SECTION 14

PULSED CIRCUITS AND WAVEFORM GENERATION

Pulsed circuits and waveform generation are very important to testing and identification in a whole range of electrical and electronic circuits and systems. There are essentially two types of such networks, those that are considered passive and the rest that can be lumped into active wave shaping (which includes those done digitally).

Passive circuits are lumped into linear and nonlinear. Linear are most commonly, single pole RC and RL networks. Nonlinear networks are usually designed around diodes with or without capacitors and inductors.

A common element used in waveform generation is the switch. Mechanical switches are cleaner giving better electrical characteristics; however, they do have serious limitations. Electronic switches can be compensated so that they can come close to the mechanical switches without the serious limitations such as contact bounce. In addition, electronic switches can be made smaller and are able to work at much higher frequencies.

Active networks are either analog or digital. Analog networks have been in use for a long period of time and still have many practical uses. Digital networks have advantages especially in the area of noise, speed, and accuracy and have successfully replaced most of the analog networks in most applications. C.A.

In This Section:

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Dynamic Behavior of Bipolar Switches

Section References:

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CHAPTER 14.1
PASSIVE WAVEFORM SHAPING

Paul G. A. Jespers

LINEAR PASSIVE NETWORKS

Waveform generation is customarily performed in active nonlinear circuits. Since passive networks, linear as well as nonlinear, enter into the design of pulse-forming circuits, this survey starts with the study of the transient behavior of passive circuits.

Among linear passive networks, the single-pole *RC* and *RL* networks are the most widely used. Their transient behavior in fact has a broad field of applications since the responses of many complex higher-order networks are dominated by a single pole; i.e., their response to a step function is very similar to that of a first-order system.

Transient Analysis of the *RC* Integrator

The step-function response of the *RC* circuit shown in Fig. 14.1.1*a*, after closing of the switch *S*, is given by

$$V(t) = E[1 - \exp(-t/T)] \tag{1}$$

where T = time constant = RC. The inverse of T is called the cutoff pulsation ω_0 of the circuit.

The Taylor-series expansion of Eq. (1) yields

$$V(t) = E\frac{t}{T}\left(1 - \frac{t}{2!T} + \frac{t^2}{3!T^2} - \cdots\right) \tag{2}$$

When the values of t are small compared with T, a first-order approximation of Eq. (2) is

$$V(t) \approx Et/T \tag{3}$$

In other words, the *RC* circuit of Fig. 14.1.1 behaves like an imperfect integrator. The relative error ϵ with respect to the true integral response is given by

$$\epsilon = -\frac{t}{2!T} + \frac{t^2}{3!T^2} - \frac{t^2}{4!T^3} + \cdots$$

The theoretical step-function response of Eq. (1) and the ideal-integrator output of Eq. (3) are represented in Fig. 14.1.1*b*.

Small values of t with respect to T correspond in the frequency domain (Fig. 14.1.1*c*) to frequency components situated above ω_0, that is, the transient signal whose spectrum lies to the right of ω_0 in the figure. In that case, the difference is small between the response curve of the *RC* filter and that of an ideal integrator

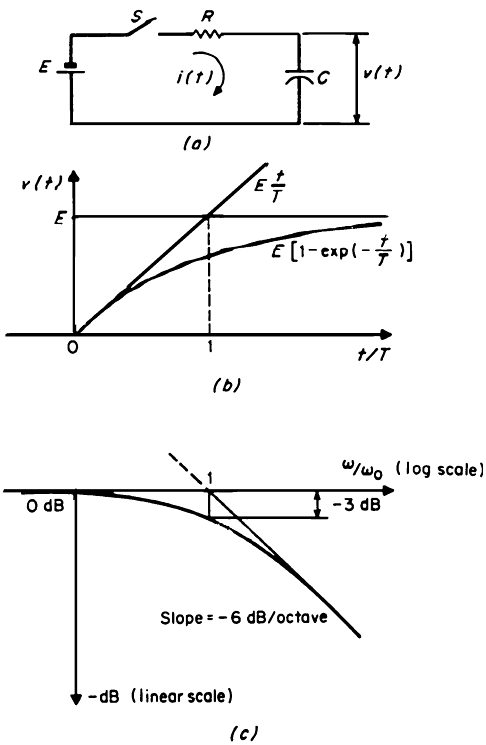

FIGURE 14.1.1 (*a*) *RC* integrator circuit; (*b*) voltage vs. time across capacitor; (*c*) attenuation vs. angular frequency.

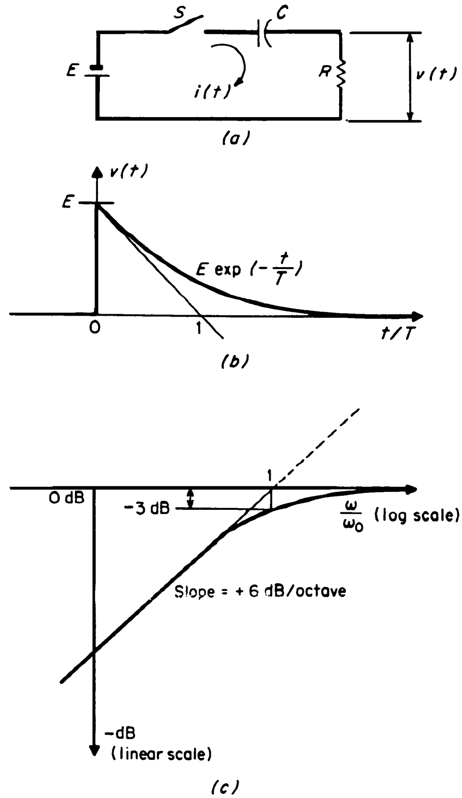

FIGURE 14.1.2 (*a*) *RC* differentiator circuit; (*b*) voltage across resistor vs. time; (*c*) attenuation vs. angular frequency.

(represented by the –6 dB/octave line in the figure). The circuit shown in Fig. 14.1.1*a* thus approximates an integrator, provided either of the following conditions is satisfied: (1) the time under consideration is much smaller than T or (2) the spectrum of the signal lies almost entirely above ω_0.

Transient Analysis of the *RC* Differentiator

When the resistor and the capacitor of the integrator are interchanged, the circuit (Fig. 14.1.2*a*) is able to differentiate signals. The step-function response (Fig. 14.1.2*b*) of the *RC* differentiator is given by

$$v(t) = E \exp(-t/T) \tag{4}$$

The time constant T is equal to the product RC, and its inverse ω_0 represents the cutoff of the frequency response of the circuit. As the values of t become large compared with T, the step-function response becomes more like a sharp spike; i.e., it increasingly resembles the delta function.

The response differs from the ideal delta function, however, because both its amplitude and its duration are always finite quantities. The area under the exponential pulse, equal to ET, is the important quantity in applications where such a signal is generated to simulate a delta function, as in the measurement of the impulse response of a system. These considerations may be transported in the frequency domain (Fig. 14.1.2*a*).

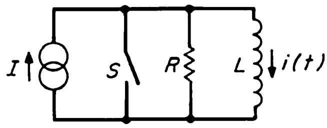

FIGURE 14.1.3 *RL* current-integrator circuit, the dual of the circuit in Fig. 14.1.1*a*.

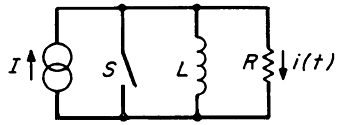

FIGURE 14.1.4 *RL* current-differentiator circuit, the dual of the circuit in Fig. 14.1.2*a*.

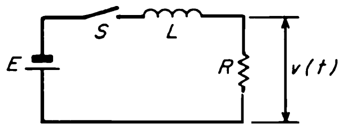

FIGURE 14.1.5 *RL* voltage integrator.

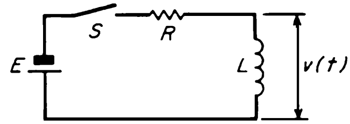

FIGURE 14.1.6 *RL* voltage differentiator.

Transient Analysis of *RL* Networks

Circuits involving a resistor and an inductor are also often used in pulse formation. Since integration and differentiation are related to the functional properties of first-order systems rather than to the topology of actual circuits, *RL* networks may perform the same function as *RC* networks. The duals of the circuits represented in Figs. 14.1.1 and 14.1.2, respectively, are shown in Figs. 14.1.3 and 14.1.4 and exhibit identical functional properties. In the first case, the current in the inductor increases exponentially from zero to I with a time constant equal to L/R, while in the second case it drops exponentially from the initial value I to zero, with the same time constant. Similar behavior can be obtained regarding voltage instead of current by changing the circuit from Fig. 14.1.3 to that of Fig. 14.1.5 and from Fig. 14.1.4 to Fig. 14.1.6, respectively. This duality applies also to the *RC* case.

Compensated Attenuator

The compensated attenuator is a widely used network, e.g., as an attenuator probe used in conjunction with oscilloscopes. The compensated attenuator (Fig. 14.1.7) is designed to perform the following functions:

1. To provide remote sensing with a very high input impedance, thus producing a minimum perturbation to the circuit under test.
2. To deliver a signal to the receiving end (usually the input of a wide-band oscilloscope) which is an accurate replica of the signal at the input of the attenuator probe. These conditions can be met only by introducing substantial attenuation to the signal being measured, but this is a minor drawback since adequate gain to compensate the loss is usually available.

Diagrams of two types of oscilloscope attenuator probes are given in Fig. 14.1.9, similar to the circuit of Fig. 14.1.7. In both cases, the coaxial-cable parallels the input capacitance of the receiver end; C_p represents the sum of both capacitances.

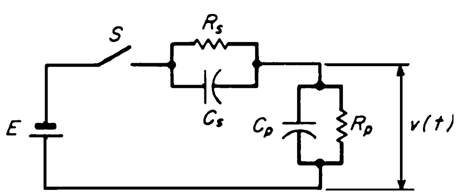

FIGURE 14.1.7 Compensated attenuator circuit.

The shunt resistor R_p has a high value, usually 1 MΩ, while the series resistor R_s is typically 9 MΩ. The dc attenuation ratio of the attenuator probe therefore is 1:10, while the input impedance of the probe is 10 times that of the receiver.

At high frequencies the parallel and series capacitors C_p and C_s play the same role as the resistive attenuator. Ideally these capacitors should be kept as low as possible to achieve a high input impedance even at high frequencies. Since it is impossible to reduce C_p below the capacitance of the coaxial cable, there is no alternative other than

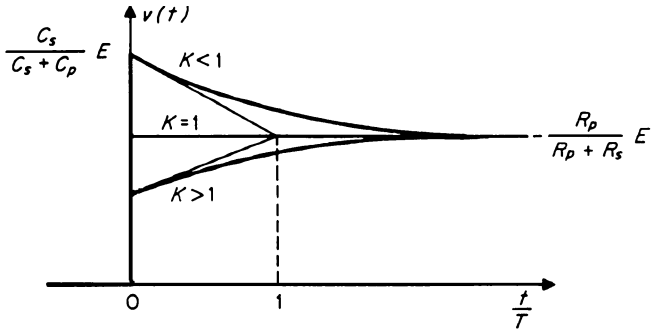

FIGURE 14.1.8 Voltage vs. time responses of attenuator, showing correctly compensated condition at $K = 1$.

to insert the appropriate value of C_s to achieve a constant attenuation ratio over the required frequency band. In consequence, as the frequency increases, the nature of the attenuator changes from resistive to capacitive. However, the attenuation ratio remains unaffected, and no signal distortion is produced. The condition that ensures constant attenuation ratio is given by

$$R_p C_p = R_s C_s \tag{5}$$

The step-function response of the compensated attenuator, which is illustrated in Fig. 14.1.8, clearly shows how distortion occurs when the above condition is not met. The output voltage $V(t)$ of the attenuator is given by

$$V(t) = \frac{C_s}{C_s + C_p}\left\{1 - (1-K)\left[1 - \exp\left(-\frac{t}{T}\right)\right]\right\}E \tag{6}$$

where K represents the ratio of the resistive attenuation factor to that of the capacitive attenuation factor

$$K = \frac{R_p}{R_p + R_s} \Big/ \frac{C_s}{C_p + C_s}$$

and

$$T = (R_p \parallel R_s)\,(C_s + C_p) \tag{7}$$

The $\parallel$ sign stands for the parallel combination of two elements, e.g., in the present case $R_p \parallel R_s = R_p R_s/(R_p + R_s)$. Only when K is equal to 1, in other words when Eq. (5) is satisfied, will no distortion occur, as shown in Fig. 14.1.8.

In all other cases there is a difference between the initial amplitude of the step-function response (which is controlled by the attenuation ratio of the capacitive divider) and the steady-state response (which depends on the resistive divider only).

A simple adjustment to compensate the attenuator consists of trimming one capacitor, either C_p and C_s, to obtain the proper step-function response. Adjustments of this kind are provided in attenuators like those shown in Fig. 14.1.9.

Compensated attenuators may be placed in cascade to achieve variable levels of attenuation. The conditions imposed on each cell are like those enumerated above, but an additional requirement is introduced, namely, the requirement for constant input impedance. This introduces a different structure compared with the compensated attenuator, as shown in Fig. 14.1.10. The resistances R_p and R_s must be chosen so that the impedance is kept constant and equal to R. The capacitor C_s is adjusted to compensate the attenuator, while C_p provides the required additional capacitance to make the input susceptance equal to that of the load.

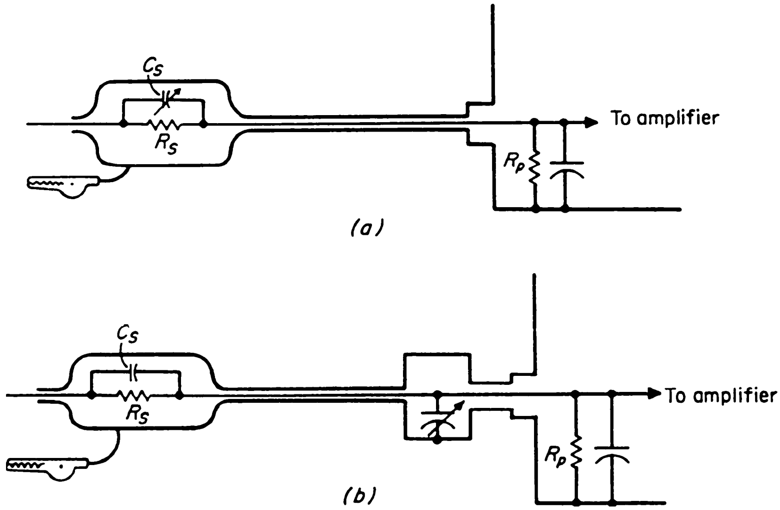

FIGURE 14.1.9 Coaxial-cable type of attenuator circuit: (*a*) series adjustment; (*b*) shunt adjustment.

Periodic Input Signals

Repetitive transients are typical input signals to the majority of pulsed circuits. In linear networks there is no difficulty in predicting the response of circuits to a succession of periodic step functions, alternatively positive and negative, since the principle of superposition holds. We restrict our attention here to two simple cases, the square-wave response of an *RC* integrator and an *RC* differentiator.

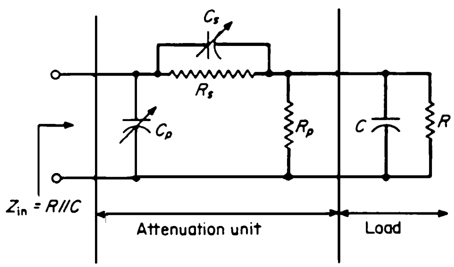

FIGURE 14.1.10 Compensated attenuator suitable for use in cascaded circuits.

Figure 14.1.11 represents, at the left, the buildup of the response of the *RC* integrator, assuming that the period τ of the input square wave is smaller than the time constant of the circuit *T*. On the right in the figure the steady-state response is shown. The triangular waveshape represents a fair approximation to the integral of the input square wave. The triangular wave is superimposed on a dc pedestal of amplitude *E*/2. Higher repetition rates of the input reduce the amplitude of the triangular wave without affecting the dc pedestal. When the frequency of the input square wave is high enough, the dc component is the only remaining signal; i.e., the *RC* integrator then acts like an ideal low-pass filter.

A similar presentation of the behavior of the *RC* differentiator is shown in Fig. 14.1.12*a* and *b*. The steady-state output in this case is symmetrical with respect to the zero axis because no dc component can flow through the series capacitor. When, as shown in Fig.14.1.12*b*, no overlapping of the pulses occurs, the steady-state solution is obtained from the first step.

Pulse Generators

The step function and the delta function (Dirac function) are widely used to determine the dynamic behavior of physical systems. Theoretically the delta function is a pulse of infinite amplitude and infinitesimal duration but having a finite area (product of amplitude and time). In practice the question of the equivalent physical impulse arises. The answer involves the system under consideration as well as the impulse itself.

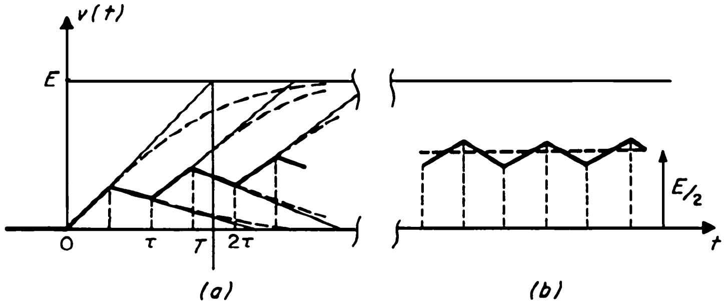

FIGURE 14.1.11 *RC* integrator with square-wave input of period smaller than *RC*: (*a*) initial buildup; (*b*) steady state.

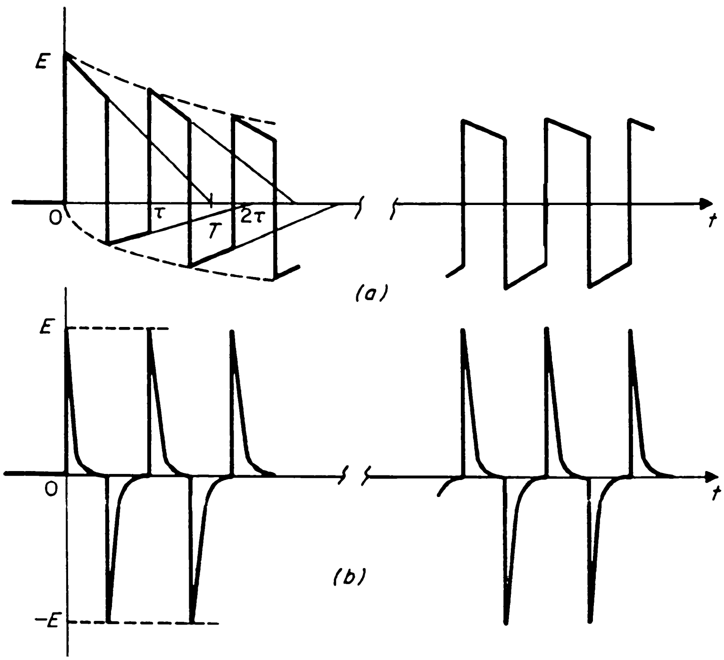

FIGURE 14.1.12 *RC* differentiator with square-wave input: (*a*) period of input signal smaller than *RC*; (*b*) input period longer than *RC*.

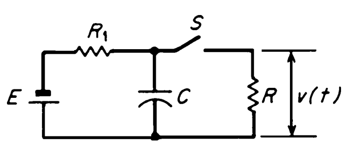

FIGURE 14.1.13 *RC* pulse-generator circuit with large series resistance R_1.

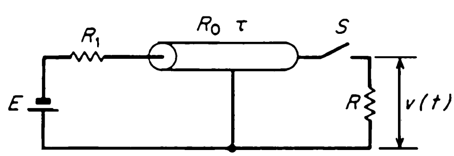

FIGURE 14.1.14 Coaxial-cable version of *RC* pulse generator.

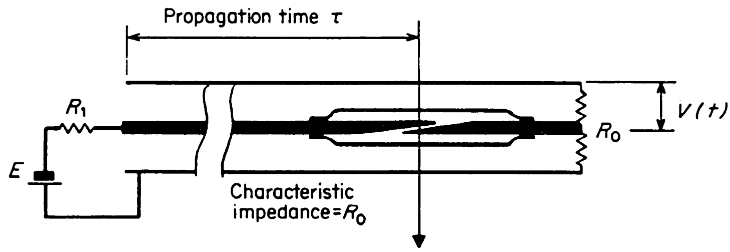

FIGURE 14.1.15 Use of mercury-wetted switch contacts in coaxial pulse generator.

The spectrum of the delta function has a constant amplitude over the whole frequency spectrum from zero to infinity. Other signals of finite area (amplitude × time) have different spectral distributions. On a logarithmic scale of frequency, the spectrum of any finite-area transient signal tends to be constant between zero and a cutoff frequency that depends on the shape of the signal. The shorter the duration of the signal, the wider the constant-amplitude portion of the spectrum.

If such a signal is used in a system whose useful frequency band is located below the cutoff frequency of the signal spectrum, the system response is indistinguishable from its delta impulse response. Any transient signal with a finite area, whatever its shape, can thus be considered as a delta function relative to the given system, provided that the flat portion of its spectrum embraces the whole system's useful frequency range. A measure of the effectiveness of a pulse to serve as a delta function is given by the approximation of useful spectrum bandwidth $B = 1/\tau$, where τ represents the midheight duration of the pulse.

Very short pulses are used in various applications in order to measure their delta-function response. In the field of radio interference, for instance, the basic response curve of the CISPR receiver* is defined in terms of its response to regularly repeated pulses. In this case, the amplitude of the uniform portion of the pulse spectrum must be calibrated, i.e., the area under the pulse must be a known constant which is a function of a limited number of circuit parameters.

The step-function response of an RC differentiator provides such a convenient signal. Its area is given by the amplitude of the input step multiplied by the time constant RC of the circuit. Moreover, since the signal is exponential in shape, its –3-dB spectrum bandwidth is equal to $1/RC$. In the circuit of Fig. 14.1.13, R_1 is much larger than R; when the switch S is open, the capacitor charges to the voltage E of the dc source. When the switch is closed, the capacitor discharges through R, producing an exponential signal of known amplitude and duration (known area).

A circuit based on the same principle is shown in Fig. 14.1.14. Here the coaxial line plays the role of energy storage source. If the line is lossless, its characteristic impedance is given by R_0, the propagation delay is equal to τ, and the Laplace transform of the voltage drop across R is

$$V(p) = (1/p)E[1 + (R_0/R)\coth p\tau]^{-1} \tag{8}$$

When the line is matched to the load, Eq. (8) reduces to

$$V(p) = (1/2p)E(1 - e^{-p2\tau}) \tag{9}$$

which indicates that $V(t)$ is a square wave of amplitude $E/2$ and duration 2τ. The area of the pulse is equal to the product of E and the time constant τ. Both quantities can be kept reasonably constant. The bandwidth is larger than that of an exponential pulse of the same area (Fig. 14.1.13) by the factor π.

Very wide bandwidth pulse generators based on this principles use a coaxial mercury-wetted switch built into the line (Fig. 14.1.15) to achieve low standing-wave ratios. A bandwidth of several GHz can be obtained in this manner.

In coaxial circuits, any impedance mismatch causes reflections to occur at both ends of the line, replacing the desired square-wave signal by a succession of steps of decreasing amplitude. The cutoff frequency of the

*International Electrotechnical Commission (IEC), "Specification, de l'apparelloge de mesure CISPR pour les frequences comprises entre 25 et 300 MHz," 1961.

spectrum is lowered thereby, and its shape above the uniform part can be drastically changed. Below cutoff frequency, however, the spectrum amplitude is given by $E\tau R/R_0$.

When the finite closing time of the switch is taken into account, it can be shown that only the width of the spectrum is reduced without affecting its value below the cutoff frequency. Stable calibrated pulse generators can also be built using electronic instead of mechanical switches.

NONLINEAR-PASSIVE-NETWORK WAVESHAPING

Nonlinear passive networks offer wider possibilities for waveshaping than linear networks, especially when energy-storage elements such as capacitors or inductors are used with nonlinear devices. Since the analysis of the behavior of such circuits is difficult, we first consider purely resistive nonlinear circuits.

Diode Networks without Storage Elements

Diodes provide a simple means for clamping a voltage to a constant value. Both forward conduction and avalanche (zener) breakdown are well suited for this purpose. Avalanche breakdown usually offers sharper nonlinearity than forward biasing, but consumes more power.

Clamping action can be obtained in many different ways. The distinction between series and parallel clamping is shown in Fig. 14.1.16. Clamping occurs in the first case when the diode conducts; in the second when it is blocked.

Since the diode is not an ideal device, it is useful to introduce an equivalent network that takes into account some of its imperfections. The complexity of the equivalent network is a trade-off between accuracy and ease of manipulation.

The physical diode is characterized by

$$I = I_s[\exp(V/V_T) - 1] \tag{10}$$

where I_S is the leakage current and $V_T = kT/q$, typically 26 mV at room temperature.

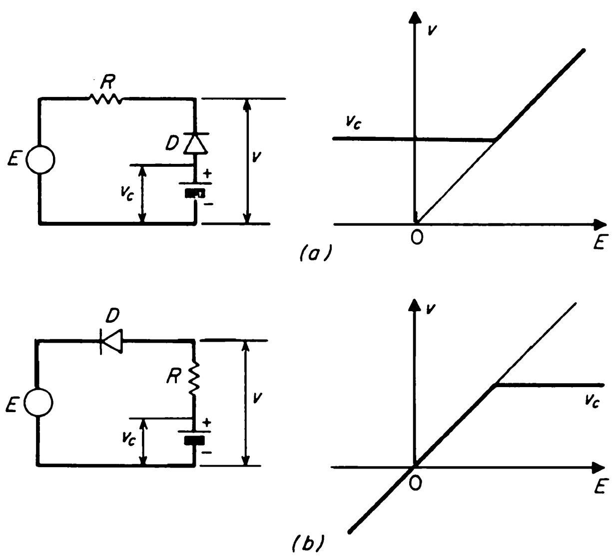

FIGURE 14.1.16 Diode clamping circuit and voltage vs. time responses: (*a*) shunt diode; (*b*) series diode.

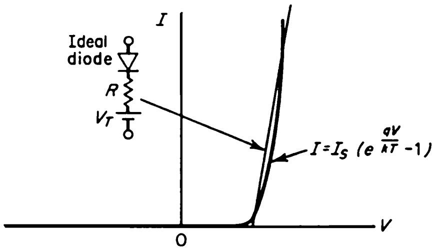

FIGURE 14.1.17 Actual and approximate current-voltage characteristics of ideal and real diodes.

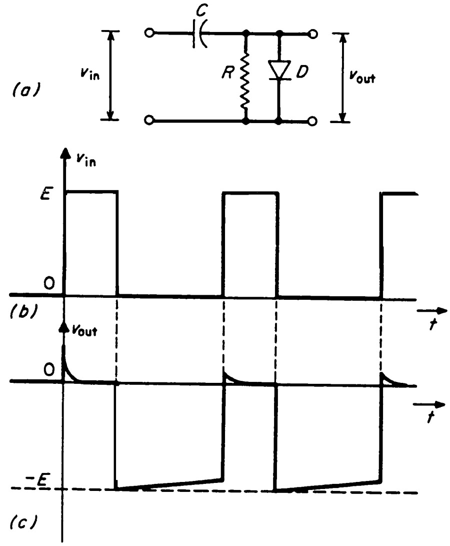

FIGURE 14.1.18 (*a*) DC restorer circuit; (*b*) input signal; (*c*) output signal.

The leakage current is usually quite small, typically 100 pA or less. Therefore, V must be at least several hundred millivolts, typically 600 mV or more, to attain values of forward current I in the range of milliamperes. A first approximation of the forward-biased real diode consists therefore of a series combination of the ideal diode and a small emf (Fig. 14.1.17). Moreover, to take into account the finite slope of the forward characteristic, a better approximation is obtained by inserting a small resistance in series.

Diode Networks with Storage Elements

There is no simple theory to represent the behavior of nonlinear circuits with storage elements, such as capacitors or inductances. Acceptable solutions can be found, however, by breaking the analysis of the circuit under investigation into a series of linear problems. A typical example is the dc restorer circuit hereafter.

The circuit shown in Fig. 14.1.18 resembles the *RC* differentiator but exhibits properties that differ substantially from those examined previously. The diode *D* is assumed to be ideal first to simplify the analysis of the circuit, which is carried out in two steps, i.e., with the diode forward- and reverse-biased. In the first step, the output of the circuit is short-circuited; in the second, the diode has no effect, and the circuit is identical to the linear *RC* differentiator.

When a series of alternatively positive and negative steps is applied at the input, after the first positive step is applied, no output voltage is obtained. The first positive step causes a large transient current to flow through the diode and charges the capacitor. Since *D* is assumed to be an ideal short circuit, the current will be close to a delta function as long as the internal impedance of the generator connected at the input is zero.

In practice, the finite series resistance of the diode must be added to the generator internal impedance, but this does not affect the load time constant significantly, since it is assumed to be much smaller than the time between the first positive step and the following negative step. This allows the circuit to attain the steady-state conditions between steps. When the input voltage suddenly returns to zero, the output voltage undergoes a large negative swing whose magnitude is equal to that of the input step. The diode is then blocked, and the capacitor discharges slowly through the resistor. If the time constant is assumed to be much larger than the period of the input wave, the output voltage swings back to zero when the second positive voltage step is applied and only a small current flows through the forward-biased diode to restore the charge lost when the diode was under the reverse-bias condition.

If the finite resistance of the diode is taken into consideration, a series of short positive exponential pulses must be added to the output signal, as shown in the lower part of Fig. 14.1.18. The first pulse, which corresponds to the initial full charge on the capacitor, is substantially higher than the next pulse, but this is of little importance in the operation of the circuit.

An interesting feature of the dc restorer circuit lies in the fact that although no dc component can flow from input to output, the output signal has a well-defined nonzero dc level, although determined only by the amplitude of the negative steps (assuming, of course, that the lost charge between two steps is negligible). This circuit is used extensively in video systems to prevent the average brightness level of the image from being affected by its varying video content. In this case, the reference steps are the line-synchronizing pulses.