BIBLIOGRAPHY


ATTENUATION MEASUREMENT

KURT HILTY
Swiss Federal Office of Metrology
Bern-Wabern, Switzerland

When a signal is sent along any transmission path, many different mechanisms degrade it. Because of finite conductivity, every cable shows a resistive loss. Furthermore, the dielectric loss and the skin effect may be significant at higher frequencies, and imperfect screening of cables leads to radiation losses which might be quite important at higher frequencies. Single-mode optical fibers have two main intrinsic loss mechanisms, scattering loss and absorption loss. Scattering loss is dominated by Rayleigh scattering, and below 1600 nm absorption losses are caused mainly by OH absorption. Fibers that are bent too tightly or cabled too poorly may have, in addition, bend losses because of nonguided modes. Connectors show losses because of nonideal contacts and imperfect impedance matching, which reflect part of the signal to the transmitter. Filters built in the transmission path have
losses in the passband caused by finite conductivity and probably dielectric losses. Wireless transmission paths, such as microwave links, satellite links, and broadcast or mobile communication, are affected by scattering caused by rain, clouds, and multiple reflections.

The two examples following clearly show how additional attenuation or losses influence a system:

In a radar system a total loss of 2 dB in the feeder system and the duplexer wastes 37% of the transmitter power. During the development of the different parts, the losses have to be measured very carefully to minimize the total loss.

Satellite systems quite often operate with cooled front ends at the receiver because the signals to be picked up are extremely weak. Therefore the front ends often operate with noise temperatures of 5 K. An additional loss of 0.1 dB from an uncooled waveguide would raise the noise temperature to 7 K.

These examples show how important it is to measure the losses of the different parts of a transmission system as accurately as possible to optimize system parameters.

Many different methods and a variety of systems for measuring attenuation have been developed. The most important techniques are described in the following sections.

1. ATTENUATION

1.1. Definition

In the field of loss measurement, the most important terms are attenuation, insertion loss, mismatch loss, and voltage loss [1,2]. These terms are discussed in the following sections.

1.1.1. Attenuation. According to Beatty [3], attenuation is defined as the decrease in power level at the load caused by inserting a device between a $Z_0$ source and load, where $Z_0$ is the characteristic impedance of the line. Figure 1 shows the basic idea of such an attenuation measurement.

Attenuation is mostly expressed by a logarithmic scale in decibels (dB) or in nepers. The attenuation of a two-port device is defined as follows [4]:

In decibels:

$$A = 10 \log \frac{\text{power delivered to a matched load}}{\text{by a matched source}} - \frac{\text{power delivered to the same load when the two-port device is inserted}}{\text{(1)}}$$

In nepers:

$$A = \frac{1}{2} \ln \frac{\text{power delivered to a matched load}}{\text{by a matched source}} - \frac{\text{power delivered to the same load when the two-port device is inserted}}{\text{(2)}}$$

Because $\log(x) = \ln(x)/\ln(10)$, the following relationship between decibels and nepers is valid:

Attenuation in decibels $= 8.6858 \times$ attenuation in nepers.

Attenuation is a property only of a two-port device.

1.1.2. Insertion Loss. In practical applications neither the source nor the load have an impedance exactly equal to $Z_0$, the characteristic impedance of a line. Therefore source and load have a reflection coefficient of $r_s$ and $r_L$, respectively. Let $P_1$ be the power delivered from the source to the load and $P_2$ be the power absorbed by the same load when the two-port device is inserted (Fig. 2).

Then the loss is defined by

$$L_1 = 10 \log \frac{P_1}{P_2}$$

(3)

The insertion loss depends on the property of the device and the reflection coefficients of the source and the load.

1.1.3. Scattering Parameters. Two-port networks, especially at radiofrequencies, are very well characterized by scattering parameters.

A two-port device inserted between a source and a load is shown in Fig. 3.
The complex wave amplitudes $a_1$, $a_2$, $b_1$, and $b_2$ shown in Fig. 3 are related as follows:

$$b_1 = s_{11}a_1 + s_{12}a_2$$  \hspace{1cm} (4)

and

$$b_2 = s_{21}a_1 + s_{22}a_2$$  \hspace{1cm} (5)

Setting up the signal flow graph for the configuration in Fig. 3 and using the nontouching loop rules of Mason [5,6], the insertion loss is given by the following expression:

$$L_I = 20 \log \left| \frac{(1 - r_s s_{11})(1 - r_L s_{22}) - r_s r_L s_{12} s_{21}}{|s_{21}| \cdot |1 - r_s r_L|} \right|$$  \hspace{1cm} (6)

For a matched system where $r_s = r_L = 0$, Eq. (6) delivers the attenuation

$$A = 20 \log \frac{1}{|s_{21}|}$$  \hspace{1cm} (7)

Equations (6) and (7) clearly show that the insertion loss $L_I$ depends on the property of the two-port device and the reflection coefficients of the source $r_s$ and the load $r_L$. Otherwise the attenuation is the pure property of the two-port device.

1.1.4. Voltage Loss. Voltage loss is used only for applications in the DC-to-UHF part of the spectrum where voltage is well defined. According to Warner [4], it is defined as follows:

$$L_V = 20 \log \left| \frac{\text{voltage at the input of the two-port device}}{\text{voltage at the output of the two-port device}} \right|$$  \hspace{1cm} (8)

and using the scattering parameters

$$L_V = 20 \log \left| \frac{(1 + s_{11})(1 - r_L s_{22}) + r_L s_{12} s_{21}}{|s_{21}(1 - r_L)|} \right|$$  \hspace{1cm} (9)

Note that $L_V$ is independent of the source reflection coefficient $r_s$.

When $s_{11} = 0$ and $r_L = 0$, the voltage loss is equal to the attenuation.

1.1.5. Mismatch Loss. At higher frequencies every real two-port device has an input and output reflection coefficient that differs from zero. Therefore, there is always a mismatch between the source and the two-port device and between the two-port device and the load (Fig. 4). This reflects part of the incoming and outgoing wave toward the source and toward the two-port device, respectively, resulting in additional losses.

The mismatch loss between the source and the two-port device is expressed as:

$$L_{m1} = 10 \log \frac{\text{power absorbed at the input of the two-port device}}{\text{maximal available power from the source}}$$  \hspace{1cm} (10)

According to Fig. 3 the mismatch loss is given by

$$L_{m1} = \frac{(1 - |r_s|^2)(1 - |r_L|^2)}{|1 - r_s r_L|^2}$$  \hspace{1cm} (11)

Similarly, the mismatch loss between the two-port device and the load is given by

$$L_{m2} = 10 \log \frac{\text{power absorbed by the load}}{\text{maximal available power at the output of the two-port device}}$$  \hspace{1cm} (12)

and with the parameters of Fig. 4

$$L_{m2} = 10 \log \frac{(1 - |r_s|^2)(1 - |r_L|^2)}{|1 - r_s r_L|^2}$$  \hspace{1cm} (13)

If several two-port devices are cascaded, the mismatch loss between them has to be calculated similarly and taken into account.

2. ATTENUATOR

Apart from the natural losses in devices and transmission paths, manufactured devices have well-defined losses. These devices, called attenuators, are used for measurement and for adjusting power levels to a defined value in transmission systems. Attenuators are probably the most important devices in measurement systems and therefore exist in a large number of different forms [7–9], such as symmetric, coaxial, waveguide, optical, fixed-value, and variable-loss. The important properties of attenuators are frequency range, attenuation accuracy, attenuation variation versus frequency, input and output impedance match (reflection coefficient), power handling capacity, and phase linearity.

2.1. Balanced-Line Attenuator

Balanced lines are used especially in telecommunications and lately in local-area networks and in-house...
communications systems. Therefore, there is a demand for balanced-line attenuators. Chains of symmetric double-T or double-II circuits, as shown in Fig. 5, are mostly used [10].

The reference handbooks [10] give formulas and tables to determine the circuit elements for a given line impedance and different element attenuations. The circuit has to be symmetric to the ground plane and well matched to the line impedance. Special techniques are given to optimize the circuit for small frequency dependence.

Variable-value attenuators are commercially available for different impedances (150Ω, 120Ω) in the frequency range from DC to several megahertz and have attenuation ranges from 0 to 132 dB.

2.2. Coaxial-Line Attenuator

2.2.1. Fixed-Value Coaxial Attenuator. Coaxial attenuators generally have multi octave bandwidth or frequently operate from DC to several gigahertz (GHz). Coaxial transmission lines are generally operated in the transverse electromagnetic (TEM) mode and therefore an obvious solution for attenuators is to use lumped elements that are small compared to the wavelength. There are four major constructions: T circuit, Π circuit, distributed lossy line, and distributed thin-film technology.

2.2.1.1. T-Circuit Attenuator. The T-attenuator circuit is shown in Fig. 6. The values of $R_1$ and $R_2$ are calculated from the following formulas, where $K$ is the transmission coefficient [7,10]:

$$ R_1 = Z_0 \frac{K-1}{K+1} \quad (14) $$

$$ R_2 = \frac{2Z_0K}{K^2 - 1} \quad (15) $$

The attenuation in decibels is given by

$$ A(\text{dB}) = 20 \log K \quad (16) $$

Resistive rods for $R_1$ are often used, and disk resistors are used for $R_2$, or film resistors are used for both $R_1$ and $R_2$.

2.2.1.2. Π-Circuit Attenuator. Techniques similar to the T circuit are used for the Π circuit shown in Fig. 7, and the corresponding formulas are as follows [7,10]:

$$ R_1 = Z_0 \frac{K+1}{K-1} \quad (17) $$

$$ R_2 = Z_0 \frac{K^2 - 1}{2K} \quad (18) $$

At higher microwave frequencies the elements of the T and Π attenuators have dimensions comparable to the wavelength, and therefore the reactance of the elements becomes more important and thus degrades the performance of the attenuator.

2.2.1.3. Lossy-Line Attenuator. Distributed lossy-line attenuators, explained by Weber [11], have very favorable performance. As long as the attenuation is not very high, they have a flat attenuation response and an excellent impedance match. The center conductor of the transmission line is made mostly of lossy material using thin film deposited on a substrate with a circular cross section. The disadvantage of lossy-line attenuators is that they have a lower frequency limit depending on their physical length.

2.2.1.4. Thin-Film Attenuator. Distributed thin-film attenuators [12] use a resistive material deposited on a dielectric substrate for the series and shunt losses (Fig. 8).

Strip transmission lines and narrow ground electrodes on the longitudinal sides are added to the input and output terminals. The characteristic impedance and the attenuation are constant and given as follows [7]:

TART numbered equation (19)

$$ Z_0 = \rho \sqrt{\frac{D-a}{4a}} \quad (\Omega) \quad (19) $$

$$ x = \sqrt{\frac{4}{a(D-a)}} \quad \text{(in nepers per unit length)} \quad (20) $$

Resistive rods for $R_1$ are often used, and disk resistors are used for $R_2$, or film resistors are used for both $R_1$ and $R_2$. 

Figure 5. Basic symmetric double-T and double-II attenuation circuits.

Figure 6. T-attenuator circuit.

Figure 7. Π-attenuator circuit.
Equation (20) shows that the attenuation is independent of the resistivity of the film as long as the film is homogenous. The attenuation depends only on the geometry and therefore is insensitive to temperature changes.

2.2.2. Variable-Value Coaxial Attenuator. There are two types of variable attenuators: continuous variable attenuators and step attenuators. Continuous variable attenuators have the advantage of being noninterruptive when changing attenuation. This feature is important for some measurements, for example, receiver sensitivity measurements. On the other hand, these attenuators may sometimes lack accuracy, settable impedance match, high insertion loss and have limited bandwidth.

Step attenuators are very accurate and have the following qualities: good reproduceability, low insertion loss, excellent impedance match, and wide operating bandwidth. But most step attenuators are interruptive when changing the attenuation value.

2.2.2.1. Continuously Variable Attenuator

2.2.2.1.1. Piston Attenuator (Waveguide Beyond Cutoff Attenuation). The piston attenuator is one of the oldest continuously variable microwave attenuators [13]. It is used especially as a precision attenuator or for handling high power levels. This type of attenuator uses a waveguide below its cutoff frequency. According to transmission-line theory, the amplitude of a wave launched into such a waveguide decays exponentially. So the attenuation is calculable. Most constructions use a circular cross section and a magnetic coupling that generates the lowest cutoff higher-order transverse electric mode \( \text{TE}_{11} \). Figure 9 shows the simplified construction of a piston attenuator. By sliding the two concentric cylinders into each other, the physical displacement of the coupling coil is changed and therefore also the attenuation. Special care must be taken to avoid unwanted higher order modes.

Attenuation as a function of wavelength is given per unit length as [10]

\[
A = \frac{2\pi \cdot 20}{\lambda_c \cdot \ln 10} \sqrt{1 - \left(\frac{\lambda_c}{\lambda}\right)^2} \text{ in decibels per unit length}
\]

\[(21)\]

where \( \lambda_c \) is the cutoff wavelength of the waveguide and \( \lambda \) is the free-space wavelength.

If the operating frequency is chosen to be much lower than the cutoff frequency, the term \( \lambda_c/\lambda \) is negligibly small for the operating frequency band, and a flat attenuation frequency response is achieved. The advantages of cutoff attenuators are that they are calculable and that high accuracy is achieved (0.001 dB/10 dB for 60 dB attenuation). For this reason, cutoff attenuators are often used as standard attenuators. Their main disadvantage is the high insertion loss (15 to 20 dB) because tight coupling has to be avoided so as not to stimulate higher modes.

2.2.2.1.2. Resistant Card, T-, or P-Type Attenuator. Several constructions of variable attenuators using resistive cards or resistive films are on the market [9]. The variable card attenuator operates like a potentiometer. A resistive film is fixed on a substrate so that the resistance between the input and the output is varied with a movable coaxial wiper, and thus the amount of attenuation changes. This type of attenuator does not have good input and output impedance matches. More sophisticated constructions use T- or P-type structures where the series and the shunt resistors are changed simultaneously. Therefore the input and output ports of these attenuators are quite well matched. The minimum insertion loss is on the order of 4 dB, and they operate up to several gigahertz.

2.2.2.1.3. Lossy-Line Attenuator. Lossy-line attenuators use a lossy center conductor partly covered with a thin sliding shield [7]. This effectively changes the length of the resistive conductor and thus the attenuation. Some constructions use microstrip lines with variable lossy walls [9]. This type of attenuator is limited in its maximal attenuation because of the length of the device.

2.2.2.1.4. Pin Attenuator. Pin attenuators change the loss in either a step or continuous mode. The series and shunt resistors are by pin diodes, electronic devices that change their conductivity. The diodes are controlled by a bias current. Various types of circuits are available, such as series, shunt, bridged T, and II [9]. Pin attenuators are electronic circuits and therefore may produce harmonics. Below about 10 MHz, the pin diode exhibits some rectifying behavior. At much higher frequencies,
the pin diode behaves more like a variable resistor. Because of the matching circuits for input and output and the bias network, most devices operate in a limited frequency band. Attenuation from 2 to 80 dB is achieved.

### 2.2.2.1.5. Step Attenuator

Step attenuators always use a set of fixed attenuators that are switched into a line by different mechanisms. The steps are mostly 0.1, 1, and 10 dB. The 0.1-dB steps generally cover a range of 1.1 dB, the 1-dB steps a range of 11 dB, and the 10-dB steps cover a range up to 110 dB. The step attenuator has excellent repeatability, covers a wide frequency range (e.g., DC to 26.5 GHz), has a good impedance match, and mostly has a flat frequency response of the attenuation value.

In the turret-type coaxial attenuator a set of coaxial fixed attenuators is placed in a cylindrical arrangement. With a rotary motion the different elements (e.g., 0 dB, 10 dB, 20 dB, ...) are switched between the junctions of the transmission line.

Another type of step attenuator uses a variety of fixed attenuation elements. The different attenuation elements are cascaded by switches or bypassed. The switches may be activated manually or electrically.

### 2.3. Waveguide Attenuator

Waveguide attenuators work mostly in the entire usable waveguide bandwidth, which is not quite half an octave. To attenuate a wave propagated in a waveguide, either the electric or the magnetic field or even both are influenced. As an example, a resistive card inserted into the waveguide parallel to the electric field attenuates it. Another technique uses lossy walls that influence the current in the waveguide wall. Most of these attenuators are not phase-invariant.

#### 2.3.1. Fixed-Value Waveguide Attenuator

The waveguide flap attenuator (Fig. 10) and the side-vane attenuator (Fig. 11) [7] are very popular.

The flap attenuator is based on a resistive card inserted in the center of the waveguide parallel to the electric field. The more the card dives into the waveguide, the more the electric field is weakened, and therefore attenuation increases. A smooth card shape is chosen to minimize the reflection caused by the discontinuity.

The side-vane attenuator (Fig. 11) influences the electric field similarly. The vane is always completely inside the waveguide, but it uses the fact that the electric field varies along the broad side. For the most popular TE_{10} mode, the electric field is zero at the sidewall and has its maximum in the center of the waveguide. Therefore the position of the resistive card defines the attenuation value. A smooth shape minimizes the reflection of the discontinuity.

Several constructions of lossy-wall attenuators exist. One version is shown in Fig. 12 [14]. Section 2 with equally spaced slots filled with a lossy material defines the attenuation. Sections 1 and 3 are configured to minimize the reflection due to the discontinuity in the wall. The lossy-wall attenuator withstands high power because the dissipated heat is transferred to any cooling system.

#### 2.3.2. Variable-Value Waveguide Attenuator

**2.3.2.1. Flap and Side-Vane Attenuator.** By adding a mechanism that changes the position of the resistive card, the fixed-value-flap and side-vane attenuator are easily transformed into a variable-value attenuator. It is often used as a settable attenuator.

**2.3.2.2. Rotary-Vane Attenuator.** The rotary-vane attenuator was invented in the early 1950s by E. A. N. Whitebread (Elliot H. Brothers, London) and A. E. Bowen (Bell Telephone Laboratories). It was proposed and developed as a precision waveguide attenuator [15–19]. The rotary-vane attenuator consists of three sections of waveguide that have a resistive film spanned across the waveguide, as shown in Fig. 13. The middle section has a circular cross section and can be rotated with respect to the two fixed-end sections. Figure 13 illustrates the principle of a rotary-vane attenuator. For clarity, the transitions from a round to a rectangular waveguide at both ends are omitted.

The electric field is perpendicular to all resistive films whenever the films are aligned. In this case no current flows in the resistive film, and therefore no attenuation occurs. If the center part is rotated by an angle $\theta$, the component $E_{\sin \theta}$ in $\theta$ produces a current flowing in the resistive film and are absorbed. Thus the resulting...
The attenuation is given by

\[ A(\text{dB}) = -40 \log(\cos \theta) + A_0 \]  

(22)

The rotary-vane attenuator has the following advantages:

- The attenuation is almost independent of the frequency.
- The phase shift is very small. The phase variations are smaller than 1° up to 40 dB attenuation.
- The input and output VSWRs are very low under all conditions.
- The attenuation is not sensitive to the resistive film as long as its attenuation is high enough.
- The attenuation is not sensitive to temperature changes.

Rotary-vane attenuators are commercially available for most of the known waveguide bands. They typically have an accuracy of 2% of the reading in decibels or 0.1 dB, whichever is greater.

2.4. Attenuator for Optical Fibers

Attenuators have to be adapted to an optical fiber system. Therefore different attenuators for single-mode or multimode applications are available and mostly operate within one or two wavelength windows. Various attenuation techniques are used to reduce the transmitting light, such as lateral or axial displacement of the fibers or optical prisms, and inserting absorbing filters.

2.4.1. Fixed- or Adjustable-Value Attenuator. A pair of lenses is used in most fiberoptic attenuators to collimate the light of the input fiber and to refocus it to the output fiber. Any attenuation mechanism, such as absorbing filters or absorbing glass with variable thickness can be inserted into the optical beam path. Attention has to be given to the attenuation mechanism so that there is very little polarization-dependent loss. Fixed-value attenuators are commercially available for a range of 3–40 dB with an accuracy of 0.5–1 dB.

2.4.2. Variable-Value Attenuator. The reflection type variable attenuator often combines a series of 10 dB steps for the high values and a continuous variable attenuation of up to 10 or 15 dB. Figure 14 shows the basic configuration of an optical section [20].

A rod lens collimates the light of the fiber into a spatial beam that passes two neutral-density filters (ND filters). The second rod lens focuses the spatial beam into the output fiber. The ND filter must have high uniformity, little attenuation change with time, and flat spectral transmittance. This is achieved by vacuum-depositing Ni...
and Cr alloys onto a glass substrate. The ND filters are the reflecting type. To avoid multiple reflections and to ensure that the light is not reflected into the input lead, the two filters are inclined at a small angle with respect to the optical axis.

Commercially available attenuators of this type have an insertion loss of about 3.5 dB, a range of up to 65 dB, and an accuracy of about ±0.5 dB per 10 dB. They are available for wavelengths of 850, 1300, and 1550 nm.

2.5. Calculable Attenuation Standards

Many attenuators have been described in the previous sections, but only a few are suitable as attenuation standards. An attenuation standard is a device that can be traced to the SI units by an unbroken chain. The application of the described standards depend on the frequency band and the technique used in the measurement system.

2.5.1. Kelvin–Varley Divider. The Kelvin–Varley divider (KVD) was described for the first time by Varley in 1866 [21]. It is a resistive voltage divider and operates from DC up to several hundred kilohertz. Figure 15 shows an example of a four-decade Kelvin–Varley divider.

Each decade consists of 11 equal resistors except for the last decade, which has only 10. The decades are connected by switches that always span two resistors. The value of the resistors of each following decade is reduced by a factor of 5. The four-decade Kelvin–Varley divider allows varying the output-to-input voltage ratio from zero to one in steps of one part in 10^4. The unloaded KVD has a constant input impedance that is independent of the switch setting, whereas the output resistance varies with the setting. The original type of KVD (Fig. 15) requires either very large or very small resistance values because each decade needs 5 times larger values. Due to stray capacitance, the divider reacts similarly to an RC filter. For example, a 100 kΩ input impedance limits the 3 dB bandwidth to about 100 kHz. To avoid large resistance values, modified constructions with a resistor shunting the decades have been developed.

Two major errors determine the accuracy: deviations of the resistors from nominal values and the resistances of the switch contacts and leads. The first three or four decades are the most sensitive, and therefore these resistors often are adjustable. Several calibration techniques have been developed and are described in the literature [22,23].

Today commercially available Kelvin–Varley dividers have up to seven decades, have an absolute linearity of ±1 part in 10^7, and long term stability of ±1 part in 10^6 per year.

2.5.2. Inductive-Voltage Divider. The inductive-voltage divider (IVD), also called a ratio transformer, is an exceptionally accurate variable attenuation standard. It consists of a number of very accurately tapped autotransformers. The autotransformers are connected together by high quality switches. The IVD operates from 10 Hz to about 100 kHz, and the greatest accuracy is achieved at about 1 kHz. In 1962 Hill and Miller [24] described a multidecade IVD with seven decades and a resolution of 1 part in 10^7. Figure 16 shows the principle of a seven-decade IVD with a setting of 0.4324785. The output-to-input voltage ratio can be set from zero to one.

The tapped autotransformers are constructed by winding exactly equal lengths of copper wire on a high permeability toroidal core. A superalloy having an extremely high permeability (>100,000) and low hysteresis loss is preferred as a core material. For an exact division of the tapped autotransformer, it is not necessary to have a 100% coupling between the ten inductors [4]. But the 10 self-inductances and the 45 mutual inductances have to be exactly equal.

![Figure 15. Principle of a four-decade Kelvin–Varley divider.](image-url)
The following error sources limit the accuracy of IVDs:

- Inequality in the series resistances and the leakage inductances of the sections in each autotransformer
- Inhomogenities in the magnetic cores
- Distributed admittances between the windings
- Internal loading caused by the later stages
- Impedances of the connecting leads and switch contacts
- Variations in the input voltage, frequency, and ambient temperature

With careful design, these errors can be minimized. Programmable binary IVDs with 30 bits, a resolution of 1 part in $10^9$, and a linearity of 0.1 ppm have been developed [25]. IVDs with eight decades and an accuracy of four parts in $10^8$ are commercially available.

### 2.5.3. Intermediate-Frequency Piston Attenuator

The IF piston attenuator is based on the same principle as the attenuator previously described for RF, but it is designed to operate at a specific, fixed frequency, mostly 30 or 60 MHz. As Eq. (21) shows, attenuation depends on the cutoff wavelength $\lambda_c$, the free-space wavelength $\lambda$, and the displacement of the two coils. The waveguide dimensions, which can be determined, define the cutoff wavelength, and the displacement can be measured very precisely. Therefore the IF piston attenuator is used as a calculable standard. Figure 17 shows a simplified diagram of an IF piston attenuator.

The standard IF piston attenuator consists of a high-precision circular cylinder that has excellent conductivity, a fixed coil, and a movable coil mounted on a piston. The piston attenuator operates in the $H_{11}$ (TE$_{11}$) mode that has the lowest attenuation. A well-designed metal strip filter in front of the fixed coil attenuates the higher modes. To allow smooth movement and to avoid any scratches, the plunger carrying the moving coil is insulated. Equation (23) expresses the attenuation per unit length more precisely

$$A = \frac{s_{11}}{r} \sqrt{1 - \left(\frac{\lambda_c}{\lambda}\right)^2} \cdot \epsilon - \frac{\delta}{r}$$  (23)

in nepers per unit length, where $\lambda_c = 2\pi r/s_{11}$, $\lambda_c$ is the cutoff wavelength, $\lambda$ is the free-space wavelength, $s_{11}$ the first zero of the Bessel function $J_1 = 1.841838$, $r$ the radius of the cylinder, and $\delta$ is the skin depth.

Highly accurate standard attenuators use a laser interferometer to accurately determine the displacement of the coil. Yell [26,27] and Bayer [28] developed extremely accurate IF piston attenuators with a dynamic range of 120 dB, a resolution of 0.0001 dB, and an accuracy of 0.0002 dB/10 dB over the linear range of 90 dB.

### 2.5.4. Rotary-Vane Attenuator

The principle of the rotary-vane attenuator is described in Section 2.3.2.2. Because the attenuation is given by the equation

$$A(\text{dB}) = -40 \log(\cos \theta) + A_0$$  (24)
where \( A_0 \) is the insertion loss at a setting of \( \theta = 0 \), the device can be used as a calculable primary standard.

The rotating angle \( \theta \) of the center vane has to be determined very precisely, especially for higher attenuation values. As an example, a rotational angle accuracy of \( \pm 0.001^\circ \) results in an attenuation accuracy of \( \pm 0.01 \) dB at a setting of 60 dB. Especially precise optical readouts have been developed by national standards laboratories [29–31] to allow an angular resolution of \( \pm 0.001^\circ \). Following are the main error sources for precision rotary-vane attenuators:

- Misalignment of the end vanes
- Insufficient attenuation of the central vane
- Incorrect readout of the rotational angle
- Eccentricity of the rotor
- Leakage of the rotating joints
- Internal reflections at the ends of the three vanes

Careful design of the attenuator results in an accuracy of \( \pm 0.0015 \) dB up to 16 dB at 10 GHz.

### 2.5.5. Comparison of Attenuation Standards

The attenuation standards mentioned previously are used in various precision measurement systems, such as RF, IF, or LF substitution. The standards have very different accuracy depending on the attenuation setting. Figure 18 shows a comparison of different precision attenuation standards used in national metrology laboratories [32].

#### 2.5.6. Optical Attenuation Standards

Imamura [33] shows one solution of a calculable optical attenuation standard that is used to calibrate precision attenuators. The key element is a rotating disk with a well-defined opening. The device operates as an optical chopping system (Fig. 19).

As long as the response of the detector is slow compared to the rotational speed \( \omega \), the ratio of \( P_1 \) to \( P_0 \) defines the attenuation. In this case the attenuation depends only on the opening angle \( \theta \) (in radians) and is given by the following equation:

\[
A = -10 \cdot \log \frac{P_1}{P_0} = -10 \cdot \log \frac{\theta}{2\pi}
\]

An opening angle of 36\(^\circ\) defines an attenuation of 10 dB. Special care has to be paid to the diffraction of light at the edges of the disk openings and the stability of the light source and the sensor. The overall accuracy is estimated to be \( \pm 0.008 \) dB for 10 dB attenuation.

### 3. MEASUREMENT OF ATTENUATION

Various kinds of measurement systems are used depending on the frequency range, the type of attenuator, the required accuracy, and the available standards. Most of the modern attenuation measurement systems are computer- or microprocessor-controlled. They use the same principle as the manual systems but operate much faster. If maximum accuracy is required, manually controlled measurement systems are often preferred.

#### 3.1. Low-Frequency Measurement Systems

Low-frequency attenuation measurements are used mainly in communication systems where voice, video, and data have to be transmitted with as little distortion as possible. Dedicated test systems have been developed for testing and adjusting the communication systems working either with coaxial lines (50 or 75 \( \Omega \)) or balanced lines (124, 150, or 600 \( \Omega \)).

#### 3.1.1. Direct Measurement

Operation of communication systems requires a great number of test systems for
which low-cost test systems that are easily handled were developed. The systems are based on current, voltage, or power measurement and operate in a frequency band from 200 Hz up to 30 MHz. The test system (Fig. 20) consists of a tuneable generator with a known constant output level and a wideband voltmeter or a high-sensitivity, selective superheterodyne receiver.

Many test systems can be switched to operate in either coaxial-line (50 or 75 Ω) or balanced-line (124, 150, or 600 Ω) configuration. In the balanced-line mode the test systems have a limited frequency range of 200 Hz to several MHz depending on the impedance selected. In the selective-level meter mode, bandwidths of 25 Hz–3.1 kHz are used, and the dynamic range achieved is on the order of 80 dB. The attenuation measurement accuracy in the frequency range indicated is about 1 dB.

3.1.2. Low-Frequency Substitution Method. The LF substitution method is based on a precisely calibrated low-frequency reference attenuator used in a parallel or a serial configuration (Fig. 21).

Because the attenuation of the device under test (DUT) is compared with that of the reference attenuator, neither the output level of the generator nor the absolute level indication of the receiver have to be known. The only requirements are that generator and receiver remain stable during measurements. The accuracy is determined mainly by the calibration of the reference attenuator.

3.2. Radiofrequency and Microwave Measurement Systems

In radiofrequency and microwave ranges many different measurement principles are known. They all have their own characteristics; one is meant to measure low values of attenuation, another to measure high values, a third to achieve highest accuracy. The most popular measurement principles are discussed in the following sections.

3.2.1. Power Ratio Method. The power ratio method [4,40] (Fig. 22) is very simple for measuring attenuation. It is commonly used as long as maximum accuracy is not required.

The method is based on the linearity of the power meter or the receiver used. First the power $P_1$ of the generator is measured without the device under test (DUT) and then $P_2$ is measured with the DUT inserted. The attenuation of the DUT is calculated by the ratio of $P_2$ to $P_1$:

$$ A(\text{dB}) = 10 \log \frac{P_2}{P_1} $$  \hspace{1cm} (26)  

To measure attenuation, the insertion points have to be matched by either tuners or matching pads. The square-law characteristics of the power sensors and the noise limit the dynamic range to about 40 dB. If a tuned receiver is used instead of a power meter, the measurement range is extended to about 100 dB.

Several error sources influence the accuracy:

- The stability of the generator and the detector system
- The frequency stability of the generator
- The matching at the insertion points
- The square-law region of the detector system
- The crosstalk for high-attenuation measurement

Commercially available systems using a tuned receiver achieve a measurement uncertainty of 0.1 dB at 50 dB attenuation. These systems are easily automated by controlling the instruments with a computer.

In national standards laboratories very sophisticated systems have been developed resulting in an accuracy of 0.06 dB at 50 dB attenuation [34,35].

3.2.2. Voltage Ratio Method. The voltage ratio method makes use of high-resolution AC digital voltmeters (AC DVMs) available now. Because the AC DVMs work up to only several MHz, the RF signals have to be downconverted to low frequency. Figure 23 shows the principle of a

![Figure 20. Principle of a direct measurement system.](image)

![Figure 22. Principle of the power ratio method.](image)

![Figure 21. Principle of a LF substitution method.](image)
voltage ratio system working at an audiofrequency of 50 kHz.

If a synthesizer locked to the same reference frequency is used as a signal generator and local oscillator, a very stable audiofrequency $f_a$ (e.g., 50 kHz) is generated. The audiofrequency signal is amplified and measured with an AC DVM. If $U_1$ is the voltage measured with the two insertion points clamped together and $U_2$ is the voltage with the DUT inserted, the attenuation is given by

$$A_{\text{dB}} = 20 \log \left( \frac{U_2}{U_1} \right) + C$$

(27)

where $C$ is the correction factor in decibels for the non-linearity of the amplifier and the DVM.

The dynamic range of the direct system is about 20–30 dB. More sophisticated systems achieve an uncertainty less than 0.001 dB for 20 dB attenuation. By adding a gauge block technique, for example, a calibrated step attenuator (10, 20, 30, 40, 50, 60, 70 dB) in series with the DUT in the RF path, the range is extended to 90 dB with excellent accuracy of 0.001 dB [32].

The error sources which limit the measurement uncertainty are

- The matching of the insertion points
- The generator output level stability
- The AF amplifier stability
- The AF amplifier and AC DVM linearity
- The mixer linearity
- The gauge block attenuator stability and reproducibility
- The crosstalk for high attenuation measurement

3.2.3. IF Substitution Method. The IF substitution method [4,40] (Fig. 24) gives good accuracy, operates over a large dynamic range, and is used up to very high frequencies. Most systems operate in a parallel substitution mode.

The signal passing through the DUT is mixed to an IF of 30 or 60 MHz. This signal is compared with the signal of the 30-MHz reference oscillator and the standard attenuator by a narrowband 30 MHz receiver (mostly with synchronous detection). In the first phase the insertion points are clamped together, and the standard attenuator is adjusted until there is no switching signal (i.e., 1 kHz) detectable any longer. The reading $A_1$ of the standard attenuator is taken. In a second phase the DUT is inserted, the standard attenuator is adjusted so that the signal of the standard attenuator equals the signal of the mixer, and the reading $A_2$ is taken. The attenuation of the DUT is given by the difference $A_2$ minus $A_1$ between readings.

A piston attenuator, an inductive-voltage divider, or a high-precision resistive attenuator can be used as a standard attenuator.

In national standards laboratories very-high-precision piston attenuators with a resolution of 0.0001 dB over a 100 dB range have been used in a parallel substitution system. The accuracy achieved is better than 0.001 dB per 10-dB step [27,32,36].

Accuracy of about 0.002 dB and a dynamic range of 100 dB have been achieved by using a seven-decade 50-kHz inductive voltage divider in a parallel substitution system [37].

Weinert [38] proposed a parallel IF complex vector substitution system using a high precision IF attenuator. The system has a single-step dynamic range of 140 dB and a display resolution of 0.001 dB.

The following error sources limit the accuracy and the dynamic range:

- The matching at the insertion points
- The level stability of the signal source
- The mixer linearity
- The noise balance
- The level stability of the IF reference oscillator
- The standard attenuator resolution and stability
- The crosstalk for high attenuation measurement
3.2.4. RF Substitution Method. In the RF-substitution method [4,40] (Fig. 25), the reference standard attenuator and the DUT operate at the same frequency. Because the attenuation of the reference standard is compared either in a series or in a parallel substitution system with the DUT, the results are independent of the receiver characteristics. A rotary-vane attenuator, a piston attenuator, or a chain of well-matched and precisely calibrated attenuators (e.g., step attenuator) is used as a reference standard.

In the first step the insertion points are clamped together, and the reference standard is adjusted to a value $A_1$ according to the estimated attenuation of the DUT. The receiver shows the reading $U_1$. In the second step the DUT is inserted, and the reference standard is adjusted to get the same reading $U_1$ at the receiver $A_2$. The attenuation of the DUT is calculated as the difference between the two decibel readings of the reference attenuator, $A_1$ minus $A_2$.

3.2.5. Scalar Measurement. All attenuation measurement systems described in the previous sections provide scalar measurements. There are many commercial scalar network analyzers available (Fig. 26) [8,39]. These analyzers measure the input reflection and the attenuation of the device under test. Because mostly wideband detectors are used, the magnitude of only two quantities can be determined. The signal of the sweep generator is divided by a power splitter or a directional coupler into reference and measurement paths. The directional bridge or coupler measures the reflected wave of the DUT. The analyzer forms the ratio $A/R$, which is proportional to the input reflection coefficient of the DUT. Using a third detector, the attenuation is measured by calculating the ratio $B/R$. Most scalar network analyzers are microprocessor- or computer-controlled and offer simple correction methods. The calibration for reflection measurements is frequently done by using open and short circuits, and a connect through normalization is used for the transmission path. Because these analyzers are broadband systems, they operate very fast and are easily expandable to highest frequencies. Commonly used scalar network analyzers operate from 10 MHz to 18 GHz or 26 GHz, and often their frequency range can be extended to 50 GHz in coaxial lines and to 110 GHz in waveguides. The dynamic range is limited to about 75 dB by the square-law characteristic of the detectors and noise. The measurement accuracy achieved is quite reasonable, for example, 0.6 dB measuring a 30-dB attenuator. The insertion points have to be well matched.

The following errors influence the measurement uncertainty:

- The harmonics of the sweep generator
- The matching of the insertion points
- The square-law characteristic of the detectors
- The sweep generator level stability

3.2.6. Vector Measurement. Vector measurements enable characterization of a two-port circuit completely. In addition to the magnitude, the phase of the scattering parameters is also determined. There are two major concepts for measuring the complex parameters of a two-port device: the vector network analyzer and the six-port technique.

Modern vector network analyzers [8,39,40] measure all four scattering parameters—$s_{11}$, $s_{21}$, $s_{12}$, and $s_{22}$—without the necessity of turning the DUT around. Therefore they are symmetrical (Fig. 27) and measure in both directions. The basic concept looks similar to that of the scalar network analyzer. The signal of the generator is divided into reference and measurement paths. In forward measurements, the directional bridge $A$ determines the reflected signal, bridge $B$ determines the transmitted signal,
and vice versa for the reverse case. Instead of using diode detectors, the signals are downconverted to an intermediate frequency and analyzed in magnitude and phase. Synthesized sweep generators and synchronous detection are being used to obtain high accuracy for magnitude and phase measurements. Because the complex signals are measured, the main errors due to component imperfections may be corrected. Frequently a 12-term error model is applied to correct the source and load match, the bridge characteristics, the transmission leakage crosstalk, and downconverter characteristics. In the first phase well-known standards (e.g., open, short, line) are measured, and the 12 error parameters are determined. In the second phase the DUT is measured and the data corrected according to the calculated error terms. Several different techniques for measuring the error parameters are used, such as open–short–load, transmission–reflect line, and line–reflect line. Each technique uses different kinds of reference standards, such as short and open circuits, well-defined lines, and known loads.

Excellent performance is achieved by using the 12-term error correction technique. For example, at 20 GHz, load and source match better than $-38$ dB return loss, transmission tracking is better than 0.06 dB and crosstalk is less than $-104$ dB. As a result a 30 dB attenuator can be measured at 20 GHz with an uncertainty of 0.07 dB.

Vector network analyzers are commercially available in coaxial configurations in frequency bands from about 100 kHz to 50 GHz and in waveguides up to 110 GHz. Some specially dedicated systems operate in waveguides up to 1000 GHz.

The measuring uncertainty is defined mainly by the following parameters:

Accuracy of the reference standards
Stability of the generator and of the detection system
Stability of the connection cables
Repeatability of the connectors
Accuracy of the built-in software to calculate the error parameters and DUT scattering parameters

The six-port technique is another method for measuring the complex scattering parameters of a device. The magnitude and the phase of the signal are calculated from four scalar power measurements made with detectors arranged as shown in Fig. 28 [8,40].

The four power sensors gather enough information to calculate the magnitude and phase of the reflection of the DUT and the power launched into it. The calibration of the six-port device is rather complicated because a set of quadratic equations has to be solved. The quadratic equations can be linearized and solved for both calibration and measurement [8,40].

The simplicity of the detection system is an advantage of the six-port device especially for wideband applications and very high operating frequencies. Compared to the vector network analyzer, the six-port device requires more calibration and more complicated mathematics.

Two six-port devices connected in the configuration shown in Fig. 29 are required to provide attenuation measurements.

The dividing circuit includes phase adjustments to obtain different ratios $b_1/b_2$ at the terminals of the DUT. Using state-of-the-art power sensors the dynamic range of a dual six-port device is as large as 60 dB.

![Figure 27. Principle of a vector network analyzer.](image)

![Figure 28. Principle of six-port technique for reflection measurement.](image)
To achieve maximum accuracy, through-connection, reflection line (TRL) calibration is frequently used. Because the six-port device determines the complex parameters during the calibration process, the test ports appear to be well matched. The measurement uncertainties are primarily limited by the calibration standards, mainly the reflection standard (short or open). Real-time operation is limited by the computing time and the response time of the power sensors.

3.3. Fiberoptics

Three main methods [41,42] are used for attenuation measurements: the insertion loss technique, the cutback technique, and the backscattering method. The first two methods perform two-point (end-to-end) measurements and the last performs a one-ended characterization. Some of the methods are standardized [43,44].

3.3.1. Insertion Loss Method. The insertion loss technique consists of a stable laser source and a stable, accurate, optical power meter. The power $P_2$ of the laser source is sent into the DUT (e.g., an optical fiber), and the power $P_3$ is measured at the far end. The attenuation is given by the ratio of the two power levels as

$$A(\text{dB}) = 10 \log \left( \frac{P_3}{P_2} \right)$$

(28)

To achieve more accurate measurements in the first phase, the power of the source is directly measured and is remeasured in the second phase with the DUT inserted. More sophisticated measuring systems use a configuration shown in Fig. 30.

A second power sensor measures the power level of the source instantaneously by a power divider. In this configuration the power stability of the source is less important because $P_1$ is always used as a reference. By using cooled detectors, a dynamic range of $\leq 90$ dB is achieved.

The accuracy of the measurements are determined by the following factors:

- The power level and wavelength stability of the source
- The calibration and stability of the power sensors
- The reproducibility of the connectors
- The linearity of the detectors

The measurement uncertainties for the insertion loss technique are on the order of 0.9 dB including the connector reproducibility. Sophisticated systems reach over a limited dynamic range of 50 dB and uncertainty of 0.2 dB.

3.3.2. Cutback Method. The cutback method [41,45] (Fig. 31) is the most accurate technique, but it is destructive. This method was developed to measure the attenuation of fibers as a function of the wavelength. Using a light source combined with a monochromator, a fiber can be tested at any wavelength from 800 to 1600 nm with a spectral width of 3 nm. The light from the source is projected into the fiber by a lens.

The power $P_2(\lambda)$ is measured at the far end of the fiber (test length $l_t$) by using a cooled detector. Then the fiber is cut back to a short length of 2–3 m without changing the projecting conditions, and the power $P_1(\lambda)$ is recorded. If the power loss in the short length of fiber is assumed to be negligible, the attenuation is given by the following equation:

$$A(\lambda) = 10 \log \left( \frac{P_2(\lambda)}{P_1(\lambda)} \right)$$

(29)

Assuming a uniform fiber, the attenuation coefficient per unit length of the fiber is given by

$$\alpha(\lambda) = \frac{A(\lambda)}{l_t - l_r}$$

(30)

where $l_t$ and $l_r$ are given in kilometers. The achieved uncertainty for cable length of several kilometers is about 0.02 dB/km for multimode fibers and 0.004 dB/km for single mode.

3.3.3. Backscattering Method. The backscattering method is a one-ended measurement based on Rayleigh scattering in an optical fiber [41,46]. Figure 32 shows
the principle of an optical time-domain reflectometer (OTDR).

A laser pulse is projected into the fiber by a coupler, and the backscattered power is measured. The backscattered power is related to the attenuation loss of the fiber, and the measured time delay is related to the distance in the fiber. The attenuation is calculated by using two values of the backscattered power at different time delays (different locations along the fiber). The OTDR has the advantage of providing attenuation and reflection information along the fiber. A typical recording is shown in Fig. 33.

The length of the pulse is responsible for the deadzone, where no measurement is possible. A lead-in fiber allows masking of the deadzone. From the measured data details in the fiber path, such as connector loss or splice loss, irregularities of attenuation and defects can be analyzed.

Commercially available OTDRs have a dynamic range of about 35 dB and cover distances up to 65 km depending on the fiber loss. A well-calibrated OTDR can produce a measurement uncertainty of about 0.02 dB/km.

4. ERRORS AND UNCERTAINITIES IN ATTENUATION MEASUREMENTS

Whenever measurements are made, the results differ from the true or theoretically correct values. The differences are the result of errors in the measurement system, and it should be the aim to minimize these errors. In practice there are limits because no measurement instruments operate perfectly. A statement of measurement uncertainty reflects the quality of the measured results, and it has to be accompanied by a statement of confidence.

The International Committee for Weights and Measures (CIPM) [47] has published a guide for expressing uncertainty in measurements which has been adopted by the European Cooperation for Accreditation of Laboratories (EA) [48]. According to the guide, uncertainty is grouped in two categories: type A and type B:

Type A evaluation is the result of statistical analysis of a series of repeated observations and therefore includes random effects.

Type B evaluation is by definition other than type A, for example, judgment based on data of calibration certificates, experiences with instruments, and manufacturers' specifications.

### 4.1. Type A Evaluation of Uncertainty Components

Random effects result in errors that vary unpredictably. For an estimate of the standard deviation $s(q_k)$ of a series of $n$ readings, $q_k$ is obtained from

$$s(q_k) = \sqrt{\frac{1}{(n-1)} \sum_{k=1}^{n} (q_k - \bar{q})^2}$$  \hspace{1cm} (31)

where $\bar{q}$ is the mean of $n$ measurements.

The random component of uncertainty is reduced by repeating the measurements. This yields the standard deviation of the mean $s(\bar{q})$

$$s(\bar{q}) = \frac{s(q_k)}{\sqrt{n}}$$  \hspace{1cm} (32)

The standard uncertainty of the input estimate $\bar{q}$ is the experimental standard deviation of the mean (for $n \geq 10$)

$$u(\bar{q}) = s(\bar{q})$$  \hspace{1cm} (33)

### 4.2. Type B Evaluation of Uncertainty Components

Systematic effects that remain constant during measurements but change if the measurement conditions are altered cannot be corrected and therefore contribute to uncertainty. Other contributions arise from errors that are not possible or impractical to correct for, such as from calibration certificates or manufacturers' specifications. Most of these contributions are adequately represented by a symmetrical distribution. In RF metrology three main distributions are of interest: normal, rectangular, and U-shaped.

![Figure 31. Principle of the cutback method.](image1)

![Figure 32. Principle of an optical time-domain reflectometer.](image2)
4.2.1. Normal Distribution. Uncertainties derived from multiple contributions are assumed to be normally distributed. Accredited calibration laboratories issue calibration certificates calculated for a normal distribution and a minimum level of confidence of 95% (approximate coverage factor \( k = 2 \)). The standard uncertainty associated with the estimate \( x_i \) is given by
\[
u(x_i) = \frac{\text{uncertainty}}{k} \tag{34}
\]

4.2.2. Rectangular Distribution. This means that there is equal probability that the true value lies between limits. This is the case for most manufacturers’ specifications that give a semi-range limit \( a_i \):
\[
u(x_i) = \frac{a_i}{\sqrt{3}} \tag{35}
\]

4.2.3. U-Shaped Distribution. This distribution is applicable to mismatch uncertainty [49]. Because the phases of the reflection coefficients (of source, DUT, load) in scalar measurement are not known, the mismatch loss has to be taken into account as an uncertainty. The mismatch uncertainty is asymmetric to the measured result, and normally the larger of the two limits \( M = 20 \log (1 - |\Gamma_G| |\Gamma_L|) \) is used. The standard uncertainty is calculated as
\[
u(x_i) = \frac{M}{\sqrt{2}} \tag{36}
\]

4.3. Combined Standard Uncertainty

The combined standard uncertainty for uncorrelated input quantities is calculated as the square root of the sum of the squares of the individual standard uncertainties:
\[
u_c(y) = \sqrt{\sum_{i=1}^{n} u_i^2(y)} \tag{37}
\]

4.4. Expanded Uncertainty

The expanded uncertainty \( U \) defines an interval in which there is the true value with a specified confidence level. Normally accredited calibration laboratories are asked to use the coverage factor \( k = 2 \) (approximately 95% confidence level), giving
\[
U = k \cdot u_c(y)
\]

4.5. Uncertainty in Attenuation Measurement

Let us assume a simple attenuation measuring setup, shown in Fig. 34, consisting of a generator, two matching circuits, and a receiver. In the first phase, when the two insertion points are clamped together, the receiver measures \( P_1(f) \) (often called a normalization). In the second phase the DUT is inserted, and the receiver reads the values \( P_2(f) \).

Attenuation as a function of the frequency is calculated from the ratio of the two sets of readings:
\[
A(f) = 10 \log \left[ \frac{P_2(f)}{P_1(f)} \right] \tag{38}
\]

The following errors contribute to the uncertainty of the measurement:

The statistical errors of \( n \) repeated measurements (type A) are given by the arithmetic experimental standard deviation:
\[
s(A) = \frac{1}{(n-1)} \sum_{k=1}^{n} (A_k - \bar{A})^2 \tag{39}
\]

(\( \bar{A} \) is the arithmetic mean of the measurements)
The standard uncertainty is calculated from
\[
u_s(A) = s(\bar{A}) = \frac{s(A)}{\sqrt{n}} \tag{40}
\]

The generator level stability \( a_G \) is taken from the manufacturer’s specification and is assumed to be rectangularly distributed. The uncertainty is calculated

Figure 33. Typical backscattering signature of a fiber.

Figure 34. Example of an attenuation measurement system.
as follows:

\[ u_G = \frac{a_G}{\sqrt{3}} \]  \hspace{1cm} (41)

The receiver level linearity and stability \( a_R \) is taken from the manufacturer’s specification. The uncertainty is calculated as

\[ u_R = \frac{a_R}{\sqrt{3}} \]  \hspace{1cm} (42)

The noise level of the receiver influences the measurement of high attenuation values. It is given in the manufacturer’s specification and contributes to the uncertainty as

\[ u_N = \frac{a_N}{\sqrt{3}} \]  \hspace{1cm} (43)

The crosstalk of the measurement system \( a_l \) is determined by measurements and regarded as limits, and therefore contributes to the uncertainty as

\[ u_l = \frac{a_l}{\sqrt{3}} \]  \hspace{1cm} (44)

Two mismatch losses have to be taken into account, one during the normalization (often also called calibration) phase and the second while measuring the DUT.

4.5.1. Normalization Phase. The maximum mismatch loss [49] is calculated from the reflection coefficients of the source and the receiver as

\[ M_C = 20 \log(1 - |r_G||r_L|) \]  \hspace{1cm} (45)

As in scalar measurements, the phases of the reflection coefficients are unknown. The mismatch loss contributes to the measurement uncertainty and is normally assumed to be U-shaped-distributed:

\[ u_C = \frac{M_C}{\sqrt{2}} \]  \hspace{1cm} (46)

4.5.2. Measurement Phase. There are two mismatch losses [49] that have to be considered: one between the generator and the input of the DUT and the other between the output of the DUT and the receiver. In addition, for small attenuation values the interaction between the input and the output connections has to be considered. The maximum limits of the mismatch loss that have to be used for the uncertainty are given by

\[ M_m = 20 \log \frac{|1 - |r_{G21}||r_{L22}|| - |r_{G11}||r_{L22}|| - |r_{G11}||r_{L21}||r_{21}||r_{12}||}{1 - |r_{G11}||r_{L22}||} \]  \hspace{1cm} (47)

The uncertainty is given by

\[ u_m = \frac{M_m}{\sqrt{2}} \]  \hspace{1cm} (48)

The total uncertainty is calculated either from linear values or from decibel values as long as they are small:

\[ u_c(A) = \sqrt{u_m^2 + u_C^2 + u_R^2 + u_N^2 + u_{L1}^2 + u_{L1}^2} \]  \hspace{1cm} (49)

The expanded uncertainty is calculated using a coverage factor \( k = 2 \) (approximately 95% confidence level) as

\[ U(A) = k \cdot u_c(A) = 2 \cdot u_c(A) \]  \hspace{1cm} (50)

The uncertainty has to be calculated for all the measurement frequencies to find the maximum value of the uncertainty.

**BIBLIOGRAPHY**


34. G. F. Engen and R. W. Beatty, Microwave attenuation measurements with accuracies from 0.0001 to 0.06 dB over a range of 0.01 to 50 dB, *J. Res. NBS* 64C:139–145 (1960).


**ATTENUATORS**

RAJI SUNDARARAJN
EDWARD PETERSON
ROBERT NOWLIN
Arizona State University East Mesa, Arizona

Attenuators are linear, passive, or active networks or devices that attenuate electrical or microwave signals, such as voltages or currents, in a system by a predetermined ratio. They may be in the form of transmission-line, stripline, or waveguide components. Attenuation is usually expressed as the ratio of input power ($P_{in}$) to output power ($P_{out}$), in decibels (dB), as

$$\text{Attenuation (dB)} = 10 \log_{10} \frac{P_{in}}{P_{out}} = 20 \log_{10} \frac{E_{in}}{E_{out}} = 20 \log_{10} \frac{E_1}{E_2}$$

(1)

This is derived from the standard definition of attenuation in Nepers (Np), as

$$\text{Attenuation (Np)} = z \Delta x = - \ln \frac{E_2}{E_1}$$

(2)

where $z$ is attenuation constant (Np/m) and $\Delta x$ is the distance between the voltages of interest ($E_1$ and $E_2$).
Figure 1 illustrates this concept. The relation between Np and dB is

\[ 1 \text{Np} = 8.686 \text{ dB} \]  

Here the load and source are matched to the characteristic impedance. The decibels are converted to the attenuation ratio as follows: \( \frac{P_{\text{in}}}{P_{\text{out}}} = \log_{10} \frac{V_{\text{in}}}{V_{\text{out}}} = 10 \text{ dB} \).

The most commonly used method in attenuators is to place resistors at the center of an electric field. Due to the electric field, there is current induced, resulting in ohmic loss.

1. APPLICATION

There are many instances when it is necessary to reduce the value, or level, of electrical or microwave signals (such as voltages and currents) by a fixed amount to allow the rest of the system to work properly. Attenuators are used for this purpose. For example, in turning down the volume on a radio, we make use of a variable attenuator to reduce the signal. Almost all electronic instruments use attenuators to allow for the measurement of a wide range of voltage and current values, such as voltmeters, oscilloscopes, and other electronic instruments. Thus, the various applications in which attenuators are used include the following:

- To reduce signal levels to prevent overloading
- To match source and load impedances to reduce their interaction
- To measure loss or gain of two-port devices
- To provide isolation between circuit components, or circuits or instruments so as to reduce interaction among them
- To extend the dynamic range of equipment and prevent burnout or overloading equipment

2. TYPES

There are various types of attenuators based on the nature of circuit elements used, type of configuration, and kind of adjustment. They are as follows:

- Passive and active attenuators
- Absorptive and reflective attenuators
- Fixed and variable attenuators

A fixed attenuator is used when the attenuation is constant. Variable attenuators have varying attenuation, using varying resistances for instance. The variability can be in steps or continuous, obtained either manually or programmably. There are also electronically variable attenuators. They are reversible, except in special cases, such as a high-power attenuator. They are linear, resistive, or reactive, and are normally symmetric in impedance. They include waveguide, coaxial, and striplines, as well as calibrated and uncalibrated versions. Figures 2–4 show fixed, manual step, and continuously variable commercial attenuators, respectively.

Based on their usage, IEEE Std 474 classifies them as:

- Class I: Standard
- Class II: Precision
- Class III: General-purpose
- Class IV: Utility

Typical commercial attenuators are listed below:

WA 1 (0–12.4 GHz), WA 2 (0–3 GHz), coaxial, fixed attenuators: 1–60 dB; 5 W av./1 kW peak
WA 115A manual step attenuators: 0–18 GHz, 0–9 dB, 1-dB steps
VA/02/100 continuously variable attenuators, resistive, 0–2 GHz, 5 W av./0.5 kW peak
HP 84904L programmable step attenuator, direct current (DC) to 40 GHz, 0–11 dB, 1-dB steps
HP 84906K programmable step attenuator, DC to 26.5 GHz, 0–90 dB, 10-dB steps
HP 84904L programmable step attenuator, DC to 40 GHz, 0–70 dB, 10-dB steps
HP 8495B manual step attenuator, DC to 18 GHz, 0–70 dB, 10-dB steps
HP 355F programmable step attenuator, DC to 1 GHz, 0–120 dB, 10-dB steps
HP 8493A coaxial fixed attenuator, DC to 12.4 GHz

Based on their utility, military attenuators are classified as:

- **Class I**: For use as a primary standard
- **Class II**: For use as a secondary standard, and in lab and precision test equipment
  - A—with lumped constant or distributed shunt and series elements
  - B—with lossy-line pads
- **Class III**: For use in general field equipment
- **Class IV**: For use in equipment where precision and stability are secondary considerations

Typical military specifications for fixed coaxial attenuators are as follows:

- Mil-A-3933/1: attenuators, fixed, coaxial line, DC to 3 GHz, class IIA, low power
- Mil-A-3933/2: attenuators, fixed, coaxial line, 1–4 GHz, class IIB, medium power
- Mil-A-3933/10: attenuators, fixed, coaxial line, DC to 18 GHz, class III, medium power
- Mil-A-3933/26: attenuators, fixed, coaxial line, 0.4–18 GHz, class IV low power

### 3. SPECIFICATIONS

To specify an attenuator, the purpose of the attenuator should be known. Attenuators are used to provide protection, reduce power, and extend the dynamic range of the test equipment. In choosing an attenuator, the frequency range of operation should be considered since the accuracy depends on the frequency. Attenuation involves placing resistive material to absorb the signal’s electric field. This means, there will always be some reflection. So, attenuators must be designed to minimize reflection. This is quantified in terms of voltage standing-wave ratio (VSWR). Another factor to be considered is the insertion loss, which is the ratio of power levels with and without the component insertion. If it is a variable step attenuator, the step size is to be known. Thus, the parameters available in the specs are as follows:

- **dB rating**
- **VSWR**
- **Accuracy**
- **Power rating**
- **Stepsize (if variable)**
- **Frequency band**
- **Degree of stability (measured by the change in attenuation due to temperature, humidity, frequency, and power level variations)**
- **Characteristic impedance of attenuator**
- **Repeatability**
- **Life**
- **Degree of resolution (difference between actual attenuation and measured value)**

The definitions of various parameters used in selecting attenuators are given below.

#### 3.1. Electrical Parameters and Definitions (from MIL-HDBK-216)

**Attenuation.** A general transmission term used to indicate a decrease in signal magnitude. This decrease in power is commonly expressed in decibels (dB) as

\[
\text{Attenuation (A)} = 10 \log_{10} \frac{P_{\text{in}}}{P_{\text{out}}}
\]

**Deviation of Attenuation from Normal.** Difference in actual attenuation from the nominal value at 23°C and an input power of 10 mW at a specified reference frequency or frequency range. When used in a frequency range, it involves the frequency sensitivity.

**Frequency Sensitivity.** This is the peak-to-peak variation in the loss of the attenuator through the specified frequency range.

**Frequency Range.** Range of frequency over which the accuracy of attenuator is specified.

**Insertion Loss.** Amount of power loss due to the insertion of the attenuator in the transmission system. It is expressed as a ratio of the power delivered to that part of the system following the attenuator, before and after the insertion.
Characteristic Insertion Loss. This is the insertion loss in a transmission line or waveguide that is reflectionless in both directions from the inserted attenuator.

Power-Handling Capabilities. Maximum power that can be applied to the attenuator under specified conditions and durations without producing a permanent change in the performance characteristics that would be outside of specification limits.

Power Sensitivity. This is the temporary variation in attenuation (dB/W) under steady-state conditions when the input power is varied from 10 mW to maximum input power.

Stability of Attenuation. Capability of attenuator to retain its parameters when subjected to various environmental conditions.

Operating Temperature Range. Temperature range of the attenuator can be operated with maximum input power.

Temperature Sensitivity. Temperature variation in attenuation [dB/(dB°C)] over the operating range.

Input VSWR. This is the level of reflected signal created at the attenuator input when the output is terminated with a load with the same characteristic impedance as the source.

Output VSWR. This is the level of reflected signal created at the attenuator output when the input is terminated with a load with the same characteristic impedance as the source.

4. PASSIVE ATTENUATORS

4.1. Resistance Networks for Attenuators

Typically T, pi, or L designs are used for attenuators. Figure 5 shows four commonly used symmetric (input and output resistors of equal value) configurations. The formulas for the resistance values in ohms for these pads when the characteristic resistance $R_0 = 1\Omega$ are given below. If $R_0$ is other than $1\Omega$, multiply each of the resistance values ($a$, $b$, $c$, $1/a$, $1/b$, and $1/c$) by $R_0$, where

\[
a = \frac{10^{\text{dB}/20} - 1}{10^{\text{dB}/20} + 1}
\]

\[
b = \frac{2 \times 10^{\text{dB}/20}}{10^{\text{dB}/10} - 1}
\]

\[
c = (10^{\text{dB}/20} - 1)
\]

Simple wirewound resistors are used in audio applications. Nonreactive wirewound resistors, such as mica card, Ayrton–Perry winding, woven resistors are used for high frequencies. For coaxial applications (over 26.5 GHz), thin-film resistors are used. For higher frequencies, distributive resistive films, such as nichrome alloy film, on a high-quality ceramic substrate, such as alumina or sapphire, is used. An unsymmetrical pad is shown in Fig. 6, and the formulas for this pad are

\[
j = \frac{R_1 - kR_2}{k + R_2}
\]

\[
k = \left[ \frac{R_1R_2}{(R_1 - R_2)} \right]^{1/2} \quad \text{where } R_1 > R_2
\]

Minimum loss (dB) = 20 log \left\{ \left[ \frac{(R_1 - R_2)}{R_2} \right]^{1/2} + \left( \frac{R_1}{R_2} \right)^{1/2} \right\}

Figure 5. Symmetric pads with matched impedances: (a) T pad; (b) pi pad; (c) bridged T pad; (d) balanced pad.

Figure 6. Unsymmetric matching L attenuator.
Typical values for the pads in Fig. 5 are shown in Table 1, and those of Fig. 6 are shown in Table 2.

For a broadband match between impedances $R_1$ and $R_2$, use the minimum-loss L pad (Fig. 6).

### 4.2. Power Dissipation within a T Pad

Table 3 lists values of power dissipation within a T pad. The values are for an input of 1 W; for other input powers, multiply the values by the input power.

### 5. INSERTION LOSS

An attenuator is used to introduce attenuation between a source and a load. Due to the introduction of the attenuator, there is change in the current. This loss is designated as insertion loss, which depends on the configuration. Usually, the load and source impedances are matched. Figure 7 illustrates this concept. If $I_{L0}$ is the load current without the attenuator pad and $I_L$ is the current with the attenuator pad, then the ratio $I_L/I_{L0}$ is called the **insertion loss**, one of the parameters of the attenuators. Figure 7a shows the source and load connected without an attenuator, and Fig. 7b shows the same system with an attenuator. (The quantities $I_L$, $R_{in}$, and $R_{out}$ depend on the attenuator configuration.) The quantities insertion loss ($I_L$), input resistance ($R_{in}$), and output resistance ($R_{out}$) depend on the attenuator configuration. The value of each of the three resistors of the T (Fig. 8) and pi (Fig. 9) attenuators can be chosen independently of others. This enables the three-design criteria of input resistance, output resistance, and insertion loss to be met. In many situations, the only function of the pad is to provide matching between source and load; and although attenua-

### Table 1. Resistance Values for Attenuator Pads When $R_0 = 1 \Omega$

<table>
<thead>
<tr>
<th>dB</th>
<th>$a$</th>
<th>$b$</th>
<th>$1/b$</th>
<th>$1/a$</th>
<th>$c$</th>
<th>$1/c$</th>
<th>$a$</th>
<th>$1/a$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.1</td>
<td>0.0057567</td>
<td>86.853</td>
<td>0.011514</td>
<td>173.71</td>
<td>0.011580</td>
<td>86.356</td>
<td>0.0057567</td>
<td>173.71</td>
</tr>
<tr>
<td>0.2</td>
<td>0.011513</td>
<td>43.424</td>
<td>0.023029</td>
<td>86.859</td>
<td>0.023294</td>
<td>42.930</td>
<td>0.011513</td>
<td>43.424</td>
</tr>
<tr>
<td>0.3</td>
<td>0.017268</td>
<td>28.947</td>
<td>0.034546</td>
<td>57.910</td>
<td>0.035143</td>
<td>28.455</td>
<td>0.017268</td>
<td>28.947</td>
</tr>
<tr>
<td>0.4</td>
<td>0.023022</td>
<td>21.707</td>
<td>0.046068</td>
<td>43.438</td>
<td>0.047128</td>
<td>21.219</td>
<td>0.023022</td>
<td>21.707</td>
</tr>
<tr>
<td>0.5</td>
<td>0.028775</td>
<td>17.362</td>
<td>0.057597</td>
<td>34.753</td>
<td>0.059254</td>
<td>16.877</td>
<td>0.028775</td>
<td>17.362</td>
</tr>
<tr>
<td>0.6</td>
<td>0.034525</td>
<td>14.465</td>
<td>0.069132</td>
<td>28.965</td>
<td>0.071519</td>
<td>13.915</td>
<td>0.034525</td>
<td>14.465</td>
</tr>
<tr>
<td>0.7</td>
<td>0.040274</td>
<td>12.395</td>
<td>0.080678</td>
<td>21.730</td>
<td>0.089647</td>
<td>10.365</td>
<td>0.040274</td>
<td>12.395</td>
</tr>
<tr>
<td>0.8</td>
<td>0.046019</td>
<td>10.842</td>
<td>0.092234</td>
<td>17.391</td>
<td>0.10918</td>
<td>9.1596</td>
<td>0.046019</td>
<td>10.842</td>
</tr>
</tbody>
</table>

$^a$If $R_0 \neq 1 \Omega$, multiply all values by $R_0$. (From Ref. data for Radio Engineers, 1985.)

$^b$For other decibel values, use formulas in text.

$^c$These values have been multiplied by $10^3$. 

456 ATTENUATORS
The T attenuator contains resistors $R_1$, $R_2$, and $R_3$; these form a T configuration, as shown in Fig. 6. Insertion loss is usually measured in dB, defined as $I_L(dB) = -20\log I_L$ or $|20\log I_L|$, the amount of attenuation required. The insertion loss $I_L$ is given as

$$I_L(dB) = \frac{I_L}{I_{0}} = \frac{R_3(R_3 + R_L)}{(R_S + R_1 + R_3)(R_2 + R_3 + R_L) - R_3^2}$$  \hspace{1cm} (10)$$

The input and the output of resistances of the attenuator are given by

$$R_{in} = R_1 + \frac{R_3(R_2 + R_L)}{R_2 + R_3 + R_L}$$  \hspace{1cm} (11)$$

and

$$R_{out} = R_2 + \frac{R_3(R_1 + R_S)}{R_1 + R_3 + R_S}$$  \hspace{1cm} (12)$$

5.1 T Attenuator Insertion Loss

The T attenuator also has to match the load and the source impedance. In this case, $R_1 = R_2 = R$ and $R_{in} = R_{out} = R_0$. Thus

$$R_0 = R + \frac{R_3(R + R_0)}{(R_3 + R + R_0)}$$  \hspace{1cm} (13)$$

and the insertion loss is given by

$$I_L = \frac{R_3}{R_3 + R + R_0}$$  \hspace{1cm} (14)$$

and

$$R = R_0 \frac{1 - I_L}{1 + I_L}$$  \hspace{1cm} (15)$$
Example 1 (T Attenuator). A T-type attenuator is required to provide 3 dB insertion loss and to match 50 Ω input and output. Find the resistor values, using the following equations:

\[
R_3 = \frac{2R_0 I_L}{1 - (I_L)^2} \tag{16}
\]

Check:

\[
I_L = \frac{R_3}{R_3 + R_0 + R_L} = \frac{141.6}{141.6 + 8.55 + 50} = 0.708
\]

5.1.1. The Pi Attenuator Insertion Loss. Figure 9 shows a pi attenuator formed by resistors \( R_a, R_b, \) and \( R_c. \) The insertion loss and conductances \( G_{in} \) and \( G_{out} \) are given by

\[
I_L = G_c \left( \frac{G_S + G_{L}}{G_S + G_a + G_c (G_b + G_c + G_L) - G_c^2} \right) \tag{17}
\]

\[
G_{in} = G_a + \frac{G_c G_b + G_c + G_L}{G_b + G_c + G_L} \tag{18}
\]

\[
G_{out} = G_b + \frac{G_c (G_a + G_S)}{G_a + G_c + G_S} \tag{19}
\]

where \( G = 1/R; \) thus \( G_{L} = 1/R_L \) and so on.

The same pi attenuator can be realized using a T attenuator with \( R_1, R_2, \) and \( R_3 \) values using the Y–Δ

Figure 7. Definition of characteristic insertion loss: (a) original setup without attenuator; (b) original setup with attenuator between source and load.

Figure 8. T attenuator configuration.

Figure 9. Pi attenuator configuration.

Figure 10. L attenuator configuration: (a) \( R_s < R_t; \) (b) \( R_s > R_t. \)
transformation:
\[ R_a = \frac{R_1 R_2 + R_1 R_3 + R_2 R_3}{R_2} \]  
(20)
\[ R_b = \frac{R_a R_2}{R_1} \]  
(21)
\[ R_c = \frac{R_a R_2}{R_3} \]  
(22)

The selection between pi and T is based on the value of resistors that can be used in practice. With matching source and load impedances, the values of the pi attenuator are

\[ R_a = R_b = R_0 \frac{1 + I_L}{1 - I_L} \]  
(23)

and

\[ R_c = R_0 \frac{1 - (I_L)^2}{2I_L} \]  
(24)

**Example 2 (Pi Attenuator).** Repeat Example 1 using a pi attenuator

\[ R_A = R_B = R_0 \frac{1 + I_L}{1 - I_L} = 50 \frac{1 + 0.708}{1 - 0.708} = 292.46 \Omega \]

\[ R_C = R_0 \frac{1 - (I_C)^2}{2I_L} = 50 \frac{1 - (0.708)^2}{2 \times 0.708} = 17.61 \Omega \]

using the following equations:

**5.1.2. The L Attenuator Insertion Loss.** An L attenuator can be derived from a T or pi attenuator by removing one resistor. As shown in Fig. 10, two configurations are obtained depending on \( R_S > R_L \) or \( R_S < R_L \). Simple circuit theory shows that for \( R_S > R_L \), we have

\[ R_S = R_{in} = R_1 + \frac{R_3 R_L}{R_3 + R_L} \]  
(25)

and

\[ R_L = R_{out} = \frac{R_3 (R_1 + R_S)}{R_3 + R_1 + R_S} \]  
(26)

from which it can be shown that

\[ R_1 = \sqrt{R_S (R_S - R_L)} \]  
(27)

and

\[ R_3 = \frac{R_S^2 - R_1^2}{R_1} \]  
(28)

and when we put \( R_2 = 0 \), the insertion loss is calculated as

\[ I_L = \frac{R_3 (R_S + R_L)}{(R_S + R_1 + R_3)(R_3 + R_L) - R_3^2} \]  
(29)

**Example 3.** Design an L attenuator to match a 300-\( \Omega \) source to a 50-\( \Omega \) load and determine insertion loss. Here \( R_S > R_L \), using the following equation:

\[ R_1 = \sqrt{R_S (R_S - R_L)} = \sqrt{300(300 - 50)} = \sqrt{300 \times 250} = 273.86 \Omega \]

Using the following equation:

\[ R_3 = \frac{R_S^2 - R_1^2}{R_1} = \frac{300^2 - 273.86^2}{273.86} = 54.775 \Omega \]

\[ R_L = \frac{R_3 (R_S + R_L)}{(R_S + R_1 + R_3)(R_3 + R_L) - R_3^2} \]

= \frac{54.775 (300 + 50)}{(300 + 273.86 + 54.775)(54.775 + 50) - (54.775)^2}

= 0.305

\[ R_L_{(dB)} = -20 \log 0.305 = 10.3 \text{ dB} \]

For \( R_S < R_1 \), we have

\[ R_{in} = \frac{R_3 (R_2 + R_L)}{R_2 + R_3 + R_L} \]  
(30)

and

\[ R_{out} = R_2 + \frac{R_3 R_S}{R_3 + R_S} \]  
(31)

and

\[ R_2 = \sqrt{R_L (R_L - R_S)} \]  
(32)

and

\[ R_3 = \frac{R_L^2 - R_2^2}{R_2} \]  
(33)

The corresponding insertion loss is

\[ I_L = \frac{R_3 (R_S + R_L)}{(R_S + R_3)(R_2 + R_3 - R_L) - R_3^2} \]  
(34)

**Example 4.** Design an L attenuator to match 50-\( \Omega \) source to 75-\( \Omega \) load and determine the insertion loss \( R_S < R_L \)
using the following equation:

\[ R_2 = \sqrt{R_L(R_L - R_S)} = \sqrt{75(75 - 50)} = 43.3 \Omega \]

using the following equations:

\[ R_3 = \frac{R^2_1 - R^2_2}{R_2} = \frac{75^2 - 43.3^2}{43.3} = 86.6 \Omega \]

\[ R_L = \frac{R_3(R_3 + R_L)}{(R_S + R_3)(R_2 + R_3 + R_L) - R^2_3} \]

\[ = 0.0123 = 38.2 \text{ dB} \]

6. FIXED ATTENUATORS

Fixed attenuators, commonly known as “pads,” reduce the input signal power by a fixed amount, such as 3, 10, and 50 dB. For example, an input signal of 10 dBm (10 mW) passing through a 3-dB fixed attenuator will exit with a power of 10 dBm - 3 dB = 7 dBm (5 mW). Figure 2 shows a fixed coaxial commercial attenuator. A typical datasheet for a fixed coaxial attenuator is as follows (courtesy of Weinschel Associates):

- **Frequency**: 0–3 GHz
- **Attenuation**: 50 dB
- **Accuracy**: ±0.10 dB (DC)
  - ±0.15 dB (0–2 GHz)
  - ±0.13 dB (0–3 GHz)
- **VSWR**: 1.15 (0–1 GHz)
  - 1.20 (1–3 GHz)
- **Input power**: 1 W av., 1 kW peak at -30°C–70°C
- **Connectors**: Type N; St. St.; m, f
- **Length**: 68 mm (2.7 in.)
- **Diameter**: 210 mm (0.83 in.)
- **Weight**: 100 g (3.6 oz)
- **Power sensitivity**: <0.005 dB/dB × W; bidirectional in power
- **Temperature stability**: <0.0004 dB/dB × °C

6.1. Applications

Fixed attenuators are used in numerous applications. In general, they can be classified into two distinct categories:

1. Reduction in signal level
2. Impedance matching of a source and a load

Those in the first category are used in the following situations:

- Operation of a detector in its square-law range for most efficient operations.
- Testing of devices in their small signal range.

Those in the second category are used in the following situations:

- Reduction of signal variations as a function of frequency. The variations here are caused by a high VSWR. The attenuator provides a reduction in these variations and a better match.
- Reduction in frequency pulling (changing the source frequency by changing the load) of solid-state sources by high reflection loads.

6.2. Types

Based on construction, fixed attenuators are available in coaxial, waveguide, and stripline configurations. The various types are:

1. Waveguide vane
2. Rotary vane (fixed)
3. Directional coupler
4. T or pi
5. Lossy line
6. Distributed resistive film

6.2.1. Coaxial Fixed Attenuators. T or pi configurations are most commonly used both at low and high frequencies. At low frequencies, normal wirewound resistors are used. At high frequencies, thin-film resistors are used. Figures 11 and 12 show T and pi fixed attenuators. Thin-film resistors designed for microwave frequencies are used in place of carbon resistors. These resistors employ a nichrome alloy film on a high-quality ceramic substrate to ensure a firmly bonded film with low-temperature coefficients. This type of construction makes the resistors extremely stable at high frequencies. The skin effect of these resistors is excellent, used extensively in the microwave applications.

The T and pi configurations are obtained by placing the resistors in series on the center conductor and in shunt, contacting both the center and outer conductor. Thus, the T configuration can be fabricated with one shunt flanked by two series resistors and the pi configuration, with one series flanked by two shunt resistors. The series resistors in the T and pi configurations have less than 1 W capacity, thereby severely limiting the use of high-power applications, unless an elaborate heatsinking is provided. Power attenuators usually have huge sinks to handle high-power applications.

6.2.2. Resistive Card Attenuator. In a fixed dissipative, waveguide-type resistive card attenuator, the card is bonded in place (Fig. 13). It is tapered at both ends to maintain a low-input and low-output VSWR over the useful waveguide band. Maximum attenuation per length is obtained when the card is parallel to the E field
and at the center, where the $TE_{10}$ mode is maximum. The conductivity and the dimensions of the card are adjusted, by trial and error, to obtain the desired attenuation, which is a function of frequency. The attenuation increases with increase in frequency. In power applications, ceramic-type absorbing materials are used instead of a resistive card.

7. VARIABLE ATTENUATORS

A variable attenuator has a range, such as 0–20 dB or 0–100 dB. The variation can be continuous or in steps, obtained manually or programmably.

7.1. Step Attenuators

A series of fixed attenuators are mechanically arranged to offer discrete-step variation. The fixed attenuators are arranged in a rotatable drum or in a slab for switching between contacts. This arrangement provides discrete values of attenuation in each position and a high reliability factor. The step size can be 0.1, 1, or 10 dB. Stationary coaxial contacts provide the input and output of the device. These are used in applications requiring broadband flatness with low VSWR and satisfactory reseatability over ranges from 0 to 120 dB. Their application range is DC to 18 GHz.
7.2. Manual Step Attenuators

Figure 3 shows a manual step attenuator. A typical datasheet looks as follows:

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Attenuation</th>
<th>Stepsize</th>
<th>VSWR</th>
<th>Connectors</th>
<th>Height</th>
<th>Depth</th>
<th>Width</th>
</tr>
</thead>
<tbody>
<tr>
<td>0–4, 0–8, 0–12.4, 0–18 GHz</td>
<td>0–9, 0–60, 0–69</td>
<td>1, 10, 1 dB, respectively, for the range given above</td>
<td>1.20, 1.25, 1.40, 1.50 for the frequency ranges given above</td>
<td>N/SMA; St. St.</td>
<td>83 mm (3.3 in.)</td>
<td>79 mm (3.1 in.) (excludes shaft and knob)</td>
<td>65, 65, 118 mm (2.6, 2.6, 4.7 in.) for the three attenuation ranges given above</td>
</tr>
</tbody>
</table>

7.3. Continuously Variable Attenuators

Figure 4 shows a continuously variable attenuator. Typical specs are

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Connectors</th>
<th>Zero loss</th>
<th>Attenuation</th>
<th>Voltage</th>
<th>Speed</th>
<th>Power</th>
<th>RF connectors</th>
<th>Shipping weight</th>
</tr>
</thead>
<tbody>
<tr>
<td>1–18 GHz, 1 W av./1 kW peak</td>
<td>St. St., M, F; type N, SMA</td>
<td>Typically 0.5–1 dB</td>
<td>0–9, 0–60, 0–69 dB</td>
<td>20–30 V</td>
<td>&lt;20 ms</td>
<td>2.7 W</td>
<td>2.4 mm, F</td>
<td>291 g (10.3 oz)</td>
</tr>
</tbody>
</table>

The various types of continuously variable attenuators are

- Lossy wall
- Movable vane (Flap)
- Rotary vane
- Variable coupler
- Absorptive type
- Coaxial resistive film
- Variable T
- Waveguide below cutoff (piston)

7.4. Programmable and Solenoid Attenuators

7.4.1. Programmable. These are rapid switching attenuators with high accuracy and repeatability, useful for remote and computerized applications. Switching speeds can be as low as 30 ns. Two varieties of the programmable attenuators are the step-controlled and voltage-controlled types. The attenuation is varied by controlling the electrical signal applied to the attenuator. These signals can be in the form of either a biasing current or binary digit. The biasing can be pulses, square waves, or sine waves. A typical datasheet for coaxial programmable step attenuator is as follows:

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Attenuation</th>
<th>Maximum VSWR</th>
<th>Insertion loss</th>
<th>Repeatability</th>
<th>Power rating average</th>
<th>Peak</th>
<th>Maximum pulse width</th>
<th>Life</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC to 40 GHz</td>
<td>0–11 dB, in steps of 1 dB</td>
<td>1.3–12.4 GHz</td>
<td>0.8 dB + 0.04 GHz</td>
<td>0.03 dB</td>
<td>1 W</td>
<td>50 W</td>
<td>10 μs</td>
<td>5 million cycles per section</td>
</tr>
</tbody>
</table>

7.4.2. Solenoid. A typical datasheet would be as follows:

<table>
<thead>
<tr>
<th>Voltage</th>
<th>Speed</th>
<th>Power</th>
</tr>
</thead>
<tbody>
<tr>
<td>20–30 V</td>
<td>&lt;20 ms</td>
<td>2.7 W</td>
</tr>
</tbody>
</table>

7.5. Lossy Wall Attenuator

Figure 14 shows lossy wall variable attenuator. It consists of a glass vane coated with a lossy material, such as aquadag or carbon. For maximum attenuation, the vane is placed in the center of the guide’s wide dimension, where the electric field intensity is the maximum. A drive mechanism with a dial then shifts the vane away from the center so that the degree of attenuation is varied. This needs calibration by a precise attenuator. To match the attenuator to the waveguide, the vane can be tapered at each end; usually a taper of \( \lambda/2 \) provides an adequate match. Thus, it is frequency sensitive and the glass dielectric introduces appreciable phase shift.

Attenuation may also be obtained by inserting a resistive element through a shutter. The plane of the element lies in the distribution of the electric field across the wide dimension of the waveguide and the result is a degree of attenuation, which increases with the depth of insertion. However, due to the discontinuity, there is reflection of energy.

7.6. Movable-Vane (Flap) Attenuator

Figure 15 shows a waveguide variable, dissipative attenuator. The card enters the waveguide through the slot in the broad wall, thereby intercepting and absorbing a portion of the TE10 wave. The card penetration, and hence the attenuation, is controlled by means of the hinge arrangement to obtain variable attenuation. The ratings are typically 30 dB and are widely used in microwave equipment. However, the attenuation is frequency sensitive and the phase of the output signal is a function of card penetration and hence attenuation. This may result in nulling when the attenuator is part of a bridge network.

Since it is not simple to calculate the loss in dB, this type of attenuator has to be calibrated against a superior standard. To overcome these drawbacks, a rotary-vane attenuator is used.

7.7. Rotary-Vane Attenuator

The rotary-vane attenuator is a direct-reading precision attenuator that obeys a simple mathematical law, \( A = -20 \log \cos^2 \theta = -40 \log \cos \theta \) dB. As such, it is frequency-independent, which is a very attractive criterion.
for an attenuator. A functional diagram illustrates the operating principle of this attenuator. It consists of three sections of waveguide in tandem as shown (Fig. 16). A rectangular-to-circular waveguide transition containing a horizontal attenuator strip is connected to a rotatable circular waveguide containing an attenuator strip. This in turn is connected to a circular-to-rectangular waveguide transition containing a horizontal attenuator strip.

The incoming TE$_{10}$ mode is transformed into the TE$_{11}$ mode in the circular waveguide by the rectangular-to-circular waveguide transition with negligible reflections. The polarization of the TE$_{11}$ mode is such that the electric field is perpendicular to the thin resistive card in the transition section. As such, this resistive card has a negligible effect on the TE$_{11}$ mode. Since the resistive card in the center can be rotated, its orientation relative to the electric field of the incoming TE$_{11}$ mode can be varied so that the amount by which this mode is attenuated is adjustable.

When all the strips are aligned, the electric field of the applied wave is normal to the strips and hence no current flows in the attenuation strips and therefore no attenuation occurs. In a position where the central attenuation strip is rotated by an angle $\theta$, the electric field of the applied wave can be resolved into two orthogonally polarized modes; one perpendicular and one parallel to the resistive card. That portion which is parallel to the resistive slab will be absorbed, whereas the portion, which is polarized perpendicular to the slab, will be transmitted.

7.8. Variable Coupler Attenuator

These are basically directional couplers where the attenuation is varied by mechanically changing the coupling between two sections. This is accomplished by varying the spacing between coupled lines. These attenuators have a large range, high power handling capability, and retain calibration over a range of ambient conditions. They have a higher insertion loss at lower frequencies (Fig. 17).

7.9. Absorptive Attenuator

Figure 18 shows an absorptive variable attenuator. Attenuation is obtained by using a lossy dielectric material. The TEM electric field is concentrated in the vicinity of the center strip of the stripline. When the absorbing material is inserted in the high-field region, a portion of the TEM wave is intercepted and absorbed by the lossy dielectric. Thus, the attenuation increases. Since the characteristic impedance of the stripline changes with the dielectric
material insertion, the SWR tends to increase as the attenuation increases. To minimize this, the ends of the lossy material are tapered to provide a smooth impedance transformation into and out of the lossy section. SWR values of > 1.5 are possible over a limited frequency range. In general, the SWR deteriorates at low frequencies. The attenuation increases with increasing frequency for a fixed setting. This is another disadvantage, since this makes the calibration a cumbersome procedure. Compensation techniques are occasionally used to reduce this variation with frequency.

7.10. Coaxial Resistive Film Attenuator

Figure 19 shows a coaxial resistive film attenuator. In this configuration, if $r$ is the RF resistance per unit length, by adjusting the length $l$, the series resistance $R = rl$ of the center conductor is changed; thus, the attenuation is variable. If $I$ is the conduction current on the center conductor, the voltage drop is $V = RI = Irl$. If $E_i$ is the input voltage, then the output voltage is $E_0 = E_i - rll$ and the attenuation is

$$A = 20 \log \frac{E_i}{E_i - rll} \text{(dB)}$$

(35)

7.11. Variable T

The variable T attenuator is the same as the fixed attenuator except that the resistors are variable (Fig. 20). All three resistors are variable simultaneously to give good input/output VSWR.

7.12. Waveguide below Cutoff or Piston Attenuator

The simple principle of cutoff below frequency is used in the piston or the cutoff attenuator. The cylindrical waveguide used is operating at a frequency below cutoff. For high-power applications, a coaxial configuration is used. A simple waveguide cutoff attenuator is shown in Fig. 21. A metal tube, acting as a waveguide, has loops arranged at each end to couple from the coaxial lines into and out of the waveguide. One of the loops is mounted on a movable plunger or hollow piston so that the distance between the loops is variable. The input coupling loop converts the incoming TEM wave into the TE<sub>11</sub> mode in the circular guide, while the output loop converts the attenuated TE<sub>11</sub> mode back to TEM. The attenuator can be matched by adding $Z_0$ resistors. The attenuation is given as

$$A \text{(dB)} = 54.6 \frac{l}{\lambda_c} \sqrt{1 - \left(\frac{f}{f_c}\right)^2} \quad (36)$$

By choosing the diameter such that $\lambda_c \ll \lambda_o$, and hence $f/f_c \ll 1$, this equation reduces to

$$A \text{(dB)} = 54.6 \frac{1}{\lambda_c} \quad (37)$$

This was obtained from

$$x = \frac{2\pi}{\lambda_o} \text{ Np/m or } x = \frac{54.6}{\lambda_o} \text{ dB/m where 1 Np = 8.686 dB} \quad (38)$$

[If $\lambda_o = 10 \text{ cm}$, and $\lambda_o$ is much greater (10 times or more—in this case, 1 m or more), the attenuation]
increases 5.45 dB per cm of outward movement of the plunger.

The sliding cylindrical conductors allow length $l$ to be varied, which varies the attenuation, since attenuation $A = 2a l$, where $a$ is the attenuation constant due to the cutoff effect, and $l$ is the length of the circular guide. The cutoff wavelength is $\lambda_c = 1.706D$, where $D$ is the diameter of the waveguide. Thus the attenuation is

$$A(\text{dB}) = 54.6 \frac{l}{\lambda_c} = 32 \frac{l}{D} \quad (39)$$

or

$$\Delta A(\text{dB}) = 32 \frac{\Delta l}{D} \quad (40)$$

The attenuation is independent of frequency; it depends only on the physical dimensions and hence can be accurately controlled by maintaining tight tolerances on the length and diameter of the circular guide. With $\Delta A$ linearly proportional to $\Delta l$, the cutoff attenuator is easily calibrated and hence particularly useful as a precision variable attenuator.

The cutoff attenuator is one of the most widely used precision variable attenuators in coaxial measurement equipment. This is a reflective-type attenuator, since the waveguide is essentially dissipationless. The signal is reflected rather than absorbed. For higher attenuation (>10 dB), the SWR at both ports is very high (>30). This can cause problems in certain applications.

This type of attenuator is very useful, but has the disadvantage of high insertion loss. Due to the nature of cutoff, the insertion loss is high, up to 15–20 dB. If this loss is overcome, piston attenuators are one of the most accurate attenuators available. Values of 0.001 dB/10 dB of attenuation over a 60 dB range are common. A good input/output match is obtained using inductive loops within the waveguides. Excellent matching is obtained over the entire range of attenuation due to inductive loop coupling. Figure 22 shows a commercially available standard variable piston attenuator and the various calibration curves. It contains an accurately dimensioned tube acting as a circular waveguide, below cutoff TE_{11} mode. Typical specifications are (courtesy of Weinschel Associates)

- Operating frequency: Dual frequency 1.25 MHz + 0.05 MHz and 30.0 MHz + 0.1 MHz
- Waveguide mode: TE_{11}, below cutoff
- Incremental attenuation range: 100 dB
- Minimum insertion loss: 10 dB nominal
- Resolution: 0.0001 dB for $\Delta$ dB, 0.002 dB for total loss
- Attenuation readout: Front panel 7-digit LED or remotely via IEEE bus
- Connectors: Type N jacks
- VSWR (input and output): 1.2 max at 1.25 and 30 MHz in 50-Ω system
- Accuracy: 0.001 dB/10 dB + 0.0005 dB between 15 and 115 dB total loss
- Weight: Net: 77 kg (170 lb); shipping: 145 kg (320 lb)
- Accessories: Power supply, controller, calibration tape, two power cables, one 22-wire power cable, instruction/maintenance manual
8. ACTIVE ATTENUATORS

8.1. pin Diode Attenuators

The normal diode junction consists of a p-type material brought together with an n-type material to form the familiar pn junction. The pin diode is distinguished from the normal pn junction type by an area called an intrinsic region sandwiched between the p⁺-doped and n⁺-doped silicon layers. This intrinsic layer has almost no doping and thus has a very large resistance. When a variable DC control voltage forward-biases the pin diode, the DC bias or control current causes it to behave as almost a pure resistance at RF frequencies, with a resistance value that can be varied over a range from 1Ω to 10 kΩ. As the bias current is increased, the diode resistance decreases. This relation makes the pin diode ideally suited as a variable

---

**Figure 22.** (a) Standard variable piston attenuator and (b–d) calibration curves. (b) Typical VSWR versus frequency of SPA-2 attenuator with frequency. (c) Typical variation of insertion loss of SPA-2 attenuator with frequency in a 50-Ω system. (d) Deviation versus indicated incremental insertion. Typical deviation from linearity for the model SPA-2 operating frequency is 30.0 MHz.
attenuator for leveling and amplitude modulating a RF signal.

These attenuators provide local oscillator, IF, and RF signal level control throughout communications, measurement, and control circuits. One example is the reduction in the output of a receive mixer in a code-division multiple-access (CDMA) base station prior to the IF amplifier. Also, to provide one step of transmit level control with little degradation of the noise figure (NF), it could be used in a CDMA handset transmit chain between the mixer (upconverter) and the bandpass filter (Fig. 24). Since the attenuator is purely passive, it produces no additive noise and the NF is essentially its insertion loss. Even in the attenuator mode, the effect on the noise figure would be minimal.

In personal communication service (PCS) systems, the base stations may be fed from multiple picocells that are physically separated from it by up to 100 ft or more of coaxial cable. The signal levels coming into the base station will vary depending on the cable length and individual transponder power. It is desirable to keep the signals at uniform levels coming into the base station; to do so, it may be necessary to attenuate the stronger signals. An attenuator can be easily inserted for this purpose.

The upper end of a receiver's linear dynamic range is determined by the largest signal it can handle without being overdriven and producing unacceptable levels of distortion caused by device nonlinearities. Inserting an attenuator before a low-noise amplifier (LNA) in the presence of strong, in-band signals produces better reception by preventing them from overdriving the receiver's front end. This effectively shifts the dynamic range upward by the amount of attenuation. It must be remembered that when inserted into the system, the attenuator will also present a load and a source impedance to the previous and succeeding stages, respectively, hence the importance of the attenuator impedance match.

RF variable attenuators are used to control the transmitting and receiving signal power levels to prevent strong–weak adjacent signals from seriously degrading the bit error rate (BER) of digital mobile communication systems, such as TDMA or CDMA. Figure 25 shows the basic RF functional block diagram of a typical digital cellular phone system, where variable attenuators are required.

8.2. Characteristics of the pin Diode

The approximate high frequency equivalent circuit of a pin diode is shown in Fig. 26. Here, $R_I$ is the effective resistance of the intrinsic ($I$) layer, given by

$$R_I = \frac{k}{I_{DC}}$$

where $I_{DC}$ is the DC bias current in mA, and $k$ and $x$ are device-dependent empirical constants. Although shown as a variable, this resistance is constant with respect to the RF signal. The high-frequency resistance function is plotted in Fig. 27 for the Hewlett-Packard HPND-4165 diode. For a specific diode design, the exponent $X$ is usually a constant. For the HPND-4165, $X$ is typically 0.92. The constant $k$ and therefore $R_I$, however, are highly dependent on the fabrication and process control and its value can vary by as much as 3:1 from diode to diode. For analog applications, such as a variable attenuator, where repeatable attenuation with bias current is desired, the variation of $R_I$ must be controlled. The HPND-4165 is...
precisely controlled in manufacturing, and resistance values at specific bias points are specified and the slope of resistance versus bias matched with narrow limits. The specification limits of these parameters are shown in Table 4.

8.3. Applications

The pin diode is ideally suited to switch and attenuate RF signals. Since the pin diode is a RF variable resistor, the logical application is that of a variable attenuator. This attenuator may be either a step or a continuously variable type. Two of the simplest circuits are the series and shunt attenuators shown in Figs. 28 and 29.

Attenuation in the series pin circuit is decreased (more power appears at the output) as the RF resistance of the diode is reduced. This resistance is reduced by increasing the forward bias control current on the diode. The opposite occurs for the shunt configuration. The attenuation in the shunt circuit is decreased when the RF resistance of the diode increases because less power is absorbed in the diode and more appears at the output. If the control bias is switched rapidly between high and low (zero) values, then the circuit acts simply as a switch. When used as a switch, the attenuation that exists when the switch is on is called

*insertion loss.* The attenuation provided when the switch is off is called *isolation.* If the diode is a pure resistance, the attenuation for the series and shunt circuit can be calculated as

\[
A(\text{series}) = 20 \log \left( 1 + \frac{R_i}{Z_0} \right) \quad (42)
\]

\[
A(\text{shunt}) = 20 \log \left( 1 + \frac{Z_0}{2R_i} \right) \quad (43)
\]

where \(Z_0 = R_G = R_L = \text{circuit, generator, and load resistance, respectively. In reviewing these equations, it is seen that the attenuation is not a function of frequency but only a ratio of circuit and diode resistances, which is a great advantage. As the bias on the diode is varied, the load resistance experienced by the source also varies. These circuits are generally referred to as reflective attenuators because they operate on the principle of reflection. Many RF systems require that the impedance at both RF ports remain essentially constant at the design value \(Z_0\). Four such circuits and their pin diode counterparts are shown in Fig. 30. All four circuits operate on the principle of absorbing the undesired RF signal power in the pin diodes. In circuits (a), (b), and (c), the control current variation through each diode is arranged in such a way that the impedance at both RF ports remain essentially constant at the characteristic impedance \(Z_0\) of the system while the attenuation can be varied over a range of less than 1 dB to greater than 20 dB. In circuit (d), the input impedance is kept constant by using a distributed structure with a large number of diodes. The impedance variation of each diode is also shaped so that the diodes in the center of the structure vary more than those near the ports. The resulting tapered impedance
structure results in an essentially constant impedance at the ports, while the overall attenuation can be varied up to a range of 40–80 dB, depending on the length of the structure.

A pin diode pi attenuator such as that in Fig. 30a is often selected when designing a variable attenuator. The basic pi fixed attenuator is shown, along with its design equations, in Fig. 31. Shunt resistors $R_1$ and the series resistor $R_3$ are set to achieve a desired value of attenuation, while simultaneously providing an input and output impedance that matches the characteristic impedance $Z_0$ of the system.

Three pin diodes can be used as shown in Fig. 32 to replace the fixed resistors of the pi circuit to create a variable attenuator. The attenuator provides good performance over the frequency range of 10 MHz to over

---

**Figure 29.** Shunt pin RF attenuator or switch: (a) complete circuit; (b) idealized RF equivalent circuit.

**Figure 30.** Constant impedance pin diode attenuators: (a) pi attenuator; (b) bridged T attenuator; (c) T attenuator; (d) resistive line attenuator.
500 MHz. However, the use of three diodes as the three variable resistors in a pi attenuator results in a complex unsymmetric bias network. If resistor $R_3$ is replaced by two diodes, as shown in Fig. 33, the resulting attenuator is symmetric and the bias network is significantly simplified. $V_+$ is a fixed voltage, and $V_c$ is the variable control voltage, which controls the attenuation of the network. The only drawback to using two series diodes in place of one is the slight increase in insertion loss. Resistors $R_1$ and $R_3$ serve as bias returns for series diodes $D_2$ and $D_3$. Resistors $R_3$ and $R_4$ are chosen to match the specific characteristics of the pin diodes used. Properly selected, they will provide the correct split of bias current between series and shunt diodes required to maintain a good impedance match over the entire dynamic range of attenuation.

The pin diode variable attenuator is an excellent circuit used to set the power level of an RF signal from a voltage control; it is used widely in commercial applications, such as cellular phones, PCN (personal communication networks), wireless LANs (local-area networks), and portable radios.

8.4. GaAs NMESFET Attenuator

The GaAs N-channel metal semiconductor field effect transistor (NMESFET) is used in microwave attenuator designs. The metal–semiconductor FET (MESFET) is a field-effect transistor that operates on the principle that the gate-to-source voltage controls the drain current. The MESFET is a device extension of a JFET, where the gate structure is a Schottky MN (metal-N semiconductor) junction.

$$R_{ds} = R_{ds0} \left( \frac{1}{1 - (V_g/V_p)} \right)$$ (44)

In GaAs NMESFET attenuator designs, the devices are operated either in the linear region where the device is modeled as a voltage variable resistor or as an ON/OFF switch in conjunction with thin-film nichrome resistors to provide appropriate levels of attenuation. The channel resistance of the GaAs NMESFET is known to follow the classical theory for a FET in the linear region of operation. With the FET biased in the linear region, the resistance varies inversely to the gate voltage as shown below:
where \( V_g \) = gate bias voltage (V), \( V_p \) = pinchoff voltage (V), and \( R_{ds0} \) = channel resistance (\( \Omega \)) with \( V_g = 0 \) V.

As the gate voltage approaches the pinchoff voltage, the resistance becomes very high (relative to 50 \( \Omega \)). Conversely, as the gate voltage approaches zero, so does the channel resistance. For each attenuator configuration, two independent gate bias voltages are used; one to control the series MESFETs and one to control the shunt MESFETs. The T attenuator configuration is shown in Fig. 34, with one voltage controlling the series resistance arms, and another the shunt resistance arm. Table 5 gives the resistor values of the series and shunt resistances in a \( Z_0 = 50 \Omega \) system. The channel resistances of the MESFETs are matched as closely as possible for these resistances. A matched condition at the input and output port to \( Z_0 \) occurs when

\[
Z_0^2 = R_1^2 + 2R_1R_2
\]  

The resulting matched attenuation is

\[
A = 20 \log \left( \frac{R_1 + R_2 + Z_0}{R_2} \right)
\]

The pi attenuator configuration is shown in Fig. 35, with one voltage controlling the shunt resistance arms, and another the series resistance arm. Table 6 gives the values of the series and shunt resistances for different levels of attenuation in a \( Z_0 = 50 \Omega \) system. Shunt resistor \( R_1 \) and series resistor \( R_2 \) provide an input and output impedance that matches the characteristic impedance \( Z_0 = 50 \Omega \) of the system, while setting the desired level of attenuation. The design equations are

\[
R_1 = \frac{Z_0}{1} \left[ \frac{K + 1}{K - 1} \right]
\]

\[
R_2 = \frac{Z_0}{2} \left[ \frac{K - 1}{K} \right]
\]

\[
A(\text{dB}) = 20 \log K
\]

where \( K \) is the input to output voltage ratio.

GaAs NMESFET digital attenuators allow a specific value of attenuation to be selected via a digital \( n \) bit programming word. In these designs, the NMESFET operates as an \textit{on}/\textit{off} switch and is used in conjunction with nichrome thin-film resistors to provide the desired level of attenuation. Figure 36 shows the circuit configurations used for individual attenuator bits. The switched bridged T attenuator consists of the classical bridged T attenuator with a shunt and series FET. These two FETs are switched on or off to switch between the two states. The attenuation in dB is given by

\[
A(\text{dB}) = 20 \log \left( \frac{Z_0 + R_2}{R_2} \right)
\]

where \( Z_0 = R_1R_2 \).

The performance is determined by the FET characteristics in the \textit{on} and \textit{off} states and the realizability limit on required resistance values and their associated parasitics. The switched T or pi attenuators are similar in principle to the switched bridged T attenuator except for the circuit topology. These attenuators are normally used for high attenuation values. To obtain smaller values of attenuation, the thin-film resistors are replaced with appropriate channel resistances.

There are GaAs NMESFET digital RF attenuators on the market with excellent performance, in both step and continuously variable types. The variable or programmable class allows a specific value of attenuation to be selected from an overall range via an \( N \)-bit programming word. They are more flexible than step attenuators, as they allow any amount of attenuation to be set, but the cost is greater circuit complexity. Both types have a bypass state when no attenuation is selected, and the attenuation is just the insertion loss of the device. An example of each type is presented.
The RF Microdevices RF 2420 is a multistage monolithic variable or programmable attenuator that has attenuation programmability over a 44 dB range in 2-dB steps. The attenuation is set by 5 bits of digital data. A functional block diagram of the RF 2420 is shown in Fig. 37. It consists of five cascaded, DC-coupled attenuator sections, each with its own logic translator. The logic translator converts the one-bit control signal, which uses logic levels approximating standard TTL logic, to the voltage levels required to switch the attenuator stage FETS. The RF input and output signal lines are biased at approximately $V_{DD}$, and therefore external DC blocking capacitors are required. An external $V_{DD}$ bypass capacitor is also required.

A functional schematic of the RF portion of one attenuator section is shown in Fig. 38. A MESFET bridges the series resistor in a resistive pi attenuator, and two more MESFETs are connected as a double-pole single-throw (DPST) RF switch connecting the shunt branches of the pi attenuator to RF ground. In the bypass state, the bridge MESFET is in its high conductance state, and the DPST switch is open, so that the pi-attenuator is effectively removed from the circuit. When the attenuator bit is selected, the bridge MESFET is put into its low conductance state or cutoff state and the shunt FETs are put into their on state, so that the pi attenuator is connected into the RF series path. This attenuator has only moderate variation across a broad band of operation from 100 to 950 MHz, as illustrated in Fig. 39.

Furthermore, the attenuation varies smoothly and consistently with attenuator switch settings. Other features of the device are single 3–6-V supply operation, and 4 dB insertion loss, and the input and output have a low-VSWR 50-Ω match. All these features make the RF 2420 an excellent component for communications systems that require RF transmit power control by digital means. Typical applications are in dual mode IS-54/55 compatible cellular transceivers and TETRA systems. Figure 40 shows the complete schematic details of the RF 2420 being employed in a typical RF/IF switching attenuator application.

The RF Microdevice RF 2421 is a GaAs MESFET-switched step attenuator. It has a single-step digitally controlled attenuation of 10 dB. A functional block diagram of the device is shown in Fig. 41. The supply voltage range required is 2.7 V to 6 V DC. The input and output of the device have a low-voltage standing-wave ratio (VSWR) 50-Ω match, and the RF output can drive up to $+16$ dBm. It has 1.0 dB of insertion loss over the specified 500 MHz–3 GHz operating frequency range. The resistors are nickel chromium (nichrome) and provide excellent temperature stability. The RF ports are reversible, which means that the input signal can be applied to either port. The attenuation control pin has an internal pulldown resistor that causes the attenuator to be turned off when it is not connected. Figure 42 illustrates the RF 2421 being used to set the RF signal level in a communications system.

### 8.5. MOSFET Attenuators

Active voltage attenuators have many useful applications in analog integrated circuit design. Some of the applications are in the feedback loops of finite gain amplifiers and in the input stages of transconductance amplifiers. In discrete circuit design, the most popular way to design a finite-gain amplifier with precisely controlled gain, high linearity, and low output noise is to use operational amplifier and a voltage attenuator in the feedback loop. Here the voltage attenuator consists of two resistors connected in series as shown in the classical noninverting and inverting op amp gain configurations of Fig. 43. Resistor attenuators are not useful in integrated circuit design because of their large areas, low input impedance, large power dissipation, and parasitic capacitances, and precise resistance values cannot be realized.

Three MOS active voltage attenuator configurations useful for the realization of finite-gain amplifiers in mono-
Lithic circuits are presented. The attenuators are two single-input attenuators and a summing attenuator that has two inputs. These attenuators are simple in structure, consisting only of MOSFETs. Therefore, they are easy to fabricate in standard CMOS semiconductor processes. The attenuation factor is precisely controlled over a wide range of gains because it ideally depends only on the ratios of the dimensions of the MOSFETs.

Attenuator I, shown in Fig. 44, is an active linear voltage attenuator consisting of two n-channel MOSFETs fabricated in a common substrate. The capability to fabricate the MOSFETs in a common substrate has several advantages. First, both n-channel and p-channel attenuators can be monolithically fabricated in a standard CMOS process. Second, the required area of the attenuator is much smaller. As seen in Fig. 44, the substrate is common for both MOSFETs and is connected to the source of the bottom transistor M1. The circuit operates as a linear voltage attenuator when M1 is in the ohmic region and M2 is in the saturation region.

The operating conditions of the MOS attenuators in this section are derived as in the following equations, where

\[
\begin{align*}
V_{in} & \quad \text{Input voltage} \\
V_{out} & \quad \text{Output voltage} \\
V_{DD} & \quad \text{Drain supply voltage} \\
V_B & \quad \text{Bias supply voltage 1} \\
V_{BB} & \quad \text{Bias supply voltage 2} \\
V_{TON} = V_{TON1} = V_{TON2} & \quad \text{Zero bias threshold voltage of M1 and M2} \\
V_I & \quad \text{Threshold voltage of M2 due to body bias effect} \\
V_1 & \quad \text{Input voltage 1} \\
V_2 & \quad \text{Input voltage 2} \\
g & \quad \text{Body effect parameter} \\
\phi & \quad \text{Barrier potential} \\
I_D & \quad \text{Drain current} \\
W & \quad \text{Width of channel} \\
L & \quad \text{Length of channel} \\
W_1, W_2 & \quad \text{Width of channels 1,2} \\
L_1, L_2 & \quad \text{Length of channels 1,2} \\
K & \quad \text{Device constant, } \mu_n \text{ CoX} \\
\mu_n & \quad \text{Mobility of electron} \\
\text{CoX} & \quad \text{Gate oxide capacitance per unit area}
\end{align*}
\]

The zero-bias threshold voltage of both MOSFETs is \(V_{TON1} = V_{TON2} = V_{TON}\). The proper operating conditions will be met, provided

\[
V_{TON} < V_{in} < V_{DD} + V_T (51)
\]
where

\[ V_{T2} = V_{TON} + \gamma \left( \sqrt{\phi + V_{out}} - \sqrt{\phi} \right) \]  (52)

Since M1 is operating in the ohmic region and M2 is in the saturation region, the drain current of each MOSFET is given by

\[ I_{D1} = K \frac{W_1}{L_1} \left( V_1 - V_{TON} - \frac{V_{out}}{2} \right) V_{out} \]  (53)

and

\[ I_{D2} = K \frac{W_2}{2L_2} (V_1 - V_{T2} - V_{out})^2 \]  (54)

Equating the two drain currents, the relationship between \( V_{in} \) and \( V_{out} \) is obtained as

\[ 2R \left( \frac{V_{in} - V_{TON} - \frac{V_{out}}{2}}{2} \right) V_{out} \]

\[ = \left\{ V_{in} - V_{TON} - \gamma \left( \sqrt{\phi + V_{out}} - \sqrt{\phi} \right) \right\}^2 \]  (55)

where

\[ R = \frac{W_1}{L_1} \frac{W_2}{L_2} \]  (56)

If each MOSFET in the attenuator is fabricated in a separate substrate and the substrate of each

---

**Figure 39.** Attenuation and frequency response characteristics of RF 2420 5-bit digital RF attenuator.

**Figure 40.** RF 2420 RF/IF switching attenuator schematic.
MOSFET is connected to its source ($\gamma = 0$), the DC transfer characteristic relating $V_{in}$ and $V_{out}$ becomes a linear equation:

$$V_{out} = z(V_{in} - V_{TON}) \quad (57)$$

where $z$ is the small-signal attenuation factor.

In this case, $z$ is

$$z = 1 - \sqrt{\frac{R}{R+1}} = 1 - \sqrt{\frac{W_1/L_1}{W_1/L_1 + W_2/L_2}} \quad (58)$$

Equation (57) is a special case of Eq. (55), when the bulk effect term due to $\gamma$ is ignored. When the substrate is separate, the small-signal attenuation factor from Eq. (58) is precisely determined by width/length ratios. If the substrate is common, the relationship between the input and output is still very linear as given by Eq. (55) even though the equation appears to be a nonlinear quadratic.

Figure 45 shows the typical DC transfer characteristic of the attenuator consisting of M1 ($12 \times 10 \mu m^2$) and M2 ($3 \times 10 \mu m^2$) when the substrate is common ($\gamma \neq 0$) and $V_{DD} = 5$ V. The DC transfer characteristic exhibits a high degree of linearity for the input range 2–5 V. The small-signal attenuation factor ($z$), which is the slope of the DC transfer characteristic is 0.07824 at an input quiescent voltage of 3.5 V.

A finite-gain amplifier consisting of an ideal op amp and attenuator I in the feedback loop is shown in Fig. 44. Since the op amp is assumed ideal, we obtain

$$V_{in} = V_{out} = zV_{in} = zV'_{in} \quad (59)$$

or

$$V'_{out} = \frac{1}{z} V'_{in} \quad (60)$$

Thus, the DC transfer function of the amplifier is the inverse function of the DC transfer function of the attenuator in the feedback loop. Thus, the transfer function between the input $V'_{in}$ and the $V'_{out}$ of the amplifier is given by Eq. (55) when $V_{out}$ is replaced by $V'_{in}$ and $V_{in}$ by $V'_{in}$.

The small-signal voltage gain

$$A_V = \frac{V'_{out}}{V'_{in}} = \frac{1}{z}$$

is the reciprocal of the attenuator’s attenuation factor in the feedback loop. Figure 46 illustrates the DC transfer characteristic of the finite-gain amplifier.

Two slightly different linear inverting voltage attenuator configurations consisting of two n-channel MOSFETs are shown in Fig. 47. These circuits operate as a linear inverting voltage attenuator when both transistors are in the saturation region. Assuming the zero-bias threshold of both of the MOSFETs is $V_{TON}$, the condition will be met, provided

$$V_{out} + V_{T2} < V_B < V_{DD} + V_{T2} \quad (61)$$

and

$$V_{TON} < V_{in} < V_{out} + V_{TON} \quad (62)$$
Under this condition, the drain currents of the transistors are given by

\[ I_{D1} = K \frac{W_1}{2L_1} (V_{in} - V_{TON})^2 \]  
\[ I_{D2} = K \frac{W_2}{2L_2} (V_B - V_{out} - V_{T2})^2 \]

where

\[ V_{T2} = V_{TON} + \gamma (\sqrt{\phi + V_0} - \sqrt{\phi}) \]  
\[ \gamma = 0 \]  
\[ V_{in} \]  
\[ V_{out} \]  
\[ R_1 = \sqrt{\frac{W_1/L_1}{W_2/L_2}} \]

If \( \gamma = 0 \) in Eq. (66), which corresponds to the case of circuit (b) in Fig. 47, where the substrate is separate, the DC transfer characteristic reduces to a linear equation:

\[ V_{out} = z V_1 + [V_B - (z + 1)V_{TON}] \]  
\[ z = -R_1 \]

which is precisely determined by the width/length ratios of the MOSFETs. From Eqs. (66) and (68), it is noted that the output DC operating voltage is controlled by \( V_{in} \), independent of the attenuation factor.

The DC transfer characteristic between \( V_{in} \) and \( V_{out} \) calculated from Eq. (66) for the common substrate case, \( R_1 = 0.1149 \) and \( V_B = 3.993 \), and the DC transfer characteristics calculated from Eq. (68) for the separate substrate case, \( R_1 = 0.1 \) and \( V_B = 3.449 \) are shown in Fig. 50 for the range restricted by Eq. (62). The parameter values (\( \gamma = 0.525 \), \( \phi = 0.6 \), and \( V_{TON} = V_{TON2} = V_{TON} = 0.777 \)) were used in the calculation. The DC transfer function given by Eq. (66) for the common substrate case appears nonlinear, but the degradation from linearity due to practical values of \( \gamma \) is not significant. The small-signal attenuation factor \( z \), the slope of transfer characteristic in Fig. 48, is \(-0.1\). The high degree of linearity supports the usefulness of both configurations in precision attenuator or finite-gain amplifier applications.

Figure 49 shows a finite-gain amplifier with attenuator II in the feedback loop of an op amp. Assuming that the op amp is ideal, we obtain

\[ V_{out}' = \frac{1}{z} V_{in}' \]  
\[ z = -R_1 \]

The transfer function of the amplifier is the inverse function of the transfer function of the attenuator in the feedback loop. The DC transfer function of the amplifier is given by Eq. (66) when \( V_{in} \) is replaced by \( V_{in}' \) and \( V_{out} \) is replaced by \( V_{out}' \). If the substrate is separate, \( V_{in} \) replaces \( V_{out} \) and \( V_{out} \) replaces \( V_{in} \) in Eq. (68); then

\[ V_{out}' = \frac{1}{z} V_{in}' - \frac{1}{z} [V_B - (z + 1)V_{TON}] \]

where the small-signal attenuation factor \( z = -R_1 \).

A summing attenuator is necessary to realize versatile multiple-input finite-gain amplifiers in integrated circuits. Figure 50 shows a two-input active linear inverting voltage summing attenuator that consists of two attenuators cascaded. For the summing attenuator, \( V_{HB} \) is used to control the output DC operating voltage, and input signals are designated as \( V_1 \) and \( V_2 \).

**Figure 43.** Op amp noninverting (a) and inverting (b) gain configuration.

**Figure 44.** (a) Circuit and block diagram of attenuator I consisting of two n-channel MOSFETs, and (b) block diagram of amplifier consisting of an op amp and attenuator.
As for the inverting attenuator, the summing attenuator works when all the MOSFETs M1–M4 are operating in the saturation region. The DC transfer characteristics are found by equating the drain currents in the saturation region for each transistor. Assuming that the zero-bias threshold voltages for the four MOSFETs are matched at $V_{TON}$, the four transistors are in the saturation region, provided

$$2(V_{TON} + \gamma(\sqrt{\phi} + V_{TON} - \sqrt{\phi})) + V_{out} < V_{BB} < V_{DD} + V_{T4}$$

(72)

$$V_{TON} < V_1 < V_0 + V_{TON}$$

(73)

$$V_{TON} < V_2 < V_B + V_{TON}$$

(74)

By equating the drain currents of M3 and M4 given by

$$I_{D3} = K \frac{W_4}{2L_3} (V_2 - V_{TON})^2$$

(75)

and

$$I_{D4} = K \frac{W_4}{2L_4} (V_{BB} - V_B - V_{T4})^2$$

(76)

where

$$V_{T4} = V_{TON} + \gamma(\sqrt{\phi} + V_0 - \sqrt{\phi})$$

(77)

The DC transfer function between $V_2$ and $V_B$ is obtained as

$$V_B + \gamma(\sqrt{\phi} + V_B - \sqrt{\phi}) = -R_2V_2 + [V_{BB} + (R_2 - 1)V_{TON}]$$

(78)

where

$$R_2 = \sqrt{\frac{W_3L_4}{L_3W_4}}$$

(79)

Similarly, it can be shown that the DC transfer function between $V_1$ and $V_0$ is obtained as

$$V_0 + \gamma(\sqrt{\phi} + V_0 - \sqrt{\phi}) = -R_1V_1 + [V_B + (R_1 - 1)V_{TON}]$$

(80)

where

$$R_1 = \sqrt{\frac{W_1L_2}{L_1W_2}}$$

(81)

If $\gamma = 0$ in Eqs. (78) and (80), the equations become linear. This is realized if each transistor is fabricated in a separate substrate and the substrate of each transistor is connected to its source. In this case, the attenuation
factors are given by \( a_1 = \frac{1}{C_0 R_1} \) and \( a_2 = \frac{1}{C_0 R_2} \). Even when \( g_a = 0 \), which is the case when the substrates are common, the transfer characteristics between \( V_1 \) and \( V_0 \) and between \( V_2 \) and \( V_0 \) are nearly linear as shown in Fig. 53 for practical values of \( g \). In the calculation of Fig. 51, \( g = 0.5255 V^{1/2} \), \( f = 0.6 V \), and \( V_{TON} = 0.777 V \) were used, which are standard for a 2\( \mu \) CMOS process and \( R_1 = 0.1149 \) and \( R_2 = 0.1290 \) were set such that the small-signal attenuation factors for \( V_1 \) and \( V_2 \) are both \( 1/C_0 \).

The operating points were set by \( V_{BB} = 5.712 V \) such that \( V_{0Q} = 2.5 V (V_{BQ} = 3.993 V) \) when \( V_{1Q} = V_{2Q} = 2.5 V \).

Summing and subtracting amplifier configurations using the inverting attenuator and the inverting summing attenuator are shown in Fig. 52.

Circuit (a) in Fig. 52 functions as a summing amplifier and the circuit (b) functions as a subtracting amplifier, with controllable weights. Assuming ideal op amps and attenuators, we obtain

\[
V_+ = x_1 V_1 + x_2 V_2 + (V_{BB} - (x_1 + x_2) + 2) V_{TON} \tag{82}
\]

\[
V_+ = x V_0 + (V_B - (x + 1)V_{TON}) \tag{83}
\]

Equating \( V_+ \) and \( V_+ \), the output is given by

\[
V_0 = \frac{x}{x_1} V_1 + \frac{x_2}{x_1} V_2 + \frac{1}{x} \frac{V_{BB} - V_B - (x_1 + x_2 - x + 1)V_{TON}}{x_1} \tag{84}
\]

From Eq. (84), the circuit in Fig. 52a is a summing amplifier with a wide range of available gain for each input. Similarly, for the circuit in Fig. 52b, we obtain

\[
V_+ = x_1 V_0 + x_2 V_2 + (V_{BB} - (x_1 + x_2) + 2) V_{TON} \tag{85}
\]

\[
V_+ = x V_0 + (V_B - (x + 1)V_{TON}) \tag{86}
\]

Equating \( V_+ \) and \( V_+ \), the output is given by

\[
V_0 = \frac{x}{x_1} V_1 + \frac{x_2}{x_1} V_2 - \frac{1}{x} \frac{V_{BB} - V_B - (x_1 + x_2 - x + 1)V_{TON}}{x_1} \tag{87}
\]

From Eq. (87), the circuit in Fig. 52b is a subtracting amplifier with a wide range of available gain for each input.

The active attenuator and the active summing attenuator have many desirable characteristics such as small size, nearly infinite impedance, low power dissipation, and precisely controllable attenuation ratio with excellent linearity. These attenuators and the finite-gain amplifiers obtained from these attenuators and op amps will find increased applications in analog integrated circuits.

### 8.6. Noise

Noise in a communication system can be classified in two broad categories, depending on its source. Noise generated by components within a communication system, such as resistive, extender, and solid-state active devices, comprise internal noise. The second category, external noise, results from sources outside a communication system, including atmospheric, manmade, and extraterrestrial sources.
External noise results from the random motion of a charge carrier in electronic components. The three types are

1. **Thermal noise**: caused by random motion of free electrons in a conductor or semiconductor excited by thermal agitation.
2. **Shot noise**: caused by random amount of discrete charge carriers in such devices as thermionic tubes or semiconductors in devices.
3. **Flicker noise**: produced by semiconductors by a mechanism not well understood and is more severe the lower the frequency.

Atmospheric noise results primarily from spurious radiowaves generated by the natural discharges within the atmosphere associated with thunderstorms. Manmade noise sources include high-voltage power-line discharge and computer-generated noise in electric motors.

Other noises include

- *Generation–recombination noise*: due to free carriers being generated and recombining in semiconductor material. They are random and can be treated as a shot noise process.
- *Temperature fluctuation noise*: the result of the fluctuating heat exchange between a small body, such as a transistor, and its environment due to the fluctuations in the radiation and heat conduction processes.

9. **RECENT TRENDS**

Figure 53 shows a 4-dB step, 28-dB variable attenuator for a 1.9-GHz personal handy phone system transmitter fabricated using silicon bipolar technology with $f_T$ of 15 GHz. The GaAs MESFET variable attenuator is configured with resistive pi attenuators and GaAs switches as shown. Step accuracy within 1.2 dB and total vector modulation error of less than 4% were realized for $-15$ dBm output. The attenuator consumes 21 mA with 2.7 V power supply and occupies $1.1 \times 0.5$ mm. This unit is being developed. This shows the technology trend.

Figure 54 shows the top view and cross section of a prototype optical microwave attenuator that can be...
controlled by illuminating the silicon substrate. The maximum attenuation is 30 dB using laser diode illumination. It is a microstrip line whose substrate consists of silicon and ferrite slabs. The ferrite slab is overlaid on the microstrip. There is a slot on the ground plane under the strip. A white light from a xenon arc lamp with a parabolic mirror is focused by a lens to the silicon surface through the slot. The intensity of the light is not uniform along the slot direction. Due to the light, electron–hole pairs are induced and the permittivity and conductivity of the silicon are changed, which vary the phase and amplitude of the microwave. With 240 mW optical power illumination, an attenuation in the range of 17–26 dB was obtained in the frequency range from 8 to 12 GHz.

**FURTHER READING**


P. S. Boecht, *FET Attenuator, 0–1 GHz Applied Microwave and Wireless*, spring 1996.


R. LaVerne Libbey, *Video Attenuator Using a Multiplier and FET*, a publication of RCA, New Jersey, 1975.

*RF Microdevices*, RF 2420 Programmable Attenuator Data Sheet, 1997.

*RF & Microwave Test Accessories Catalog*, Hewlett-Packard, 1997/98.


*RF Microdevices*, RF 2421, 10 dB Switched Attenuator Data Sheet, 1997.


