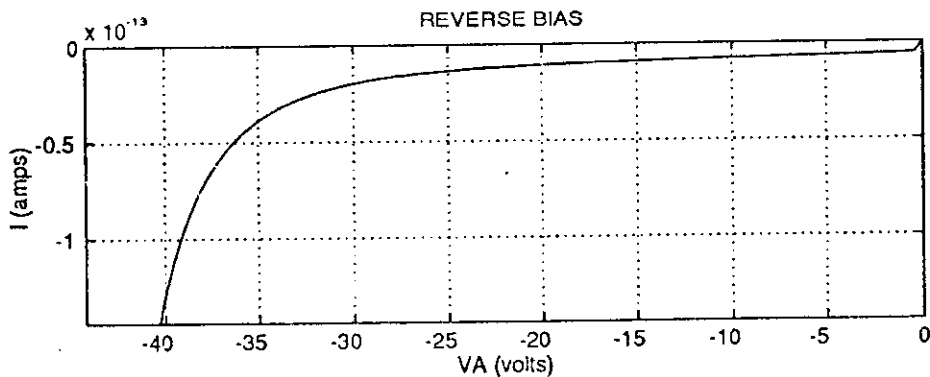
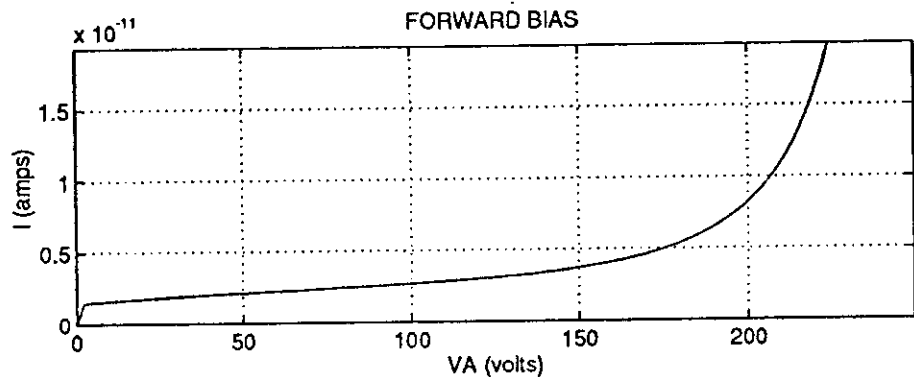
Print-outs of the I - V characteristics generated by the P_11_19b.m program (included on the Instructor's disk) are displayed below.



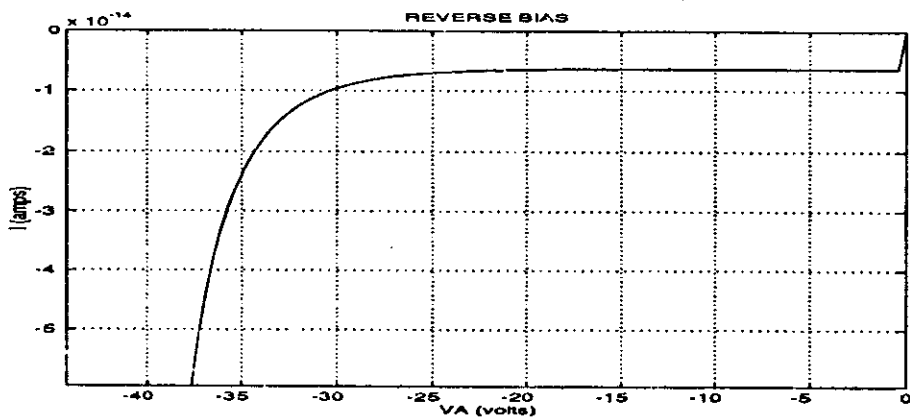
Conf. (a)



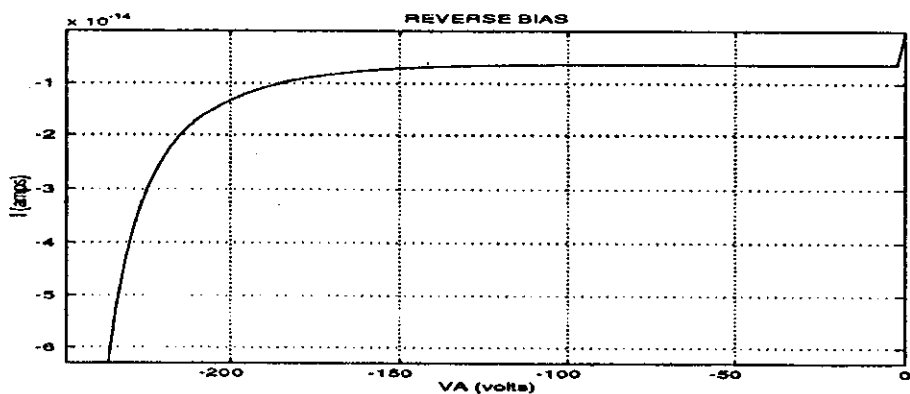
Conf. (b)



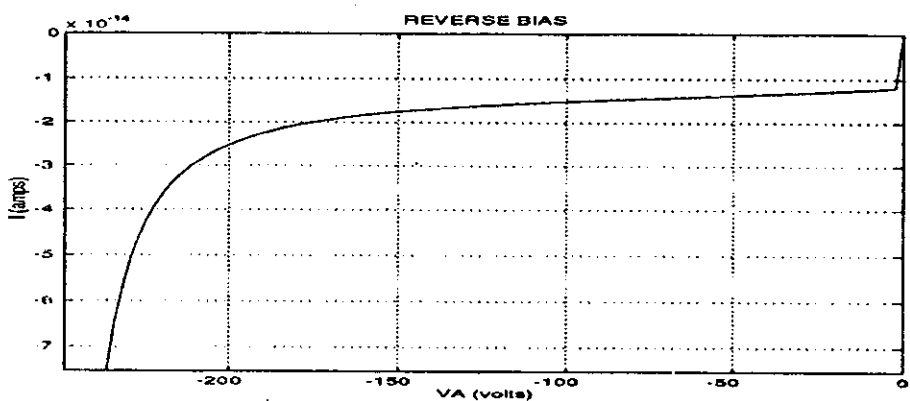
Conf. (c)



Conf. (d)



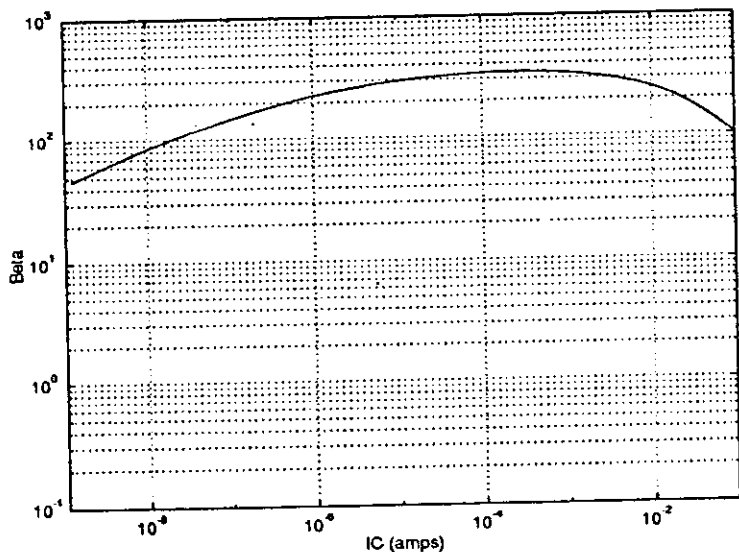
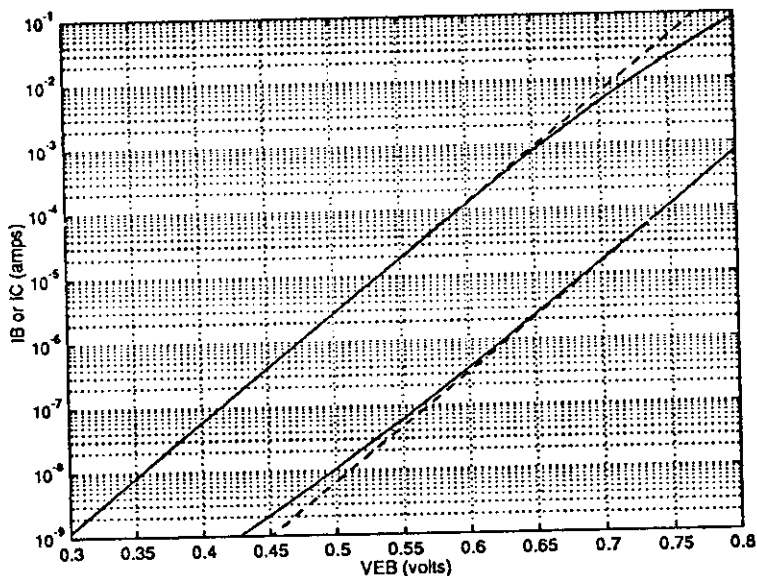
Conf. (e)



Conf. (f)

11.20

(a)/(b) The Gummel and gain plots produced per the instructions in the problem statement are displayed below. The plot-generating MATLAB program is listed after the plots. Although hardly perfect, the gain plot is a credible match to the experimental data presented in Fig. 11.16.



MATLAB program script...

%Gummel and Gain Plots...Problem 11.20

%(Note: subprogram BJT0 is a runtime requirement.)

%Initialization

clear; close

%Input Constant and Material Parameters

BJT0

n2=1.5;

I02=1.0e-14;

VHi=0.75;

%Calculation of IB and IC

VEB=linspace(0.3, 0.85);

IBideal=IF0.*(1-aF).*(exp(VEB./kT)-1);

IRG=I02.*(exp(VEB/(n2*kT))-1);

IB=IBideal+IRG;

ICideal=aF.*IRO.*(exp(VEB./kT)-1);

IC=ICideal./(1+exp((VEB-VHi)./(2*kT)));

%Gummel Plot

semilogy(VEB,IBideal,':w',VEB,ICideal,':w');

axis([.3,.8,1.0e-9,1.0e-1])

hold on

semilogy(VEB,IB,'-m',VEB,IC,'-c'); grid;

xlabel('VEB (volts)'); ylabel('IB or IC (amps)')

hold off

pause

%Gain Plot

Beta=IC./IB;

loglog(IC,Beta)

axis([1.0e-9,1.0e-1,1.0e-1,1.0e3]); grid

xlabel('IC (amps)'); ylabel('Beta')

11.21

The primary cause of the increase in β_{dc} with increasing V_{EC} is *base-width modulation*. As noted in the text, increasing V_{EC} increases the width of the C-B depletion region in the base, which in turn decreases W and hence increases β_{dc} . The observed I_C of course increases with β_{dc} per Eq. (11.50).

Supplemental Observation #1

The same device was used to produce Figs. 11.5(d) and P11.21. The BJTplus program results displayed in Exercise 11.10 provide a reasonable match to the Fig. 11.5(d) data, and should therefore be applicable to the Fig. P11.21 data. In Prob. 11.18 we found the linear region of the Exercise 11.10 characteristics could be accurately fit by the relationship

$$I_C = (1.87 I_B)(V_{EC} + 132) \quad \dots V_{EC} \text{ in volts; } I_B, I_C \text{ in amps}$$

or for the given device,

$$\beta_{dc} = I_C / I_B = 1.87(V_{EC} + 132)$$

One computes $\beta_{dc} = 256, 266$, and 275 if respectively $V_{EC} = 5V, 10V$, and $15V$. The cited β_{dc} fit is only good of course for I_C values in the low mA range, but does exhibit the correct general dependence.

Supplemental Observation #2

The MATLAB program developed for Prob. 11.20 can be readily revised to produce a plot similar to Fig. P11.21. The plot one obtains is reproduced below. The revised program is included on the Instructor's disk as m-file P_11_21.m. Note that the predicted variation is somewhat smaller than that observed experimentally.

