Tunnel diode

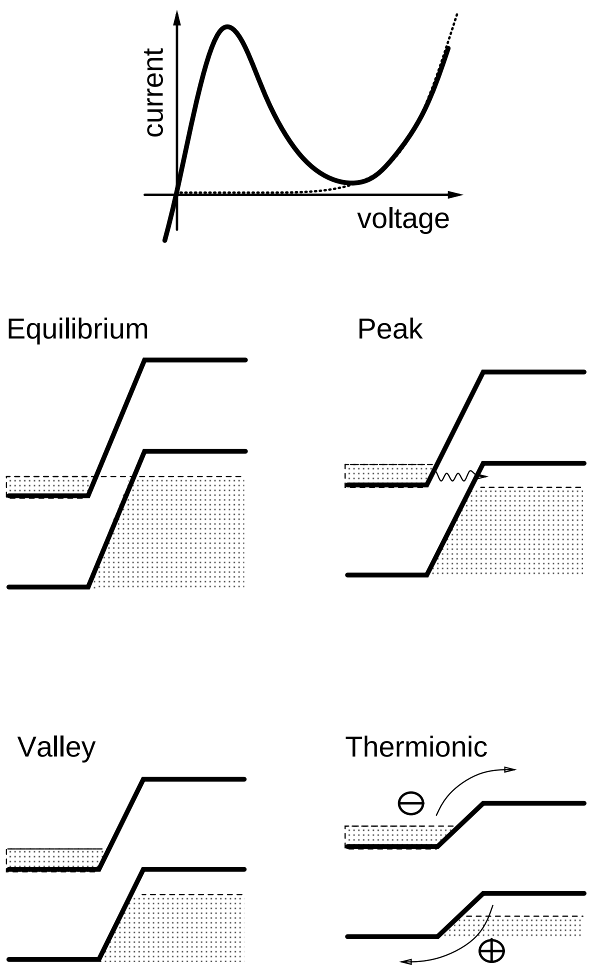


Figure 22: Tunnel (Esaki) diode

Backward diode

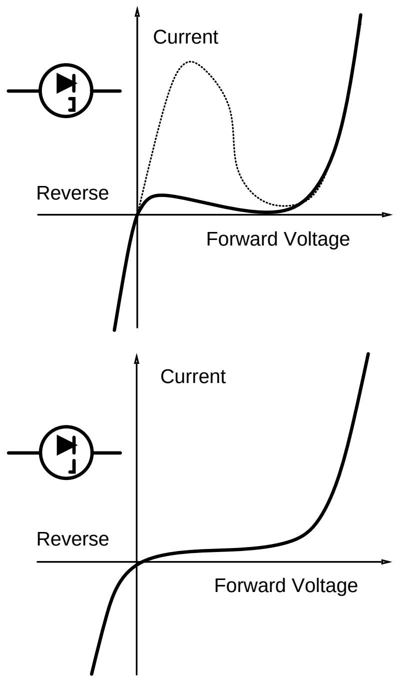


Figure 22a: Backward diode. Doping not quite as degenerate as with the tunnel diode. May not posess a region of NDR.

Turns on earlier in reverse than in forward direction.

Can be used for rectification of small signal and micrwave mixing.

Good frequency response (no minority-carrier storage effect).

Insensitive to *T* variations (also radiation resiatant!)

Has very low $\frac{1}{f}$ noise.

Can also be used for emitter-base transistor contacts

Z. S. Gribnikov and S. Luryi, "Article comprising a bipolar transistor with a floating base", US Pat. **5,461,245** (October 1995)]

A. Zaslavsky, S. Luryi, R. W. Johnson, and C. A. King, "Multiemitter Si/*GS* heterojunction bipolar transistor with no base contact and enhanced logic functionality", *IEEE Electron Device Lett.* **EDL-18**, pp. 453-455 (1997)

Double-barrier diode

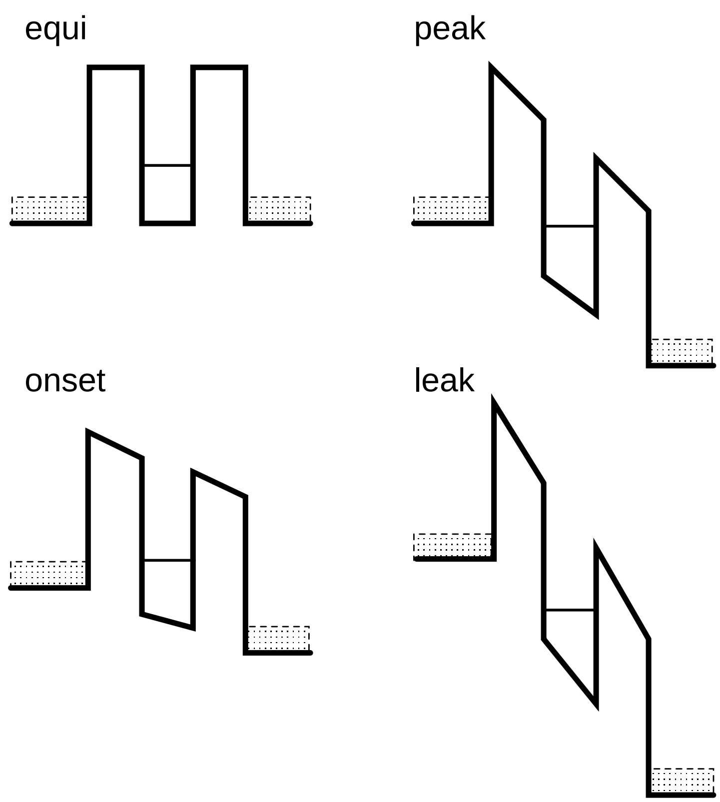


Figure 22: Resonant tunneling diode, band diagram

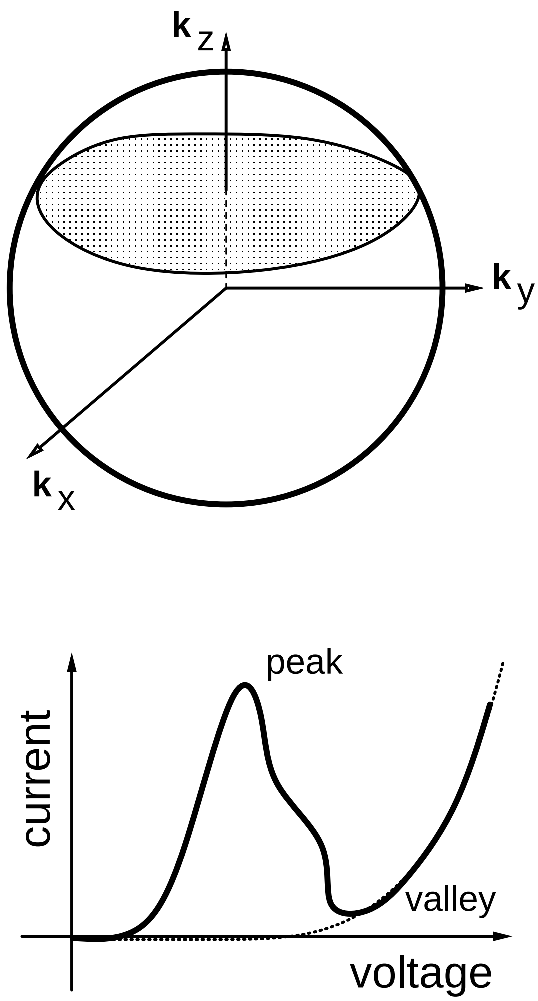


Figure 24: Mechanism of operation of double-barrier resonant tunneling diode.† Lateral components k_x, k_y of the wavevector **k** are conserved in tunneling.

---

† S. Luryi, "Mechanism of operation of double-barrier resonant-tunneling oscillators", *1985 International Electron Device Meeting*, Tech. Digest: IEDM 85, pp. 666-669 (1985).

S. Luryi and A. Zaslavsky, "Quantum-Effect and Hot-Electron Devices" Chap. 5 in *Modern Semiconductor Device Physics*, ed. by S. M. Sze (Wiley Interscience, ISBN 0-471-15327-4) pp. 253-342 (1997)

Heterojunctions

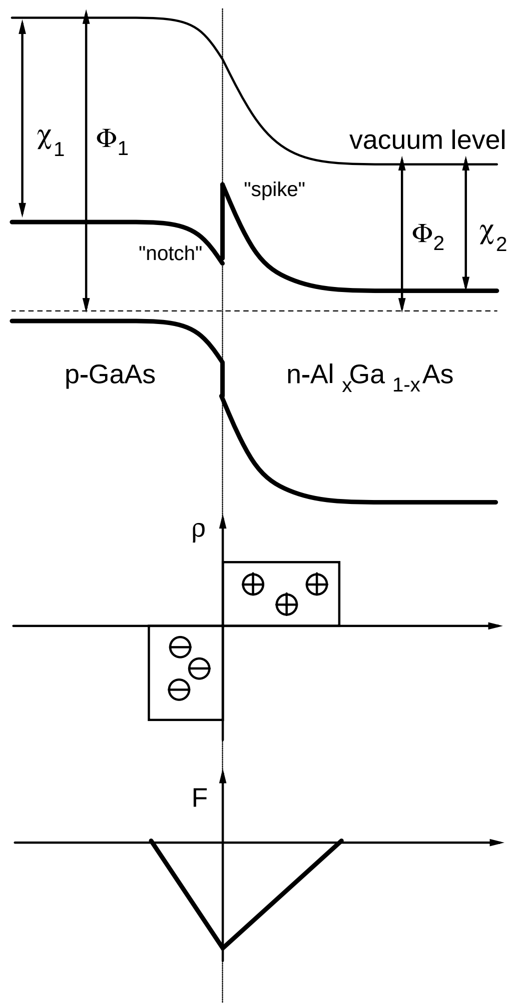


Figure 25: Heterojunctions: GaAs/$Al_xGa_{1-x}As$ lattice-matched heterostructure.

Band alignment is an empirical number. Affinity rule does not work due to an "unpredictable" chemical dipole layer at the junction. Many efforts to predict the unpredictable can be found in the literature, some quite good, empirically, none universally accepted.

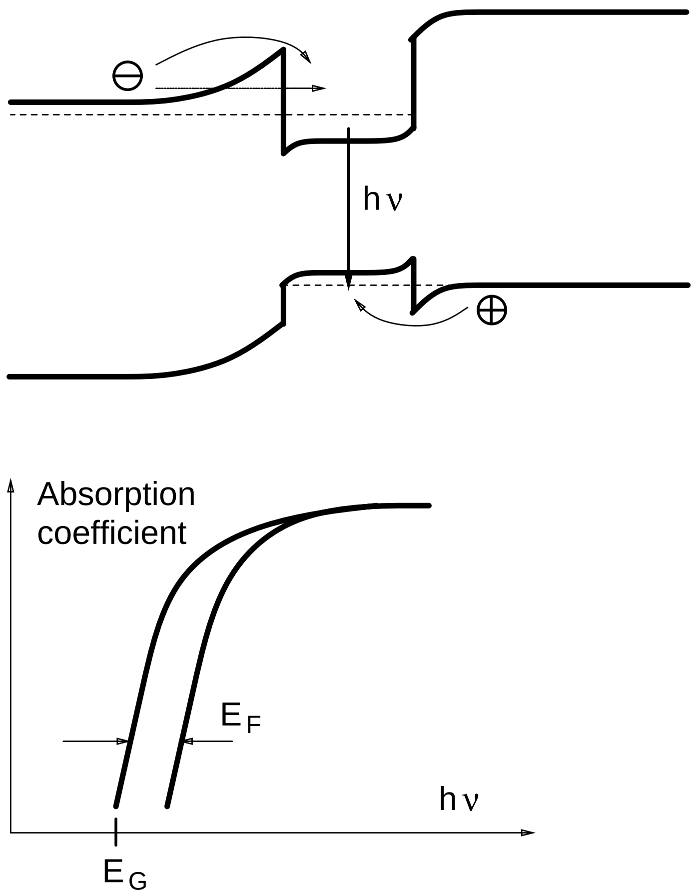


Figure 26: Double heterostructure laser structure (schematically).

Both electron and holes are injected into the narrow-gap material where they constitute ***degenerate*** ensembles.

Optical gain occurs because of the Fermi degeneracy, which suppresses the absorption of light of frequency ν (Moss-Burstein shift in the absorption spectra).

Fundamental condition for positive gain:

For $gain(\nu) > 0$ one must have

$$E_{Fn} - E_{Fp} > h\nu$$

Junction Transistor

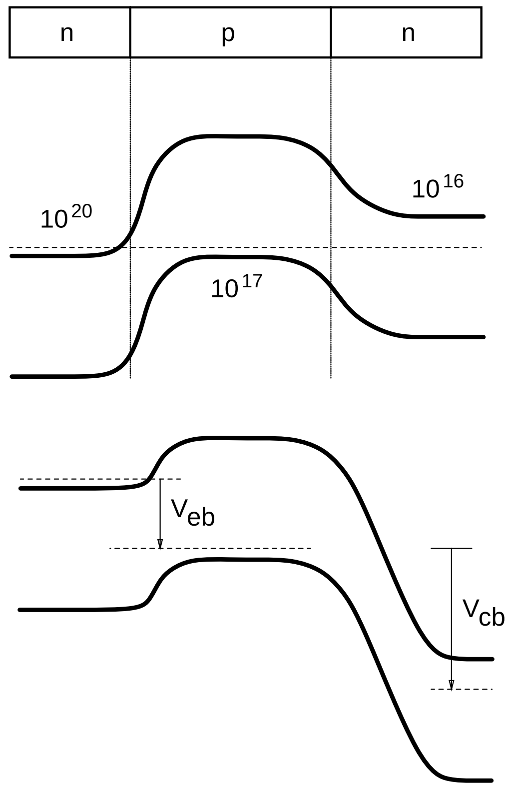


Figure 1: Schematic diagram of a *npn* transistor in equilibrium and under applied bias. By Kirchhoff law:

$$I_E = I_C + I_B \approx I_C$$

Neglecting recombination in the base and parasitic injection of holes into emitter,† the collector current flows through a much larger impedance than the emitter current, ***whence the power gain.***

---

† In a homojunction transistor, injection of holes into the emitter is suppressed compared to the useful injection of electrons into the base only by the factor $\frac{n_{p0}}{p_{n0}} = \frac{N_D\,(\text{emitter})}{N_A\,(\text{base})}$.

This means that the base must be lightly doped compared to the emitter and hence base resistance is a concern.

The fundamental trade-off: thicker base for lower base resistance, thinner base for faster diffusion across the base.

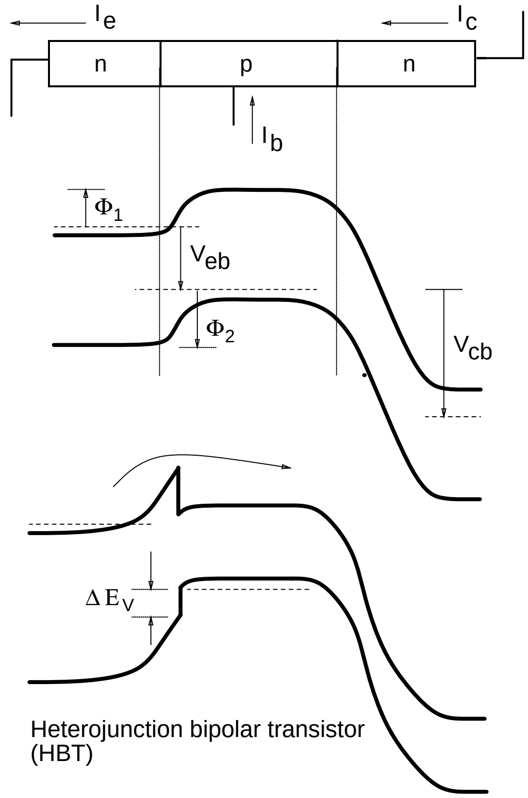


Figure 2: Homojunction and heterojunction *npn* transistor

In homojunction transistor at the base-emitter junction

$$J_E^{(e)} \propto e^{-\beta\Phi_1}$$
$$J_E^{(h)} \propto e^{-\beta\Phi_2}$$

$$\frac{J_E^{(h)}}{J_E^{(e)}} = \frac{N_A\,(\text{base})}{N_D\,(\text{emitter})}$$

Emitter efficiency $\eta \equiv \dfrac{J_E^{(e)}}{J_E^{(e)} + J_E^{(h)}} \approx 1 - \text{doping ratio}$.

In heterojunction transistor:

$$\frac{J_E^{(h)}}{J_E^{(e)}} = \frac{N_A\,(\text{base})}{N_D\,(\text{emitter})} \times e^{-\beta\Delta E_V}$$

Back to (homo)junction transistor.

We understand why the base doping must be much lower than emitter doping. Now why the collector doping must be much lower than emitter doping ?

$$N_D\,(\text{emitter}) \gg N_A\,(\text{base}) \gg N_D\,(\text{collector})$$

Threefold answer:

— For W_B to have little dependence on V_{cb} (need high output impedance)

— to lower base-collector capacitance C_{cb}

— to lower the field in base-collector junction (increase breakdown voltage)

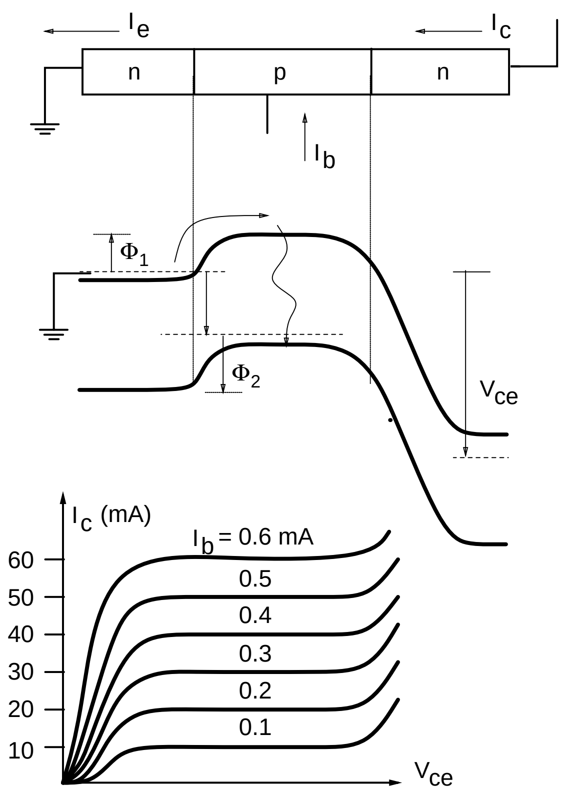


Figure 3: Common-emitter transistor characteristics. Base current is stepped up by increment of 0.1 mA and the emitter current increases by much larger amounts. Here current gain $\beta \approx 100$.

Base transport factor and current gain.

From the continuity equation:

$$n'' - \frac{n}{L_D^2} = 0$$

$$n(x) = A\,e^{\frac{x}{L_D}} + B\,e^{-\frac{x}{L_D}}$$

Boundary conditions at x and at W:

$$n(0) = \frac{n_i^2}{N_A}\,e^{\beta V_{eb}}$$

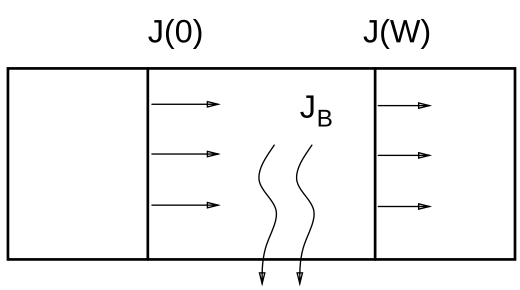
$$n(W) = \frac{n_i^2}{N_A}\,e^{\beta V_{cb}} \approx 0$$

☛ $$n(x) = A \sinh\left[(W-x)/L_D\right] = \frac{n(0)\sinh\left[(W-x)/L_D\right]}{\sinh(W/L_D)}$$

☛ $$J_n(x) = e\,D\,\frac{\partial n}{\partial x} = \frac{e\,D\,n(0)}{L_D}\,\frac{\cosh\left[(W-x)/L_D\right]}{\sinh(W/L_D)}$$

The base transport factor alpha:

$$\alpha \equiv \frac{J_n(W)}{J_n(0)} = \frac{1}{\cosh(W/L_D)} \approx 1 - \frac{W^2}{2L_D^2}$$



Current gain: $\beta = \frac{\partial I_C}{\partial I_B}$. By Kirchhoff's law, $\beta = \frac{\alpha}{1-\alpha}$

Combined with emitter efficiency η, we take instead of α the product $\alpha\eta$.

Frequency dependence of the current gain.

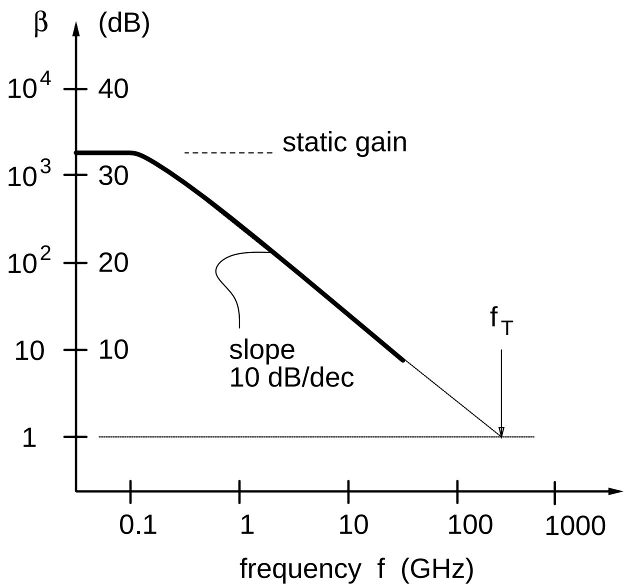


Figure 5: High-frequency gain. At high frequencies the current gain rolls-off at 10 dB/per decade (i.e. as $1/f$). This behavior is quite universal and has nothing to do with either recombination or emitter efficiency.

The characteristic cut-off frequency f_T is defined by the condition of unity current gain ($\beta \rightarrow 1$) and is mainly due to the propagation delay τ of minority carriers† through the base.

$$f_T = \frac{1}{2\pi\tau}$$

In general, $\tau = W/v$, where v is the average velocity of minority carrier propagation. For diffusive transport,

$$\tau_D = \frac{W^2}{2D} \ll \tau_C \quad .$$

† Do not confuse this τ with the minority carrier lifetime (which is typically much longer). Let us denote the minority-carrier lifetime by τ_C ("capture" time).

Let us derive an expression for $\alpha(f)$. Begin with the continuity equation:

$$\frac{\partial n}{\partial t} = D \frac{\partial^2 n}{\partial x^2} - \frac{n}{\tau_C}$$

Take

$$n(t) = n_0 + \delta n \, e^{i\omega t}$$

where $n_0 = n_0(x)$ is the static solution and $\delta n = \delta n(x)$ is the harmonic variation amplitude at frequency $\omega \equiv 2\pi f$.

Both the static equation for n_0 and the dynamic equation for δn are of similar form:

$$\frac{\partial^2 n_0}{\partial x^2} - L_D^{-2} n_0 = 0$$

$$\frac{\partial^2 \delta n}{\partial x^2} - L^{-2} \delta n = 0$$

where

$$L_D^2 \equiv D \tau_C$$

$$L^{-2} = L_D^{-2} (1 + i\omega\tau_C) \underset{\omega\tau_C \gg 1}{\approx} \frac{i\omega}{D}$$

The solutions to both equations are similar too (in form):

$$n_0(x) = A \sinh\left[\frac{W - x}{L_D}\right]$$

$$\delta n(x, \omega) = B \sinh\left[\frac{W - x}{L}\right]$$

whence we find (in analogy to $\alpha_0 = \dfrac{1}{\cosh(W/L_D)}$

$$\alpha(\omega) \equiv \frac{\partial J(W, \omega)}{\partial J(0, \omega)} = \frac{1}{\cosh(W/L)}$$

At sufficiently high ω's (nothing "spectacular", just $\omega\tau_C \gg 1$) we have

$$\frac{W}{L} = \quad \text{where} \quad \tau_D = \frac{W^2}{2D}$$

$$= \quad \text{remember} \quad = e^{i\pi/4} = \frac{1 + i}{\text{v'-1.075m'}}$$

$$\alpha(\omega) = \frac{1}{\cosh(W/L)} = \begin{cases} (1 + i\omega\tau_D)^{-1} & \text{for} \quad \omega\tau_D \ll 1 \\ 2\,e^{-\omega\tau_D}\, e^{-i\omega\tau_D} & \text{for} \quad \omega\tau_D \gtrsim 1 \end{cases}$$

Im modern transistors, typically, $W \lesssim 1{,}000\,A$ and $D \sim 50\ \text{cm}^2\ \text{s}^{-1}$. Therefore, at frequencies easily accessible to the measurement (up to, say 20 GHz) one has, typically, $\tau_D \lesssim 10^{-12}$ s and $\omega\tau_D \ll 1$.

$$\alpha(\omega) \approx \frac{1}{1 + i\,\omega\tau_D} \approx e^{-i\omega\tau_D}$$

$$\beta(\omega) = \frac{\alpha}{1-\alpha} = \frac{e^{-i\omega\tau_D}}{1 - e^{-i\omega\tau_D}} = \frac{-i\ e^{-i\omega\tau_D/2}}{2\sin(\omega\tau_D/2)}$$

$$|\beta(\omega)| = \frac{1}{2\ |\sin(\omega\tau_D/2)|} \approx \frac{1}{\omega\tau_D}$$

Uisng heterostructures, it is possible to design an HBT such that $\alpha(\omega)$ does not spiral in significantly, even for large ω's – remaing close to the circle $\exp(i\omega\tau)$ for phase angles $\phi = \omega\tau$ as large as $\phi = 2\pi$. Such ***coherent transistors***,† are capable of "life after death", showing current gain above f_T and power gain above f_{max}.

† A. A. Grinberg and S. Luryi, "Coherent transistor", *IEEE Trans. Electron Devices* **ED-40**, pp. 1512-1522 (1993).

S. Luryi, A. A. Grinberg, and V. B. Gorfinkel, "Heterostructure bipolar transistor with enhanced forward diffusion of minority carriers", *Appl. Phys. Lett.* **63**, pp. 1537-1539 (1993).

Base shrinkage and finite output conductance.

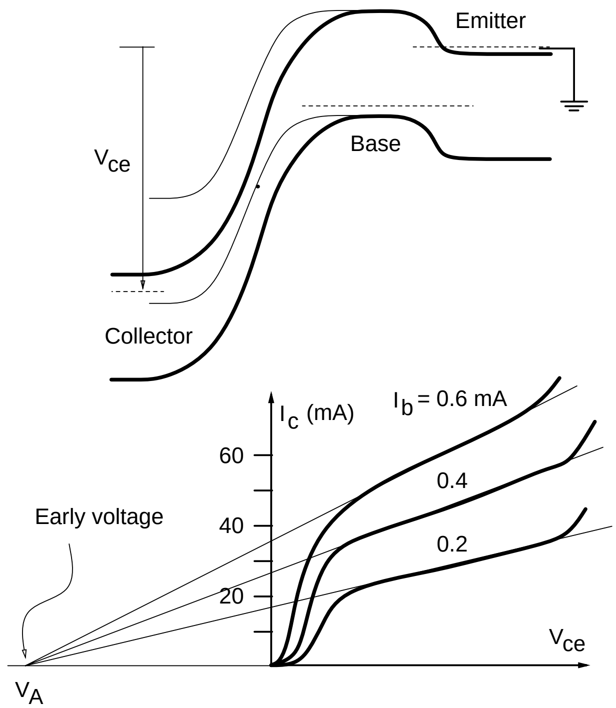


Figure: Base shrinkage and Early effect.

Neglecting recombination, the injected minority current must be constant in the base, $\vec{\nabla} \cdot \vec{J} = 0$. If the current is by diffusion only, then

$$\frac{\mathrm{d}n}{\mathrm{d}x} = \mathrm{const} = \frac{n(0) - n(W)}{W} \approx \frac{n_{p0}\, e^{\beta V_{\mathrm{eb}}}}{W}$$

W shrinks with the collector bias. Hence, at a fixed base current, one has an increasing collector current (Early effect)

$$I_{\mathrm{C}} \propto \left[1 + \frac{V_{\mathrm{cb}}}{V_{\mathrm{A}}} \right]$$

Finite output conductance is detrimental; we live better without it.

The Early voltage depends on the base width W and the Gummel number $n_{\mathrm{G}} = N_{\mathrm{A}}\, W$. For $W \gg$ than the BC junction depletion width, one has

$$V_{\mathrm{A}} \approx \frac{e\, N_{\mathrm{A}}\, W^2}{\varepsilon} \equiv \frac{e\, n_{\mathrm{G}}\, W}{\varepsilon}$$