Chapter 1:
Diode

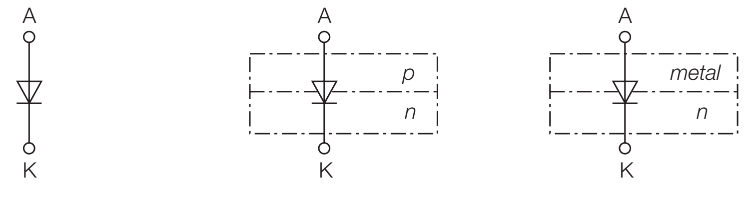
The diode is a semiconductor component with two connections, which are called the *anode* (A) and the *cathode* (K). Distinction has to be made between discrete diodes, which are intended for installation on printed circuit boards and are contained in an individual case, and integrated diodes, which are produced together with other semiconductor components on a common semiconductor carrier (*substrate*). Integrated diodes have a third connection resulting from the common carrier. It is called the *substrate* (S); it is of minor importance for electrical functions.

Construction: Diodes consist of a pn or a metal-n junction and are called pn or Schottky diodes, respectively. Figure 1.1 shows the graphic symbol and the construction of a diode. In pn diodes the p and the n regions usually consist of silicon. Some discrete diode types still use germanium and thus have a lower forward voltage, but they are considered obsolete. In Schottky diodes the p region is replaced by a metal region. This type also has a low forward voltage and is therefore used to replace germanium pn diodes.

In practice the term *diode* is used for the silicon pn diode; all other types are identified by supplements. Since the same graphic symbol is used for all types of diodes with the exception of some special diodes the various types of discrete diodes can be distinguished only by means of the type number printed on the component or the specifications in the data sheet.

Operating modes: A diode can be operated in the *forward, reverse* or *breakthrough mode*. In the following Section these operating regions are described in more detail.

Diodes that are used predominantly for the purpose of rectifying alternating voltages are called *rectifier diodes*; they operate alternately in the forward and reverse region. Diodes designed for the operation in the breakthrough region are called *Zener diodes* and are used for voltage stabilization. The *variable capacitance diodes* are another important type. They are operated in the reverse region and, due to the particularly strong response of the junction capacitance to voltage variations, are used for tuning the frequency in resonant circuits. In addition, there is a multitude of special diodes which are not covered here in detail.



Graphical symbol pn diode Schottky diode

Fig. 1.1. Graphical symbol and diode construction

1.1
Performance of the Diode

The performance of a diode is described most clearly by its characteristic curve. This shows the relation between current and voltage where all parameters are *static* which means that they do not change over time or only very slowly. In addition, formulas that describe the diode performance sufficiently accurately are required for mathematical calculations. In most cases simple equations can be used. In addition, there is a model that correctly reflects the *dynamic performance* when the diode is driven with sinusoidal or pulse-shaped signals. This model is described in Sect. 1.3 and knowledge of it is not essential to understand the fundamentals. The following Sections focus primarily on the performance of silicon pn diodes.

1.1.1
Characteristic Curve

Connecting a silicon pn diode to a voltage $V_D = V_{AK}$ and measuring the current I_D in a positive sense from A to K results in the characteristic curve shown in Fig. 1.2. It should be noted that the positive voltage range has been enhanced considerably for reasons of clarity. For $V_D > 0\,\mathrm{V}$ the diode operates in the forward mode, i.e. in the *conducting state*. In this region the current rises exponentially with an increasing voltage. When $V_D > 0.4\,\mathrm{V}$, a considerable current flows. If $-V_{BR} < V_D < 0\,\mathrm{V}$ the diode is in the reverse-biased state and only a negligible current flows. This region is called the *reverse region*. The *breakthrough voltage* V_{BR} depends on the diode and for rectifier diodes amounts to $V_{BR} = 50 \ldots 1000\,\mathrm{V}$. If $V_D < -V_{BR}$, the diode breaks through and a current flows again. Only Zener diodes are operated permanently in this *breakthrough region*; with all other diodes current flow with negative voltages is not desirable. With germanium and Schottky diodes a considerable current flows in the forward region even for $V_D > 0.2\,\mathrm{V}$, and the breakthrough voltage V_{BR} is $10 \ldots 200\,\mathrm{V}$.

In the forward region the voltage for typical currents remains almost constant due to the pronounced rise of the characteristic curve. This voltage is called the *forward voltage* V_F

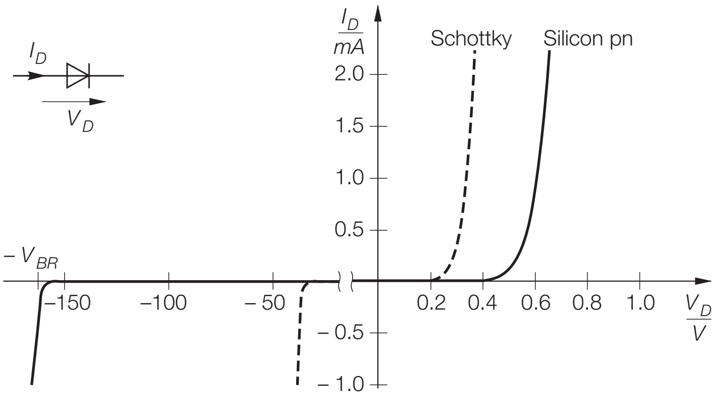


Fig. 1.2. Current-voltage characteristic of a small-signal diode

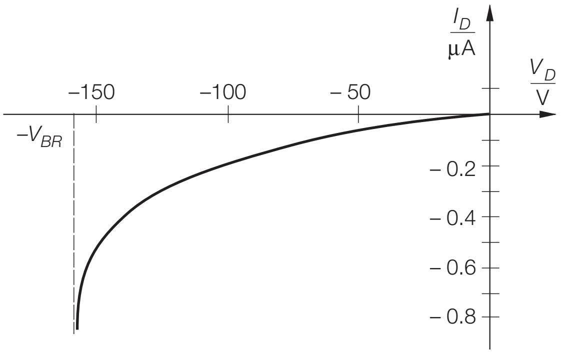


Fig. 1.3. Characteristic curve of a small-signal diode in the reverse region

and for both germanium and Schottky diodes lies at $V_{F,Ge} \approx V_{F,Schottky} \approx 0.3 \ldots 0.4\,\mathrm{V}$ and for silicon pn diodes at $V_{F,Si} \approx 0.6 \ldots 0.7\,\mathrm{V}$. With currents in the ampere range as used in power diodes the voltage may be significantly higher since in addition to the *internal* forward voltage a considerable voltage drop occurs across the spreading and connection resistances of the diode: $V_F = V_{F,i} + I_D R_B$. In the borderline case of $I_D \to \infty$ the diode acts like a very low resistance with $R_B \approx 0.01 \ldots 10\,\Omega$.

Figure 1.3 shows the enlarged reverse region. The *reverse current* $I_R = -I_D$ is very small with a low reverse voltage $V_R = -V_D$ and increases slowly when the voltage approaches the breakthrough voltage while it shoots up suddenly at the onset of the breakthrough.

1.1.2
Description by Equations

Plotting the characteristic curve for the region $V_D > 0$ in a semilogarithmic form results approximately in a straight line (see Fig. 1.4); this means that there is an exponential relation between I_D and V_D due to $\ln I_D \sim V_D$. The calculation on the basis of semiconductor physics leads to [1.1]:

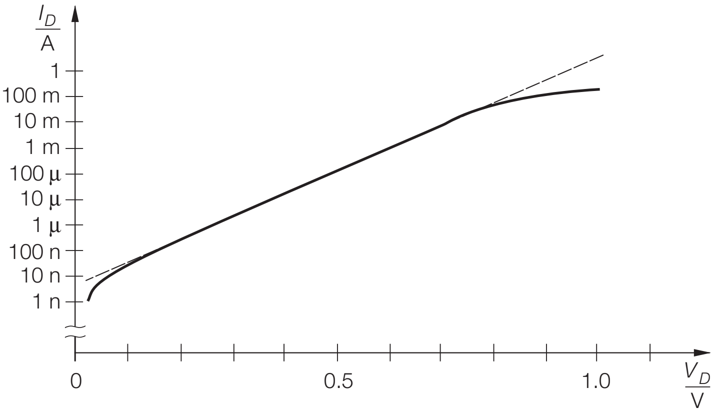


Fig. 1.4. Semilogarithmic representation of the characteristic curve for $V_D > 0$

$$I_D(V_D) = I_S\left(e^{\frac{V_D}{V_T}} - 1\right) \qquad \text{for } V_D \geq 0$$

For the correct description of a real diode a correction factor is required which enables the slope of the straight line in the semilogarithmic representation to be adapted [1.1]:

$$\boxed{I_D = I_S\left(e^{\frac{V_D}{nV_T}} - 1\right)} \tag{1.1}$$

Here, $I_S \approx 10^{-12} \dots 10^{-6}$ A is the *reverse saturation current*, $n \approx 1 \dots 2$ is the *emission coefficient* and $V_T = kT/q \approx 26\,\text{mV}$ is the *temperature voltage* at room temperature.

Even though (1.1) actually applies only to $V_D \geq 0$ it is sometimes used for $V_D < 0$. For $V_D \ll -nV_T$ this results in a constant current $I_D = -I_S$ which is generally much smaller than the current that is actually flowing. Therefore, only the qualitative statement that a small negative current flows in the reverse region is correct. The shape of the current curve as shown in Fig. 1.3 can only be described with the help of additional equations (see Sect. 1.3).

$V_D \gg nV_T \approx 26 \dots 52\,\text{mV}$ applies to the forward region and the approximation

$$I_D = I_S\, e^{\frac{V_D}{nV_T}} \tag{1.2}$$

can be used. Then the voltage is:

$$V_D = nV_T \ln \frac{I_D}{I_S} = nV_T \ln 10 \cdot \log \frac{I_D}{I_S} \approx 60 \dots 120\,\text{mV} \cdot \log \frac{I_D}{I_S}$$

This means that the voltage increases by $60 \dots 120\,\text{mV}$ when the current rises by a factor of 10. With high currents the voltage drop $I_D R_B$ at the spreading resistance R_B must be taken into account, which occurs in addition to the voltage at the pn junction:

$$V_D = nV_T \ln \frac{I_D}{I_S} + I_D R_B$$

In this case it cannot be described in the form $I_D = I_D(V_D)$.

For simple calculations the diode can be regarded as a switch that is opened in the reverse region and is closed in the forward region. Given the assumption that the voltage is approximately constant in the forward region and that no current flows in the reverse region, the diode can be replaced by an ideal voltage-controlled switch and a voltage source with the forward voltage V_F (see Fig.1.5a). Figure 1.5b shows the characteristic curve of this equivalent circuit which consists of two straight lines:

$$\begin{aligned} I_D &= 0 && \text{for } V_D < V_F && \rightarrow \text{switch open (a)} \\ V_D &= V_F && \text{for } I_D > 0 && \rightarrow \text{switch closed (b)} \end{aligned}$$

When the additional spreading resistance R_B is taken into consideration, we have:

$$I_D = \begin{cases} 0 & \text{for } V_D < V_F \quad \rightarrow \text{switch open (a)} \\ \dfrac{V_D - V_F}{R_B} & \text{for } V_D \geq V_F \quad \rightarrow \text{switch closed (b)} \end{cases}$$

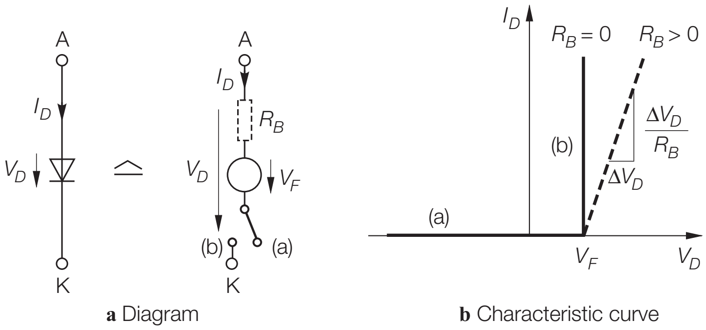


Fig. 1.5. Simple equivalent circuit diagram for a diode without (-) and with (- -) spreading resistance

The voltage V_F is $V_F \approx 0.6\,\text{V}$ for silicon pn diodes and $V_F \approx 0.3\,\text{V}$ for Schottky diodes. The corresponding circuit diagram and characteristic curve are shown in Fig. 1.5 as dashed lines. Different cases must be distinguished for both variations, that is, it is necessary to calculate with the switch open *and* closed and to determine the situation in which there is no contradiction. The advantage is that either case leads to linear equations which are easy to solve. In contrast, when using the e function according to (1.1), it is necessary to cope with an implicit nonlinear equation that can only be solved numerically.

Example: Figure 1.6 shows a diode in a bridge circuit. To calculate the voltages V_1 and V_2 and the diode voltage $V_D = V_1 - V_2$ it is assumed that the diode is in the reverse state, that is, $V_D < V_F = 0.6\,\text{V}$ and the switch in the equivalent circuit is open. In this case, V_1 and V_2 can be determined by the voltage divider formula $V_1 = V_b R_2/(R_1 + R_2) = 3.75\,\text{V}$ and $\text{V}_2 = V_b R_4/(R_3 + R_4) = 2.5\,\text{V}$. This results in $V_D = 1.25\,\text{V}$, which does not comply with the assumption. Consequently the diode is conductive and the switch in the equivalent circuit is closed; this leads to $V_D = V_F = 0.6\,\text{V}$ and $I_D > 0$. From the nodal equations

$$\frac{V_1}{R_2} + I_D = \frac{V_b - V_1}{R_1} \quad , \quad \frac{V_2}{R_4} = I_D + \frac{V_b - V_2}{R_3}$$

it is possible to eliminate the unknown elements I_D and V_1 by adding the equations and inserting $V_1 = V_2 + V_F$; this leads to:

$$V_2 \left(\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} + \frac{1}{R_4} \right) = V_b \left(\frac{1}{R_1} + \frac{1}{R_3} \right) - V_F \left(\frac{1}{R_1} + \frac{1}{R_2} \right)$$

This results in $V_2 = 2.76\,\text{V}$, $V_1 = V_2 + V_F = 3.36\,\text{V}$ and in $I_D = 0.52\,\text{mA}$ by substitution in one of the nodal equations. The initial condition $I_D > 0$ has been fulfilled, that is, there is no contradiction and the solution has been found.

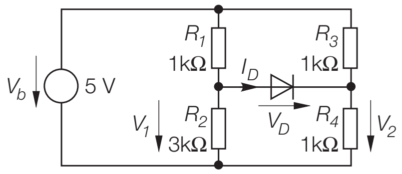


Fig. 1.6. Example for the demonstration of the use of the equivalent circuit of Fig. 1.5

1.1.3
Switching Performance

In many applications the diodes operate alternately in the forward mode and in the reverse mode, for example when rectifying alternating currents. The transition does not follow the static characteristic curve as the parasitic capacitance of the diode stores a charge that builds up in the forward state and is discharged in the reverse state. Figure 1.7 shows a circuit for determining the *switching performance* with an ohmic load ($L = 0$) or an ohmic-inductive load ($L > 0$). Applying a square wave produces the transitions shown in Fig. 1.8.

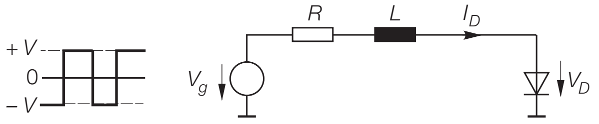


Fig. 1.7. Circuit for determining the switching performance

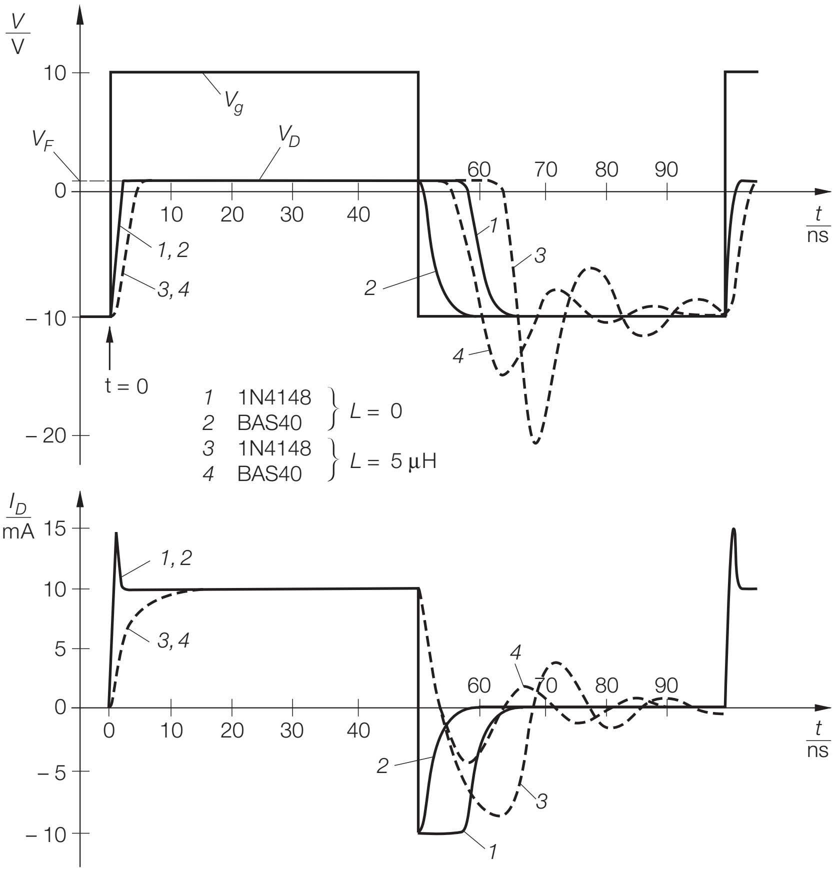
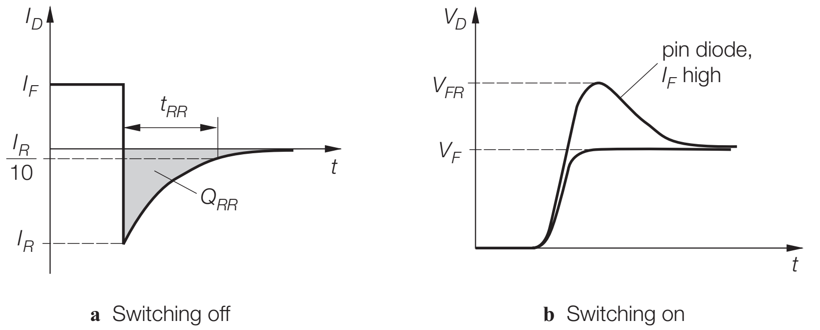


Fig. 1.8. Switching performance of the silicon diode 1N4148 and the Schottky diode BAS40 in the measuring circuit of Fig. 1.7 with $V = 10\,\text{V}$, $f = 10\,\text{MHz}$, $R = 1\,\text{k}\Omega$ and $L = 0$ or $L = 5\,\mu\text{H}$



a Switching off **b** Switching on

Fig. 1.9. Illustration of switching performance

Switching performance with ohmic load: With an ohmic load ($L = 0$) a current peak caused by the charge built up in the capacitance of the diode occurs when the circuit is activated. The voltage rises during this current peak from the previously existing reverse voltage to the forward voltage V_F which terminates the switch-on process. In pin diodes[1] higher currents may cause a voltage overshoot (see Fig. 1.9b) as these diodes initially have a higher spreading resistance R_B at the switch-on point. Subsequently the voltage declines to the static value in accordance with the decrease of R_B. When switching off there is a current in the opposite direction until the capacitance is discharged; then the current returns to zero and the voltage drops to the reverse voltage. Since the capacitance of Schottky diodes is much lower than that of silicon diodes of the same size, their turn-off time is significantly shorter (see Fig. 1.8). Therefore, Schottky diodes are preferred for rectifier diodes in switched power supplies with high cycle rates ($f > 20\,\text{kHz}$), while the lower priced silicon diodes are used in rectifiers for the mains voltage ($f = 50\,\text{Hz}$). When the frequency becomes so high that the capacitance discharge process is not completed before the next conducting state starts, the rectification no longer takes place.

Switching performance with ohmic-inductive load: With an ohmic-inductive load ($L > 0$) the transition to the conductive state takes longer since the increase in current is limited by the inductivity; no current peaks occur. While the voltage rises relatively fast to the forward voltage, the current increases with the time constant $T = L/R$ of the load. During switch-off the current first decreases with the time constant of the load until the diode cuts off. Then, the load and the capacitance of the diode form a series resonant circuit, and the current and the voltage perform damped oscillations. As shown in Fig. 1.8 high reverse voltages may arise which are much higher than the static reverse voltage and consequently require a high diode breakthrough voltage.

Figure 1.9 shows the typical data for *reverse recovery (RR)* and *forward recovery (FR)*. The *reverse recovery time* t_{RR} is the period measured from the moment at which the current passes through zero until the moment at which the reverse current drops to 10 %[2] of its maximum value I_R. Typical values range from $t_{RR} < 100\,\text{ps}$ for fast Schottky diodes to $t_{RR} = 1 \ldots 20\,\text{ns}$ for small-signal silicon diodes or $t_{RR} > 1\,\mu\text{s}$ for rectifier diodes. The *reverse recovery charge* Q_{RR} transported during the capacitance discharge corresponds to

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[1] pin diodes have a nondoped (*intrinsic*) or slightly doped layer between the p and n layers in order to achieve a higher breakthrough voltage.

[2] With rectifier diodes the measurement is sometimes taken at 25 %.

the area below the x axis (see Fig. 1.9a). Both parameters depend on the previously flowing forward current I_F and the cutoff speed; therefore the data sheets show either information on the measuring conditions or the measuring circuit. An approximation is $Q_{RR} \sim I_F$ and $Q_{RR} \sim |I_R| t_{RR}$ [1.2]; this means that in a first approximation the reverse recovery time is proportional to the ratio of the forward and reverse current: $t_{RR} \sim I_F / |I_R|$. However, this approximation only applies to $|I_R| < 3 \ldots 5 \cdot I_F$, in other words, t_{RR} can not be reduced endlessly. In pin diodes featuring a high breakdown voltage, the high cutoff speed may even cause the breakdown to occur far below the static breakdown voltage V_{BR} if the reverse voltage at the diode increases sharply before the weakly doped i-layer is free of charge carriers. With the transition to the forward state the *forward recovery voltage* V_{FR} occurs, which also depends on the actual switching conditions [1.3]; data sheets quote a maximum value for V_{FR}, typically $V_{FR} = 1 \ldots 2.5\,\mathrm{V}$.

1.1.4
Small-Signal Response

The performance of the diode when controlled by *small* signals around an operating point characterized by $V_{D,A}$ and $I_{D,A}$ is called the *small-signal response*. In this case, the nonlinear characteristic given in (1.1) can be replaced by a tangent to the operating point; with the small-signal parameters

$$i_D = I_D - I_{D,A} \quad , \quad v_D = V_D - V_{D,A}$$

one arrives at:

$$i_D = \left. \frac{dI_D}{dV_D} \right|_A v_D = \frac{1}{r_D} v_D$$

From this the *differential resistance* r_D of the diode is derived:

$$\boxed{r_D = \left. \frac{dV_D}{dI_D} \right|_A = \frac{nV_T}{I_{D,A} + I_S} \overset{I_{D,A} \gg I_S}{\approx} \frac{nV_T}{I_{D,A}}} \qquad (1.3)$$

Thus, the equivalent small-signal circuit for the diode consists of a resistance with the value r_D; with large currents r_D becomes very small and an additional spreading resistance R_B must be introduced (see Fig. 1.10).

The equivalent circuit shown in Fig. 1.10 is only suitable for calculating the small-signal response at low frequencies ($0 \ldots 10\,\mathrm{kHz}$); therefore, it is called the *DC small-signal equivalent circuit*. For higher frequencies it is necessary to use the AC small-signal equivalent circuit given in Sect. 1.3.3.

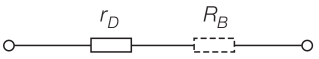


Fig. 1.10. Small-signal equivalent circuit of a diode

1.1.5
Limit Values and Reverse Currents

The data sheet for a diode shows limit values that must not be exceeded. These are the limit voltages, limit currents and maximum power dissipation. In order to deal with positive values for the limit data the reference arrows for the current and the voltage are reversed in their direction for reverse-biased operation and the relevant values are given with the index R (*reverse*); the index F (*forward*) is used for forward-biased operation.

Limit Voltages

Reaching the *breakthrough voltage* $V_{(BR)}$ or V_{BR} causes the diode to break through in the reverse mode and the reverse current rises sharply. Since the current already increases markedly when approaching the breakthrough voltage, as shown in Fig. 1.3, a *maximum reverse voltage* $V_{R,max}$ is specified up to which the reverse current remains below a limit value in the µA range. Higher reverse voltages are permissible when driving the diode with a pulse chain or a single pulse; they are called the *repetitive peak reverse voltage* V_{RRM} and the *peak surge reverse voltage* V_{RSM}, respectively, and they are chosen so that the diode remains undamaged. The pulse frequency is considered to be $f = 50\,\mathrm{Hz}$ since it is assumed that it will be used as a mains rectifier. Due to the reversed direction of the reference arrow all voltages are positive and are related in the following way:

$$V_{R,max} < V_{RRM} < V_{RSM} < V_{(BR)}$$

Limit Currents

For forward-biased operation a *maximum steady-state forward current* $I_{F,max}$ is specified. It applies to situations in which the diode case is kept at a temperature of $T = 25\,°\mathrm{C}$; at higher temperatures the permissible steady-state current is lower. Higher forward currents are permissible when driving the diode with several pulses or a single pulse; they are called the *repetitive peak forward current* I_{FRM} and the *peak surge forward current* I_{FSM}, respectively, and they depend on the duty cycle or the pulse duration. The currents are related:

$$I_{F,max} < I_{FRM} < I_{FSM}$$

With very short single pulses $I_{FSM} \approx 4 \ldots 20 \cdot I_{F,max}$. The current I_{FRM} is of particular importance for rectifier diodes because of their pulsating periodic current (see Sect. 16.2); in this case the maximum value is much higher than the mean value.

For the breakthrough region a *maximum current-time area* I^2t is quoted which may occur at the breakthrough caused by a pulse:

$$I^2t = \int I_R^2 dt$$

Despite its unit A^2s it is often referred to as the *maximum pulse energy*.

Reverse Current

The *reverse current* I_R is measured at a reverse voltage below the breakthrough voltage and depends largely on the reverse voltage and the temperature of the diode. At room temperature $I_R = 0.01 \ldots 1\,\mu\mathrm{A}$ for a small-signal silicon diode, $I_R = 1 \ldots 10\,\mu\mathrm{A}$ for a small-signal Schottky diode and a silicon rectifier diode in the Ampere range and $I_R > 10\,\mu A$ for a Schottky rectifier diode; at a temperature of $T = 150\,°C$ these values are increased by a factor of 20 . . . 200.

Maximum Power Dissipation

The power dissipation of the diode is the power converted to heat:

$$P_V = V_D I_D$$

This occurs at the junction or, with large currents, at the spreading resistance R_B. The temperature of the diode increases up to a value at which, due to the temperature gradients, the heat can be dissipated from the junction through the case to the environment. Section 2.1.6 describes this in more detail for bipolar transistors; the same results apply to the diode when P_V is replaced by the power dissipation of the diode. Data sheets specify the *maximum power dissipation* P_{tot} for the situation in which the diode case is kept at a temperature of $T = 25\,°C$; P_{tot} is lower at higher temperatures.

1.1.6
Thermal Performance

The thermal performance of components is described in Sect. 2.1.6 for bipolar transistors; the parameters and conditions described there also apply to the diode when P_V is replaced by the power dissipation of the diode.

1.1.7
Temperature Sensitivity of Diode Parameters

The characteristic curve of a diode is heavily dependent on the temperature; an explicit statement of the temperature sensitivity means for the silicon pn diode [1.1]

$$I_D(V_D, T) = I_S(T) \left(e^{\frac{V_D}{nV_T(T)}} - 1 \right)$$

with:

$$V_T(T) = \frac{kT}{q} = 86.142 \frac{\mu V}{K}\, T \overset{T=300\,K}{\approx} 26\,mV$$

$$I_S(T) = I_S(T_0)\, e^{\left(\frac{T}{T_0}-1\right)\frac{V_G(T)}{nV_T(T)}} \left(\frac{T}{T_0}\right)^{\frac{x_{T,I}}{n}} \quad \text{with } x_{T,I} \approx 3 \tag{1.4}$$

Here, $k = 1.38 \cdot 10^{-23}$ VAs/K is *Boltzmann's constant*, $q = 1.602 \cdot 10^{-19}$ As is the *elementary charge* and $V_G = 1.12\,V$ is the *gap voltage* of silicon; the low temperature sensitivity of V_G may be ignored. The temperature T_0 with the respective current $I_S(T_0)$ serves as a reference point; usually $T_0 = 300\,K$ is used.

In reverse mode the reverse current $I_R = -I_D \approx I_s$ flows; with $x_{T,I} = 3$ this yields the temperature coefficient of the reverse current:

$$\frac{1}{I_R}\frac{dI_R}{dT} \approx \frac{1}{I_S}\frac{dI_S}{dT} = \frac{1}{nT}\left(3 + \frac{V_G}{V_T}\right)$$

In this region $n \approx 2$ applies to most diodes, resulting in:

$$\frac{1}{I_R}\frac{dI_R}{dT} \approx \frac{1}{2T}\left(3 + \frac{V_G}{V_T}\right) \overset{T=300\,\mathrm{K}}{\approx} 0.08\,\mathrm{K}^{-1}$$

This means that the reverse current doubles with a temperature increase of 9 K and rises by a factor of 10 with a temperature increase of 30 K. In practice there are often lower temperature coefficients; this is caused by surface and leakage currents which are often higher than the reverse current of the pn junction and have a different temperature response.

The temperature coefficient of the current at constant voltage in forward-bias operation is calculated by differentiation of $I_D(V_D, T)$:

$$\left.\frac{1}{I_D}\frac{dI_D}{dT}\right|_{V_D=\mathrm{const.}} = \frac{1}{nT}\left(3 + \frac{V_G - V_D}{V_T}\right) \overset{T=300\,\mathrm{K}}{\approx} 0.04\ldots0.08\,\mathrm{K}^{-1}$$

By means of the total differential

$$dI_D = \frac{\partial I_D}{\partial V_D}\,dV_D + \frac{\partial I_D}{\partial T}\,dT = 0$$

the temperature-induced change of V_D at constant current can be determined:

$$\boxed{\left.\frac{dV_D}{dT}\right|_{I_D=\mathrm{const.}} = \frac{V_D - V_G - 3V_T}{T} \overset{\substack{T=300\,\mathrm{K}\\ V_D=0.7\,\mathrm{V}}}{\approx} -1.7\,\frac{\mathrm{mV}}{\mathrm{K}}} \tag{1.5}$$

This means that the forward voltage decreases when the temperature rises; a temperature increase of 60 K causes a drop in V_D of approximately 100 mV. This effect is used in integrated circuits for measuring the temperature.

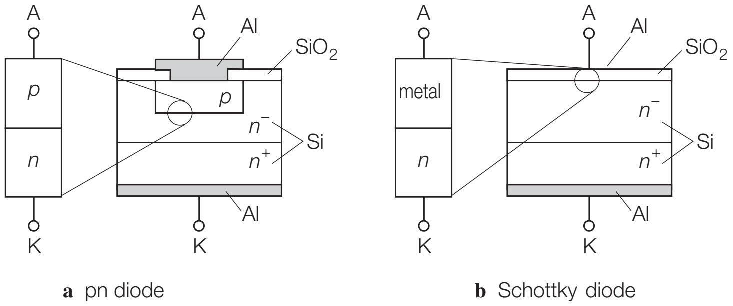
These results also apply to Schottky diodes when setting $x_{T,I} \approx 2$ and replacing the gap voltage V_G by the voltage that describes the energy difference between the n and metal regions: $V_{Mn} = (W_{Metal} - W_{n\text{-}Si})/q$; thus $V_{Mn} \approx 0.7\ldots0.8\,\mathrm{V}$ [1.1].

1.2
Construction of a Diode

Diodes are manufactured in a multi-step process on a semiconductor *wafer* that is then cut into small *dies*. On one chip there is either a discrete diode or an *integrated circuit* (*IC*), comprising several components.

1.2.1
Discrete Diode

Internal design: Discrete diodes are mostly produced using epitaxial-planar technology. Figure 1.11 illustrates the construction of a pn and a Schottky diode where the active areas are particularly emphasized. Doping is heavy in the n^+ layer, medium in the p layer and low in the n^- layer. The special arrangement of differently doped layers helps to minimize the spreading resistance and to increase the breakthrough voltage. Almost all pn diodes are designed as *pin diodes*, in other words, they feature a middle layer with little or no doping



a pn diode **b** Schottky diode

Fig. 1.11. Construction of a semiconductor chip with one diode

and with a thickness that is roughly proportional to the breakthrough voltage; in Fig. 1.11a this is the n^- layer. For practical purposes diodes are referred to as *pin diodes* only if the lifetime of the charge carriers in the middle layer is very high, thus producing a particular characteristic; this will be described in more detail in Sect. 1.4.2. In Schottky diodes the weakly doped n^- layer is required for the Schottky contact (see Fig. 1.11b); in contrast a junction between metal and a layer of medium or heavy doping produces an inferior diode effect or no effect at all, in which case it behaves rather like a resistor (*ohmic contact*).

Case: To mount a diode in a case the bottom side is soldered to the cathode terminal or connected to a metal part of the case. The anode side is connected to the anode terminal via a fine gold or aluminum *bond wire*. Finally the diode is sealed in a plastic compound or mounted in a metal case with screw connector.

For the various diode sizes and applications there is a multitude of case designs that differ in the maximum heat dissipation capacity or are adapted to special geometrical requirements. Figure 1.12 shows a selection of common models. Power diodes are provided

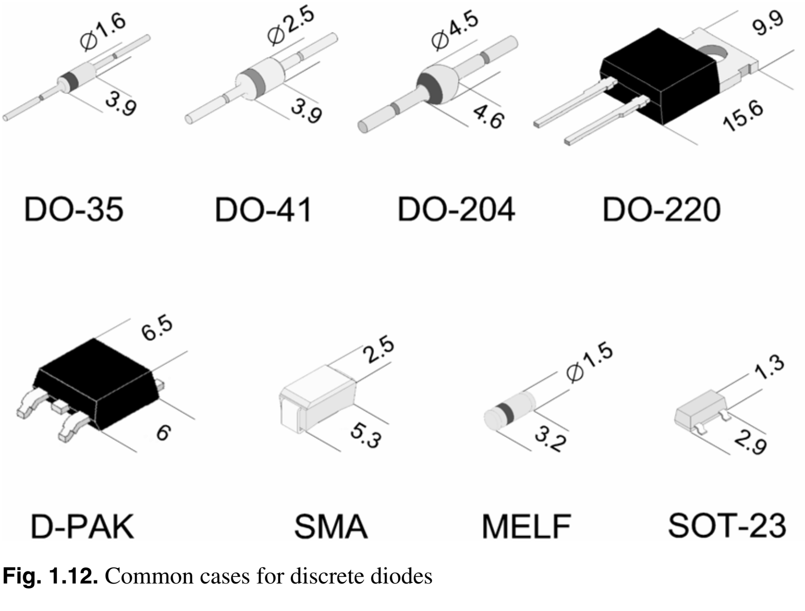


Fig. 1.12. Common cases for discrete diodes

with a heat sink for their installation; the larger the contact surface, the better the heat dissipation. Rectifier diodes are often designed as *bridge rectifiers* consisting of four diodes to serve as full-wave rectifiers in power supply units (see Sect. 1.4.4); the *mixer* described in Sect. 1.4.5 is also made of four diodes. High-frequency diodes require special cases because in the GHz frequency range their electrical performance depends on the case geometry. Often, the case is omitted altogether and the diode chip is soldered or bonded directly to the circuit.

1.2.2
Integrated Diode

Integrated diodes are also produced using epitaxial-planar technology. Here, all connections are located at the top of the chip and the diode is electrically isolated from other components by a reverse-biased pn junction. The active region is located in a very thin layer at the surface. The depth of the chip is called the *substrate* (*S*) and forms a common connection for all components of the integrated circuit.

Internal construction: Figure 1.13 illustrates the design of an integrated pn diode. The current flows from the p layer through the pn junction to the n^- layer and from there via the n^+ layer to the cathode; a low spreading resistance is achieved by means of the heavily doped n^+ layer.

Substrate diode: The equivalent circuit diagram in Fig. 1.13 shows an additional substrate diode located between the cathode and the substrate. The substrate is connected to the negative supply voltage so that this diode is always in the reverse mode to act as isolation relative to other components and the substrate.

Differences between integrated pn and Schottky diodes: In principle an integrated Schottky diode can be built like an integrated pn diode by simply omitting the p junction at the anode connection. However, for practical applications this is not so easy as different metals must be used for the Schottky diodes and for the component wiring, and in most manufacturing processes for integrated circuits the necessary steps are not intended.

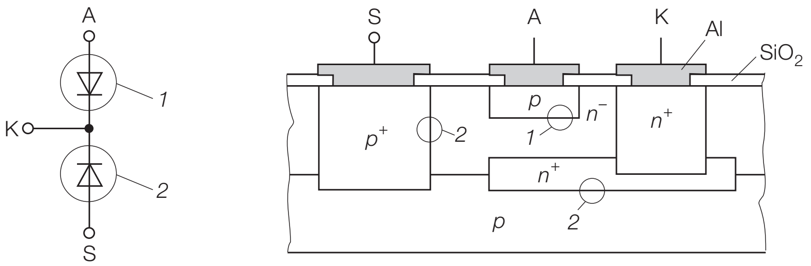


Fig. 1.13. Equivalent circuit and construction of an integrated pn diode with useful diode (1) and parasitic substrate diode (2)

1.3
Model of a Diode

Section 1.1.2 describes the static performance of the diode using an exponential function; but this neglects the breakthrough and the second-order effects in the forward operation. For computer-aided circuit design a model is required that considers all of these effects and, in addition, correctly reflects the *dynamic* performance. The *dynamic small-signal model* is derived from this *large-signal model* by linearization.

1.3.1
Static Performance

The description is based on the ideal diode equation given in (1.1) and also takes other effects into account. A standardized diode model like the the Gummel-Poon model for bipolar transistors does not exist; some of the CAD programs therefore have to use several diode models to describe a real diode with all of its current components. The diode model is almost unnecessary for the design of integrated circuits since here the base-emitter diode of a bipolar transistor is usually used as a diode.

Range of Medium Forward Currents

In pn diodes the *diffusion current* I_{DD} dominates in the range of medium forward currents; this follows from the ideal diode theory and can be described according to (1.1):

$$I_{DD} = I_S \left(e^{\frac{V_D}{nV_T}} - 1 \right) \tag{1.6}$$

The model parameters are the *saturation reverse current* I_S and the *emission coefficient* n. For the ideal diode $n = 1$; for real diodes $n \approx 1 \ldots 2$. This range is called the *diffusion range*.

In Schottky diodes the emission current takes the place of the diffusion current. But since both current conducting mechanisms lead to the same characteristic curve (1.6) can also be used for Schottky diodes [1.1, 1.3].

Other Effects

With very small and very high forward currents as well as in reverse operation there are deviations from the *ideal* performance according to (1.6):

- High forward currents produce the *high-current effect,* which is caused by a sharp rise in the charge carrier concentration at the edge of the depletion layer [1.1]; this is also referred to as a *strong injection*. This also affects the diffusion current and is described by an extension to (1.6).
- Because of the recombination of charge carriers in the depletion layer a *leakage* or *recombination current* I_{DR} occurs in addition to the diffusion current which is described by a separate equation [1.1].
- The application of high reverse voltages causes the diode to break through. The *breakthrough current* I_{DBR} is also described in an additional equation.

The current I_D thus comprises three partial currents:

$$I_D \;=\; I_{DD} + I_{DR} + I_{DBR} \tag{1.7}$$

High-current effect: The high-current effect causes the emission coefficient to rise from n in the medium current range to $2n$ for $I_D \to \infty$; it can be described by an extension to (1.6) [1.4]:

$$I_{DD} \;=\; \frac{I_S \left(e^{\frac{V_D}{nV_T}} - 1 \right)}{\sqrt{1 + \dfrac{I_S}{I_K} \left(e^{\frac{V_D}{nV_T}} - 1 \right)}} \;\approx\; \begin{cases} I_S \, e^{\frac{V_D}{nV_T}} & \text{for } I_S \, e^{\frac{V_D}{nV_T}} < I_K \\ \sqrt{I_S I_K} \, e^{\frac{V_D}{2nV_T}} & \text{for } I_S \, e^{\frac{V_D}{nV_T}} > I_K \end{cases} \tag{1.8}$$

An additional parameter is the *knee-point current* I_K, which marks the beginning of the *high-current region.*

Leakage current: Based on the ideal diode theory the following is applicable to the leakage current [1.1]:

$$I_{DR} \;=\; I_{S,R} \left(e^{\frac{V_D}{n_R V_T}} - 1 \right)$$

This equation only describes the recombination current accurately enough for forward operation. Setting $V_D \to -\infty$ yields a constant current $I_{DR} = -I_{S,R}$ in the reverse region, while in a real diode the recombination current rises with an increasing reverse voltage. A more accurate description is achieved by taking into account the voltage sensitivity of the width of the depletion layer [1.4]:

$$I_{DR} \;=\; I_{S,R} \left(e^{\frac{V_D}{n_R V_T}} - 1 \right) \left(\left(1 - \frac{V_D}{V_{Diff}} \right)^2 + 0.005 \right)^{\frac{m_J}{2}} \tag{1.9}$$

Additional parameters are the *leakage saturation reverse current* $I_{S,R}$, the *emission coefficient* $n_R \geq 2$, the *diffusion voltage* $V_{Diff} \approx 0.5 \ldots 1\,\text{V}$ and the *capacitance coefficient* $m_J \approx 1/3 \ldots 1/2$.[3] From (1.9) it follows that:

$$I_{DR} \;\approx\; -\, I_{S,R} \left(\frac{|V_D|}{V_{Diff}} \right)^{m_J} \qquad \text{for } V_D < -\, V_{Diff}$$

The magnitude of the current rises as the reverse voltage increases; its actual curve depends on the capacitance coefficient m_J. In the forward mode the additional factor given in (1.9) has almost no effect since in this case the exponential dependence of V_D is dominant.

Since $I_{S,R} \gg I_S$, the recombination current is larger than the diffusion current at low positive voltages; this region is called the *recombination region.* For

$$V_{D,RD} \;=\; V_T \, \frac{n n_R}{n_R - n} \, \ln \frac{I_{S,R}}{I_S}$$

both currents have the same value. With larger voltages the diffusion current becomes dominant and the diode operates in the diffusion region.

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[3] V_{Diff} and m_J are primarily used to describe the depletion layer capacitance of the diode (see Sect. 1.3.2).

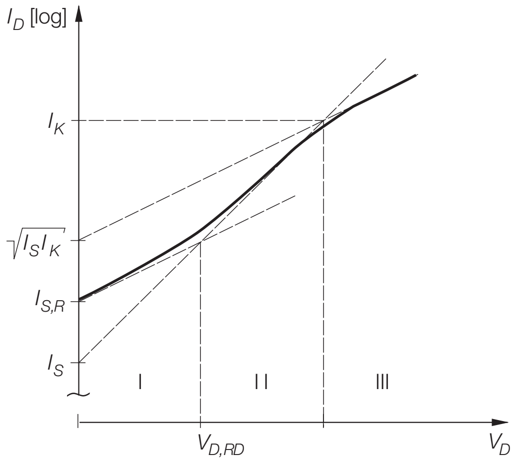

Fig. 1.14. Semi-logarithmic diagram of I_D in forward mode: (I) recombination, (II) diffusion, (III) high-current regions

Figure 1.14 is the semilogarithmic presentation of I_D in the forward region and shows the importance of parameters I_S, $I_{S,R}$ and I_K. In some diodes the emission coefficients n and n_R are almost identical. In such cases the semilogarithmic characteristic curve has the same slope in the recombination and diffusion regions and can be described for both regions using *one* exponential function.[4]

Breakthrough: For $V_D < -V_{BR}$ the diode breaks through; the flowing current can be approximated by an exponential function [1.5]:

$$I_{DBR} \; = \; - \, I_{BR} \, e^{-\frac{V_D + V_{BR}}{n_{BR} V_T}} \tag{1.10}$$

For this, the *breakthrough voltage* $V_{BR} \approx 50 \ldots 1000\,\mathrm{V}$, the *breakthrough knee-point current* I_{BR} and the *breakthrough emission coefficient* $n_{BR} \approx 1$ are required. For $n_{BR} = 1$ and $V_T \approx 26\,\mathrm{mV}$ the current is:[5]

$$I_D \; \approx \; I_{DBR} \; = \; \begin{cases} -\, I_{BR} & \text{for } V_D = -\, V_{BR} \\ -\, 10^{10} I_{BR} & \text{for } V_D = -\, V_{BR} - 0.6\,\mathrm{V} \end{cases}$$

Quoting I_{BR} and V_{BR} is not a clear definition since the same curve can be described with different value sets (V_{BR}, I_{BR}); therefore, the model for a certain diode may have different parameters.

Spreading Resistance

The spreading resistance R_B is necessary for the full description of the static performance; according to Fig. 1.15 it is comprized of the resistances of the various layers and it is represented in the model by a series resistor. A distinction has to be made between the *internal diode voltage* V_D' and the *external diode voltage*

$$V_D = V_D' + I_D R_B \,. \tag{1.11}$$

In the equations for I_{DD}, I_{DR} and I_{DBR} voltage V_D must be replaced by V_D'. The spreading resistance is between 0.01 Ω for power diodes and 10 Ω for small-signal diodes.

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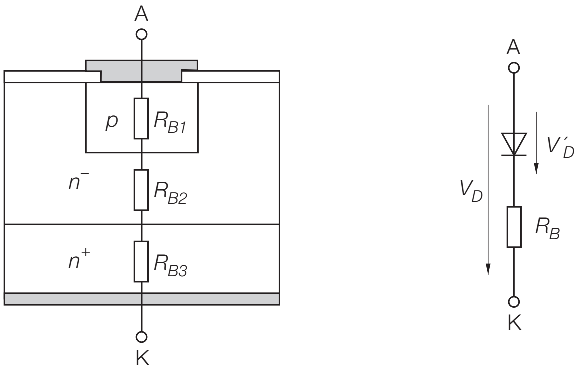
[4] Figure 1.4 shows the characteristic curve of such a diode.

[5] Based on $10 V_T \ln 10 = 0.6\,\mathrm{V}$.



a In the diode **b** In the model

Fig. 1.15. Spreading resistance of a diode

1.3.2
Dynamic Performance

The response to pulsating or sinusoidal signals is called the *dynamic performance*, and it cannot be derived from the characteristic curves. The reasons for this are the nonlinear *junction capacitance* of the pn or metal-semiconductor junction and the *diffusion charge* that is stored in the pn junction and determined by the *diffusion capacitance*, which is also nonlinear.

Junction Capacitance

A pn or metal–semiconductor junction has a voltage-dependent *junction capacitance* C_J that is influenced by the doping of the adjacent layers, the doping profile, the area of the junction and the applied voltage V'_D. The junction can be visualized as a plate capacitor with the capacitance $C = \epsilon A/d$; where A represents the junction area and d the junction width. A simplified view of the pn junction gives $d(V) \sim (1 - V/V_{Diff})^{m_J}$ [1.1] and thus:

$$C_J(V'_D) = \frac{C_{J0}}{\left(1 - \dfrac{V'_D}{V_{Diff}}\right)^{m_J}} \qquad \text{for } V'_D < V_{Diff} \tag{1.12}$$

The parameters are the *zero capacitance* $C_{J0} = C_J(V'_D = 0)$, the *diffusion voltage* $V_{Diff} \approx 0.5 \ldots 1\,\text{V}$ and the *capacitance coefficient* $m_J \approx 1/3 \ldots 1/2$ [1.2].

For $V'_D \to V_{Diff}$ the assumptions leading to (1.12) are no longer met. Therefore, the curve for $V'_D > f_C V_{Diff}$ is replaced by a straight line [1.5]:

$$C_J(V'_D) = C_{J0} \begin{cases} \dfrac{1}{\left(1 - \dfrac{V'_D}{V_{Diff}}\right)^{m_J}} & \text{for } V'_D \le f_C V_{Diff} \\[2ex] \dfrac{1 - f_C\,(1 + m_J) + \dfrac{m_J V'_D}{V_{Diff}}}{(1 - f_C)^{(1+m_J)}} & \text{for } V'_D > f_C V_{Diff} \end{cases} \tag{1.13}$$

where $f_C \approx 0.4 \ldots 0.7$. Figure 2.32 on page 70 shows the curve of C_J for $m_J = 1/2$ and $m_J = 1/3$.

Diffusion Capacitance

In forward operation the pn junction contains a stored diffusion charge Q_D that is proportional to the diffusion current flowing through the pn junction [1.2]:

$$Q_D = \tau_T I_{DD}$$

The parameter τ_T is the *transit time*. Differentiation of (1.8) produces the *diffusion capacitance*:

$$C_{D,D}(V_D') = \frac{dQ_D}{dV_D'} = \frac{\tau_T I_{DD}}{nV_T} \, \frac{1 + \dfrac{I_S}{2I_K} \, e^{\frac{V_D'}{nV_T}}}{1 + \dfrac{I_S}{I_K} \, e^{\frac{V_D'}{nV_T}}} \tag{1.14}$$

For the diffusion region $I_{DD} \gg I_{DR}$ and thus $I_D \approx I_{DD}$, meaning that the diffusion capacitance can be approximated by:

$$C_{D,D} \approx \frac{\tau_T I_D}{nV_T} \, \frac{1 + \dfrac{I_D}{2I_K}}{1 + \dfrac{I_D}{I_K}} \overset{I_D \ll I_K}{\approx} \frac{\tau_T I_D}{nV_T} \tag{1.15}$$

In silicon pn diodes $\tau_T \approx 1 \ldots 100\,\text{ns}$; in Schottky diodes the diffusion charge is negligible, since $\tau_T \approx 10 \ldots 100\,\text{ps}$.

Complete Model of a Diode

Figure 1.16 shows the complete model of a diode; it is used in CAD programs for circuit simulation. The diode symbols in the model represent the diffusion current I_{DD} and the recombination current I_{DR}; the breakthrough current I_{DBR} is shown as a controlled current source. Figure 1.17 contains the variables and equations. The parameters are listed in Fig. 1.18; in addition the parameter designations used in the circuit simulator *PSpice*[6] are shown. Figure 1.19 indicates the parameter values of some selected diodes taken from the component library of *PSpice*. Parameters not specified are treated differently by *PSpice*:

- A standard value is used:
  $I_S = 10^{-14}\,\text{A}$, $n = 1$, $n_R = 2$, $I_{BR} = 10^{-10}\,\text{A}$, $n_{BR} = 1$, $x_{T,I} = 3$, $f_C = 0.5$, $V_{Diff} = 1\,\text{V}$, $m_J = 0.5$
- The parameter is set to zero: $I_{S,R}$, R_B, C_{J0}, τ_T
- The parameter is set to infinity: I_K, V_{BR}

As a consequence of the values zero and infinite the respective effects are removed from the model [1.4].

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[6] *PSpice* is an *OrCAD* product.

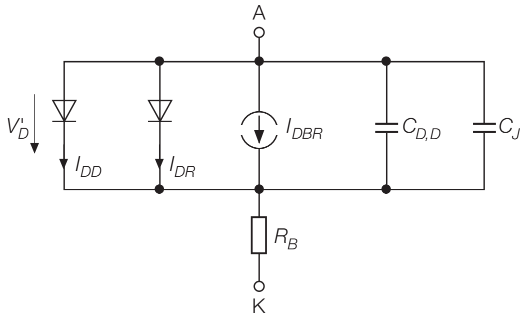

Fig. 1.16. Full model of a diode

Variable	Designation	Equation
I_{DD}	Diffusion current	(1.8)
I_{DR}	Recombination current	(1.9)
I_{DBR}	Breakthrough current	(1.10)
R_B	Spreading resistance	
C_J	Junction capacitance	(1.13)
$C_{D,D}$	Diffusion capacitance	(1.14)

Fig. 1.17. Variables of the diode model

Parameter	PSpice	Designation
Static performance		
I_S	IS	Saturation reverse current
n	N	Emission coefficient
$I_{S,R}$	ISR	Leakage saturation reverse current
n_R	NR	Emission coefficient
I_K	IK	Knee-point current for strong injection
I_{BR}	IBV	Breakthrough knee-point current
n_{BR}	NBV	Emission coefficient
V_{BR}	BV	Breakthrough voltage
R_B	RS	Spreading resistance
Dynamic performance		
C_{J0}	CJO	Zero capacitance of the depletion layer
V_{Diff}	VJ	Diffusion voltage
m_J	M	Capacitance coefficient
f_C	FC	Coefficient for the variation of the capacitance
τ_T	TT	Transit time
Thermal performance		
$x_{T,I}$	XTI	Temperature coefficient of reverse currents account to (1.14)

Fig. 1.18. Parameters in the diode model [1.4]

Parameter	PSpice	1N4148	1N4001	BAS40	Unit
I_S	IS	2.68	14.1	0	nA
n	N	1.84	1.98	1	
$I_{S,R}$	ISR	1.57	0	254	fA
n_R	NR	2	2	2	
I_K	IK	0.041	94.8	0.01	A
I_{BR}	IBV	100	10	10	µA
n_{BR}	NBV	1	1	1	
V_{BR}	BV	100	75	40	V
R_B	RS	0.6	0.034	0.1	Ω
C_{J0}	CJO	4	25.9	4	pF
V_{Diff}	VJ	0.5	0.325	0.5	V
m_J	M	0.333	0.44	0.333	
f_C	FC	0.5	0.5	0.5	
τ_T	TT	11.5	5700	0.025	ns
$x_{T,I}$	XTI	3	3	2	

1N4148 small-signal diode; 1N4001 rectifier diode; BAS40 Schottky diode

Fig. 1.19. Parameters of some diodes

1.3.3
Small-Signal Model

The linear *small-signal model* is derived from the nonlinear model by linearization at an operating point. The *static small-signal model* describes the small-signal response at low frequencies and is therefore called the *DC small-signal equivalent circuit*. The *dynamic small-signal model* also describes the dynamic small-signal response and is required for calculating the frequency response of a circuit; it is called the *AC small-signal equivalent circuit*.

Static Small-Signal Model

Linearization of the static characteristic curve given in (1.11) leads to the small-signal resistance:

$$\left.\frac{dV_D}{dI_D}\right|_A \;=\; \left.\frac{dV_D'}{I_D}\right|_A + R_B \;=\; r_D + R_B$$

It is made up of the spreading resistance R_B and the *differential resistance* r_D of the inner diode (see Fig. 1.10). Resistance r_D comprises three portions corresponding to the three current components I_{DD}, I_{DR} and I_{DBR}:

$$\frac{1}{r_D} \;=\; \left.\frac{dI_D}{dV_D'}\right|_A \;=\; \left.\frac{dI_{DD}}{dV_D'}\right|_A + \left.\frac{dI_{DR}}{dV_D'}\right|_A + \left.\frac{dI_{DBR}}{dV_D'}\right|_A$$

The differentiation of (1.6), (1.9) and (1.10) produces complex expressions; for practical purposes the following approximations may be used:

$$\frac{1}{r_{DD}} \;=\; \left.\frac{dI_{DD}}{dV_D'}\right|_A \;\approx\; \frac{I_{DD,A}+I_S}{nV_T}\,\frac{1+\dfrac{I_{DD,A}}{2I_K}}{1+\dfrac{I_{DD,A}}{I_K}} \;\overset{I_S \ll I_{DD,A} \ll I_K}{\approx}\; \frac{I_{DD,A}}{nV_T}$$

$$\frac{1}{r_{DR}} \;=\; \left.\frac{dI_{DR}}{dV_D'}\right|_A \;\approx\; \begin{cases} \dfrac{I_{DR,A}+I_{S,R}}{n_R V_T} & \text{for } I_{DR,A} > 0 \\[2ex] \dfrac{I_{S,R}}{m_J V_{Diff}^{m_J} |V_{D,A}'|^{1-m_J}} & \text{for } I_{DR,A} < 0 \end{cases}$$

$$\frac{1}{r_{DBR}} \;=\; \left.\frac{dI_{DBR}}{dV_D'}\right|_A \;=\; -\,\frac{I_{DBR,A}}{n_{BR} V_T}$$

Thus, the differential resistance r_D is:

$$r_D \;=\; r_{DD} || r_{DR} || r_{DBR}$$

For operating points that are in the diffusion region and below the high-current region $I_{D,A} \approx I_{DD,A}$ and $I_{D,A} < I_K$;[7] the following approximation can be used:

$$r_D \;=\; r_{DD} \;\approx\; \frac{nV_T}{I_{D,A}}\,. \tag{1.16}$$

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[7] This region is also called the *range of medium forward currents.*

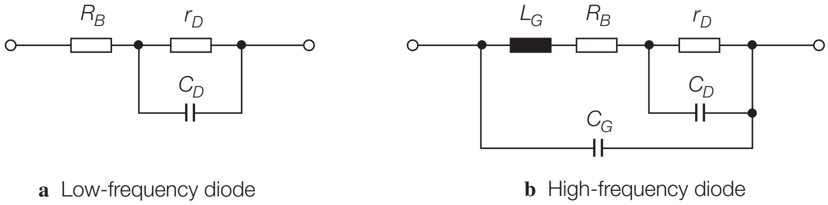


Fig. 1.20. Dynamic small-signal model

This equation corresponds to (1.3) in Sect. 1.1.4. As an approximation it may be used for all operating points in forward mode; in the high-current and recombination regions it provides values that are too low by a factor of 1 ... 2. Setting $n = 1 \ldots 2$ results in:

$$I_{D,A} = 1 \left\{ \begin{array}{c} \mu\text{A} \\ \text{mA} \\ \text{A} \end{array} \right\} \overset{V_T = 26\,\text{mV}}{\Rightarrow} r_D = 26 \ldots 52 \left\{ \begin{array}{c} \text{k}\Omega \\ \Omega \\ \text{m}\Omega \end{array} \right\}$$

With small-signal diodes in reverse mode the diffusion resistance is $r_D \approx 10^6 \ldots 10^9\,\Omega$; in the Ampere region of rectifier diodes this value is reduced by a factor of 10 ... 100.

The small-signal resistance in the breakthrough region is required only for Zener diodes since only in Zener diodes an operating point in the breakthrough range is permissible; the resistance is therefore called r_z. For $I_{D,A} \approx I_{DBR,A}$ its value is:

$$r_Z = r_{DBR} = \frac{n_{BR} V_T}{|I_{D,A}|} \tag{1.17}$$

Dynamic Small-Signal Model

Complete model: From the static small-signal model as shown in Fig. 1.10 the dynamic small-signal model according to Fig. 1.20a is derived by adding the junction capacitance and the diffusion capacitance; with reference to Sect. 1.3.2 the following applies:

$$C_D = C_J(V_D') + C_{D,D}(V_D')$$

In high-frequency diodes the additional parasitic influences of the case must be taken into consideration: Figure 1.20b shows the extended model with a case inductivity $L_G \approx 1 \ldots 100\,\text{nH}$ and a case capacitance of $C_G \approx 0.1 \ldots 1$ pF [1.6].

Simplified model: For practical calculations the spreading resistance R_B can be ignored and approximations can be used for r_D and C_D. From (1.15), (1.16) and the estimation $C_J(V_D') \approx 2C_{J0}$ the values for forward operation are:

$$r_D \approx \frac{n V_T}{I_{D,A}} \tag{1.18}$$

$$C_D \approx \frac{\tau_T I_{D,A}}{n V_T} + 2C_{J0} = \frac{\tau_T}{r_D} + 2C_{J0} \tag{1.19}$$

For reverse operation r_D is ignored, that is, $r_D \to \infty$ and $C_D \approx C_{J0}$.

1.4
Special Diodes and Their Application

1.4.1
Zener Diode

A *Zener diode* has a precisely specified breakthrough voltage that is rated for continuous operation in the breakthrough region; it is used for voltage stabilization or limitation. In Zener diodes the breakthrough voltage V_{BR} is called the *Zener voltage* V_Z and amounts to $V_Z \approx 3 \ldots 300\,\mathrm{V}$ in standard Zener diodes. Figure 1.21 shows the graphic symbol and the characteristic for a Zener diode. The current in the breakthrough region is given by (1.10):

$$I_D \;\approx\; I_{DBR} \;=\; -\,I_{BR}\, e^{-\frac{V_D+V_Z}{n_{BR}V_T}}$$

The Zener voltage depends on the temperature. The *temperature coefficient*

$$TC \;=\; \left.\frac{dV_Z}{dT}\right|_{T=300\,\mathrm{K},\, I_D=\mathrm{const.}}$$

determines the voltage variation at a constant current:

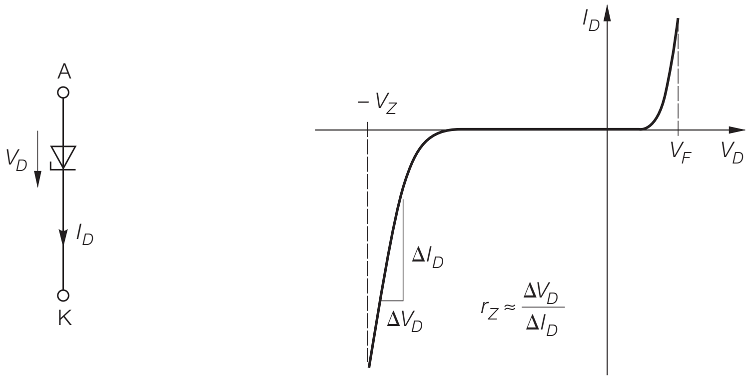
$$V_Z(T) \;=\; V_Z(T_0)\,(1 + TC\,(T - T_0)) \quad \text{with } T_0 = 300\,\mathrm{K}$$

When the Zener voltage is below 5 V, the Zener effect dominates with a negative temperature coefficient, while higher voltages produce the avalanche effect with a positive temperature coefficient; typical values are $TC \approx -6 \cdot 10^{-4}\,\mathrm{K}^{-1}$ for $V_Z = 3.3\,\mathrm{V}$, $TC \approx 0$ for $V_Z = 5.1\,\mathrm{V}$ and $TC \approx 10^{-3}\,K^{-1}$ for $V_Z = 47\,\mathrm{V}$.

The differential resistance in the breakthrough region is denoted by r_Z and corresponds to the reciprocal of the slope of the characteristic; from (1.17) it follows that:

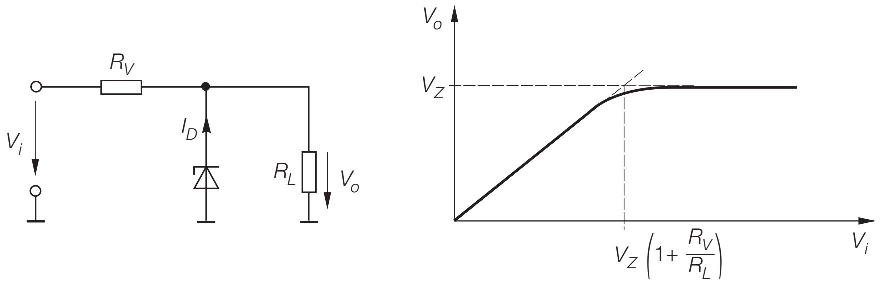
$$r_Z \;=\; \frac{dV_D}{dI_D} \;=\; \frac{n_{BR}V_T}{|I_D|} \;=\; -\,\frac{n_{BR}V_T}{I_D} \;\approx\; \frac{\Delta V_D}{\Delta I_D}$$

The differential resistance depends largely on the emission coefficient n_{BR} that reaches a minimum of $n_{BR} \approx 1 \ldots 2$ with $V_Z \approx 8\,\mathrm{V}$ and increases with lower or higher Zener voltages; typical values are $n_{BR} \approx 10 \ldots 20$ for $V_Z = 3.3\,\mathrm{V}$ and $n_{BR} \approx 4 \ldots 8$ for



a Graphical symbol **b** Characteristic

Fig. 1.21. Zener diode



a Circuit diagram **b** Characteristic

Fig. 1.22. Voltage stabilization with Zener diode

$V_Z = 47\,\text{V}$. The voltage-stabilizing effect of the Zener diode is based on the fact that the characteristic is very steep in the breakthrough region so that the differential resistance is very low; Zener diodes are best suited with $V_Z \approx 8\,\text{V}$ since here the characteristic shows the steepest slope due to the minimum value of n_{BR}. For $|I_D| = 5\,\text{mA}$ the resistance is $r_Z \approx 5 \ldots 10\,\Omega$ for $V_Z = 8.2\,\text{V}$ and $r_Z \approx 50 \ldots 100\,\Omega$ for $V_Z = 3.3\,\text{V}$.

Figure 1.22a displays a typical circuit for voltage stabilization. For $0 \le V_o < V_Z$ the Zener diode is reverse-biased and the output voltage is generated by voltage division with resistors R_V and R_L:

$$V_o \;=\; V_i\,\frac{R_L}{R_V + R_L}$$

$V_o \approx V_Z$ applies to the Zener diode in the conductive state. For the characteristic curve shown in Fig. 1.22b this means that:

$$V_o \;\approx\; \begin{cases} V_i\,\dfrac{R_L}{R_V + R_L} & \text{for } V_i < V_Z\left(1+\dfrac{R_V}{R_L}\right) \\ V_Z & \text{for } V_i > V_Z\left(1+\dfrac{R_V}{R_L}\right) \end{cases}$$

In order to render the stabilization effective, the operating point must be in the region in which the characteristic is almost horizontal. From the nodal equation

$$\frac{V_i - V_o}{R_V} + I_D \;=\; \frac{V_o}{R_L}$$

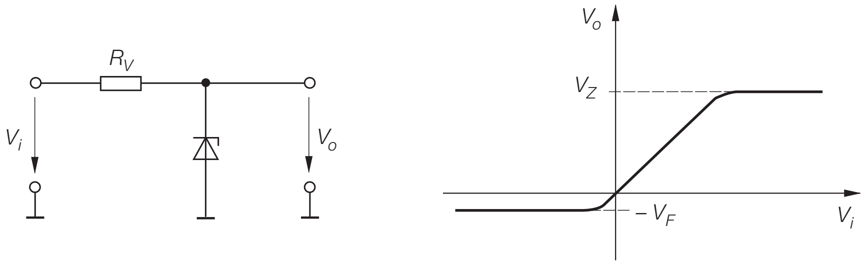
differentiation by V_o generates the *smoothing factor*

$$G \;=\; \frac{dV_i}{dV_o} \;=\; 1 + \frac{R_V}{r_Z} + \frac{R_V}{R_L} \;\overset{r_Z \ll R_V, R_L}{\approx}\; \frac{R_V}{r_Z} \tag{1.20}$$

and the *stabilization factor* [1.7]:

$$S \;=\; \frac{\dfrac{dV_i}{V_i}}{\dfrac{dV_o}{V_o}} \;=\; \frac{V_o}{V_i}\,\frac{dV_i}{dV_o} \;=\; \frac{V_o}{V_i}\,G \;\approx\; \frac{V_o R_V}{V_i r_Z}$$

Example: In a circuit with a supply voltage $V_b = 12\,\text{V} \pm 1\,\text{V}$ a section A is to be provided with the voltage $V_A = 5.1\,\text{V} \pm 10\,\text{mV}$; it requires a current $I_A = 1\,\text{mA}$. One can regard



a Circuit diagram **b** Characteristic

Fig. 1.23. Voltage limitation with Zener diode

this circuit section as a resistor $R_L = V_A/I_A = 5.1\,\mathrm{k}\Omega$ and use the Zener diode circuit in Fig. 1.22 with $V_Z = 5.1\,\mathrm{V}$ if $V_i = V_b$ and $V_o = V_A$. The series resistor R_V must be selected in such a way that $G = dV_i/dV_o > 1\,\mathrm{V}/10\,\mathrm{mV} = 100$; therefore from (1.20) it follows that $R_V \approx Gr_Z \geq 100r_Z$. The nodal equation leads to

$$- I_D = \frac{V_i - V_o}{R_V} - \frac{V_o}{R_L} = \frac{V_b - V_A}{R_V} - I_A$$

and (1.17) leads to $-I_D = n_{BR}V_T/r_Z$; by setting $R_V = Gr_Z$, $G = 100$ and $n_{BR} = 2$ the resistor R_V is:

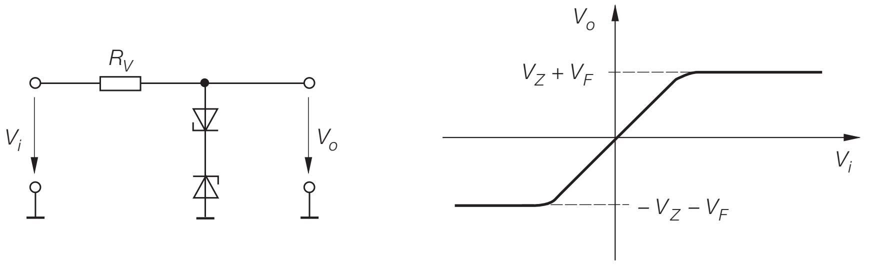
$$R_V = \frac{V_b - V_A - Gn_{BR}V_T}{I_A} = 1.7\,\mathrm{k}\Omega$$

Then the currents are $I_V = (V_b - V_A)/R_V = 4.06\,\mathrm{mA}$ and $|I_D| = I_V - I_A = 3.06\,\mathrm{mA}$. It can be seen that the Zener diode causes the current to be much higher than the current consumption I_A for the circuit section to be supplied. Therefore, this type of voltage stabilization is suitable only for partial circuits with a low current input. Circuits with a higher current input require a voltage regulator that may be more expensive but, as well as lower power losses, it also offers a better stabilization effect.

The circuit shown in Fig. 1.22a can also be used for voltage limitation. Removing the resistor R_L in Fig. 1.22a leads to the circuit in Fig. 1.23a with the characteristic shown in Fig. 1.23b:

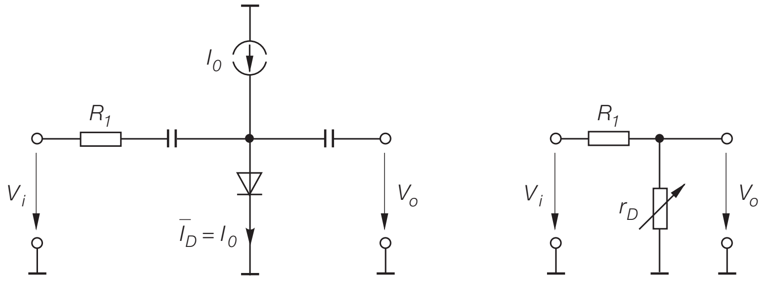
$$V_o \approx \begin{cases} -V_F & \text{for } V_i \leq -V_F \\ V_i & \text{for } -V_F < V_i < V_Z \\ V_Z & \text{for } V_i \geq V_Z \end{cases}$$

In the medium range the diode is reverse-biased, that is, $V_o = V_i$. For $V_i \geq V_Z$ the diode breaks through and limits the output voltage to V_Z. For $V_i \leq -V_F \approx 0.6\,\mathrm{V}$ the diode operates in the forward mode and limits negative voltages to the forward voltage V_F. The circuit in Fig. 1.24a allows a symmetrical limitation with $|V_o| \leq V_Z + V_F$; in the event of limitation one of the diodes is forward-biased and the other breaks through.



a Circuit diagram **b** Characteristic

Fig. 1.24. Symmetrical voltage limitation with two Zener diodes



a Circuit diagram **b** Equivalent circuit

Fig. 1.25. Voltage divider for alternating voltages with pin diode

1.4.2
Pin Diode

In *pin diodes*[8] the life cycle τ of the charge carriers in the nondoped i layer is particularly long. Since a transition from the forward-biased to the reverse-biased mode occurs only after recombination of almost all charge carriers in the i layer, a conductive pin diode remains in the forward mode even with short negative voltage pulses of a pulse duration $t_P \ll \tau$. The diode then acts as an ohmic resistor, with a value that is proportional to the charge in the i layer and thus proportional to the mean current $\overline{I}_{D,pin}$ [1.8]:

$$r_{D,pin} \approx \frac{n V_T}{\overline{I}_{D,pin}} \qquad \text{with } n \approx 1 \ldots 2$$

On the basis of this property the pin diode may be used with alternating voltages of a frequency $f \gg 1/\tau$ as a *DC-controlled AC resistance*. Figure 1.25 shows the circuit and the small-signal equivalent circuit of a simple variable voltage divider using a pin diode. In high frequency circuits mostly π *attenuators* with three pin diodes are used (see Fig. 1.26); a variable attenuation and a matching of both sides to a certain resistance of usually 50 Ω is then achieved by means of suitable control signals. The capacitances and inductances in Fig. 1.26 result in a separation of the DC and AC circuit paths. Typical pin diodes have $\tau \approx 0.1 \ldots 5\,\mu\text{s}$; this makes the circuit suitable for frequencies $f > 2 \ldots 100\,\text{MHz} \gg 1/\tau$.

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[8] Most pn diodes are designed as pin diodes, so that a high reverse voltage is reached across the i layer. The term *pin diode* is used only for diodes with lower impurity concentrations and a correspondingly higher life cycle of the charge carriers in the i layer.

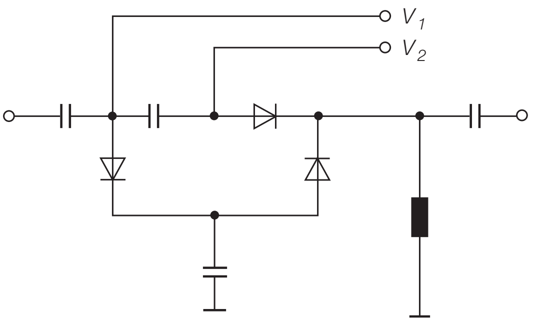

Fig. 1.26. π attenuator with three pin diodes for RF applications

Another important feature of pin diodes is the low junction capacitance due to a relatively thick i layer. This allows pin diodes to be used also for high frequency switches that provide a good off-state attenuation because of the low junction capacitance when the switch is open ($\overline{I}_{D,pin} = 0$). The typical circuit of an RF switch corresponds largely to the attenuator circuit shown in Fig. 1.26 which is designed as a short-series-short-switch with a particularly high off-state attenuation.

1.4.3
Varactor Diodes

Due to the voltage sensitivity of the junction capacitance a diode can be used as a variable capacitor (varactor); in this case the diode is operated in reverse mode and the junction capacitance is controlled by the reverse voltage. Equation (1.12) shows that the region in which the capacitance can be varied depends to a large degree on the capacitance coefficient m_J and increases as m_J increases. A particularly large range of $1 : 3 \ldots 10$ is reached in diodes with *hyperabrupt doping* ($m_J \approx 0.5 \ldots 1$) in which the impurity concentration increases close to the pn border just at the junction to the other region [1.8]. Diodes with this doping profile are called *variable-capacitance diodes* (*tuning diodes, varicap*) and are used predominantly for frequency tuning in LC oscillator circuits. Figure 1.27 shows the graphic symbol of a varactor diode and the curve of the junction capacitance C_J for some typical diodes. Although the curves are similar only diode BB512 shows the particular characteristic of a steeply decreasing junction capacitance. The capacitance coefficient m_J can be derived from the slope in the double logarithmic diagram; therefore Fig. 1.27 also depicts the slopes for $m_J = 0.5$ and $m_J = 1$.

In addition to the curve of the junction capacitance C_J, the quality factor Q is an important measure for the performance of a varactor diode. From the quality definition[9]

$$Q = \frac{|\mathrm{Im}\,\{Z\}\,|}{\mathrm{Re}\,\{Z\}}$$

and the impedance of the diode

$$Z(s) = R_B + \frac{1}{sC_J} \overset{s=j\omega}{=} R_B + \frac{1}{j\omega C_J}$$

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[9] This quality definition applies to all reactive components.

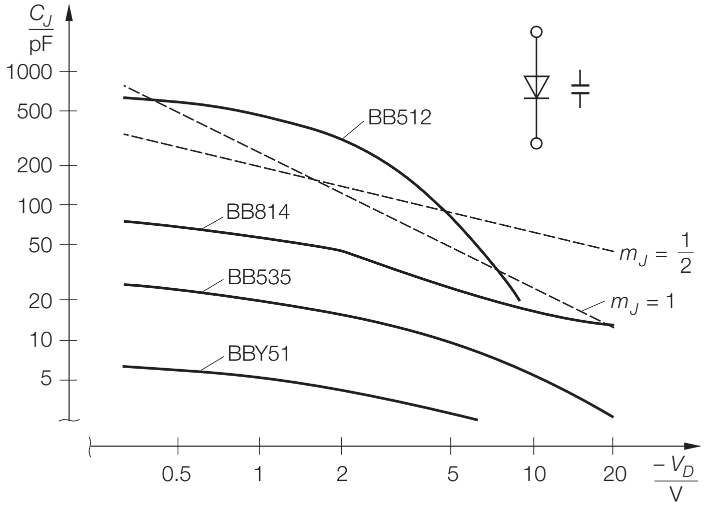


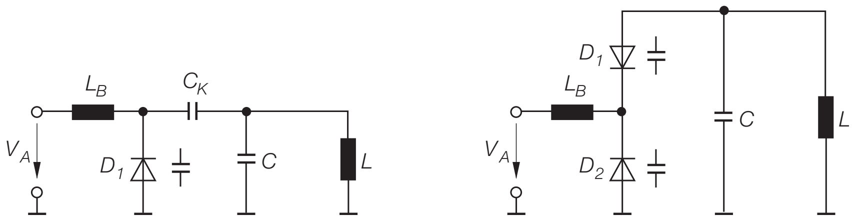
Fig. 1.27. Graphical symbol and capacitance curve of a varactor diode

Q is derived as [1.8]:

$$Q = \frac{1}{\omega C_J R_B}$$

For a given frequency, Q is inversely proportional to the spreading resistance R_B. Therefore, a high performance level is equivalent to a low spreading resistance and corresponds to low losses and a low damping when used in resonant circuits. Typical diodes have a quality factor of $Q \approx 50 \ldots 500$. As it is principally the spreading resistance that is needed for simple calculations and for circuit simulations new data sheets often specify R_B only.

In most cases, the circuits shown in Fig. 1.28 are used for frequency tuning in LC resonant circuits. In the circuit depicted in Fig. 1.28a both the junction capacitance C_J of the diode and the coupling capacitance C_K are connected in series and arranged in parallel with the parallel resonant circuit consisting of L and C. The tuning voltage $V_A > 0$ is provided via the inductivity L_B; with respect to the AC voltage this isolates the resonant circuit from the voltage source V_A and prevents the resonant circuit from being short-circuited by the voltage source. It is essential that $L_B \gg L$ is chosen to ensure that L_B does not affect the resonant frequency. The tuning voltage may also be provided via a resistor which, however, is an additional load to the resonant circuit and thus reduces the



a With one diode **b** With two diodes

Fig. 1.28. Frequency tuning in LC circuits with varactor diodes

quality of the circuit. The coupling capacitance C_K prevents the voltage source V_A from being short-circuited by the inductance L of the resonant circuit. Provided that $L_B \gg L$, the resonant frequency is:

$$\omega_R = 2\pi f_R = \frac{1}{\sqrt{L\left(C + \dfrac{C_J(V_A)\,C_K}{C_J(V_A)+C_K}\right)}} \overset{C_K \gg C_J(V_A)}{\approx} \frac{1}{\sqrt{L\,(C + C_J(V_A))}}$$

The tuning range depends on the characteristic of the junction capacitance and its relation to the resonant circuit capacitance C. The maximum tuning range is achieved with $C = 0$ and $C_K \gg C_J$.

In the circuit depicted in Fig. 1.28b a series connection of two junction capacitances is arranged in parallel to the resonant circuit. Here, too, the inductivity $L_B \gg L$ prevents a high-frequency short-circuit of the resonant circuit by the voltage source V_A. A coupling capacitance is not required since both diodes are in reverse mode so that no DC current can flow into the resonant circuit. In this case, the resonant frequency is:

$$\omega_R = 2\pi f_R = \frac{1}{\sqrt{L\left(C + \dfrac{C_J(V_A)}{2}\right)}}$$

Here, again, the tuning range is maximum for $C = 0$; however, only half the junction capacitance is effective so that compared to the circuit shown in Fig. 1.28a either the junction capacitance or the inductance must be twice as high for the same resonant frequency. A material advantage of the symmetrical diode arrangement is the improved linearity with high amplitudes in the resonant circuit; this largely offsets the decrease in the resonant frequency with increasing amplitudes that is caused by the nonlinearity of the junction capacitance [1.3].

1.4.4
Bridge Rectifier

The circuit shown in Fig. 1.29 made up of four diodes is called a *bridge rectifier* and is used for full-way rectification in power supplies and AC voltmeters. Bridge rectifiers for power supplies are divided into high-voltage bridge rectifiers, which are used for direct rectification of the mains voltage and must therefore have a high breakdown voltage ($V_{BR} \geq 350\,\text{V}$), and low-voltage bridge rectifiers, which are used on the secondary side of a line transformer; Sect. 16.5 describes this in more detail. Of the four connections two are marked with $\sim$ and one each with $+$ and $-$.

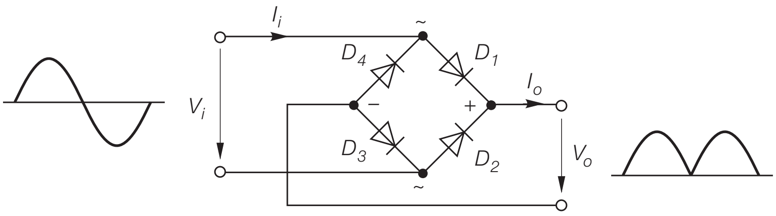
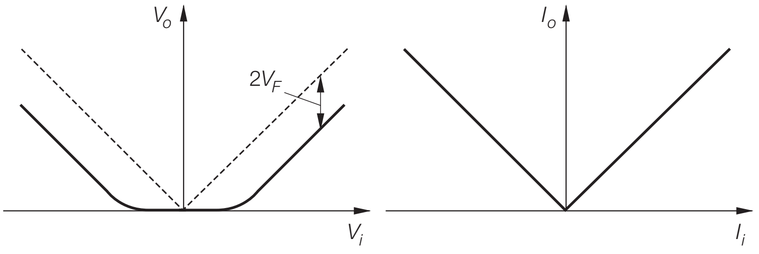


Fig. 1.29. Bridge rectifier



a Voltage characteristic **b** Current characteristic

Fig. 1.30. Characteristic curves of a bridge rectifier

With a positive input voltage D_1 and D_3 are conductive while D_2 and D_4 are reverse-biased; with a negative input voltage D_2 and D_4 are conductive while D_1 and D_3 are reverse-biased. Since at any given moment the current flows through two conductive diodes, the rectified output voltage is lower (by $2V_F \approx 1.2 \ldots 2\,\mathrm{V}$) than the magnitude of the input voltage:

$$V_o \approx \begin{cases} 0 & \text{for } |V_i| \leq 2V_F \\ |V_i| - 2V_F & \text{for } |V_i| > 2V_F \end{cases}$$

Figure 1.30a shows the voltage characteristic. A peak reverse voltage of $|V_D|_{max} = |V_i|_{max}$, which must be lower than the breakthrough voltage of the diodes, occurs across the diodes in reverse mode.

Unlike the voltages the magnitudes of the currents are in a linear relationship (see Fig. 1.30b):

$$I_o = |I_i|$$

This fact is used in meter rectifiers; the AC voltage to be measured is fed through a voltage-to-current converter and the resulting current is rectified in a bridge rectifier.

1.4.5
Mixer

Mixers are used in communication systems for frequency conversion. There are *passive mixers*, which use diodes or other passive components, and *active mixers*, which use transistors. In the case of passive mixers, the *ring modulator* consisting of four diodes and two transformers with centre tabs is most frequently used. Figure 1.31 shows a ring modulator in *downconverter* configuration with diodes $D_1 \ldots D_4$ and transformers $L_1 - L_2$ and $L_3 - L_4$ [1.9]. The circuit converts the input signal V_{RF} with the frequency f_{RF} by means of the *local oscillator voltage* V_{LO} with a frequency f_{LO} to an *intermediate frequency* $f_{IF} = |f_{RF} - f_{LO}|$. The output voltage V_{IF} is supplied to a resonant circuit in tune with the intermediate frequency in order to strip the signal from additional frequency components generated in the conversion process. The local oscillator provides a sinusoidal or rectangular voltage with an amplitude $\hat{v}_{LO}$; V_{RF} and V_{IF} are sinusoidal voltages of the amplitudes $\hat{v}_{RF}$ and $\hat{v}_{IF}$ respectively. In normal operation $\hat{v}_{LO} \gg \hat{v}_{RF} > \hat{v}_{IF}$ applies; in other words, the voltage of the local oscillator determines which diodes are conductive;

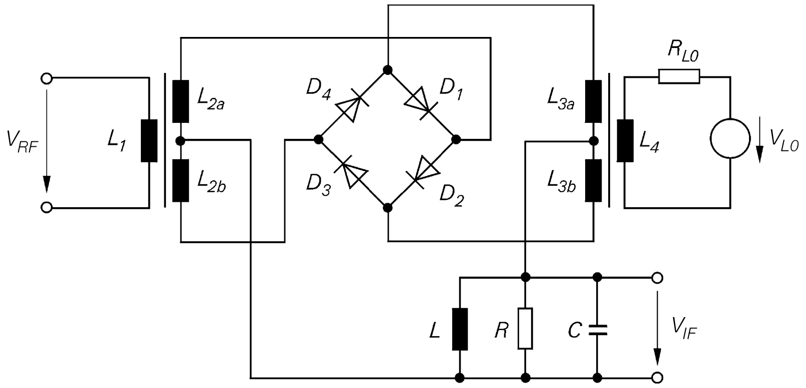


Fig. 1.31. Ring modulator in downconverter configuration

the following applies when using a 1:1 transformer with $L_4 = L_{3a} + L_{3b}$:

$$\left.\begin{array}{c} V_{LO} \ge 2V_F \\ -2V_F < V_{LO} < 2V_F \\ V_{LO} < -2V_F \end{array}\right\} \Rightarrow \left\{\begin{array}{l} D_1 \text{ and } D_2 \text{ are conductive} \\ \text{No diode is conductive} \\ D_3 \text{ and } D_4 \text{ are conductive} \end{array}\right.$$

V_F is the forward voltage of the diodes. Due to their better switching performance Schottky diodes with $V_F \approx 0.3\,\mathrm{V}$ are used exclusively; the current through the diodes is limited by the internal resistance R_{LO} of the local oscillator.

When D_1 and D_2 are conductive a current caused by V_{RF} flows through L_{2a} and $D_1 - L_{3a}$ or $D_2 - L_{3b}$ in the IF resonant circuit; when D_3 and D_4 are conductive, the current flows through L_{2b} and $D_3 - L_{3b}$ or $D_4 - L_{3a}$. The polarity of V_{IF} is different from that of V_{RF} so that the local oscillator and the diodes cause a polarity change at the frequency f_{LO} (see Fig. 1.32). If V_{LO} is a square wave signal with $\hat{v}_{LO} > 2V_F$ the change in polarity occurs suddenly; that is, the ring modulator multiplies the input signal with the square wave signal. The IF filter extracts the desired components with $m = 1, n = -1$ or $m = -1$ and $n = 1$ from the generated frequency components in the form $|mf_{LO} + nf_{RF}|$ with any integer value for m and $n = \pm 1$.

The ring modulator is available as a component with six connections, two each at the RF, LO, and IF sides [1.9]. Furthermore, there are integrated circuits containing only the diodes, and therefore have only four connections. In this context it must be noted that, despite their similarity in form, the mixer and the bridge rectifier differ from one another in terms of the arrangement of the diodes, as shown by a comparison of Figs. 1.31 and 1.29.

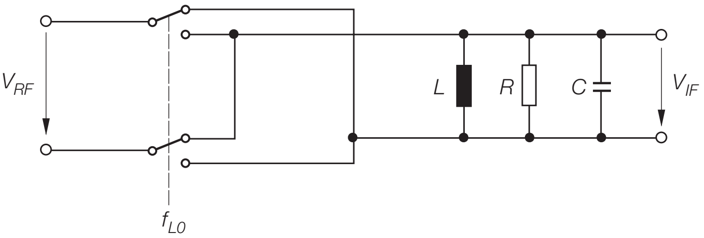


Fig. 1.32. Functioning of a ring modulator