มาตรฐานผลิตภัณฑ์อุตสาหกรรม

THAI INDUSTRIAL STANDARD

มอก. 1863 – 2552

IEC 60747– 4(2007)

อุปกรณ์สารกึ่งตัวนำ - อุปกรณ์ไม่รวมหน่วย

เล่ม 4 ไดโอดและทรานซิสเตอร์ไมโครเวฟ

SEMICONDUCTOR DEVICES - DISCRETE DEVICES

PART 4: MICROWAVE DIODES AND TRANSISTORS

สำนักงานมาตรฐานผลิตภัณฑ์อุตสาหกรรม

กระทรวงอุตสาหกรรม

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มาตรฐานผลิตภัณฑ์อุตสาหกรรม
อุปกรณ์สารกึ่งตัวนำ - อุปกรณ์ไม่รวมหน่วย
เล่ม 4 ไดโอดและทรานซิสเตอร์ไมโครเวฟ

มอก. 1863 – 2552

สำนักงานมาตรฐานผลิตภัณฑ์อุตสาหกรรม
กระทรวงอุตสาหกรรม ถนนพระรามที่ 6 กรุงเทพฯ 10400
โทรศัพท์ 02 202 3300

ประกาศในราชกิจจานุเบกษา ฉบับประกาศและงานทั่วไป เล่ม 127 ตอนพิเศษ 92ง
วันที่ 30 กรกฎาคม พุทธศักราช 2553

มาตรฐานผลิตภัณฑ์อุตสาหกรรมอุปกรณ์สารกึ่งตัวนำ – อุปกรณ์ไม่รวมหน่วย เล่ม 4 ไดโอดและทรานซิสเตอร์ไมโครเวฟ ได้ประกาศใช้ครั้งแรกโดยรับ IEC 747-4(1991-04) Semiconductor Devices-Discrete devises-Part4: Microwave diodes and transistors มาใช้ในระดับเหมือนกันทุกประการ (Identical) โดยใช้ IEC ฉบับภาษาอังกฤษเป็นหลัก โดยประกาศในราชกิจจานุเบกษา ฉบับประกาศทั่วไป เล่มที่ 117 ตอนพิเศษที่ 134ง วันที่ 29 ธันวาคม พุทธศักราช 2543

เนื่องจาก IEC ได้แก้ไขปรับปรุงมาตรฐาน IEC 747-4(1991-04) เป็น IEC 60747-4 (2007) จึงได้ยกเลิกมาตรฐานเดิมและกำหนดมาตรฐานใหม่โดยรับ IEC 60747-4 (2007) Semiconductor devices - Discrete devices- Part 4: Microwave diodes and transistors มาใช้ในระดับเหมือนกันทุกประการโดยใช้มาตรฐาน IEC ฉบับภาษาอังกฤษเป็นหลัก

คณะกรรมการมาตรฐานผลิตภัณฑ์อุตสาหกรรมได้พิจารณามาตรฐานนี้แล้ว เห็นสมควรเสนอรัฐมนตรีประกาศตามมาตรา 15 แห่งพระราชบัญญัติมาตรฐานผลิตภัณฑ์อุตสาหกรรม พ.ศ. 2511

ประกาศกระทรวงอุตสาหกรรม

ฉบับที่ 4206 (พ.ศ. 2553)

ออกตามความในพระราชบัญญัติมาตรฐานผลิตภัณฑ์อุตสาหกรรม

พ.ศ. 2511

เรื่อง ยกเลิกและกำหนดมาตรฐานผลิตภัณฑ์อุตสาหกรรม

อุปกรณ์สารกึ่งตัวนำ – อุปกรณ์ไม่รวมหน่วย

เล่ม 4 ไดโอดและทรานซิสเตอร์ไมโครเวฟ

โดยที่เป็นการสมควรปรับปรุงมาตรฐานผลิตภัณฑ์อุตสาหกรรม อุปกรณ์สารกึ่งตัวนำ – อุปกรณ์ไม่รวมหน่วย เล่ม 4 ไดโอดและทรานซิสเตอร์ไมโครเวฟ มาตรฐานเลขที่ มอก.1863–2542

อาศัยอำนาจตามความในมาตรา 15 แห่งพระราชบัญญัติมาตรฐานผลิตภัณฑ์อุตสาหกรรม พ.ศ. 2511 รัฐมนตรีว่าการกระทรวงอุตสาหกรรมออกประกาศยกเลิกประกาศกระทรวงอุตสาหกรรม ฉบับที่ 2744 (พ.ศ.2543) ออกตามความในพระราชบัญญัติมาตรฐานผลิตภัณฑ์อุตสาหกรรม พ.ศ.2511 เรื่อง กำหนดมาตรฐานผลิตภัณฑ์อุตสาหกรรม อุปกรณ์สารกึ่งตัวนำ – อุปกรณ์ไม่รวมหน่วย เล่ม 4 ไดโอดและทรานซิสเตอร์ไมโครเวฟ ลงวันที่ 9 ตุลาคม พ.ศ.2543 และออกประกาศกำหนดมาตรฐานผลิตภัณฑ์อุตสาหกรรม อุปกรณ์สารกึ่งตัวนำ – อุปกรณ์ไม่รวมหน่วย เล่ม 4 ไดโอดและทรานซิสเตอร์ไมโครเวฟ มาตรฐานเลขที่ มอก.1863–2552 ขึ้นใหม่ ดังมีรายละเอียดต่อท้ายประกาศนี้

ทั้งนี้ให้มีผลตั้งแต่วันถัดจากวันที่ประกาศในราชกิจจานุเบกษา เป็นต้นไป

ประกาศ ณ วันที่ 7 เมษายน พ.ศ. 2553

ชาญชัย ชัยรุ่งเรือง

รัฐมนตรีว่าการกระทรวงอุตสาหกรรม

มาตรฐานผลิตภัณฑ์อุตสาหกรรม
อุปกรณ์สารกึ่งตัวนำ - อุปกรณ์ไม่รวมหน่วย
เล่ม 4 ไดโอดและทรานซิสเตอร์ไมโครเวฟ

มาตรฐานผลิตภัณฑ์อุตสาหกรรมนี้กำหนดขึ้นโดยรับ IEC 60747-4 (1991) Amendment 1: 1993, Amendment 2:1999, Semiconductor devices - Discrete devices- Part 4: Microwave devices มาใช้ในระดับเหมือนกันทุกประการ (identical) โดยใช้ IEC ฉบับภาษาอังกฤษเป็นหลัก

มาตรฐานผลิตภัณฑ์อุตสาหกรรมนี้มีวัตถุประสงค์เพื่อกำหนดให้เกิดมาตรฐานสำหรับอุปกรณ์ไม่รวมหน่วยต่อไปนี้

- ไดโอดความจุไฟฟ้าแปรเปลี่ยนและสแนป-ออฟไดโอด (snap-off diode) (สำหรับการปรับความถี่ การแปลงผันขึ้นหรือการคูณฮาร์มอนิก การสวิตช์ การจำกัดขีดจำกัด การเลื่อนเฟส การขยายพาราเมตริก ฯลฯ)
- ไดโอดผสมคลื่นและไดโอดตรวจจับ
- ไดโอดแบบพังทลาย (สำหรับการทำให้เกิดฮาร์มอนิกโดยตรง การขยาย ฯลฯ
- กันน์ไดโอด (Gunn diode) (สำหรับการทำให้เกิดฮาร์มอนิกโดยตรง)
- ทรานซิสเตอร์ผลสนาม (สำหรับการขยาย การแกว่ง)

รายละเอียดให้เป็นไปตาม IEC 60747-4 (2007)

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CONTENTS

RECOMMENDED METHODS OF MEASUREMENT ON RECEIVERS FOR TELEVISION BROADCAST TRANSMISSIONS –

Part 3: Electrical measurements on multichannel sound television receivers using subcarrier systems

CHAPTER I: GENERAL

SECTION ONE – INTRODUCTION

1 Scope

The methods of measuring the electrical characteristics described in this standard apply particularly to broadcast television receivers designed for the reception of multichannel sound systems using subcarriers.

NOTE – Currently two systems are in operation: The FM-FM and BTSC systems, which are described in ITU-R Report BS.795-3*

2 Object

The object of this standard is to standardize the methods of measurement for the more important electrical characteristics of receivers, within the scope of this part of the standard.

SECTION TWO – GENERAL EXPLANATION OF TERMS

3 Definitions

The following definitions apply for the purpose of this standard.

3.1
main channel
the main channel is an audio channel which carries the main audio signal directly by frequency-modulating the main sound carrier of a television system

NOTE – This channel is compatible with the audio channel in the related monophonic television broadcasting system.

3.2
left (right) channel
the left (right) channel is an audio channel which carries the left (right) audio signal in the stereophonic transmission

3.3
stereo sum channel
the stereo sum channel is a channel which carries the sum signal (L + R) of the left audio signal (L) and the right audio signal (R)

The main channel is used for the stereo sum channel

* ITU-R: International Telecommunication Union Radiocommunication Sector.

3.4
stereo subchannel
the stereo subchannel is a channel which carries the difference signal (L – R) of the left audio signal and the right audio signal on a subcarrier, the frequency of which is equal to twice the line-scan frequency. Modulation of the subcarrier is FM in the FM-FM system, and AM-DSB-SC with a pilot signal at the line-frequency in the BTSC system

3.5
the second channel
the second channel is an additional audio channel which carries a second audio signal on a FM subcarrier. In the FM-FM system, the same subcarrier is used for both the stereo subchannel and the second channel

3.6
SAP (Second Audio Programme) subchannel
the SAP subchannel is the second channel in the BTSC system. The subcarrier frequency is equal to five times the line-scan frequency

3.7
dual-sound mode
the dual-sound mode is a transmission mode in which one audio signal S is carried by the main channel and another by the second (SAP) channel

3.8
stereo mode
the stereo mode is a transmission mode in which the left and right audio signals are carried by the stereo sum channel and the stereo subchannel

3.9
stereo and SAP mode
the stereo and SAP mode is a transmission mode in which the left and right audio signals and the second audio programme signal are carried simultaneously. This mode is used only in the BTSC system

3.10
mode identification
mode identification is a function which identifies the transmission modes

The mode identification is performed by using the control signal in the FM-FM system and the pilot signal and the amplitude of the second subcarrier in the BTSC system

3.11
compander
the compander is a noise reduction system in which programme signals are compressed by a compressor at the input of the sound modulator and expanded by an expander at the output of the sound demodulator

In the BTSC system, compressors (encoders) are mandatorily used in the stereo subchannel and the SAP subchannel.

SECTION THREE – GENERAL NOTES ON MEASUREMENTS

4 General conditions

Unless otherwise specified, measurements shall be carried out under the conditions described in IEC 60107-1 and IEC 60107-2.

5 Reference frequency and reference modulation factor

The reference frequency for measurements and adjustments of audio channels shall be 1 000 Hz in the FM-FM system and 300 Hz in the BTSC system.

The modulation factor is defined as the frequency deviation of the main sound carrier or a subcarrier referred to the rated maximum system deviation and expressed in percentage.

The rated maximum system deviations for both the systems are given in Appendix A.

The following modulation factors shall be used for measurements and adjustments of audio channels as the reference:

Main channel:	FM-FM system	30 %
	BTSC system	14 %
Left (right) channel:	FM-FM system	30 %
Stereo sum channel:	BTSC system	14 % by L or R signal 14 % by L and R signals
	FM-FM system	15 % by L or R signal
Stereo subchannel:	BTSC system	14 % by L or R signal at 300 Hz 14 % by L and –R signals at 300 Hz
	FM-FM system	15 %, L or R signal
The second channel:	FM-FM system	30 %
	BTSC system	14 % at 300 Hz

NOTE 1 – In the FM-FM system, the left (right) input signal determines the modulation factor for a stereo signal.

The encoder (compressor) included in the stereo subchannel of the BTSC system causes the subchannel modulation factor to be frequency dependent as well as disproportionately level dependent. As a result, the modulation factor for a stereo signal is determined by both the stereo sum-signal and the stereo subchannel signal.

NOTE 2 – In the BTSC stereo and SAP subchannels, at 300 Hz modulating frequency, the modulation factor corresponding to an unaffected encoder level is 14,1 % (–17,0 dB). Stereo modulation (L, R or L and R) at frequencies other than 300 Hz should be monitored by the modulation factor of the stereo sum-signal. The same sum-signal modulation factor should be maintained at other frequencies.

Stereo modulation (L = –R) or SAP modulation at frequencies other than 300 Hz should be maintained at an input signal to the encoder equal to that at 300 Hz.

The devices for these applications are defined as follows:

Diodes for tuning

Diodes which are used to vary the frequency of a tuned circuit.

These diodes are usually characterized a frequency of resonance much higher than the frequency of use and have a known capacitance/voltage relationship.

Diodes for harmonic multiplication

These diodes must have a non-linear capacitance/voltage relationship at the frequency of operation and a high ratio of cut-off frequency to operating frequency.

Diodes for switching (including limiting)

These diodes exhibit a fast transition from a high impedance state to a low impedance state and vice versa and can be used to modulate or control the power level in microwave systems.

Diodes for parametric amplification

These diodes are intended to handle small amplitude signals and are most often used in low-noise amplifiers.

3.1.2 Terminology and letter symbols

See 3.1.3.3.

3.1.3 Essential ratings and characteristics

3.1.3.1 General

3.1.3.1.1 Rating conditions

Variable capacitance diodes may be specified either as ambient rated or case rated devices or, where appropriate, as both.

The ratings listed in 3.1.3.2 should be stated at the following temperatures:

– *ambient-rated devices:*

at an ambient temperature of 25 °C and at one higher temperature.

– *case-rated devices:*

at a reference point temperature of 25 °C and at another reference point temperature.

3.1.3.1.2 Application categories

The essential ratings and characteristics to be stated for each category of diode are marked with a + sign in the following table:

– column 1: tuning applications;
– column 2: harmonic multiplication applications;
– column 3: switching (including limiting) applications;
– column 4: parametric amplification applications.

	Categories			
3.1.3.2 Ratings (limiting values)				
The following ratings should be stated:	1	2	3	4
3.1.3.2.1 Temperatures				
Range of operating temperatures	+	+	+	+
Range of storage temperatures	+	+	+	+
3.1.3.2.2 Voltages and currents				
Maximum peak reverse voltage	+	+	+	+
Maximum mean forward current, where appropriate		+	+	+
Maximum peak forward current, where appropriate		+	+	+
3.1.3.2.3 Power dissipation				
Maximum dissipation, under stated conditions, over the operating temperature range	+	+	+	+
3.1.3.3 Electrical characteristics				
Unless otherwise specified, the following characteristics should be given at 25 °C (see Figure 1)				
3.1.3.3.1 Stray capacitance (C_p)				
Typical value under specified conditions	+	+	+	+
3.1.3.3.2 Series inductance (L_s)				
Typical value and, where appropriate, maximum value under specified conditions	+	+	+	+
3.1.3.3.3 Terminal capacitance (C_{tot})				
a) Minimum and maximum values, at a specified bias voltage and at a specified frequency (note 2)	+	+	+	+
b) Typical curve showing the relationship between terminal capacitance and bias voltage	+	+	+	+
3.1.3.3.4 Junction capacitance (C_j)				
Minimum and maximum values at a specified bias voltage (notes 2 and 3). When the order of magnitude of C_p is the same as that of the terminal capacitance C_{tot}, a typical value should be given for C_j instead of minimum and maximum values	+	+	+	+
3.1.3.3.5 Effective quality factor (Q)				
Minimum values at two or more specified frequencies under specified bias conditions (note 4)	+			

Item	Cat. 1	Cat. 2	Cat. 3	Cat. 4
3.1.3.3.6 Cut-off frequency				
Minimum value under specified conditions (notes 4 and 5)		+	+	+
3.1.3.3.7 Series resistance (r_s)				
Maximum and/or typical values under specified conditions (note 4)	+	+	+	+
3.1.3.3.8 Reverse current				
Maximum value at a specified reverse voltage	+	+	+	+
3.1.3.3.9 Thermal resistance				
Maximum value between junction and ambient, or between the junction and a specified reference point	+	+	+	
3.1.3.3.10 Switching time				
Typical value under specified conditions			+	
3.1.3.3.11 Stored charge or minority carrier life time				
Typical value, for either stored charge under specified conditions including bias, or minority carrier life time under specified conditions		+	+	
3.1.3.3.12 Transition time				
Typical value, under specified conditions, together with a specified measurement circuit (note 1)		+		

NOTE 1 See definition in 3.2.2.

NOTE 2 For categories 1, 2 and 3, the specified bias voltage should be –6 V and for category 4, the specified bias voltage should be 0 V.

NOTE 3 The relationship between the junction capacitance and bias voltage should be represented either by a typical curve or by a mathematical form. The mathematical form should be as follows:

$$C_j = K\,(V + \phi)^{\gamma}$$

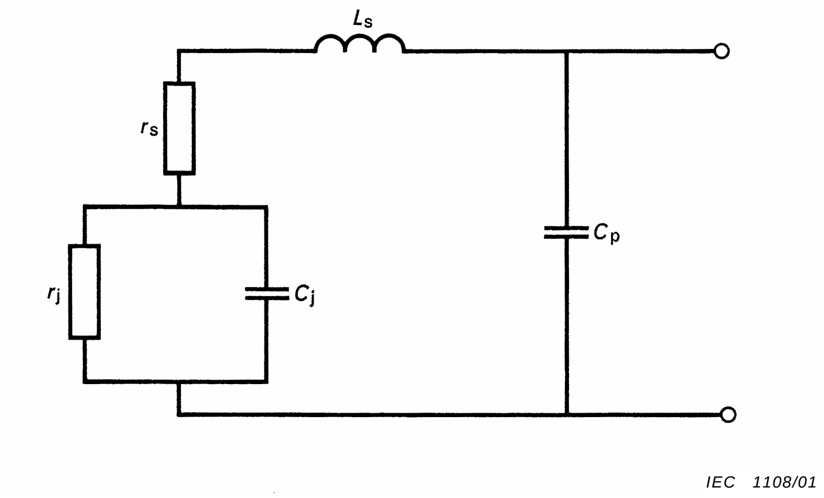
where V is the magnitude of the applied reverse voltage and K, ϕ and γ are three constants. The manufacturer should specify the typical values for K, ϕ and γ.

NOTE 4 If the Q value and the series resistance are not specified for category 1, then the cut-off frequency must be specified.

NOTE 5 The cut-off frequency f_c is defined as:

$$f_c = \frac{1}{2\pi\, r_s\, C_j}$$

where r_s is the series resistance and C_j is the capacitance of the junction measured at a specified bias point r_s is determined by the equivalent circuit shown in Figure 1 below; its value depends on the measuring method used and on the bias voltage.



Key

C_j junction capacitance

r_s series resistance

r_j low frequency resistance of the junction

C_p stray capacitance

L_s series inductance

In general, r_j is sufficiently high to be neglected.

Figure 1 – Equivalent circuit

3.1.3.4 Application data

For harmonic multiplication applications, the *efficiency* should be stated.

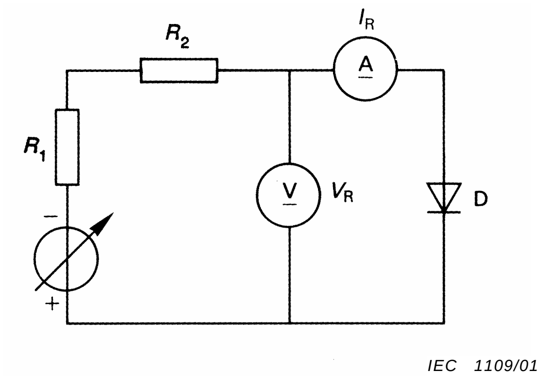
3.1.4 Measuring methods

3.1.4.1 Reverse current I_R

a) *Purpose*

To measure the reverse current of a diode under specified reverse voltage.

b) *Circuit diagram*



Key

D diode being measured

Figure 2 – Circuit for the measurement of reverse current I_R

c) *Circuit description and requirements*

R_1 is a calibrated resistor (pulse measurement only).

R_2 is a protective resistor.

If a pulse measurement is required, the variable voltage generator is replaced by a voltage pulse generator, the voltmeter is replaced by a peak-reading instrument and the ammeter is replaced by a peak-reading voltmeter across the calibrated resistor R_1.

d) *Measurement procedure*

The temperature is set to the specified value.

The variable voltage generator is adjusted to obtain the specified value of reverse voltage V_R across the diode.

The reverse current I_R is read from the ammeter A.

e) *Specified conditions*

– Ambient, case or reference-point temperature (t_{amb}, t_{case}, t_{ref}).

– Reverse voltage (V_R).

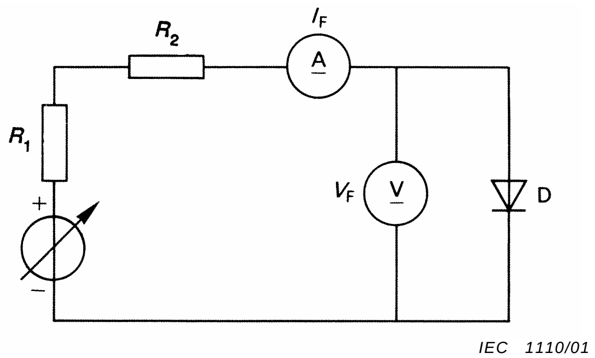
– Pulse width and duty cycle, where applicable.

3.1.4.2 Forward voltage V_F

a) *Purpose*

To measure the forward voltage across a signal or switching diode under specified conditions.

b) *Circuit diagram*



Key
D diode being measured

Figure 3 – Circuit for the measurement of forward voltage V_F

c) *Circuit description and requirements*

R_1 is a calibrated resistor (pulse measurement only).

R_2 is a high value resistor.

If a pulse measurement is required, the variable voltage generator is replaced by a voltage pulse generator, the voltmeter is replaced by a peak-reading instrument and the ammeter is replaced by a peak-reading voltmeter across the calibrated resistor R_1.

d) *Measurement procedure*

The temperature is set to the specified value.

The variable voltage generator is adjusted to obtain the specified value of forward current I_F.

The forward voltage V_F is read from the voltmeter V.

e) *Specified conditions*

– Ambient or case temperature (t_{amb}, t_{case}).

– Forward current (I_F).

– Pulse width and duty cycle, where applicable.

3.1.4.3 Capacitance C_{tot}

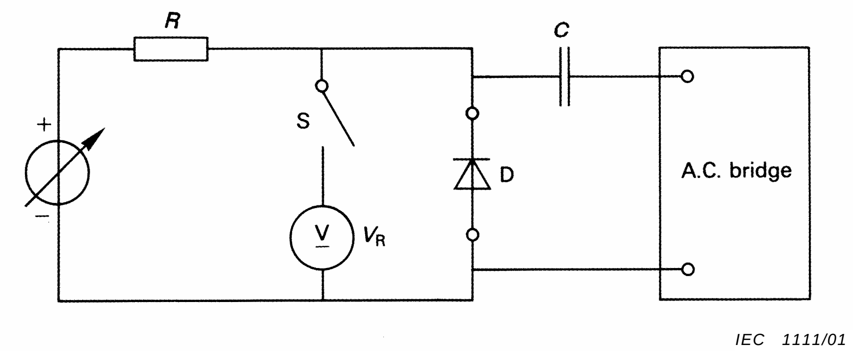
The measurement of total capacitance ($C_{tot} = C_j + C_p$) should be made at a sufficiently low frequency (below microwave frequencies) so that the effects of the lead inductance may be neglected. Under these conditions, the measured value of terminal capacitance is independent of frequency.

The total capacitance at a given bias condition is obtained by the method stated hereafter.

a) *Purpose*

To measure the total capacitance of a diode under specified conditions.

b) *Circuit diagram*



Key
D diode being measured

Figure 4 – Circuit for the measurement of capacitance C_{tot}

c) *Circuit description and requirements*

The conductance of resistor *R* should be low compared with the admittance of the diode being measured.

The capacitor *C* must be able to withstand the reverse bias voltage of the diode and should present a short circuit at the frequency of measurement.

d) *Precautions to be observed*

The bridge shall be able to withstand the reverse bias voltage of the diode without affecting the accuracy of the measurement. If the measured capacitance is very small, the mounting conditions will affect the accuracy of the results and they should be specified.

e) *Measurement procedure*

The temperature is set to the specified value.

The voltage across the diode is adjusted to the specified value V_R. Then the voltmeter V is taken out of the circuit and the capacitance of the diode being measured is determined using the a.c. bridge by subtracting the value without the diode in its mounting from the value with the diode in its mounting.

f) *Specified conditions*

Ambient or case temperature (t_{amb}, t_{case}).

– Reverse voltage (V_R).

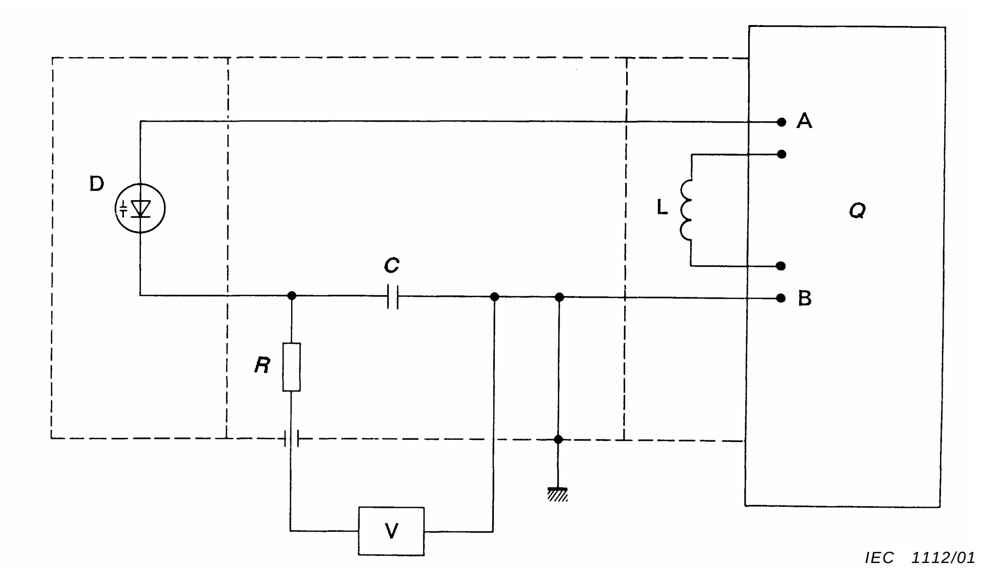
– Measurement frequency, if different from 1 MHz.

– Mounting conditions of the diode, if necessary.

NOTE The variation of total capacitance with bias voltage may be found by measurements as described above, made at a number of bias points.

3.1.4.4 Effective quality factor *Q*

The effective quality factor *Q* of a variable capacitance diode can be measured using a "*Q*-meter" or an impedance bridge (see Figure 5).



Key
D diode being measured
V voltage source
Q *Q*-meter

Figure 5 – Circuit for the measurement of effective quality factor

Description

a) The voltage source should present a high impedance at the frequency of measurement compared to that of the capacitor *C*; this is obtained by means of series resistor *R*.

b) *C* is a decoupling capacitor having a low impedance at the frequency of measurement.

c) L is an inductor chosen to resonate with the parallel circuit capacitor at the frequency of measurement.

d) It is assumed that there is a low resistance path through the *Q*-meter between points A and B.

The basic circuit of such a meter consists of a signal generator of negligible output impedance driving a high Q inductance in series with a high-quality variable capacitance. The factor Q of this circuit can be measured at a given frequency by tuning the variable capacitance for resonance.

Q is given by the ratio of the voltage across the capacitance to the voltage supplied by the generator. In order to measure the factor Q of a variable capacitance diode, it shall be connected in parallel with the variable capacitance in the Q-meter. DC isolating components shall be used so that the desired bias voltage may be applied to the diode being measured, but the biasing circuit must remain connected to the Q-meter throughout the measurement.

Four measurements are made: Q and C_1, the factor Q of the circuit and the magnitude of the variable capacitance with the diode not in circuit; and Q_2 and C_2, the factor Q of the circuit and the value of the variable capacitance for resonance at the same frequency with the diode connected to the circuit.

The factor Q of the diode is then calculated using the expression:

$$Q = \left(\frac{Q_1 Q_2}{Q_1 - Q_2}\right)\left(\frac{C_1 - C_2}{C_1}\right)$$

Two precautions are necessary:

1) The measurement shall be made at a frequency at which the reactance of the self-inductance of the diode is negligible compared with the reactance of the capacitor.
2) The magnitude of the signal applied to the variable capacitance diode shall be kept relatively small so that only a small excursion is made over the non-linear capacitance characteristic. The result must be independent of the signal level.

   NOTE

$$Q = \frac{1}{2\pi f \times C_{\mathrm{j}} \times r_{\mathrm{s}}} = \frac{f_{\mathrm{c}}}{f}$$

   Since $C_{\mathrm{p}} \leq C_{\mathrm{j}}$ for these diodes, C_{t} and C_{j} can be used interchangeably in this section.

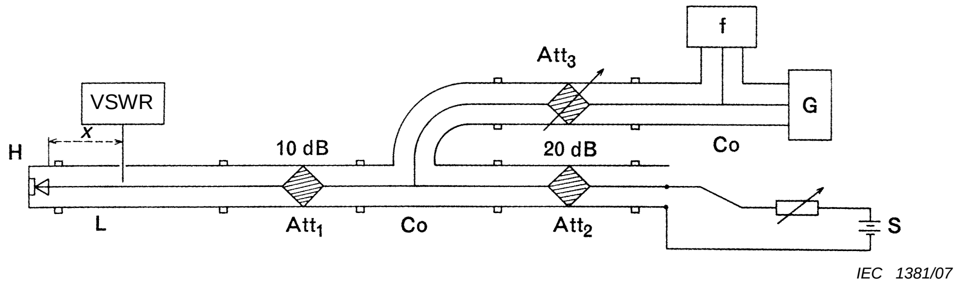
3.1.4.5 Series resistance r_{s}

The effective value of series resistance r_{s} can be deduced from the values of C_{j} and f using the formula given in 3.1.4.4.

3.1.4.6 Series inductance L_{s}

Measurements should be conducted in the frequency region where the effect of stray capacitance C_{p} relative to the terminal impedance of the diode can be neglected.

The diode is inserted in the measuring head as shown in Figure 6 which is set on the tip of the inner conductor of the coaxial slotted line.



Key

VSWR voltage standing wave ratio meter
x distance
H diode head
L slotted line
Att attenuator
Co coupler
G microwave generator
S bias supply
f frequency meter

Figure 6 – Circuit for the measurement of series inductance

Measurements are as follows:

First, determine position x_m where the standing wave voltage is minimum as measured at a bias voltage in the forward region where the terminal capacitance becomes independent of the change of bias voltage. This bias voltage should be sufficiently high so that an increase of this voltage would not affect the result of the measurement. (This condition may be satisfied when about 5 mA forward current flows.)

Next, without any break in the impedance of the line, a metal block is inserted in the measuring head in place of the diode. This is done in order to provide a short-circuit at the reference plane position which is defined and should be specified by the manufacturer of the diode. In this condition, position x_s nearest to x_m and larger than x_m is found where the standing wave voltage is minimum.

The reactance of the diode is obtained by the following equation:

$$X = Z_0 \tan \frac{2\pi(x_s - x_m)}{\lambda}$$

where

Z_0 is the characteristic impedance of the coaxial line;

λ is the wavelength of the measuring frequency.

The series inductance L_s can be obtained by use of the following equation:

$$L_s = \frac{X}{2\pi f}$$

NOTE The structure of some devices may prevent this method of measurement from giving correct results. In this case, a value for the inductance will have to be given by the manufacturer.

3.1.4.7 Thermal resistance R_{th}

3.1.4.7.1 Purpose

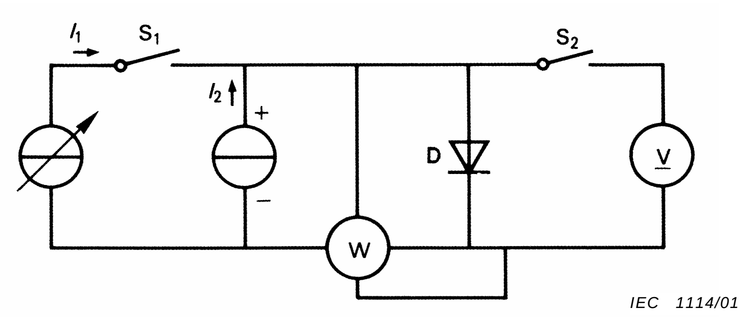
To measure the thermal resistance between the junction and a reference point (preferably at the case) of the device being measured.

3.1.4.7.2 Principle of the method

The temperatures T_1 and T_2 of the reference point of the device are measured for two different power dissipations P_1 and P_2 and cooling conditions causing the same junction temperature. The forward voltage at a reference current is used to verify that the same junction temperature has been reached.

$$R_{th} = \frac{T_1 - T_2}{P_2 - P_1}$$

3.1.4.7.3 Basic circuit diagram



Key
D device being measured

Figure 7 – Circuit for the measurement of thermal resistance R_{th}

3.1.4.7.4 Circuit description and requirements

I_1 = load current generating the power loss P in the junction, either a d.c. current or an a.c. current

I_2 = reference d.c. current monitored when the load current I_1, is interrupted periodically for short time gaps

W = wattmeter to indicate the power loss P in the junction caused by the load current I_1; for the a.c. method, W measures the average power dissipated in the device being measured

S_1 = electronic switch to interrupt periodically the load current I_1; for the d.c. method, switch S_1 is not mandatory

S_2 = electronic switch, which is closed when the load current I_1, is interrupted

V = null-method voltmeter

3.1.4.7.5 Precautions to be observed

Voltage transients occur due to excess charge carriers when switching from the load current I_1, to the reference current I_2. Additional voltage transients occur if the case of the device under test contains ferromagnetic material. The switch S_2 should not be closed before these transients have disappeared.

NOTE The load current I_1 defined in 3.1.4.7.4 may be zero, in which case the power loss P_1 is also zero and the virtual junction temperature is the same as the reference-point temperature T_1.

3.1.4.7.6 Measurement procedure

The device being measured is clamped onto a heat sink maintained at a fixed temperature. A thermocouple is fixed at the reference point to measure the temperature of the device being measured. The measurement is carried out in two steps:

a) The heat sink is maintained at an elevated temperature. A low load current I_1, is applied causing the power loss P_1, in the junction. After reaching thermal equilibrium, the null-method voltmeter V is adjusted for zero balance.

   The reference-point temperature T_1 is recorded.

b) The heat sink is maintained at a lower temperature. The load current I_1, is raised until the power loss P_2 warms up the junction to the same temperature as in the preceding step. This is indicated by zero balance of the null-method voltmeter V.

   The reference-point temperature T_2 of the case is recorded.

   The thermal resistance R_{th} is calculated using the expression:

$$R_{th} = \frac{T_1 - T_2}{P_2 - P_1}$$

3.1.4.8 Transient thermal impedance Z_{th}

3.1.4.8.1 Purpose

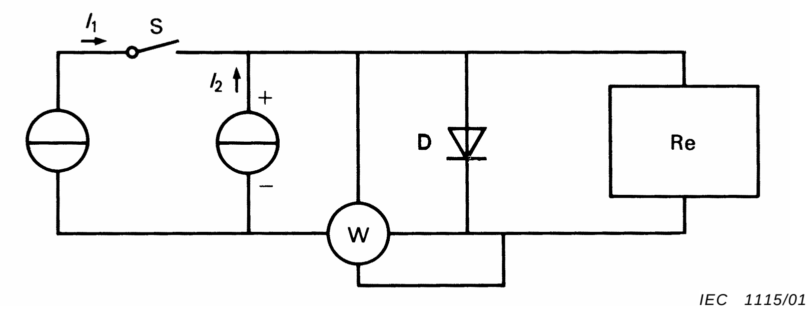
To measure the transient thermal impedance between the junction and a reference point (preferably at the case) of the device being measured.

3.1.4.8.2 Principle of the method

After applying the heating current and waiting until thermal equilibrium is reached, the power dissipated in the device is recorded. The heating current is then interrupted and the forward voltage at the reference current together with the reference-point temperature are recorded as a function of time.

The virtual junction temperature as a function of time is then calculated by means of the calibration curve obtained for the same reference current.

3.1.4.8.3 Basic circuit diagram



Key
D device being measured

Figure 8 – Circuit for the measurement of transient thermal impedance Z_{th}

3.1.4.8.4 Circuit description and requirements

I_1 = load current generating the power loss P in the junction

I_2 = reference d.c. current

S = switch to interrupt the load current I_1

W = wattmeter to indicate the power loss P in the junction caused by the load current I_1

Re = recording equipment, e.g. an oscillograph, to record the time variation of the forward voltage caused by I_2

3.1.4.8.5 Measurement procedure

1) A calibration curve is prepared by measuring the on-state or forward voltage generated by the reference current I_2 as a function of the virtual junction temperature by varying the device temperature externally e.g. by means of an oil bath.
2) The device being measured is clamped onto a heat sink maintained at a fixed temperature. A thermocouple is fixed at the reference point to measure the reference point temperature T_c of the device being measured. The heating current I_1 is applied generating the power loss P in the device being measured until thermal equilibrium is reached.
3) The heating current I_1, is interrupted by opening the switch S. The forward voltage generated by the reference current I_2 is recorded as a function of the cooling time by the recording equipment Re. The reference point temperature is recorded during this time.
4) The curve of the recorded forward voltage is converted to the virtual junction temperature T_{vj} by means of the calibration curve. The transient thermal impedance $Z_{(th)t}$ is calculated using the expression:

$$Z_{(th)t} = \frac{\left[T_{vj}\,(0) - T_c\,(0)\right] - \left[T_{vj}\,(t) - T_c\,(t)\right]}{P}$$

where

T_{vj} (0) and T_c (0) are the temperatures at the time t = 0 when opening switch S;

T_{vj} (t) and T_c (t) are the temperatures at the time t.

3.1.4.9 Case of varactor diodes

The following methods of measurement are recommended for use as appropriate to the intended conditions of operation and structure of the type of diode to be measured.

In the case of the measurement of the effective factor Q of the diode, it is recommended that, when a value of Q is quoted, the particular method of measurement used to obtain that value should be stated. This is necessary because it is possible to obtain different values of Q for a given diode when using the two given methods.

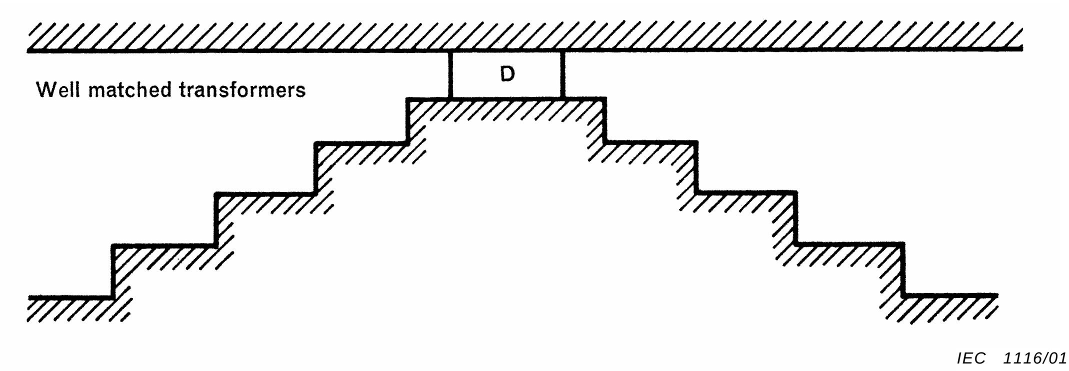
3.1.4.9.1 Transmission line measurements

These measurements are suitable for evaluating the main properties of microwave diodes which may be used in a wide range of applications, particularly those diodes which are unencapsulated, or those diodes whose package shunt capacitance has a reactance value larger than the value of diode series resistance at the series resonant frequency.

3.1.4.9.1.1 Theory

Observation is made of the effect on the transmission characteristics of any non-radiating transmission system by the introduction of a shunt impedance, in this case a diode.

The diode is mounted in shunt with the transmission line so that the mounting arrangement provides a minimum of excess reactance; for example, when using a waveguide transmission system, the diode is fitted as given in Figure 9.



Key
D diode being measured

Figure 9 – Waveguide mounting

Measurements of transmission loss introduced by the diode in the region of the series resonant frequency enable the elements of the diode equivalent circuit to be evaluated and also permit the capacitance law as a function of bias to be determined.

The equivalent circuit of the mounted diode is shown in Figure 10.

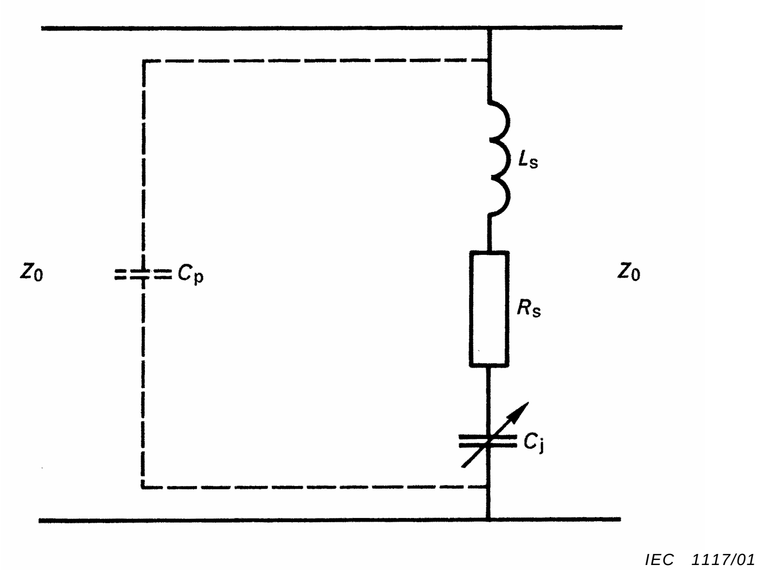


Figure 10 – Equivalent circuit of mounted diode

where

Z_0 is the characteristic impedance of the transmission line;

C_p is the package capacitance;
L_s is the series inductance;
R_s is the series resistance;
C_j is the junction capacitance.

Near series resonance, the effect of the package capacitance (C_p) is negligible and may be ignored.

Four measurements, namely:

a) transmission loss at the series resonant frequency at zero bias;

b) the bandwidth of the transmission characteristic;

c) the value of the series resonant frequency;

d) the variation of the series resonant frequency with bias;

enable the four unknown quantities:

1) series resistance (R_s);

2) junction capacitance (C_j);

3) series inductance (L_s);

4) variation of junction capacitance with bias to be determined.

3.1.4.9.1.2 Circuit diagram

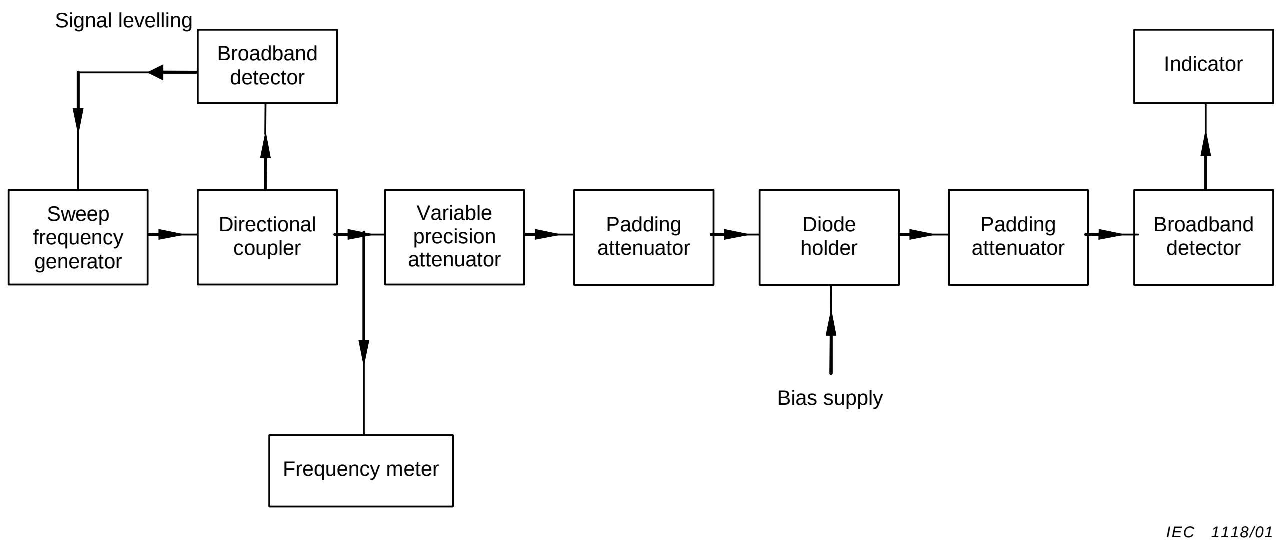


Figure 11 – Block diagram of transmission loss measurement circuit

3.1.4.9.1.3 Circuit description and requirements

The test equipment should be assembled using good microwave transmission line engineering techniques. All components, such as directional couplers, frequency measuring apparatus, attenuators and detectors, should be checked to ensure proper matching and operation over the required frequency and power test conditions.

The components of the system should be sufficiently broadband to ensure that only negligible variations or errors over the band of frequencies used for the measurement are introduced.

The RF signal generator should be capable of stable operation at a signal level equivalent to the normal small-signal conditions of the diode.

The diode holder should conform with the specified mount details.

A typical arrangement comprises a tapered mount with a choke on one face to enable bias to be applied. The tapered mount usually is a requisite feature to ensure that only the diode characteristics are being measured. In this way, the complication of using inductive posts for mounting the diode is avoided (see Figure 9).

3.1.4.9.1.4 Measurement procedure

The diode is inserted into the specified holder which is connected in a transmission system equivalent to that shown in Figure 11.

3.1.4.9.1.4.1 Series resonant frequency

The series resonant frequency may easily be obtained by operating the diode at the required bias voltage and observing the indicated transmitted power, in front of and behind the diode, as the frequency is swept over a suitable frequency range. The series resonant frequency is indicated by the point of minimum transmitted RF power. The incident RF power level on the diode shall be kept constant during the sweep.

3.1.4.9.1.4.2 Transmission loss (*T*)

The transmitted signal level at resonance with zero bias (or any other required value) applied to the diode is recorded. The diode is then removed from the holder and the precision attenuator adjusted to give the same indicated transmitted signal level as the one recorded initially. The change in the attenuator setting then gives the transmission loss (T) at resonance. It is essential that the incident RF power level on the diode shall be kept constant during this measurement.

Alternatively, the transmission loss introduced by the diode at the series resonant frequency may be obtained by firstly observing the power level incident on the matched detector at a frequency remote from the resonant frequency. The frequency is then changed to the resonant value and the precision attenuator adjusted to return the indicated power level to the same value as that obtained when the frequency was remote from resonant value. The change in attenuator reading will provide the transmission factor (T) (see Figure 12).

3.1.4.9.1.4.3 Series resistance

If the frequency of measurement chosen is equal to the series resonant frequency (f_s) given by:

$$f_s = \frac{1}{2\pi\sqrt{L_s C_j}} \tag{1}$$

where

L_s is the series inductance;

C_j is the effective capacitance of the PN junction having a required applied bias voltage.

The loss in a transmission may be measured as in 3.1.4.9.1.4.2 and the effective shunt resistance derived from:

$$R_s = \frac{Z_0}{2\sqrt{T} - 1} \tag{2}$$

where

Z_0 is the characteristic impedance of the transmission line in the vicinity of the loss element. In the case of a waveguide mount, the power/voltage definition should be used;

T is the ratio of available power incident on the diode being measured to that transmitted past the diode.

The variation of R_S with bias may be obtained by the adjustment of the measuring frequency to the corresponding series resonant value obtained for each bias value used and measuring the transmission factor (T) in each case.

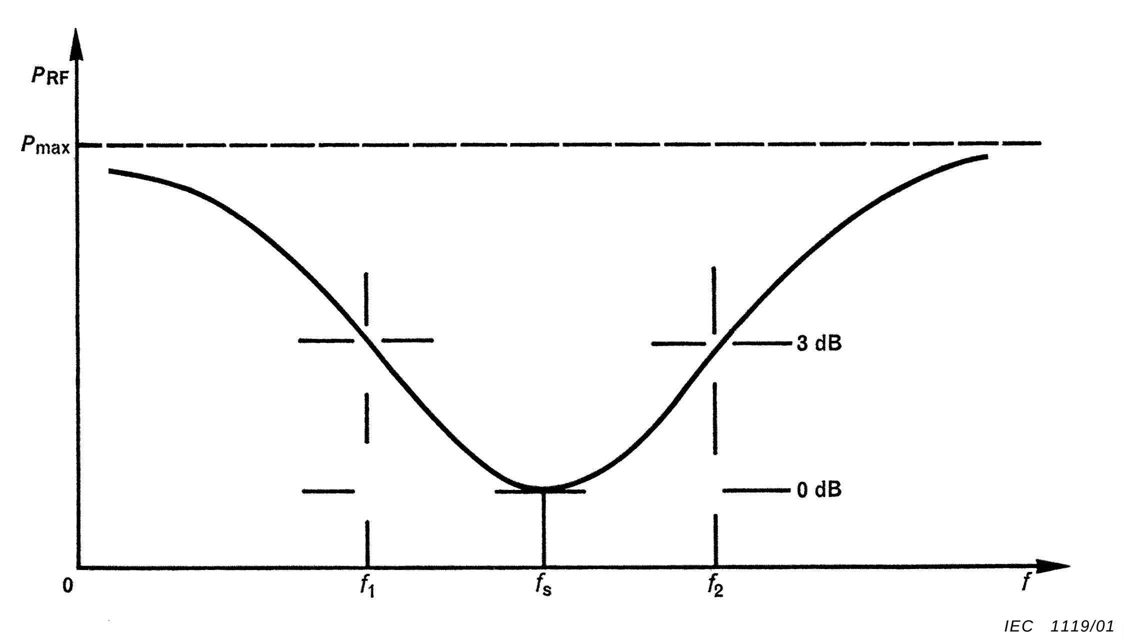
A measure of the change in the effective Q value with bias may also be obtained.

3.1.4.9.1.4.4 Effective *Q* value

a) *First method*

The effective Q value at a given bias voltage may be obtained by varying the measuring frequency to values on either side of the series resonant frequency and observing the value of those frequencies which cause the power transmitted to be twice the one obtained at the resonant frequency (see Figure 12). If the frequencies at which this is achieved are f_1 and f_2, then:

$$Q = \frac{f_S}{|f_1 - f_2|}$$



Key
f_s series resonant frequency

Figure 12 – Curve indicating transmitted power versus frequency

Alternatively, since

$$C_j \sim \frac{1}{f_s^2}$$

from equation (1), a plot of $\frac{1}{f_s^2}$ against bias will provide a plot of KC versus bias, where K is a constant.

If the frequency is adjusted to the series resonant frequency at zero bias, the forward bias voltage (V_1) and the reverse bias voltage (V_2) required to double the transmitted power are obtained.

Using the plot of KC_j versus bias, corresponding values of KC_{j1} and KC_{j2} may be found. The value of Q may then be derived from:

$$Q = \frac{KC_{j1} + KC_{j2}}{KC_{j1} - KC_{j2}}, \text{ i.e. : } \frac{f_{s2}^2 + f_{s1}^2}{f_{s2}^2 - f_{s1}^2}$$

If $f_{s2} - f_{s1}$, is small, this can be reduced to:

$$\frac{f_s}{f_{s2} - f_{s1}}$$

without serious error.

b) *Alternative method*

The effective Q value may also be obtained by the transformed impedance measurement as given in 3.1.4.10.

3.1.4.9.1.4.5 Cut-off frequency

The cut-off frequency (f_c) at zero bias may then be obtained from:

$$f_{c(0V)} = Qf_{s(0V)} \tag{3}$$

3.1.4.9.1.4.6 Junction capacitance at zero bias

This may be obtained using the value of the cut-off frequency (f_c) from equation (3) and R_s from equation (2):

$$C_{j(0V)} = \frac{1}{2\pi R_s f_{c(0V)}}$$

The value of junction capacitance at zero bias may then be used to calibrate the plot of $\frac{1}{f_s^2}$ versus bias in terms of C_j (see 3.1.4.9.1.4.4).

3.1.4.9.1.4.7 Series inductance

If the series inductance value is required, this may be obtained from:

$$f_{s(0V)} = \frac{1}{2\pi\sqrt{L_s C_{j(0V)}}}$$

3.1.4.9.1.4.8 Capacitance variation coefficient (γ)

The capacitance variation coefficient is defined as the normalized capacitance change over a defined range of operating conditions of forward current and reverse voltage.

The bias voltage which is required to provide the defined value of forward current is determined. Then using this forward voltage (V_F) and the defined reverse voltage (V_{-x}), corresponding values of C_j may be obtained from the $\frac{1}{f_s^2}$ or KC_j plot against the bias voltage (see 3.1.4.9.1.4.4).

If $C_{j(v_F)}$ and $C_{j(v_{-x})}$ are the capacitance values respectively, then:

$$y = \frac{C_{j(v_F)} - C_{j(v_{-x})}}{2\left[C_{j(v_F)} + C_{j(v_{-x})}\right]}$$

3.1.4.9.2 Cavity method

This method is satisfactory for measuring varactors having an effective quality factor which exceeds 15 at the measuring frequency; the results are not affected by changes in the series resistance with bias.

NOTE It is considered that this method is usable up to a measurement frequency of 15 GHz (whereas method 1, described in 3.1.4.9.1, is more practical for measurement above 6 GHz).

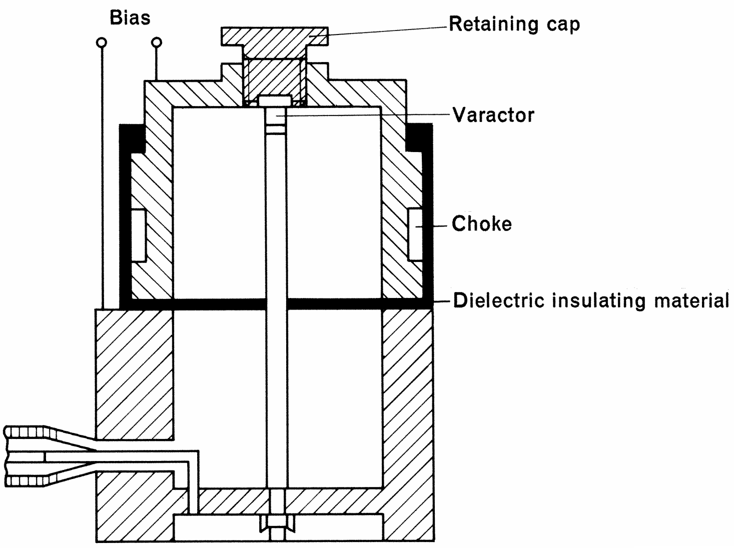


Figure 13 – Example of cavity

3.1.4.9.2.1 Theory

In this method, the effect is evaluated of a varactor diode on the resonant frequency and Q-factor of a coaxial cavity resonator about half-wavelength at the operating frequency. The diode is mounted between the centre conductor and the plane wall of the cavity resonator as shown in Figure 13.

In order to limit the range of variation of the cavity resonant frequency when the junction capacitance or the diode is changed, it is essential to use a resonator having a large ratio of external to internal conductor diameter (high-characteristic impedance of the coaxial cavity).

This method will determine the junction capacitance C_{jo} and the cut-off frequency f_{co} at bias voltage V_o.

These quantities enable the determination of the series resistance r_s and the effective Q-factor Q_{eff} of the varactor.

The following characteristics must be determined for the cavity:

f_{ro} is the resonant frequency of the cavity with the varactor at the bias voltage V_o;

Q_{ro} is the loaded Q-factor of the cavity with the varactor at the bias voltage V_o;

C_p is the stray capacitance of the varactor case;

$C_T(V)$ is the variation of the total low-frequency capacitance of the varactor (junction capacitance) versus the bias voltage around V_o;

$f_r(V)$ is the variation of the resonant frequency of the cavity with the varactor versus bias voltage around V_o;

f' is the resonant frequency of the cavity when the varactor is replaced by a metallic dummy diode with the same dimensions as the diode being measured;

Q' is the unloaded-Q of the cavity when the varactor is replaced by the dummy diode.

From the knowledge of C_T(V) and $f_r(V)$, a curve can be derived which represents f_r versus C_T. This curve enables a quantity "a" to be evaluated, "a" being the slope of the curve at $C_T = C_T(V_o)$.

This junction capacitance, deduced at the bias voltage V_o is given by:

$$C_{jo} = C_T(V_o) - C_p$$

and the cut-off frequency:

$$f_{co} = 2\ a\ k\ Q_{ro}\ C_{jo}$$

where k is a correction factor introduced to take into account losses in the cavity walls; for the second TEM resonance frequency, it is given by:

$$k = \frac{1}{1 - \dfrac{Q_{ro}}{Q'}\sqrt{\dfrac{f'}{f_{ro}}}}$$

Series resistance, at the bias voltage V_o, is given by:

$$r_{so} = \frac{1}{2\pi f_{co} C_{jo}}$$

and the effective Q_{eff} at bias voltage V_o is given by:

$$Q_{eff} = \frac{f_{co}}{f_{ro}}$$

3.1.4.9.2.2 Measurements

3.1.4.9.2.2.1 Circuit diagram

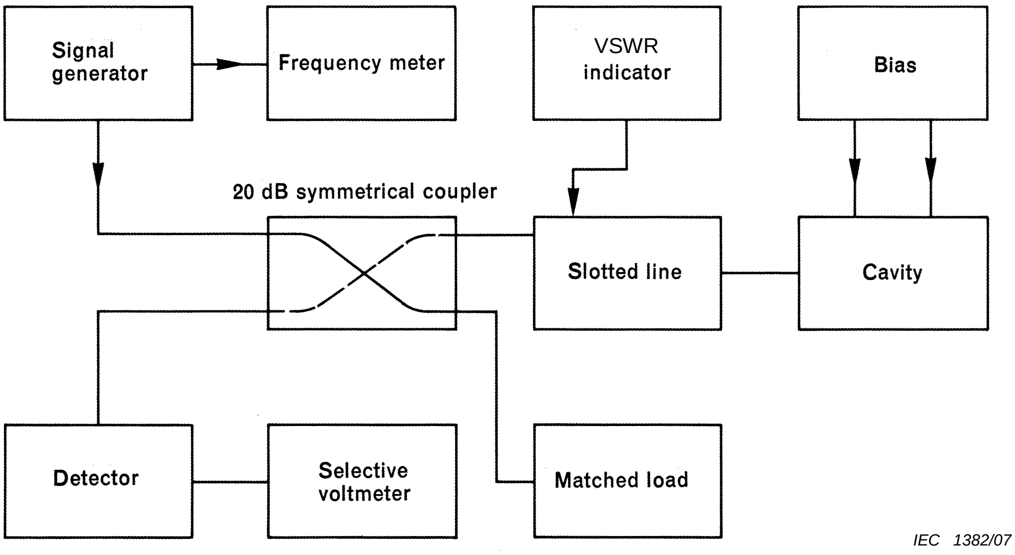


Figure 14 – Block diagram for the measurement of effective *Q* in cavity method

3.1.4.9.2.2.2 Circuit description and requirements

The RF signal shall be of high-frequency stability and modulated at a low frequency appropriate to the selective voltmeter and VSWR indicator and is applied to the cavity through a 20 dB directional coupler.

The amplitude of the peak RF signal V_p must be low enough to ensure that non-linearity does not occur at the operating point of the characteristic.

The incident power at the cavity input shall not exceed the value given by the expression:

$$P = \frac{\pi \ (r+1)^2}{4\,r} \ \frac{f_{ro}^2}{f_{co}} \ C_{jo} \, V_p^2$$

where r indicates the VSWR in the slotted line at the input of the cavity. Since the limitation on the incident power is not critical, an estimated value can be used for f_{co}.

3.1.4.9.2.2.3 Measurement procedure

a) *Measurement of $f_r(V)$ and f_{ro}*

The measurement of the resonant frequency $f_r(V)$ is performed at a number of bias points around V_o (e.g. if V_o = –6 V, f_r can be measured at the following voltages: –4 V; –4,5 V; –5 V; –5,5 V; –6 V; –6,5 V; –7 V; –7,5 V).

The measurement is performed by varying the signal frequency and observing the value for minimum reflected power. To ensure a high accuracy, it is better to determine f_r as the average between two frequencies adjacent to f_r which have the same power from the cavity.

b) *Measurement of Q_{ro}*

Set the bias voltage to V_o and determine the value of the loaded Q_{ro} by means of VSWR measurement.

c) *Measurement of f' and Q'*

These values are obtained in the same way as f_r and Q_{ro} after the varactor has been replaced by a dummy diode.

These are fundamental characteristics of the cavity.

3.1.4.9.2.3 Measurement of $C_T(V)$ and C_p

The total capacitance of the varactor diode:

$$C_T(V) = C_j(V) + C_p$$

is obtained by a conventional low-frequency bridge measurement.

The value of C_p can be deduced using the expression:

$$C_p = \frac{(\varphi - V_2)^n\, C_T(V_2) - (\varphi - V_1)^n\, C_T(V_1)}{(\varphi - V_2)^n - (\varphi - V_1)^n}$$

where

V_1 and V_2 are the two values of bias voltage; for reverse bias, V_1 and V_2 will be negative terms;

φ is the contact potential difference (e.g. 0,7 V for silicon diodes);

n is the factor of non-linear dependence of C_j on V.

3.1.4.9.2.4 Direct measurement of C_p

C_p can be measured directly when the ohmic contact between the internal metallic lead and the semiconductor chip has been interrupted in a varactor.

3.1.4.10 Transformed impedance method

This method is satisfactory for the measurement of diodes which are only to operate within that part of the diode characteristic in which the value of series resistance is sensibly independent of the value of the bias voltage.

3.1.4.10.1 Theory

The normalized impedance (Z) at any place in a lossless transmission line is related to the reflection coefficient (ρ) at that place by the expression:

$$Z = \frac{1+\rho}{1-\rho}$$

The form of this relation indicates that the normalized impedance and the reflection coefficient at any plane are bilinearly related; hence it may be shown that, for a lossless transformation between two impedance planes Z_1 and Z_2, one can write:

$$Z_2 = \alpha Z_1 + \mathrm{j}\beta \tag{4}$$

where α and β are real numbers.

If two values of impedance (Z^a and Z^b) at one place which only differ in the value of their reactive components are then considered, corresponding impedance at a second plane may be written as:

$$Z_2^a = \alpha (R_1 + \mathrm{j}X_1) + \mathrm{j}\beta = \alpha R_1 + \mathrm{j}(\alpha X_1 + \beta) \tag{5}$$

and $$Z_2^b = \alpha (R_1 + \mathrm{j}X_1 + \Delta X) + \mathrm{j}\beta = \alpha R_1 + \mathrm{j}(\alpha X_1 + \alpha \Delta X + \beta) \tag{6}$$

From equation (4), it will be seen that circles of constant resistance on a Smith chart at one plane transform into the same family of circles at another, but that the resistance value is changed in the ratio α.

This transformation is pertinent to the reactance values, so that the ratio $\frac{\Delta X}{R}$ as obtained from equations (5) and (6), is seen to be independent of the transformation constants α and β.

Thus, for a transmission line which is terminated in an impedance whose reactive component is varied, the impedance locus at a plane in the measuring line which corresponds to the terminal plane also lies on a circle of constant resistance.

If the impedance plane of Z_1 is taken as being that of the diode element itself, then $\frac{\Delta X}{R}$ in any corresponding plane is the same as the value ΔQ of the diode.

Hence if:

$$\Delta X_{12} = |X_1 - X_2|, \text{ then } \frac{\Delta X_{12}}{R} = \Delta Q_{12} \text{ i.e. } |Q_1 - Q_2|$$

where the subscripts 1 and 2 correspond to the value of the parameter which is obtained at the bias voltages V_1 and V_2 respectively.

Now the effective quality factor Q at any required point may be given by:

$$Q = \Delta Q \times \sigma \tag{7}$$

where σ is a constant factor relating two bias points, for a given type of diode.

One of the two bias points (V_1 or V_2) can be the required value.

Derivation of σ

The value of σ_{12} may be obtained from the expression:

$$\sigma_{12} = \frac{C_1}{\Delta C_{12}} = \frac{1}{1 - \left(\frac{\phi - V_1}{\phi - V_2}\right)^{\eta}} \tag{8}$$

where

ϕ is the quasi-contact potential difference;

η is the factor of non-linear dependence of *C* on *V*.

EXAMPLE: For silicon varactors made by a diffusion process, η is usually given as $\frac{1}{3}$ and ϕ is of the order of 0,5 V.

If V_1 = –4,5 V and V_2 = –6 V, then σ = 10, i.e. $Q = 10\ \Delta Q$.

The value of σ may be obtained experimentally by the measurement of the capacitance variation between three closely grouped bias points, say 1, 2 and 3, to give ΔC_{12} and ΔC_{23}.

The value of σ may then be obtained from:

$$\sigma = \frac{1 + \frac{\Delta C_{23}}{\Delta C_{12}}}{1 - \frac{\Delta C_{23}}{\Delta C_{12}} \frac{\Delta Q_{12}}{\Delta Q_{23}}} \tag{9}$$

The corresponding values of ΔQ_{12} and ΔQ_{23} may be obtained using equations such as (10), (11), (12) or (13) as convenient.

Derivation of ΔQ

The value of ΔQ may be obtained from using either of the two equations (10) or (12) as shown below:

$$\Delta Q_{12} = \sqrt{\frac{(\eta_1 - \eta_2)^2 + (\eta_1^2 - 1)(\eta_2^2 - 1)\sin^2 \Delta\Psi}{\eta_1 \eta_2}} \tag{10}$$

where

10 log (η_1^2) is the power standing wave ratio (dB) at bias value 1;

10 log (η_2^2) is the power standing wave ratio (dB) at bias value 2:

$$\Delta\Psi = \frac{M_1 - M_2}{\lambda_g} \times 360°$$

where M_1 and M_2 are positions of minimum at bias values 1 and 2 respectively.

For very large values of η_1 and η_2 (viz. values usually obtained in the case of high-quality diodes), equation (10) may be simplified to:

$$\Delta Q_{12} = \sqrt{\eta_1 \eta_2} \sin \Delta\Psi \tag{11}$$

$$\Delta Q_{12} = \sqrt{A_1\, A_2 - B_1\, B_2\, \cos\theta - 2} \tag{12}$$

where $A_1\ A_2$ and $B_1\ B_2$ are of the general form:

$$A_x = \frac{r_x^2 + 1}{r_x \sqrt{2}} \quad \text{and} \quad B_x = \frac{r_x^2 - 1}{r_x \sqrt{2}}$$

where

r_x is the VSWR at bias value x;

θ is the phase change of reflection coefficient between bias values 1 and 2.

This formula in practice may, for an accuracy better than 1 %, be reduced to:

$$\Delta Q_{12} = \sin\frac{\theta}{2}\ \sqrt{r_1\, r_2} \tag{13}$$

Considering equation (12) for the particular case when adjusting for r_1 = 1 (i.e. matched condition at required bias voltage), the formula reduces to:

$$\Delta Q_{12} = \frac{r_2 - 1}{\sqrt{r_2}} \tag{14}$$

If adjustments are made to provide matched conditions at the required bias voltage, the impedance in the measuring plane will be coincidental with the unit resistance circle on the Smith chart. It follows that this defines the plane in which the impedance Z_{in} is given by:

$$Z_{in} = Z_0\left(1 + j\frac{\Delta X}{R}\right) = Z_0(1 + j\Delta Q) \tag{15}$$

where

Z_0 is the characteristic impedance of the transmission line.

The change in reactance ΔX is then measured "relative to R" as ΔQ, to give:

$$\Delta Q = \frac{X_{(v)} - X_0}{R} = Q - Q_0 \tag{16}$$

Q_0 may be found by replacing the non-linear element with a short-circuit at the plane of the diode; whence:

$$\frac{|Z_{in}|}{R_0} = \frac{|X_0|}{R} = Q_0. \tag{17}$$

The value of Q at any other bias may then be found using equation (16).

3.1.4.10.2 Measurement procedure

The diode is inserted into a specified holder and connected to a circuit equivalent to the one shown in Figure 15.

The transformation between the diode and slotted line is obtained using a variable short-circuit behind the diode and tuning stubs in front of the diode (e.g. an E-H tuner).

The bias voltage is then adjusted to the required value and transformation adjustment made so that the measured normalized impedance point is in the central region of a Smith Chart where the scale is the most open, for example to provide a match in the measuring line at the required bias value. The effect on accuracy, for non-matched conditions over a substantial impedance range about the centre of the Smith chart, is small.

The bias voltage is then adjusted to other bias voltage points as required and, maintaining a fixed tuner adjustment, the resultant normalized impedance values are plotted on the Smith chart.

The value of Q may then be obtained using the measured VSWR values; change in reflection phase and Equations (7) to (14) as appropriate.

It is possible to obtain the effective quality factor without the derivation of σ as given in equations (8) and (9).

The transformation is made to match the diode impedance into the transmission line at the required bias, and so as to obtain an impedance point at the centre of the Smith chart. The diode bias is then changed to other values, and corresponding impedance points on the Smith chart obtained. This means that any reactance change in the impedance of the diode will, in the measuring plane, be coincidental with the unit resistance circle on the Smith chart.

The diode is then replaced by an effective short circuit and the normalized impedance is measured using the same reference plane as for the diode. An example of a diode plot is given in Figure 16. The value of Q may then be derived using equations (16) and (17).

An effective short circuit may be approximated by the use of a diode encapsulation in which the semiconductor material has been replaced by a highly conductive material having identical geometry. In some cases, the impedance of the non-linear element (diode) can approach zero with a high forward current and, as a consequence, be acceptable as an effective short -circuit.

The plotted points obtained at the various bias values are rotated round the centre of the Smith chart so that they coincide with the unit resistance circle. The short-circuit point is similarly treated. (Note that the normalized impedance points for high forward current fall on a constant reactance line in the plot.)

As this variation in the method depends on

a) the effectiveness of the short circuit,

b) the ability to obtain a match condition in the measuring plane at the standard bias voltage, and

c) the effect of the tuning element losses,

it becomes difficult to accurately determine the real part of the *Q* value of the diode. It is therefore recommended that this form of measurement be restricted to diodes having a low *Q* factor and those diodes which operate in the lower microwave frequencies.

3.1.4.10.3 Precautions to be observed

a) The variable transformer and the mount losses shall be minimized. As the losses depend on the field pattern in the vicinity of the transforming elements, which in turn depend on the diode being measured, satisfactory correction is not readily achieved.

b) If accurate values are to be obtained, the line losses, etc. which can cause serious decrease in the measured values of standing wave ratios used in Equations (10) and (11), shall be determined. The transmission line length is the length between the standing wave probe position at the nearest voltage minimum and the plane of the active region of the varactor diode seated in its mount. In addition, the mount and connector loss shall be taken into account.

c) It should be verified that the series resistance is independent of varactor bias over an adequate range of the characteristic by checking that the impedance plot lies on the $\frac{R}{R_0}$ = 1 circle. However, deviation from a circle may be caused by losses. An estimate of the significance of the combined losses can be made by comparing measurements using different settings of the transforming elements and different match bias values.

   It is possible to transform points in one experimental plot to points close to the centre of the chart. If losses are negligible, the results will agree. For example, in Figure 16, the losses are negligible and the results for –9,0 V and –4,0 V, when matched at –6,0 V, should be shown by points marked by crosses.

   An alternative method to verify the dependence of the series resistance (R_s) on bias is to calculate the values of ΔQ_{12}, ΔQ_{23} and ΔQ_{13} for the three bias values as given for equation (9) and then examine whether the values satisfy the following relation:

$$\Delta Q_{13} = \Delta Q_{12} + \Delta Q_{23} \tag{18}$$

   If this relation is satisfied within acceptable limits, then it can be assumed that the series resistance value is sensibly independent of the bias voltage. Equations (7) and (9) may then be used to evaluate *Q*.

Capacitance measurement

This measurement is made usually at non-microwave frequencies. To obtain the capacitance of the non-linear element, the cartridge capacitance shall be subtracted from the total varactor capacitance.

The simplest and most direct method of obtaining the package capacitance is to substitute a unit in which there is no contact to the semiconductor. Another method can be used, if the form of the relation between capacitance and voltage is known (see, for example, equation (19)). The cartridge capacitance may be deduced by measuring the total capacitance at an appropriate number of bias points, which yields $C_c + C(V)$ and, since the form of $C(V)$ is known, both C_c and $C(V)$ can be obtained.

$$C(V) = C(V')\left(\frac{\phi - V'}{\phi - V}\right)^{1/n} \quad \text{with } n > 0 \tag{19}$$

NOTE Although the measurements in this subclause may be made using a standing wave detector, they may also be made by the use of an automatic impedance plotting instrument, an example of which is the automatic Smith Chart display unit. Because the value of Q is given by the normalized reactance change in a plane corresponding to the diode element for any lossless transformation, the Smith Chart may be adapted to give direct readings of Q as given in 3.1.4.11.

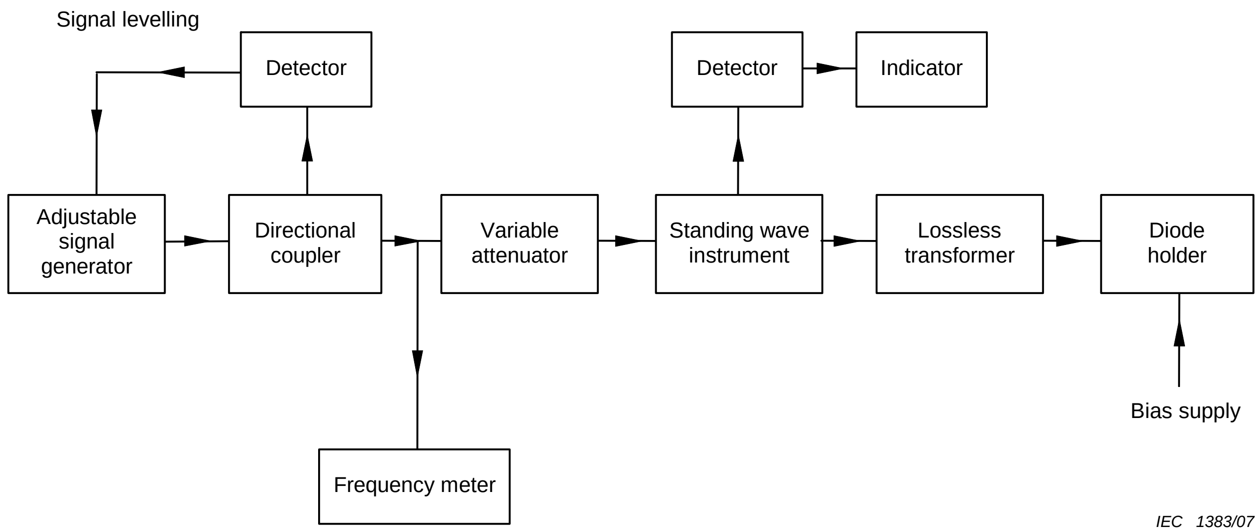
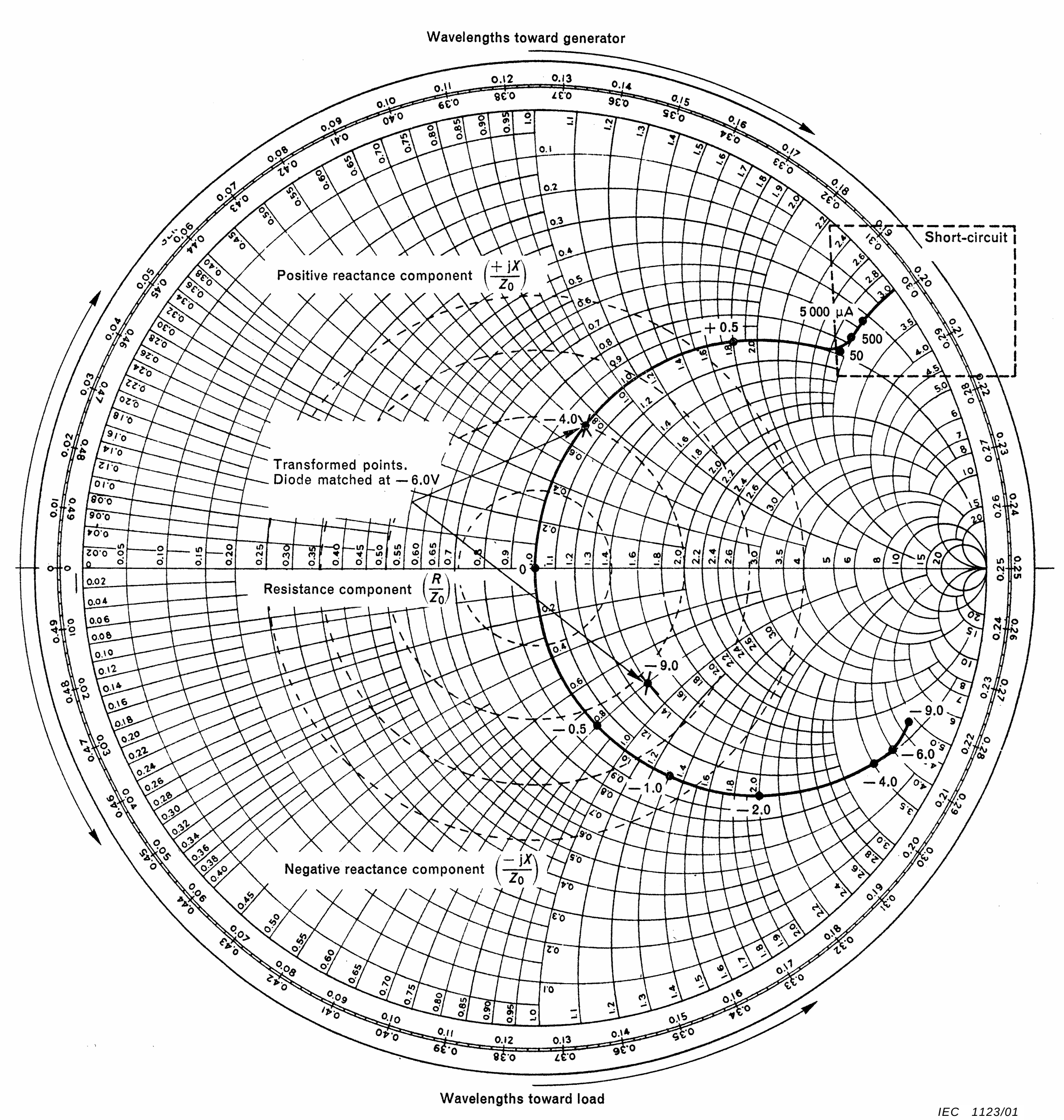


Figure 15 – Block diagram of transformed impedance measurement circuit



IEC 1123/01

Key

Frequency 10 GHz

Diameter ∅ 0,5 V

Figure 16 – Example of plot of diode impedance as a function of bias

3.1.4.11 Method of constant quality factor circles

As it has been shown in 3.1.4.10.1 (Theory), ΔQ is given by the normalized reactance change in a plane corresponding to the diode element for any lossless transformation; it follows that the Smith Chart may therefore be adapted to give direct readings of ΔQ from two impedance measurements for any arbitrary transformation. This may be done by introducing a grid of lines to represent fixed values of $\frac{X}{R}$ (i.e. Q).

The normalized impedance (Z) for any measuring plane is given in terms of the complex reflection coefficient (ρ) by:

$$Z = R + \mathrm{j}X = \frac{1+\rho}{1-\rho}$$

thus:

$$R = \frac{1-\rho\rho^*}{1+\rho\rho^* - \rho - \rho^*}$$

and:

$$\mathrm{j}X = \frac{\rho - \rho^*}{1+\rho\rho^* - \rho - \rho^*}$$

hence:

$$\frac{\mathrm{j}X}{R} = \frac{\rho - \rho^*}{1-\rho\rho^*} = \mathrm{j}Q$$

from which:

$$\rho\rho^* + \frac{\mathrm{j}}{Q}\rho^* - \frac{\mathrm{j}}{Q}\rho - 1 = 0 \qquad (20)$$

Equation (20) represents the equation of a circle and, when comparing it with the general equation for a circle, viz.:

$$(\rho - a)\,(\rho^* - a^*) = K^2$$

it can be deduced that the radius K is given from equation (20) by:

$$K = \sqrt{1 + \frac{1}{Q^2}}$$

and having a centre displaced from the origin of coordinates by a vector value of $\frac{\mathrm{j}}{Q}$.

A family of circles representing constant Q may thus be constructed on a Smith chart and these, together with the family of constant resistance circles, are sufficient to determine ΔQ. An example of the resulting chart is shown in Figure 17.

When applying the chart for the diode measurement, the diagram is orientated so that the measured normalized impedance points, corresponding to the two bias conditions, appear on a constant resistance circle. The corresponding *Q* values are then obtained.

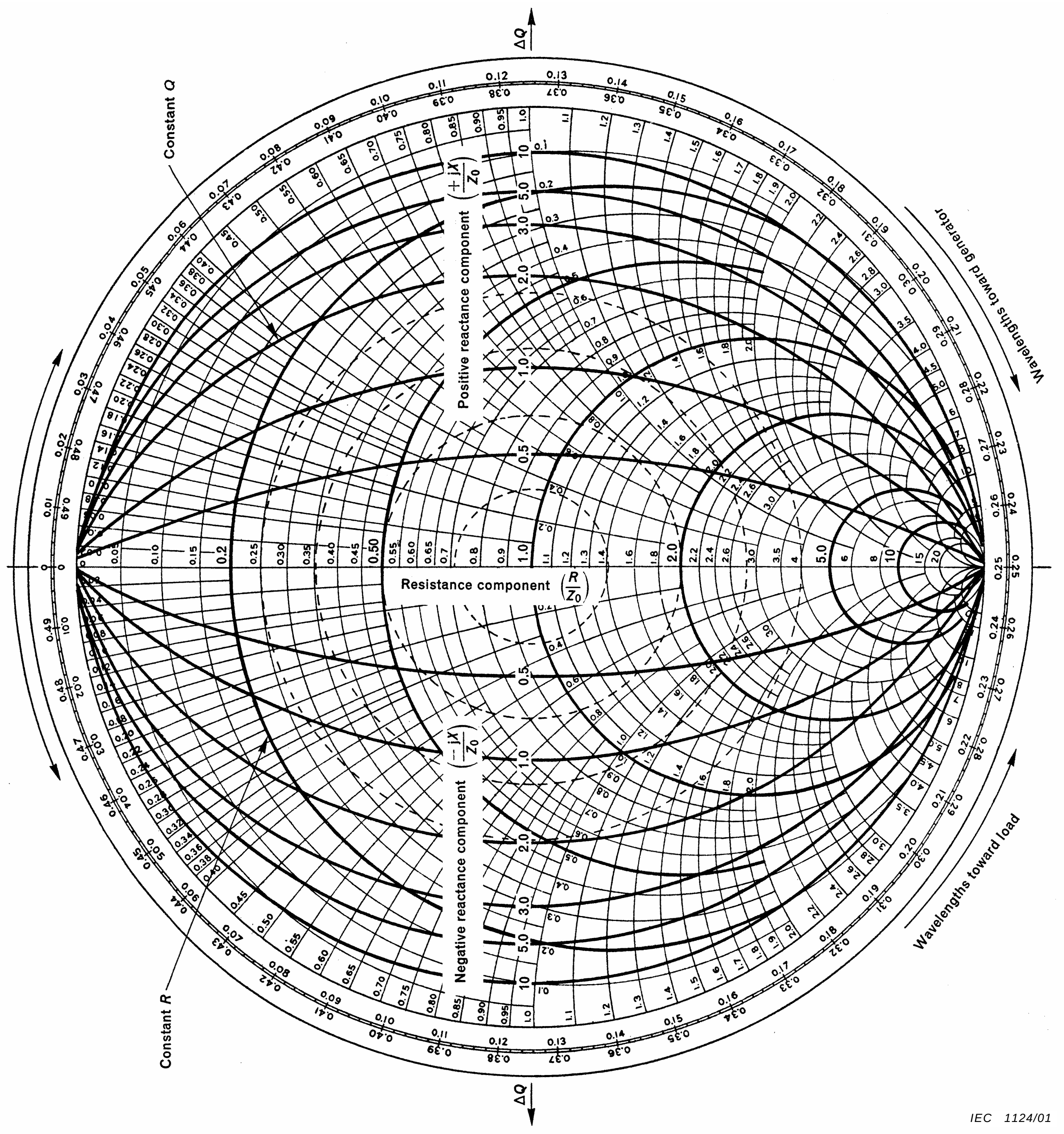


Figure 17 – Modified Smith chart indicating constant *Q* and constant *R* circles

3.2 Snap-off diodes, Schottky diodes

3.2.1 General

Snap-off (step recovery) diodes may be specified either as ambient rated or case rated or, where appropriate, as both.

3.2.2 Terminology and letter symbols

Transition time (of a snap-off diode) t_t

The time taken for the voltage across a snap-off diode to change from a specified low fraction of the total voltage step ($V_F + |V_{RM}|$) to a specified high fraction of the voltage step, when the diode is switched from forward current to reverse voltage (see Figure 18).

NOTE Values of 20 % and 80 % are preferred.

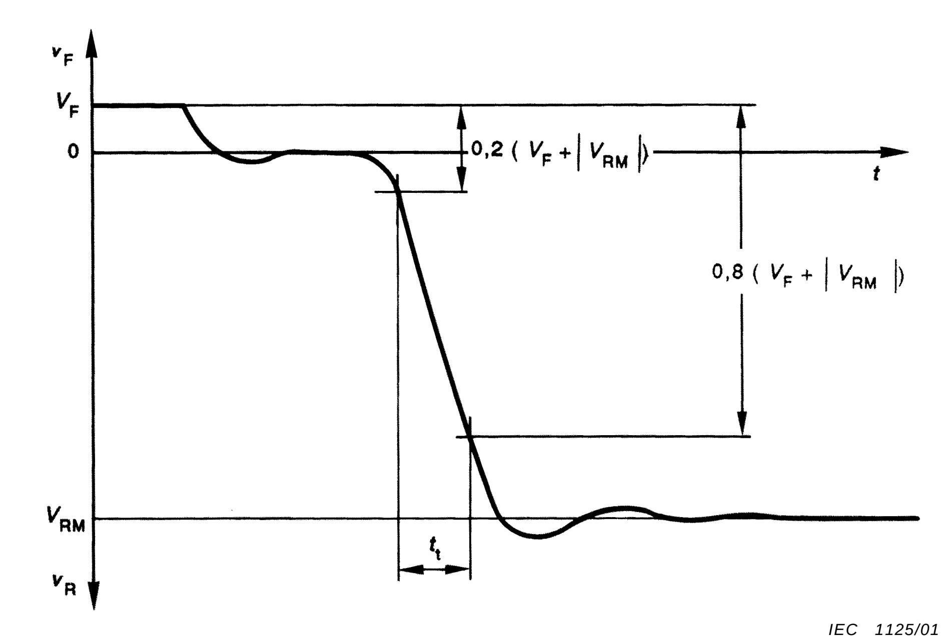


Figure 18 – Transition time t_t

For other parameters: see 3.2.3.

3.2.3 Essential ratings and characteristics

3.2.3.1 General

See 3.1.3.1.

3.2.3.2 Ratings (limiting values)

The following ratings should be stated:

3.2.3.2.1 Temperatures

3.2.3.2.1.1 Range of operating temperatures

3.2.3.2.1.2 Range of storage temperatures

3.2.3.2.2 Voltage and currents

The following ratings must be valid for the whole range of operating conditions as stated for the particular device.

3.2.3.2.2.1 Maximum peak reverse voltage.

3.2.3.2.2.2 Maximum mean forward current, where appropriate

3.2.3.2.2.3 Maximum peak forward current, where appropriate

3.2.3.2.3 Power dissipation

Maximum dissipation, under stated conditions, over the operating temperature range.

3.2.3.3 Electrical characteristics

Unless otherwise specified, the following characteristics should be stated at 25 °C:

3.2.3.3.1 Series inductance (L_s)

Typical value under specified conditions.

3.2.3.3.2 Terminal capacitance ($C_{tot} = C_j + C_p$)

Minimum and maximum values at specified bias voltage and specified frequency.

3.2.3.3.3 Junction capacitance (C_j)

Minimum and maximum values at a specified bias voltage and specified frequency. If the order of magnitude of C_p is the same as that of the terminal capacitance C_{tot}, a typical value for C_j should be given instead.

3.2.3.3.4 Series resistance

Typical or maximum value, as appropriate, under specified conditions.

3.2.3.3.5 Reverse current

Maximum value at a specified reverse voltage.

3.2.3.3.6 Stored charge or minority carrier storage time

Maximum value, under specified conditions, which may be stated either as a stored charge per unit of current (e.g. in $\frac{\text{picocoulombs}}{\text{milliamperes}}$), or as minority carrier storage time (e.g. in nanoseconds), the test circuit being also specified.

3.2.3.3.7 Transition time

Maximum value under specified conditions, the test circuit also being specified.

3.2.3.3.8 Forward voltage (where appropriate)

Maximum value for specified forward current.

3.2.3.3.9 Efficiency (where appropriate)

Minimum value, under specified conditions, of input power, input frequency, output frequency and test circuit.

3.2.3.4 Application data

– Relationship between junction capacitance and bias voltage. Typical value, in either a graphical or mathematical form.
– Cut-off frequency.

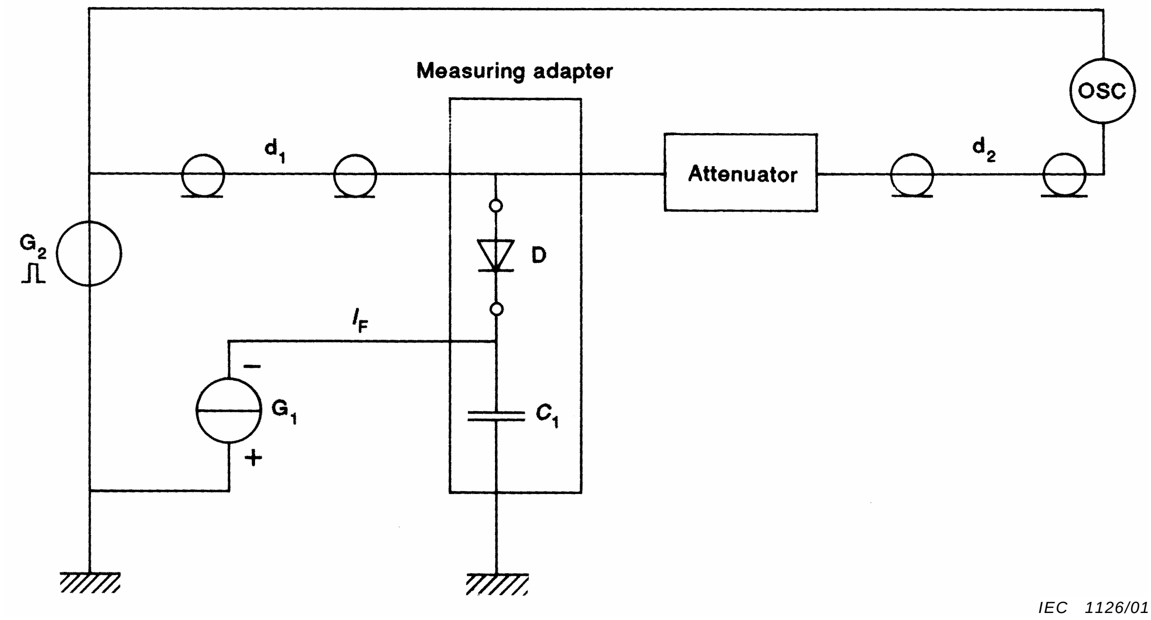
3.2.4 Measuring methods

3.2.4.1 Transition time (t_t)

a) *Purpose*

To measure the transition time t_t, for snap-off diodes.

b) *Circuit diagram*



Key

G_1	is the forward current generator (I_F)
G_2	is the pulse generator
d_1 and d_2	are the matched transmission lines
OSC	is the sampling oscilloscope

Figure 19 – Circuit for the measurement of transition time (t_t)

c) *Circuit description and requirements*

The rise time of the pulse from the generator G_2 should be equal or lower than:

$$0,5\,\frac{Q_s\,Z_0}{V_{RM}}$$

where

Q_s is the stored charge;

Z_0 is the characteristic impedance of the line (50 Ω) at which the measurement is carried out.

The pulse duration of the reverse voltage should meet the following requirement:

$$t_{VRM} \geq 1{,}5\,\frac{Q_s\,Z_0}{V_{RM}}$$

The measuring adapter should be in the form of a line with the characteristic impedance, Z_0 equal to 50 Ω; and should have a good matching at the input and the output in the frequency range from zero to $f > 0{,}5/t_t$.

Inductance of the measuring adapter L_s which includes inductances of the elements connecting of the diode D and the capacitor C_1 should meet the following requirement:

$$L_s \leq \frac{t_t\,Z_0}{5{,}6} - L_{sc}$$

where

L_{sc} is the case inductance of the diode being measured.

The capacitance of the capacitor C_1 should meet the following requirement:

$$C_1 \geq \frac{75\,Q_s}{V_{RM}}$$

A segment length of line d_1 should meet the requirement:

$$d_1 > c\,\frac{Q_s\,Z_0}{\sqrt{\varepsilon_r}\,V_{RM}}$$

where

c is the rate of propagation of electromagnetic oscillations in the vacuum;

ε_r is the relative dielectric permeability of the line segment.

The transition time of the diode being measured is calculated using the expression:

$$t_t = \sqrt{(t_{t1})^2 - (0{,}64\,t_r)^2} \qquad (21)$$

where

t_{t1} is the transient rise time measured on the oscilloscope between levels 0,2 and 0,8 of the voltage step;

t_r is the rise time of the oscilloscope between 10 % and 90 %.

d) *Measurement procedure*

The forward d.c. current I_F is applied to the diode being measured from the forward current source (G_1) and a voltage pulse with the amplitude $V_{RM} + V_F$ is set in the reverse direction.

The time interval (t_{t1}) between levels 0,2 and 0,8 of the total voltage step ($V_F + |V_{RM}|$) is read on the screen of the oscilloscope (see Figure 20).

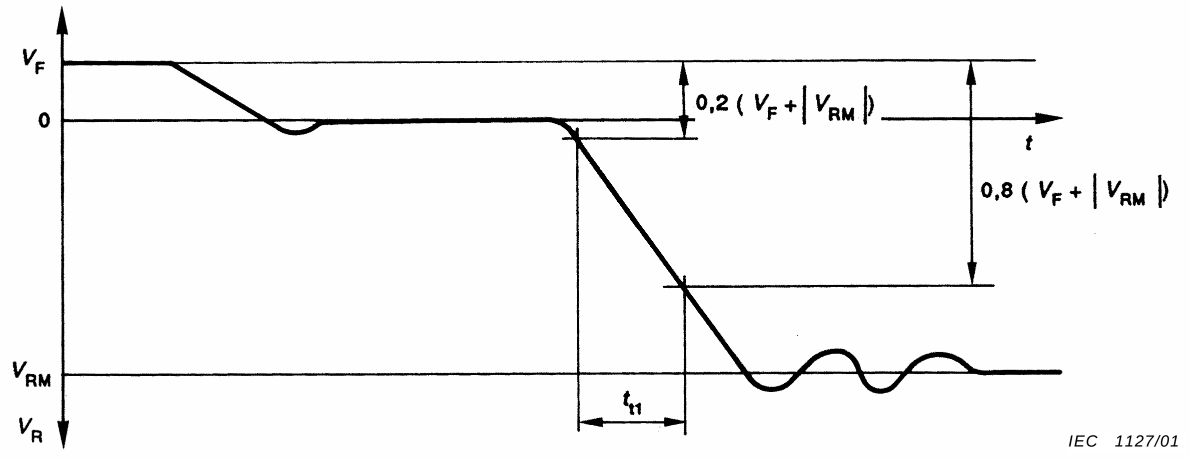


Figure 20 – Time interval (t_{t1})

Using the measured time interval, the transition time is calculated from equation (21).

If 0,64 $t_r \leq 0,3\ t_{t1}$, to, the time interval read on the oscilloscope is sufficiently close to the transition time of the diode.

e) *Specified conditions*

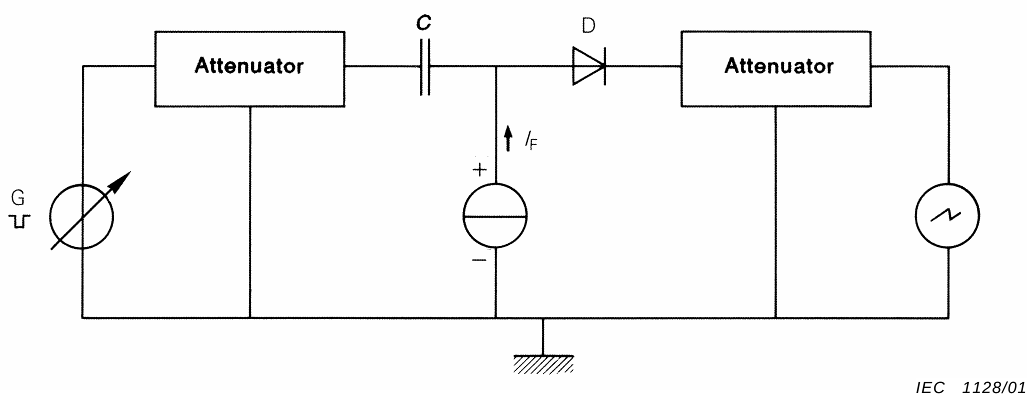
- Forward current (I_F)
- Reverse voltage (V_{RM})
- Characteristic impedance of the line (Z_0), if different from 50 Ω.
- Case inductance of the diode (L_{sc}).
- Stored charge (Q_s).

3.2.4.2 Reverse recovery time (with I_{RM} specified)

a) *Purpose*

To measure the reverse recovery time of a fast diode, e.g. with reverse recovery time less than 100 ns.

b) *Circuit diagram*



Key
D diode being measured

Figure 21 – Circuit for the measurement of reverse recovery time

c) *Circuit description and requirements*

The output impedance of the generator G and the input impedance of the oscilloscope are equal to 50 Ω unless otherwise specified. The rise times of the generator and of the oscilloscope should be small compared with t_{rr}.

The pulse width should he larger than 3 t_{rr} max.

Attenuators should have a characteristic impedance of 50 Ω unless otherwise specified, and an attenuation higher than or equal to 6 dB and should be able to carry d.c. current.

The time constant R_L C_L should be lower than 1/10 t_{rr} max., unless otherwise specified, with:

R_L = real part of the total impedance, as seen by the diode;

C_L = total capacitance of the circuit including the diode.

C should be high compared with t_{rr} max./R_L.

The impedance Z_i of the current generator should be greater than R_L.

d) *Precautions to be observed*

No special precaution.

e) *Measurement procedure*

The temperature is set to the specified value.

The current generator delivers the specified forward current I_F to the diode.

Pulses, delivered by generator G, are applied to me diode; the magnitude of the pulses is increased until the specified peak reverse current I_{RM} is reached.

The reverse recovery time t_{rr} to is the time interval between the instant at which the current passes through zero and the instant when the current is reduced from I_{RM} to the specified recovery current i_{rr} (see Figure 22).

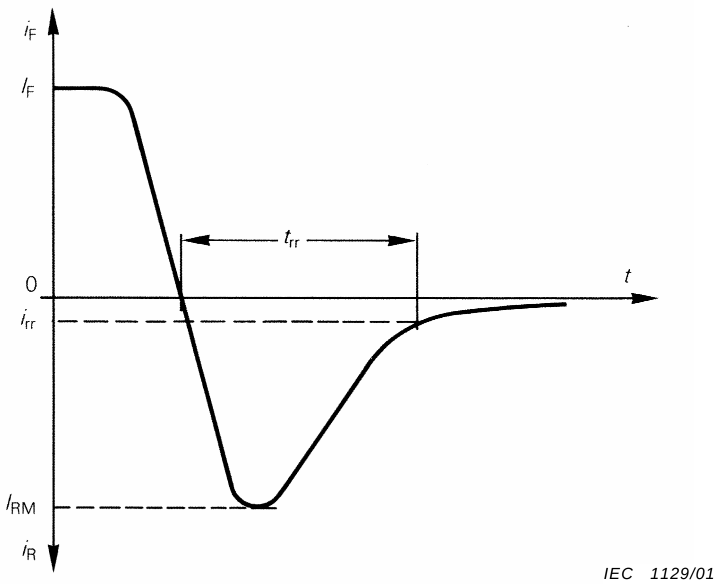


Figure 22 – Reverse recovery time t_{rr}

f) *Specified conditions*

- Ambient or reference-point temperature (t_{amb}, t_{ref})
- Forward current (I_F)
- Peak reverse current (I_{RM})
- Reverse recovery current (i_{rr})
  - Example of specified conditions: I_F = 10 mA
    I_{RM} = 10 mA
    i_{rr} = 1 mA

3.2.4.3 Measuring method of the excess carrier effective lifetime of diodes for fast-switching applications (snap-off diodes and Schottky diodes)

3.2.4.3.1 Purpose

To measure the excess carrier lifetime of diodes (following the Krakauer method, for example).

NOTE The conventional method has been modified so as to separate clearly the carrier lifetime owing to the carriers in excess, from the charges and discharges of the capacitance of the diode under test and of the parasitic elements (diode, case, mounting).

The carrier lifetime of a fast diode (Schottky, for example) has a very low value (theoretically zero for a Schottky diode). In practice, this measurement is generally made to determine the value of the forward current for which the parasitic elements of the diode under test contribute markedly to the carrier lifetime (guard-ring injection, etc.).

3.2.4.3.2 Principle of measurement

The diode to be measured is connected in series with a resistor, the set "diode + resistor" is supplied by a sinusoidal waveform generator (frequency ω).

The peak value of the forward current is compared with the reverse current, taking into account:

a) the charge and discharge current of the parasitic capacitance in parallel with the diode under test;

b) the electrostatic junction voltage ψ of the diode.

The values of the forward and reverse currents are calculated from the voltage measured across the resistor.

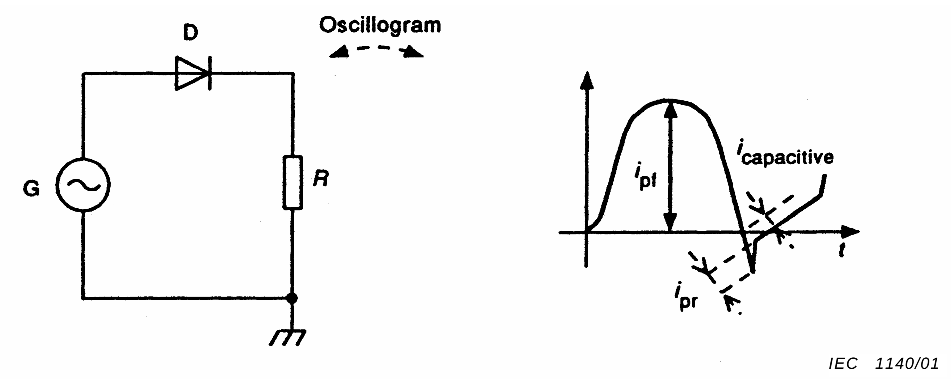


Figure 23 –Principle of the measurement of the excess carrier effective lifetime

It can be shown that:

$$\omega\tau = \frac{i_{pr}}{i_{pf}}(1-\frac{\Psi}{E_p}) \quad \text{and so:} \quad \tau = \frac{1}{2\pi f}\frac{i_{pr}}{i_{pf}}(1-\frac{\Psi}{E_p})$$

under the condition that both: $\omega\tau$ <0,3 and $i_{pr} > i_{capacitive}$

where

τ is the carrier lifetime of the diode under test;

f is the frequency of the sinusoidal waveform generator, G;

i_{pr} is the peak value of the reverse current without capacitive effects;

i_{pf} is the peak value of the forward current;

ψ is the forward voltage;

E_p is the peak value of open-circuit voltage of the sinusoidal waveform generator.

3.2.4.3.3 Circuit diagram

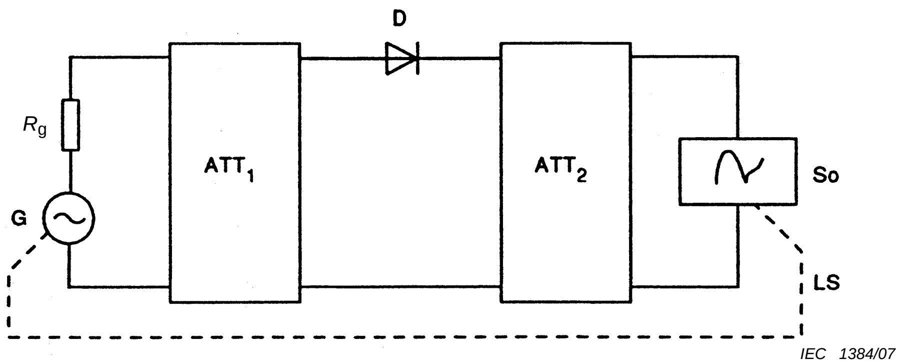


Figure 24 – Circuit for the measurement of the excess carrier effective lifetime

where

G is the sinusoidal waveform voltage generator with frequency, f;

R_g is the internal resistance of the sinusoidal wave generator;

ATT_1 is the attenuator, $Z_0 = R_g$;

D is the diode under test;

ATT_2 is the attenuator, $Z_0 = R_g$;

So is the oscilloscope, $Z_{input} = R_g$;

LS is the synchronization for the oscilloscope.

3.2.4.3.4 Measurement procedure

The diode to be measured is put in the test fixture similar to that shown in Figure 24 (circuit diagram).

The frequency of the sinusoidal waveform generator is set and the output level adjusted to the specified value of i_{pf}.

NOTE 1 When the oscilloscope is calibrated in voltage, the value of i_{pf} can be calculated from the voltage at the input of the oscilloscope and the input impedance of the oscilloscope ($R_e = R_g$), taking into account the attenuation of ATT_2.

The value of i_{pr} corresponds to the difference between the peak value of the reverse current and the extrapolated value of the capacitive current. The value of τ is calculated with the formula in 3.2.4.3.2, after having determined the value of E_p and the value of ψ (for example, with a curve tracer).

NOTE 2 For the circuit diagram of Figure 24, the value of e_{rms} can be measured directly on the calibrated oscilloscope when the attenuators ATT_1 and ATT_2 are directly connected (for example, a short circuit across the diode).

$$E_p = 2\sqrt{2} \times e_{rms}\left(e_{rms} = \frac{V_{rms}}{10^{(ATT1/20)}}\right)$$

Caution: It should be ascertained that:

$\omega\tau < 0,3$

$i_{pr} > i_{capacitive}$

Example: Given

R_g = 50 Ω

V_{rms} = 10 V

f = 54 MHz

ATT_1 = 10 dB, ATT_2 = 20 dB (these values are generally sufficient to reduce the effect of the mismatch appearing during the cycle after the carrier lifetime measurement).

The value of i_{pf} is read directly from the oscilloscope, and taking into account the attenuation of ATT_2, the following values are obtained:

i_{pf} = 75 mA, ψ = 1,35 V and E_p = 8,9 V.

From the oscilloscope the ratio of i_{pr} to i_{pf} is derived:

$$\frac{i_{pr}}{i_{pf}} = \frac{1}{5}$$

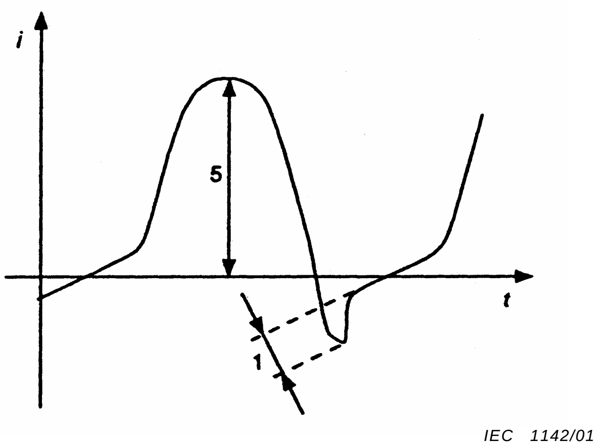


Figure 25 – Ratio of i_{pr} to i_{pf}

From the formula in 3.2.4.3.2, it follows: τ = 500 ps.

3.2.4.3.5 Requirements

The value of τ shall be within the limits specified in the relevant specification.

3.2.4.3.6 Specified conditions

– Ambient or case temperature.

– Frequency of sine wave generator, *f.*

– Peak value of forward current, $i_{pf.}$

– Values of the elements of the circuit, and the circuit diagram with values of the elements, if different from Figure 24.

4 Mixer diodes and detector diodes

4.1 Mixer diodes used in radar applications

4.1.1 General

Although a number of measurements may be carried out on a diode by itself, it is necessary for an assessment of diode performance to provide a standard mounting arrangement to enable satisfactory measurements to be made. Usually, the mounting takes the form of a standard holder designed for the purpose. If the holder is adjustable, all adjustments are made to conform with given measurement requirements.

In the case of reversible diodes, it may be necessary for measurements to be made with the diode connected to provide an assessment of performance in each polarity.

All microwave components used in the measuring equipment shall be checked to ensure satisfactory operation over the required frequency band. It is recommended that this be performed using swept frequency techniques.

The time constants, or pulse response characteristics of any indicating instrument, shall not affect the measurement results when modulation is used.

It is essential to ensure that stray electromagnetic fields do not significantly affect the accuracy of measurement.

In addition, any signal source used as part of the measuring circuitry shall be capable of stable operation at a signal level equivalent to the small-signal conditions of the diode being measured.

Unless otherwise stated, it is recommended that the maximum tolerances permitted for microwave signal levels should be:

a) ±0,1 % for low-power measurements required under small-signal conditions, and

b) ±1 % for high-power measurements.

Where a specified temperature is required, the temperature of the body of the measuring mount shall be measured when equilibrium conditions have been reached.

4.1.2 Terminology and letter symbols

See 4.1.3.3.

4.1.3 Essential ratings and characteristics

4.1.3.1 General

The essential ratings and characteristics for each category of diode are marked with a + sign in the following table:

– Category 1: pulse applications in systems where very short duration pulses are incident upon the diode.
– Category 2: c.w. applications or longer pulse systems.

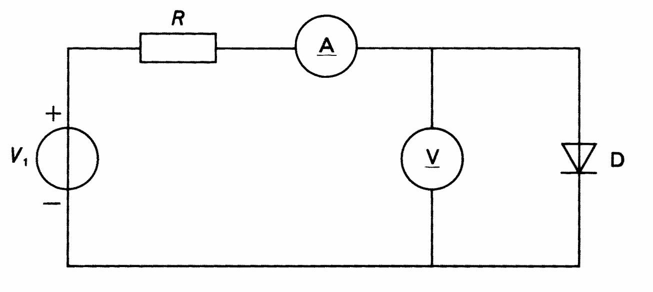
	Categories	
4.1.3.2 Ratings (limiting values)		
The following ratings should be stated:	1	2
4.1.3.2.1 Temperatures		
4.1.3.2.1.1 Range of operating temperatures	+	+
4.1.3.2.1.2 Range of storage temperatures	+	+
4.1.3.2.2 Power dissipation (including burn-out energy)		
4.1.3.2.2.1 Maximum c.w. power under specified conditions at 25 °C	+	+
4.1.3.2.2.2 Maximum peak value of pulsed RF power under specified conditions at 25 °C	+	
4.1.3.2.2.3 Burn-out energy by single pulse (or multiple pulses) under specified conditions at 25 °C	+	
4.1.3.3 Electrical characteristics		
Unless otherwise specified, the following characteristics should be stated at 25 °C:		
4.1.3.3.1 Voltage standing wave ratio		
Maximum value, when operating in a specified microwave circuit, under specified conditions	+	+
4.1.3.3.2 IF impedance		
Minimum and maximum values under specified conditions	+	+
4.1.3.3.3 Conversion loss		
Maximum value under specified conditions	+	+
4.1.3.3.4 Overall noise factor		
Maximum value, under specified operating conditions, using a specified microwave circuit, followed by a specified i.f. amplifier (under image frequency matched conditions)	+	+
4.1.3.3.5 1/*f* noise		
Maximum value, under specified operating conditions, using a specified microwave circuit followed by a specified i.f. amplifier (Doppler applications only)		+

		Categories 1	Categories 2
4.1.3.4	**Application data**		
4.1.3.4.1	**Maximum external circuit d.c. resistance**	+	+
4.1.3.4.2	**Maximum and recommended values of mean forward current**	+	+
4.1.3.4.3	**Minimum current at specified continuous (direct) forward voltage**	+	+
4.1.3.4.4	**Maximum current at specified continuous (direct) reverse voltage**	+	+
4.1.3.4.5	**Maximum value of noise/temperature ratio under the conditions given in 3.3.4**	+	+
4.1.3.4.6	**Typical curve of overall noise factor versus RF input power (expressed as rectified current) under specified operating conditions**	+	+
4.1.3.4.7	**Typical curve of diode admittance versus frequency, the admittance being given as a normalized value in terms of a specified transmission line impedance**	+	+
4.1.3.4.8	**Typical curve of overall noise factor versus temperature over a specified temperature range**	+	+

4.1.4 Measuring methods

4.1.4.1 Forward current (I_F)

4.1.4.1.1 Circuit diagram



IEC 1130/01

Key
R protective resistor
D diode being measured

Figure 26 – Circuit for the measurement of forward current (I_F)

It is essential to use a high-impedance voltmeter.

4.1.4.1.2 Measurement procedure

The specified conditions are applied and the current through the diode is measured by means of an ammeter.

4.1.4.2 Reverse current (I_R)

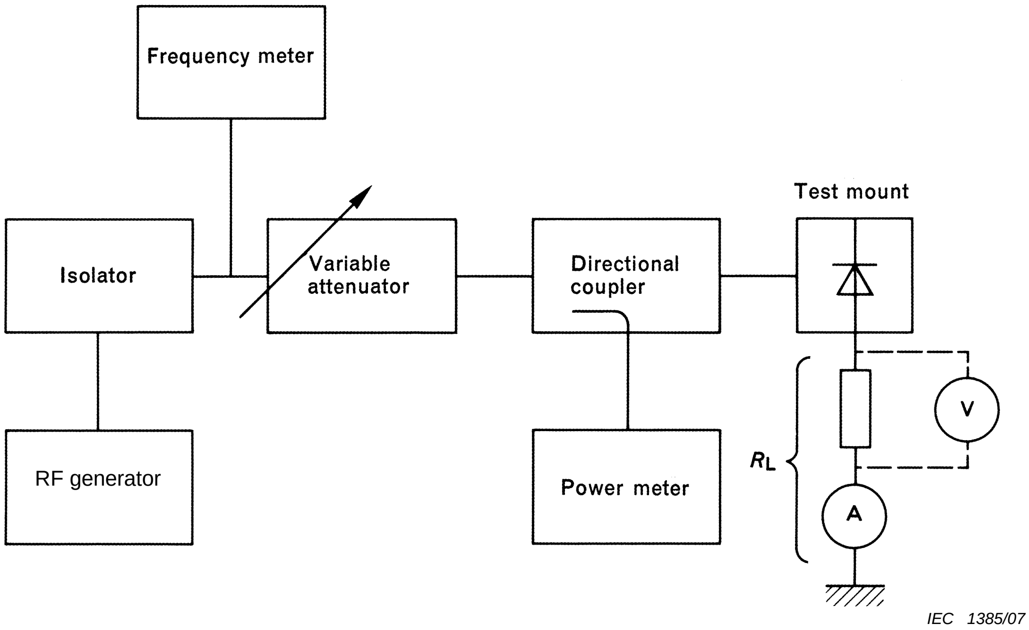
The reverse current is measured with the diode operating under given conditions by the method stated in 3.1.4.1.

4.1.4.3 Rectified current (I_0)

4.1.4.3.1 Purpose

To measure the rectified current of a microwave diode under specified conditions.

4.1.4.3.2 Circuit diagram



Key
R_L diode load resistance of specified value

Figure 27 – Circuit for the measurement of rectified current (I_0)

4.1.4.3.3 Circuit description and requirements

The frequency meter is loosely coupled to the line, the power meter and its associated coupler are selected to measure the specified power level incident upon the diode.

To reduce self-biasing effects, the value of the load resistance R_L, which includes the resistance of the ammeter shall be as low as possible and normally less than 100 Ω.

The value of the rectified current I_0 can be measured on the meter A or by use of a high-impedance voltmeter across the load resistance as shown by the dotted lines.

4.1.4.3.4 Measurement procedure

The diode is inserted into the measuring mount.

The RF power incident upon the diode is increased to the specified value and the rectified current I_0 is measured.

4.1.4.4 Intermediate frequency impedance (Z_{if})

Purpose

To measure the intermediate frequency impedance of a microwave diode under specified conditions.

4.1.4.4.1 Method 1: Impedance bridge method

4.1.4.4.1.1 Circuit diagram

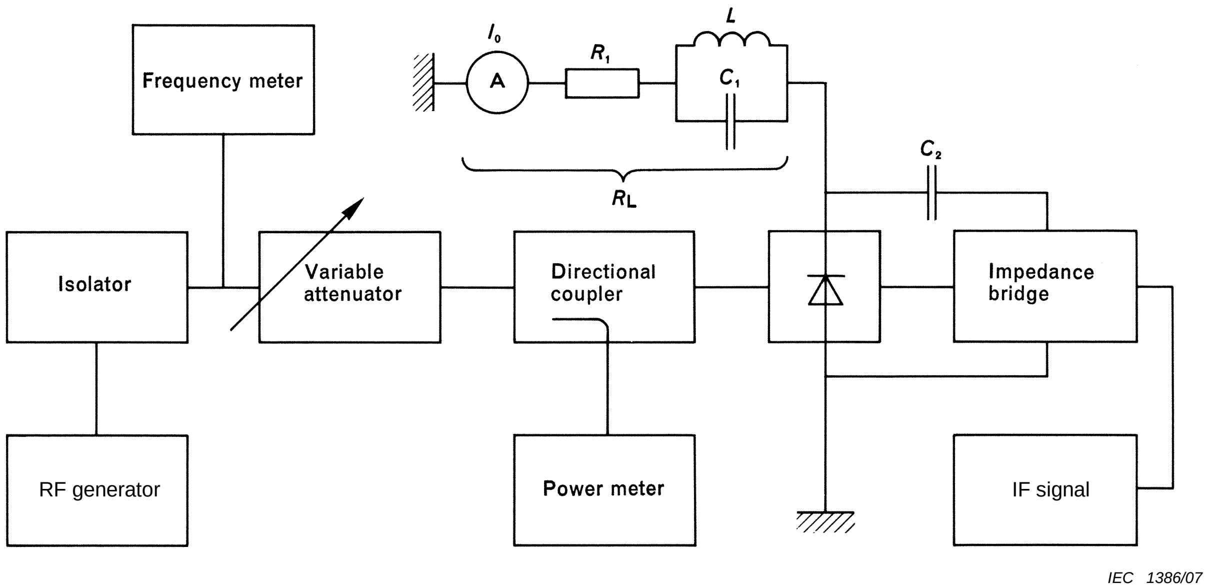


Figure 28 – Circuit for the measurement of intermediate frequency impedance (Z_{if}) in method 1

4.1.4.4.1.2 Circuit description and requirements

The RF generator shall be capable of operating at the signal frequency and the impedance bridge shall be capable of operating at the required intermediate frequency.

The frequency meter is loosely coupled to the line, the power meter and its associated coupler are selected to measure the specified power level incident upon the diode. Ammeter A measures the rectifier current I_0.

The values of L and C_1 are chosen so that the L C_1 circuit has a high impedance at the specified intermediate frequency.

The circuit comprising L, C_1, R_1 and ammeter A shall have a d.c. load value equal to the specified load R_L. Capacitor C_2 shall present a short circuit at the intermediate frequency.

The IF signal level into the bridge shall not cause more than a 1 % increase in the rectified current.

4.1.4.4.1.3 Precautions to be observed

The measurement frequency shall be sufficiently low so that the diode IF impedance can be assumed to be wholly resistive.

4.1.4.4.1.4 Measurement procedure

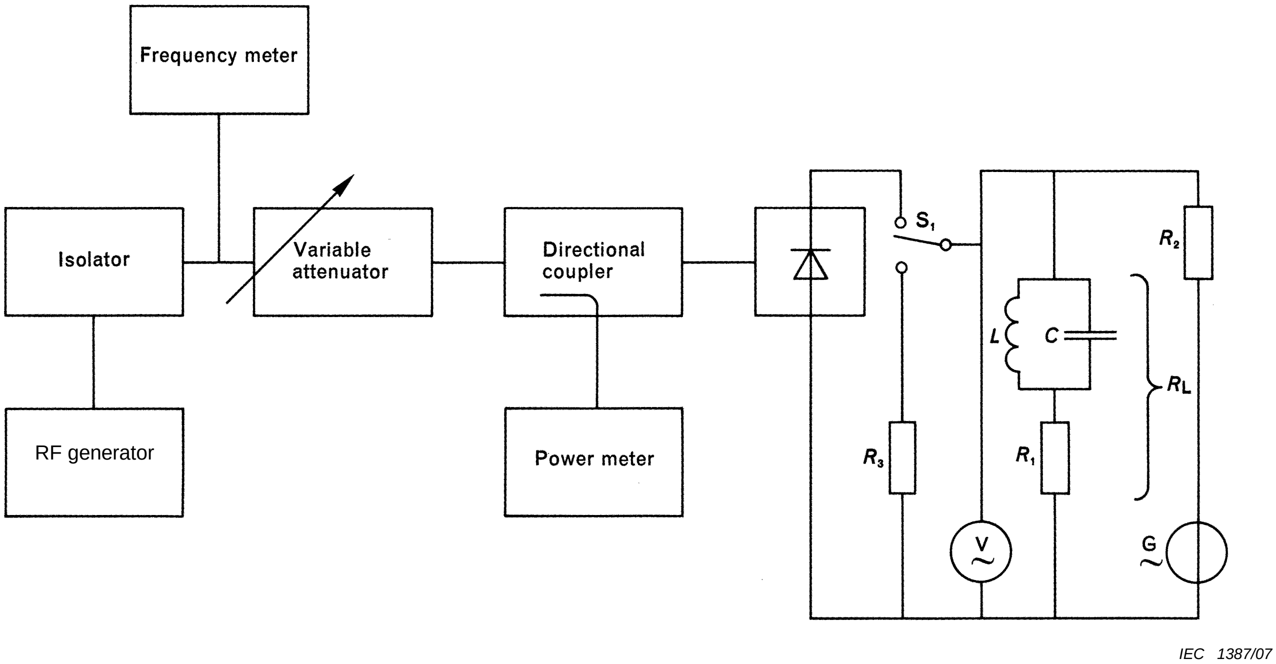
The diode is inserted into the test mount.

The bias current, where specified, is adjusted to the required value.

The signal generator is set to the required frequency and the RF power output is increased to the required power level. The impedance bridge is adjusted to the specified intermediate frequency and the diode impedance is measured.

4.1.4.4.2 Method 2: Substitution or comparison method

4.1.4.4.2.1 Circuit diagram



Key
G low-frequency generator

Figure 29 – Circuit for the measurement of intermediate frequency impedance (Z_{if}) in method 2

4.1.4.4.2.2 Circuit description and requirements

The RF generator shall be capable of operating at the intermediate frequency.

The frequency meter is loosely coupled to the line, the power meter and its associated coupler are selected to measure the specified power level incident upon the diode. The values of L and C are chosen to be in resonance at the low frequency of measurement and, together with R_1, provide an equivalent IF impedance, the same as the specified load R_L at d.c.

The low-frequency generator is coupled to the load R_L with a high resistance R_2 whose value is much greater than the diode IF impedance to provide a constant current a.c. source.

The resistor R_3 shall have a resistance of the order of the diode IF impedance. Resistors shall be non-inductive at the intermediate frequency. The a.c. voltmeter V shall have a high-input impedance.

4.1.4.4.2.3 Precautions to be observed

The power output of the low-frequency generator shall not exceed the small-signal capabilities of the diode being measured.

4.1.4.4.2.4 Measurement procedure

The diode is inserted into the test mount.

The bias current, if specified, is adjusted to the required value.

The signal generator is set to the requested intermediate frequency and the RF output is increased to the required power level.

The constant current low-frequency signal from the generator is adjusted to the required value.

The voltage across the diode is recorded. The low-frequency voltage is then switched from the diode using S_1 to a reference resistor (R_3) whose value is within the IF impedance values given. As the voltage across the diode is proportional to its output resistance, the measured voltage reading meter may be calibrated in terms of Z_{if}.

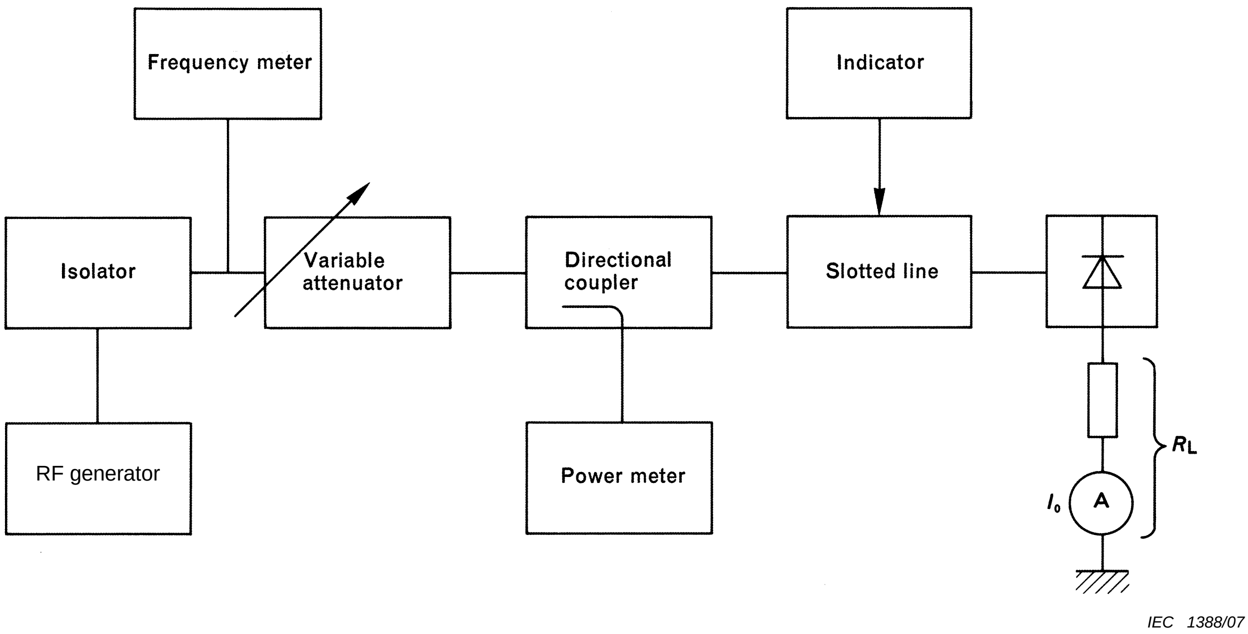
Alternatively, in place of the switch S_1 and reference resistor R_3, a number of calibrating resistors having appropriate values in the required IF impedance range may be introduced into the diode envelope and the output voltmeter calibrated accordingly.

4.1.4.5 Voltage standing wave ratio

4.1.4.5.1 Purpose

To measure the voltage standing wave ratio of a microwave diode under specified conditions.

4.1.4.5.2 Circuit diagram



Key
R_L stated load resistance

NOTE There must be sufficient filtering to prevent the indicator from responding to harmonics generated by the diode under test.

Figure 30 – Circuit for the measurement of voltage standing wave ratio

4.1.4.5.3 Circuit description and requirements

The frequency meter is loosely coupled to the line, the power meter and its associated coupler are selected to give a convenient power reading.

The coupling between the indicator probe and the slotted line shall be as loose as possible so that the field within the line is not significantly affected.

The values of the VSWR measured are dependent upon the characteristic of the detector used in the indicator; its response to varying power levels shall be checked and calibrated.

Ammeter A measures the rectified current I_0. The load resistor R_L includes the meter resistance.

4.1.4.5.4 Measurement procedure

The diode is inserted into the measuring mount and rotated, if necessary, to optimize the performance. The RF power incident upon the diode is increased to the specified value. The values of V_{max} and V_{min} as measured on the indicator are obtained by adjustment of the slotted line.

Then:

$$\mathrm{VSWR} = \frac{V_{max}}{V_{min}}$$

Alternatively, an assessment of the VSWR may be made by an inspection of the energy which is incident upon and reflected from the diode (reflectometer). If P_i is the incident microwave power applied to the diode and P_r the reflected power, then:

$$\text{VSWR} = \frac{P_i^{1/2} + P_r^{1/2}}{P_i^{1/2} - P_r^{1/2}} \tag{22}$$

Return loss is the ratio P_i/P_r, which may be obtained directly from an attenuator located immediately in front of the detector on the reflected-power side arm of the directional coupler used in the reflectometer. Using a waveguide switch or gate, the mixer is replaced by a short circuit in order to reflect all of the incident power. A properly tuned lossless system is insensitive to the phase of the reflection, so that the indicator reading is essentially unchanged by the motion of a moving short circuit replacing the mixer.

In this method, it is necessary to ensure that the couplers used provide a high directivity, preferably not less than 25 dB. In addition, the coefficient of coupling of the couplers shall be taken into account when using equation (22). The effects of the coefficients of coupling and the directivity of the couplers may be checked by the replacement of the diode and mount by a good quality matched termination.

If it is required to express the VSWR as a magnitude of impedance or admittance, it is essential to give a reference plane within the transmission system. A reference short-circuit may be readily achieved by the use of a metallic dummy diode having the same dimensions as the diode being measured.

4.1.4.6 Overall noise factor

Theory

The noise factor (F) of any network is given by the expression:

$$F = \frac{N_1}{k\,T_0\,B\,G}$$

where

N_1 is the output noise power;

k is the Boltzmann's constant = 1,38 × 10^{-23} J K^{-1};

T_0 is the absolute temperature. in kelvins (taken for convenience as 293 ± 5 K);

B is the effective bandwidth of the network;

G is the power gain of network.

When a signal of available input power N' is applied to the input of the network, the output noise N_2 becomes:

$$N_2 = F(k\,T_0\,B\,G) + N'G$$

and hence:

$$\frac{N_2}{N_1} = 1 + \frac{N'}{FkT_0B}$$

and:

$$F = \frac{N'}{kT_0B}\left[\frac{1}{\frac{N_2}{N_1}-1}\right] \quad (23)$$

If the available input power N' applied to the network to change the value of the output noise from the network, maintaining a constant gain and bandwidth of the network, the values of gain and bandwidth are removed from the expression and any measurement becomes independent of the amplifier characteristics providing they remain stable.

Circuit diagram

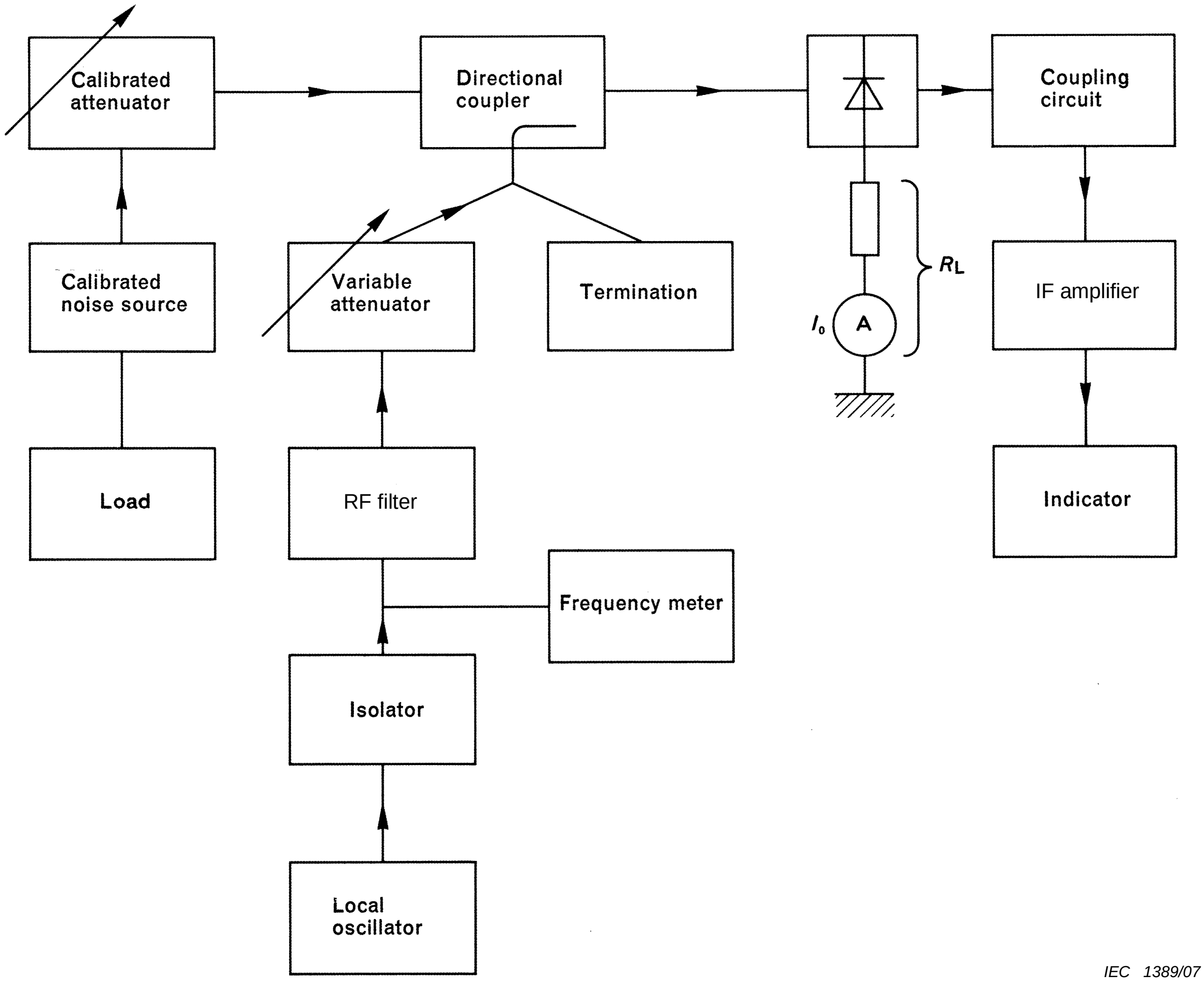


Figure 31 – Circuit for the measurement of overall noise factor

If the effective noise temperature of a noise power source is T in degrees kelvin, the available input power N' is by:

$$N' = k\,T_0\left[\frac{T}{T_0}-1\right]B$$

hence
$$F = \left| \frac{T}{T_0} - 1 \right| \left[\frac{1}{\frac{N_2}{N_1} - 1} \right] \tag{24}$$

Circuit description and requirements

The RF filter shall have a high factor Q at the local oscillator frequency, in order to provide a minimum stated rejection of the noise sidebands generated in the local oscillator (see 4.1.4.10).

The diode holder shall be as specified and the coupling circuit designed to match the diode output impedance to the IF amplifier input over the IF amplifier bandpass.

The amplifier shall have stable characteristics of gain and bandwidth.

In the case of a gas discharge noise source, it is preferable that the termination connected to the calibrated noise source be matched rather than provide a short circuit, owing to the errors which may be introduced by the attenuation of the reflected noise power by the gas plasma. To remove errors which may be introduced by a change in the noise source match conditions between a "hot" and "cold" noise tube, it is preferable to "switch in" the noise source by means of a calibrated attenuator so that the noise tube may be continually in one state.

The bias supplies may need adjustment.

Measurement procedure

The measurement of overall noise factor may be made using one of the following methods:

4.1.4.6.1 Doubling the output method

The diode is fitted into a mount connected to the input of a specified amplifier.

The operating conditions are adjusted to the specified values. The calibrated attenuator is set to provide maximum attenuation so that negligible power from the noise source is received by the diode. The gain of the amplifier is adjusted to provide a convenient level of output power, as shown on the indicator. The calibrated attenuator is then adjusted so that sufficient power from the noise source is applied to the diode, to provide a reading on the indicator which is double the original value. The noise power from the noise source is then equal to the noise in the network, hence the value of network noise power can be determined directly.

The doubling of the output noise power may be conveniently checked by the use of a 3 dB attenuator in the IF amplifier. Care shall be taken to ensure that the 3 dB attenuator is matched into the circuit.

Thus: $N_2 = 2\,N_1$, and equation (23) becomes: $F_0 = \dfrac{N'}{k\,T_0\,B}\left[\dfrac{1}{a_{rf}}\right]$

where a_{rf} is the RF attenuation

Since:
$$N' = k\,T_0\,B\left[\frac{T}{T_0} - 1\right],$$

then: $$F_o = \frac{\frac{T}{T_o} - 1}{a_{rf}}$$

When the noise source power is available at both signal and image frequencies, the noise figure will be equal to twice the applied noise power N', i.e.:

$$F_o = \frac{[1 + r]\left[\frac{T}{T_o} - 1\right]}{a_{rf}}$$

where $r = \frac{\text{gain at image frequencies}}{\text{gain at signal frequencies}} = 1$

which, expressed in decibels, becomes:

$$F_o \text{ (dB)} = 10 \log_{10} (1 + r) + 10 \log_{10}\left[\frac{T}{T_o} - 1\right] - a_{rf}$$

where a_{rf} is expressed in decibels.

4.1.4.6.2 IF attenuation method

The diode is fitted as in 4.1.4.6.1.

In this method, an IF attenuator is included in the circuit (usually between the main and preamplifier sections which make up the amplifier as shown in Figure 31). Care shall be taken to ensure the correct match of the IF attenuator into the circuit.

The method is similar to that in 4.1.4.6.1 except that, when the power from the noise source is applied to the diode by adjustment of the calibrated RF attenuator so as to double the output power, the RF attenuator is adjusted to zero attenuation and the IF attenuator is adjusted to return the power output level shown on the indicator to its original value.

Hence: $$N_1 = \frac{N_2}{a_{if}}$$

where a_{if} is the IF attenuation

Then: $$F_o = \left[\frac{T}{T_o} - 1\right]\left[\frac{1}{a_{if} - 1}\right]$$

If the noise source power is available at both signal and image frequencies, then:

$$F_o = [1 + r]\left[\frac{T}{T_o} - 1\right]\left[\frac{1}{a_{if} - 1}\right]$$

which, expressed in decibels, gives:

$$F_o \text{ (dB)} = 10 \log_{10} (1 + r) + 10 \log_{10}\left[\frac{T}{T_o} - 1\right] - 10 \log_{10} [a_{if} - 1]$$

where a_{if} is expressed in decibels.

4.1.4.6.3 Output power method

The diode is fitted as in 4.1.4.6.1 and the procedure is similar, except that the doubling of the power output is not required. Any suitable power levels may be shown on the indicator provided that the indicated output power ratio N_2/N_1 is measurable within the square law part of the detector characteristic.

In this case, equation (24) may be used directly.

Thus:

$$F_o = \left[\frac{T}{T_o} - 1\right]\left[\frac{1}{\frac{N_2}{N_1} - 1}\right]$$

If the noise source power is available at both signal and image frequencies, then:

$$F_o = [1 + r]\left[\frac{T}{T_o} - 1\right]\left[\frac{1}{\frac{N_2}{N_1} - 1}\right]$$

which, expressed in decibels, gives:

$$F_o\ (\text{dB}) = 10 \log_{10} (1 + r) + 10 \log_{10}\left[\frac{T}{T_o} - 1\right] - 10\log_{10}\left[\frac{N_2}{N_1} - 1\right].$$

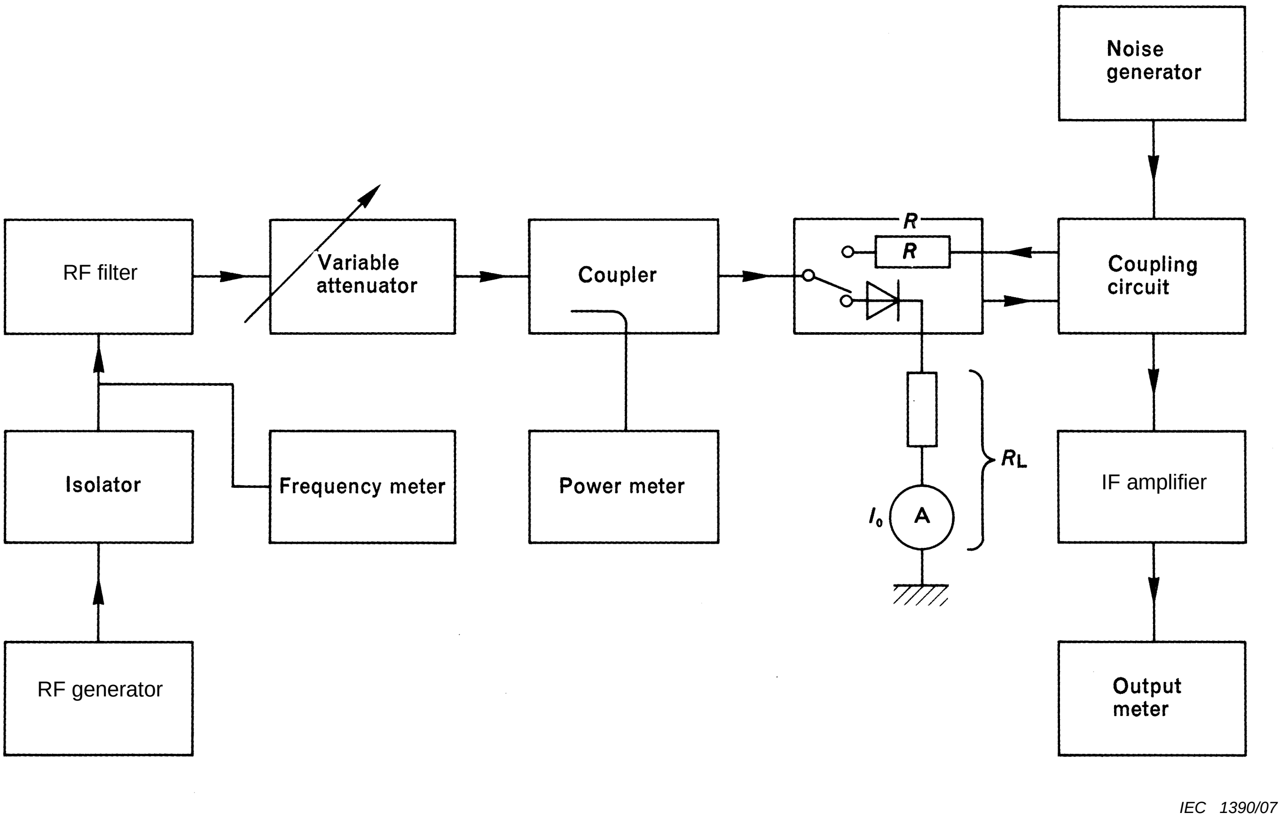
4.1.4.7 Output noise ratio

4.1.4.7.1 Direct measurement method

4.1.4.7.1.1 Purpose

To measure the output noise ratio of a microwave diode under specified conditions of frequency and bias.

4.1.4.7.1.2 Diagram



Key
R non-inductive reference resistor

Figure 32 – Circuit for the measurement of output noise ratio

4.1.4.7.1.3 Circuit description and requirements

The RF filter and diode measurement mount must conform to the requirements given in 4.1.4.6. If more accuracy is required, means should be provided to tune the coupling circuit for each individual diode. The coupling circuit shall be non-dissipative.

The gain and bandwidth of the IF amplifier shall be specified. The noise factor of the amplifier shall be lower than the expected noise output value of the diode being measured.

The output conductance of the mixer shall be equal to that of the diode being measured. For convenience, a number of reference resistors (covering the IF impedance range expected of the diodes under measurement) shall be available; they shall be mounted in a position which is physically and electrically equivalent to that of the diode.

4.1.4.7.1.4 Measurement procedure

The diode is mounted in a specified measurement mount, under specified operating conditions. The noise output from the diode is applied to the amplifier having known characteristics and measured in the output meter. This noise is compared with the injected noise developed across the reference standard resistor mounted in the measuring mount in place of the diode, all other conditions remaining the same.

The output meter may be calibrated in terms of output noise ratio by applying various values of noise current through the reference standard resistor. A temperature-limited saturated thermoionic diode may be used to generate the noise current.

The output noise ratio is given by:

$$N_r = \frac{e\,I}{2\,k\,g\,T_o}\left(\frac{N_2}{N_1} - 1\right) + 1$$

where

e is the elementary charge ($1{,}6 \times 10^{-19}$ C);

k is the Boltzmann's constant ($1{,}38 \times 10^{-23}$ J K^{-1});

I is the average (d.c.) anode current of noise diode (A);

T_o is the reference noise temperature (K);

g is the mixer IF output conductance (usually expressed in siemens or ohms of its reciprocal);

N_2 is the indicated output with diode input;

N_1 is the indicated output with reference resistor input.

If $N_2 = 2\,N_1$:

$$N_r = \frac{e\,I}{2\,k\,g\,T_o} + 1$$

4.1.4.7.2 Calculated value

As the measured value, as given in 4.1.4.7.1, is so dependent on stringent measurement circuit requirements, it is sometimes more accurate to derive the output noise ratio (N_r) from the measurements of overall noise factor (F_o) and conversion loss (L_c), provided the latter has been measured, in which case:

$$N_r = \frac{F_o}{L_c} - F_{if} + 1$$

where F_{if} is the noise factor of IF amplifier.

4.1.4.8 Conversion loss

Purpose

To measure the conversion loss of a microwave diode under specified conditions.

4.1.4.8.1 DC incremental method

4.1.4.8.1.1 Circuit diagram

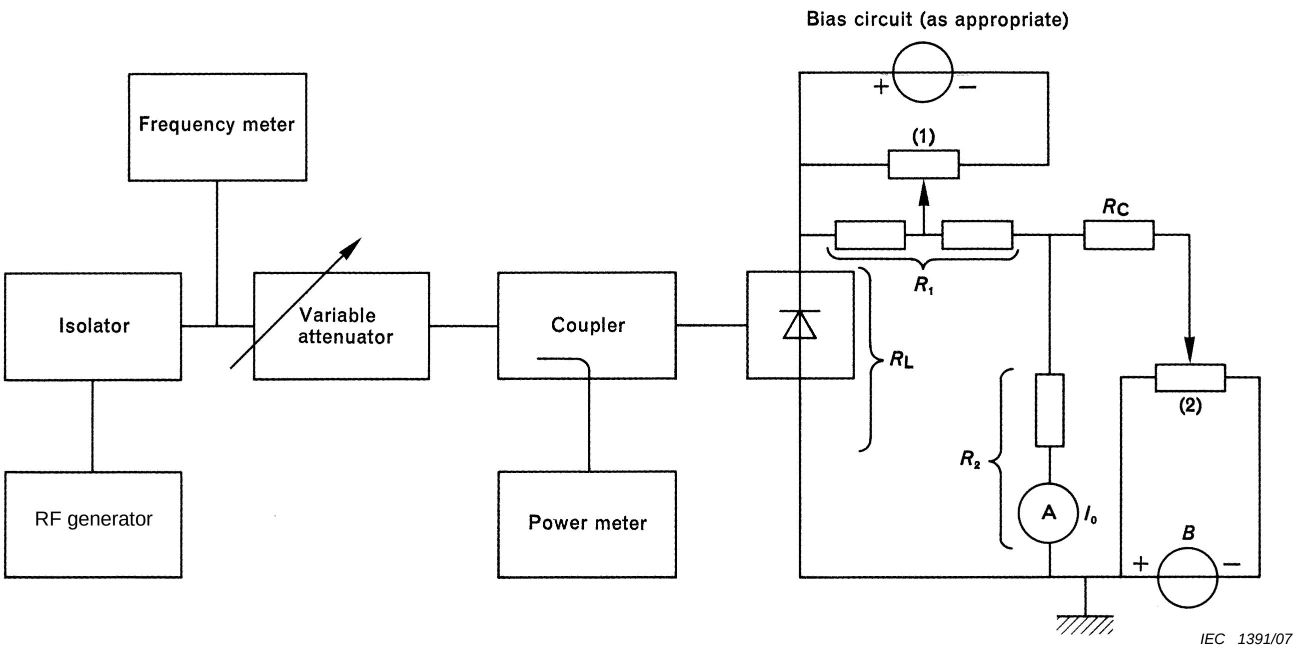


Figure 33 – Circuit for the measurement of conversion loss in d.c. incremental method

4.1.4.8.1.2 Circuit description and requirements

$R_L = R_1 + R_2$ = specified IF load resistance

R_1 = specified d.c. load resistance

Both the microwave power meter and the ammeter used to read the rectified current (I_o) shall be capable of accurate indication of small changes. Alternatively, a calibrated variable attenuator may be used to produce an accurately known change of microwave power.

R_c shall be larger than about 100 kΩ. Alternatively, en electronically regulated constant-current source may be substituted for this resistor, the battery, and potentiometer 2. The resistance between the mixer and the tap of potentiometer 1 is much lower than R_1.

4.1.4.8.1.3 Precautions to be observed

No special precautions.

4.1.4.8.1.4 Measurement procedure

The diode is inserted into the measurement circuit as shown in Figure 33 and operated under specified conditions. A known low level of microwave power (P) is applied to the diode. The potentiometer (2) is adjusted so that the current supplied from the d.c. source (B) reduces the indicated rectified current (I_o) to zero. The applied microwave power is then changed by a small amount (ΔP) and the resulting small change in rectified current (ΔI_o) is measured.

The conversion loss is then derived from the expression:

$$L_c = \frac{2}{P_0 R_L^2 \frac{1}{Z_{if}} \left[\frac{\Delta I_0}{\Delta P}\right]^2 \left[\frac{1}{R_L} + \frac{1}{Z_{if}}\right]^2}$$

where

ΔP is the small change in microwave power;

ΔI_0 is the corresponding change in rectified current;

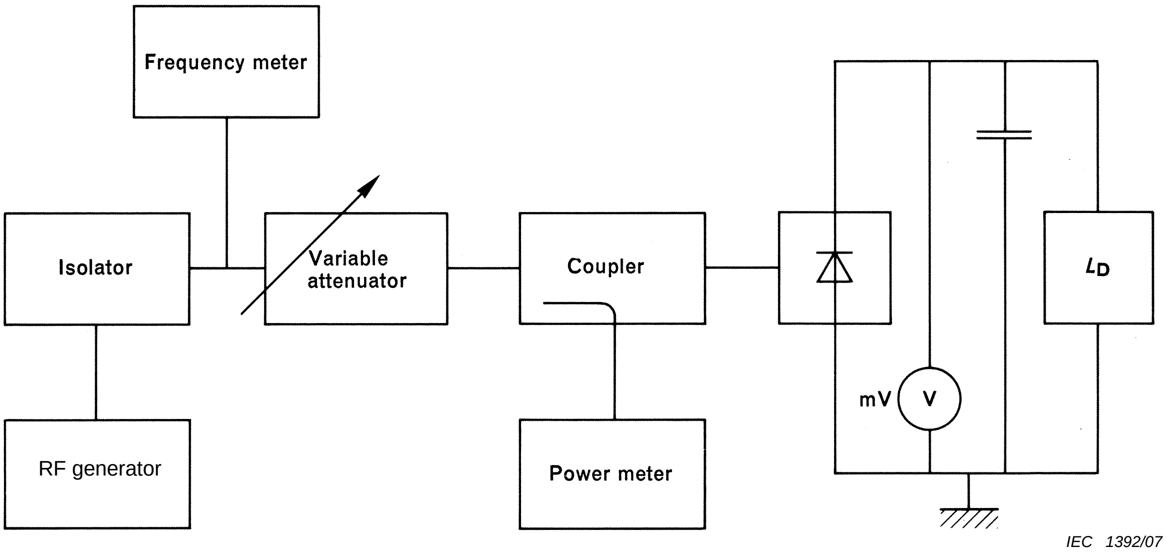
P_0 is the average power $P + \frac{\Delta P}{2}$;

Z_{ir} is the IF impedance of standard diode.

The d.c. incremental method is not readily adapted for rapid production testing, but it is used mainly to establish an absolute calibration of standard diodes. These standard diodes may then be used to calibrate the amplitude modulation method (see 4.1.4.8.2).

4.1.4.8.2 Amplitude modulation method

4.1.4.8.2.1 Circuit diagram



Key
L_D diode circuit providing d.c. and modulation frequency impedances as specified

Figure 34 – Circuit for the measurement of conversion loss in amplitude modulation method

4.1.4.8.2.2 Circuit description and requirements

R_L is the d.c. load resistance to comply with the following requirements:

$$\frac{R_L}{2} < Z_{if} < 2 R_L$$

(this approximation leads to an error smaller than 0,5 dB).

The modulator may be realized by means of an adjustable attenuator driven by a PIN diode. It may be replaced by a direct modulation of the signal source, but in this case, as an accurate assessment of the value of the modulation coefficient is essential, some difficulties may be encountered. The bias supplies may need adjustment.

4.1.4.8.2.3 Precaution

Modulation coefficient shall not exceed 10 %.

4.1.4.8.2.4 Measurement procedure

The diode is inserted into the measurement circuit as shown in Figure 34, and operated under specified conditions. The microwave power incident upon the diode is modulated in amplitude at a specified low frequency by a specified amount. The modulation envelope is detected by the diode and an a.c. voltage developed across the diode load. Knowing the percentage of modulation and the signal source power, the conversion loss may be obtained from:

$$L_c = \frac{m^2 P R_L}{V^2}$$

where

m is the modulation coefficient (not more than 10 %);

P is the mean power on the diode;

R_L is the load resistance of the diode;

V is the low-frequency a.c. voltage across R_L.

This method may be used as an absolute measurement or as a relative comparison method. In the case of a relative assessment, standard diodes as obtained using the d.c. incremental method (4.1.4.8.1) may be used.

As m^2, P and R_L are constants of the measurement circuit, there will be no need to assess their value when making relative measurements.

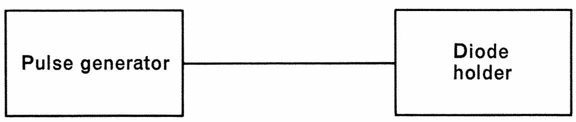
However, if the amplitude modulation method is used to obtain absolute measurements, it is essential to obtain an accurate measurement of m, P and R_L.

4.1.4.9 Burnout energy

4.1.4.9.1 Purpose

To determine the change in noise output of a microwave diode caused by application of RF or pulse energy.

4.1.4.9.2 Circuit diagram



IEC 1139/01

Figure 35 – Block diagram of burnout energy measurement circuit

4.1.4.9.3 Circuit description and requirements

The pulse or RF generator is matched to the transmission line connected to the diode measuring mount. The specified power or pulse energy is the power or energy available from the transmission line. The actual dissipation within the diode depends on the transmission line impedance and on the measuring mount geometry, both of which must be specified.

4.1.4.9.4 Precautions to be observed

Where mechanical switch contacts are used, care must be taken to ensure that possible variations in contact resistance do not affect the severity of the test conditions.

If the test is carried out using low repetition rates, less than 200 Hz, a mercury-wetted relay type switch is recommended.

4.1.4.9.5 Measurement procedure

The diode shall be subjected to one of the following tests as appropriate; after test, the diode shall be measured to determine the change in the noise output produced.

4.1.4.9.5.1 Burnout by repetitive pulses

The diode is subjected to a specified number of pulses of specified duration (which is shorter than the thermal time constant of the diode junction), specified repetition rate, having a specified energy content. The pulse polarity shall be that which causes a current to flow in the direction providing the most severe effect.

4.1.4.9.5.2 Burnout by single pulse

The diode is subjected to a pulse having a specified energy content and having a specified duration (shorter than the thermal time constant of the diode junction). Alternatively, the pulse generating circuit may include a pulse-forming network which is charged to a given voltage (corresponding to the required energy value) and a contact to the diode is made by any suitable means so that a current flows through the diode in that direction which produces the most severe effect.

4.1.4.9.5.3 Burnout by continuous wave (CW) or by RF pulses

The diode is fitted into the specified measuring mount and operated under specified conditions. The specified CW or the specified RF pulse power is applied to the diode for the specified period of time. This power must be matched to the diode input.

4.1.4.10 *Q* value of cavity required to provide a stated reduction of the noise power which a local oscillator provides at a mixer crystal output

If the noise output from a local oscillator is given by a value N (W/MHz) for a stated drive level, the equivalent (NTR) noise temperature ratio (t_o) at the mixed output may be expressed by:

$$t_o = \frac{N\ (\mathrm{W/MHz})}{L_c \times kTB}$$

where L_c is the conversion loss of the mixer as power ratio.

A 290 K and at 4 × 10–15 (W/MHz):

$$t_o = \frac{N}{L_c \times 4 \times 10^{-15}}$$

If, for example, the conversion loss (L_c) of the diode is 6,3 dB, this is equivalent to 4,3 times in power, and

$$t_o = \frac{N}{17{,}2 \times 10^{-15}}$$

The overall noise factor (F), with no local oscillator noise, is given by:

$$F = L_c\ (N_{if} - 1 + t_r)$$

where

N_{if} is the receiver noise factor value;

t_r is the mixer diode N.T.R. value;

both expressed as power ratios.

This expression may be written as $L_c\ t_1$, where t_1 is equal to the term within brackets.

If it is desired to reduce the noise power contributed by the local oscillator in the diode output by n times, the new value of t_o, say t_o^1, is given by t_1/n.

Hence:

$$\frac{t_o}{t_o^1} = \frac{Nn}{4L_c t_1 \times 10^{-15}} \tag{25}$$

Now the power response of a cavity is given by:

$$\frac{1}{1 + \left[\dfrac{2\, f_{if}\, Q_L}{f_o}\right]^2}$$

where

f_{if} is the IF frequency of receiver;

Q_L is the loaded Q of cavity;

f_o is the cavity resonant frequency.

Combining this expression with equation (25) gives:

$$\frac{1}{1 + \left[\dfrac{2\, f_{if}\, Q_L}{f_o}\right]^2} = \frac{4\, L_c\, t_1 \times 10^{-15}}{Nn}$$

and $$Q_L = \frac{f_o}{2f_{if}}\left[\frac{Nn}{4L_c t_1 \times 10^{-15}} - 1\right]^{1/2}$$

EXAMPLE:

Required to reduce the noise contributed by a local oscillator by 10 times.

Noise power from local oscillator (N) = 8 × 10^{-13} W/MHz for a given diode drive.

Diode conversion loss (L_c) = 6,3 dB = 4,3 times.

Diode NTR (t_r) = 1,2 times.

Receiver noise factor = 2 dB (N_{if}) = 1,6 times.

Receiver IF frequency (f_{if}) = 30 MHz.

Local oscillator operates at 10 GHz (f_o).

Then: $$t_1 = 1{,}6–1 + 1{,}2 = 1{,}8$$

$$Q_L = \frac{10^{10}}{2 \times 30 \times 10^6}\left[\frac{8 \times 10^{-13} \times 10^1}{4{,}3 \times 4 \times 10^{-15} \times 1{,}8} - 1\right]^{1/2} = 2\,680$$

4.1.4.11 IF amplifier noise figure

Measurement procedure

The noise figure of the IF amplifier may be determined by the procedure given in 4.1.4.6.1. The circuit arrangement, centre frequency and bandwidth are determined by the IF frequency required.

The noise figure of the amplifier can be expressed as follows:

$$F_{if} = \frac{N'}{k\,T_o\,B}\left[\frac{1}{\frac{N_2}{N_1} - 1}\right]$$

which, when $N_2 = 2N_1$ adjustment as in 4.1.4.6.1, becomes:

$$F_{if} = \frac{N'}{k\,T_o\,B}$$

If a temperature limited noise diode is used as the input noise applied through the amplifier input resistor, then:

$$N' = \frac{eIRB}{2}$$

Then:

$$F_{if} = \frac{eIR}{kT_o} = 20\,IR \text{ when } T_o = 290K$$

and where

I is the noise diode current, in amperes;

R is the amplifier input resistance, in ohms.

4.2 Mixer diodes used in communication applications

4.2.1 General

See 4.1.1.

4.2.2 Terminology and letter symbols

See 4.1.2.

4.2.3 Essential ratings and characteristics

4.2.3.1 General

The essential ratings and characteristics for each category of diode are marked with a + sign in the following table.

– Category A: discrete diode.

– Category B: diode element mounted on a substrate or integrated with a waveguide.

NOTE This includes those circuits which contain passive elements, such as d.c. bias supply circuits, d.c. protection circuits, directional couplers, filters. etc.; balanced types are also included. Circuits containing active elements, such as transistors, oscillator diodes, etc. are not included.

– Subcategory P: point contact diode.

– Subcategory S: Schottky barrier diode.

	Categories			
	A		B	
	P	S	P	S
4.2.3.2 Ratings (limiting values)				
The following ratings should be stated:				
4.2.3.2.1 Temperatures				
4.2.3.2.1.1 Range of operating temperatures	+	+	+	+
4.2.3.2.1.2 Range of storage temperatures	+	+	+	+
4.2.3.2.2 Current				
Maximum mean forward current under specified conditions at 25 °C		+		+
4.2.3.2.3 Power dissipation (including burn-out energy)				
4.2.3.2.4 Maximum c.w. power under specified conditions at 25 °C	+	+	+	+
4.2.3.2.5 Maximum burn-out energy by single pulse under specified conditions at 25 °C	+	+	+ (note 1)	+ (note 1)
4.2.3.3 Electrical characteristics				
Unless otherwise specified, the following characteristics should be stated at 25 °C:				
4.2.3.3.1 Terminal capacitance				
Typical value under specified conditions	+	+		

	Categories			
	A		B	
	P	S	P	S
4.2.3.3.2 Forward current				
Minimum value at specified voltage		+		
4.2.3.3.3 Reverse current				
Maximum value at specified reverse voltage		+		
4.2.3.3.4 Ratio of the forward current at two specified bias voltages, or coefficient *n* (note 2)				
Typical value		+		
4.2.3.3.5 Voltage standing wave ratio				
Maximum value when fitted on mount and operating under specified conditions (note 3)	+	+	+ (note 5)	+ (note 5)
Maximum value(s) under specified conditions (note 4)				
4.2.3.3.6 IF impedance				
Minimum and maximum values under specified conditions (note 3)	+	+	+ (note 5)	+ (note 5)
4.2.3.3.7 Conversion loss				
Typical value under specified conditions (note 3)	+	+	+ (note 5)	+ (note 5)
4.2.3.3.8 Overall noise factor				
Maximum value under specified conditions (notes 3 and 6)	+	+	+	+
4.2.3.3.9 Isolation				
The ratio of the value of the local oscillator signal measured at the local oscillator to the value measured at the input signal port				
Typical value, expressed in decibels, under specified conditions			+	+
4.2.3.4 Application data				
4.2.3.4.1 Recommended mean forward current, under specified operating conditions	+	+	+	+
4.2.3.4.2 Series inductance				
Typical value	+	+		
4.2.3.4.3 Output noise temperature ratio				
Typical value	+	+	+	+

NOTE 1 If the diode is integrated with the elements of the d.c. circuits such as bias supply circuit or bias protection circuit, the overall value should include the effect of this d.c. circuit.

NOTE 2 "*n*" is defined by:

$$i = i_s \left(e^{\frac{qV}{nkT}} - 1\right)$$

where

i is the diode forward current;

i_s is the reverse saturation current;

q is the electron charge;

V is the applied bias voltage;

k is the Boltzmann's constant;

T is the absolute temperature.

NOTE 3 The device holder should be specified by the manufacturer.

NOTE 4 If input signal and local oscillator terminals are separated, the voltage standing wave ratio for each terminal should be given. In this case, the frequencies should be specified for each terminal.

NOTE 5 In addition, it is necessary to give the frequency response, typical value, in either numerical or graphical representation.

NOTE 6 The noise factor F_o should be determined for an assumed or actual F_{if} of 1,5 dB.

4.2.4 Measuring methods

See 4.1.4.1 to 4.1.4.9.

4.3 Detector diodes

(Under consideration.)

5 Impatt diodes

5.1 Impatt diodes amplifiers

5.1.1 General

(To be defined.)

5.1.2 Terms and definitions

For the purposes of this clause, the following terms and definitions apply.

5.1.2.1 Terms and letter symbols

5.1.2.1.1 Temperature

Storage temperature T_{stg}

Case operating temperature T_{case}

Resonant structure ambient operating temperature T_{amb}

Resonant structure body operating temperature T_{MB}

5.1.2.1.2 Voltage

Breakdown voltage $V_{(BR)}$

Operating voltage V_{OP}

5.1.2.1.3 Current

Reverse current I_R

Operating current I_{OP}

Continuous current I_A

Peak transient current I_{AM}

5.1.2.1.4 Power

Power dissipation P_D

Output power (for a defined oscillator structure) P_o

Added output power (for a defined amplifier structure) $P_{o\ add}$

Output power change with current $\Delta P_{o(\Delta I)}$

Spurious output power P_{sp}

5.1.2.1.5 Capacitance and resistance

Case capacitance C_{case}

Junction capacitance C_j

Total capacitance C_{tot}

Junction-to-case thermal resistance $R_{th(j\text{-}c)}$

5.1.2.1.6 Frequency

Minimum frequency of mechanical tuning range f_{minM}

Maximum frequency of mechanical tuning range f_{maxM}

Minimum frequency of electrical tuning range f_{minE}

Maximum frequency of electrical tuning range f_{maxE}

Frequency change with current $\Delta f_{(\Delta I)}$

Frequency change at turn-on Δf_{on}

Change of frequency with temperature $\Delta f_{(\Delta T)}$

Change of output power with temperature $\Delta P_{o(\Delta T)}$

Change of frequency with load impedance (of an oscillator) $\Delta f_{(\Delta Z)}$

Change of output power with load impedance $\Delta P_{o(\Delta Z)}$

Injection locking range

The range of frequencies of an injected signal to which the oscillator will lock

Minimum injection locking frequency f_{minL}

Maximum injection locking frequency f_{maxL}

Oscillator conversion efficiency η, η_{OSC}

The ratio of RF power output to the d.c. input power

Power-added efficiency (for amplifiers or locked oscillators in an FM system) η, η_{add}

The ratio of the difference between the RF output power and the input power to the d.c. input power

5.1.2.1.7 Other parameters

FM spot noise figure F_{fm}

Double sideband AM spot noise figure F_{am}

RF sum-on time t_{onRF}

5.1.2.2 Complementary definitions

5.1.2.2.1 Temperature

resonant-structure ambient operating temperature
T_{amb}
air temperature measured under operating conditions below the resonant structure in an environment of substantially uniform temperature, cooled only by natural air convection and not materially affected by reflective and radiant surfaces

resonant-structure body operating temperature
T_{MB}
temperature measured under operating conditions at a specified reference point on the body surface of the resonant structure

5.1.2.2.2 Voltage

operating voltage (of an Impatt diode)
V_{OP}
voltage across the terminals that results from the flow of operating current

5.1.2.2.3 Current

operating current (of an Impatt diode)
I_{OP}
current in the avalanche region at which the diode operates

5.1.2.2.4 Power

output power (for a defined amplifier structure)
P_o
power delivered to a matched termination at the output terminals of the oscillator structure

added output power (for a defined amplifier structure)
$P_{o\ add}$
power contributed by the amplifier structure to the output power (i.e. RF input power is excluded)

spurious output power
P_{sp}
total integrated output power excluding the power at the fundamental frequency

5.1.2.2.5 Capacitance

case capacitance
C_{case}
capacitance between the terminals of the diode case with no die installed

5.1.2.2.6 Frequency

frequency change at turn-on (for an oscillator)
Δf_{on}
frequency change from an initial value immediately following the application of power to the value finally reached

NOTE For specification purposes, it must be indicated how the initial value of frequency is defined.

5.1.2.2.7 Other parameters

FM spot noise figure
F_{fm}

ratio of:

a) FM output noise power unit bandwidth (spectral density) at a single output frequency when the noise temperature of all input terminations is equal to the reference noise temperature T_0 at all frequencies that contribute to the output noise,

to:

b) that part of item a) caused by the noise of the signal-input termination at the signal-input frequency

NOTE 1 FM noise is that part of the total noise that is detected by a system that responds only to frequency modulation.

NOTE 2 The word "spot" has been introduced in the title to be consistent with 702-08-57 of IEC 60050-702:1992.

double sideband AM spot noise figure
F_{am}

ratio of:

1) AM output noise power per unit bandwidth (spectral density) at a single output frequency when the noise temperature of all input terminations is equal to the reference noise temperature T_0 at all frequencies that contribute to the output noise,

to:

2) that part of item 1) caused by the noise of the signal input termination at the signal input frequency

NOTE 3 AM noise is that part of the total noise that is detected by a system that responds only to amplitude modulation.

RF turn-on time
t_{onRF}

for a free-running oscillator, the time taken from switch-on to reach a specified frequency

for a locked oscillator or amplifier, the time taken from switch-on to reach a specified output power

5.1.3 Essential ratings and characteristics

5.1.3.1 General

The ratings of the electrical characteristics should be stated, either at ambient-rated temperature, or at case-related temperature.

5.1.3.2 Ratings (limiting values)

5.1.3.2.1 Temperature

5.1.3.2.1.1 Storage temperature

Minimum and maximum values.

5.1.3.2.1.2 Case operating temperature

Minimum and maximum values.

NOTE One of the temperatures given in 3.4.1 may be specified as an alternative.

5.1.3.2.2 Power dissipation

Maximum value at specified case temperature.

5.1.3.2.3 Continuous current

Maximum value in the avalanche region.

5.1.3.2.4 Peak transient current

Maximum value in the avalanche region, for specified pulse duration.

5.1.3.3 Characteristics

The following characteristics should be given at a case temperature of 25 °C.

5.1.3.3.1 Breakdown voltage

Minimum and maximum values, at specified reverse current.

5.1.3.3.2 Reverse current

Maximum value, at specified reverse voltage below the minimum breakdown voltage.

5.1.3.3.3 Total capacitance

Minimum and maximum values, under specified bias voltage conditions.

5.1.3.3.4 Junction-to-case thermal resistance

Maximum value.

5.1.3.4 Additional ratings and characteristics

When the diode is specified for use in a defined resonant structure, the following additional ratings and characteristics should be given.

5.1.3.4.1 Ratings (limiting values)

Either:

5.1.3.4.1.1 Resonant structure ambient operating temperature

Minimum and maximum values.

Or:

5.1.3.4.1.2 Resonant structure body operating temperature

Minimum and maximum values at a specified reference point.

5.1.3.4.2 Characteristics

The following characteristics of the avalanche diode and its defined resonant structure should be given at an ambient or structure body temperature of 25 °C unless otherwise stated.

5.1.3.4.2.1 Operating voltage

Maximum value, at specified operating current(s).

5.1.3.4.2.2 Output power (for a defined oscillator structure)

Minimum and, where appropriate, maximum values, at specified diode operating current(s) and, if applicable, over the specified mechanical or electrical tuning range and at a specified voltage standing wave ratio (VSWR).

5.1.3.4.2.3 Added output power (for a defined amplifier structure)

Minimum and, where appropriate, maximum values, at specified diode operating current(s) and, if applicable, over the specified mechanical or electrical tuning range and at a specified VSWR.

The following characteristics should be given where applicable.

5.1.3.4.2.4 Tuning range (mechanical)

Minimum and maximum frequencies at specified operating current.

5.1.3.4.2.5 Tuning range (electrical)

Minimum and maximum frequencies over a specified tuning diode voltage range and at a specified operating current.

5.1.3.4.2.6 Frequency change with operating current

Minimum or maximum value over a specified operating current range.

5.1.3.4.2.7 Output power change with current

Maximum value over a specified operating current range.

5.1.3.4.2.8 FM noise figure

Maximum value under specified conditions, preferably selected from the following:

Bandwidth: 1 Hz, 10 Hz, 100 Hz or 1 kHz
and 10 kHz, 100 kHz, 1 MHz or 10 MHz for carrier.

5.1.3.4.2.9 Double sideband AM noise figure

Maximum value under specified conditions, preferably selected from those in 5.1.3.4.2.8.

5.1.3.4.2.10 Frequency change at turn-on

Maximum value under specified conditions.

5.1.3.4.2.11 Spurious output power(s)

Maximum value(s), at specified frequencies and under specified conditions.

5.1.3.4.2.12 Frequency change with temperature

Maximum value over the operating temperature range, under specified conditions.

5.1.3.4.2.13 Output power change with temperature

Maximum value over the operating temperature range, under specified conditions.

5.1.3.4.2.14 Frequency change with load impedance variation

Maximum value for a specified voltage standing wave ratio (VSWR), all phases.

5.1.3.4.2.15 Output power change with load impedance variation

Maximum value for a specified VSWR all phases.

5.1.3.4.2.16 RF turn-on time

Maximum value under specified conditions.

5.1.3.4.2.17 Injection locking range

Minimum and maximum injection locking frequencies at specified injection locking power.

5.1.3.4.2.18 Continuous (direct) to RF conversion efficiency (for oscillators)

Minimum value under specified conditions.

5.1.3.4.2.19 Power-added efficiency (for amplifiers)

Minimum value under specified conditions.

5.1.3.5 Supplementary information

5.1.3.5.1 Case capacitance

Typical value.

5.1.3.5.2 Junction capacitance

Typical value under specified conditions.

5.2 Impatt diodes oscillators

(Under consideration).

6 Gunn diodes

6.1 General

(Under consideration).

6.2 Terms and definitions

For the purposes of this clause, the following terms and definitions apply.

6.2.1 pulse breakdown voltage (of a Gunn diode)

$V_{(BR)}$

lowest value of the voltage under specified pulse conditions at which the resistance of the diode decreases suddenly

6.3 Essential ratings and characteristics

(Under consideration).

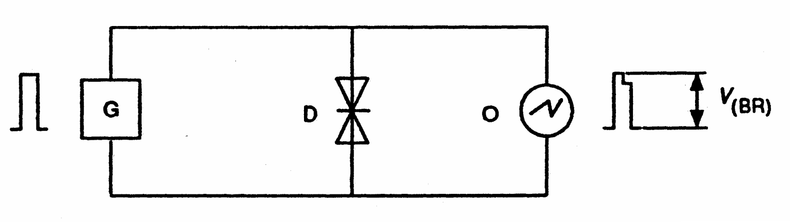
6.4 Measuring methods

6.4.1 Pulse breakdown voltage $V_{(BR)}$

6.4.1.1 Purpose

To measure the threshold voltage of a Gunn diode under the specified conditions.

6.4.1.2 Circuit diagram



Key

$V_{(BR)}$ pulse breakdown voltage

Figure 36 – Circuit for the measurement of pulse breakdown voltage

6.4.1.3 Circuit description and requirements

D is the diode being measured

G is the nanosecond pulse generator

O is the oscilloscope

6.4.1.4 Measurement procedure

The input pulse is applied to the diode by the pulse generator, G.

Increase the amplitude of the input pulse until the back-porch of the pulse observed on the oscilloscope drops abruptly.

Immediately prior to this time the voltage across the diode is the threshold voltage of the Gunn diode.

6.4.1.5 Specified conditions

Ambient temperature

Pulse duration

Pulse duty factor

6.4.1.6 Precautions

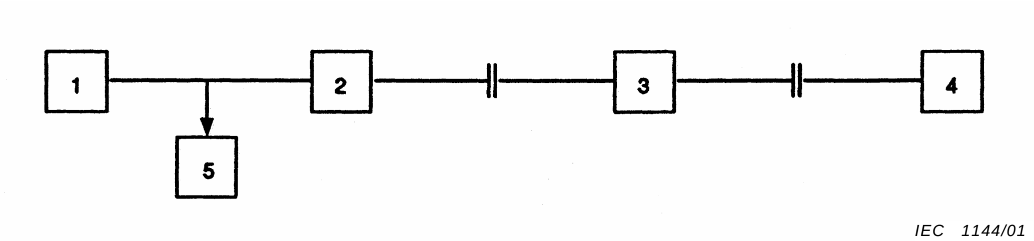
The pulse duration and duty cycle shall be chosen to prevent the permanent change of characteristics or destruction of the device.

6.4.2 Threshold voltage

6.4.2.1 Purpose

To measure the threshold voltage of a Gunn diode under specified conditions.

6.4.2.2 Circuit diagram



Key
1 voltage generator
2 cavity with the diode being measured
3 variable attenuator
4 oscillation indicator (power meter or detector chamber with a microammeter or a stroboscopic oscilloscope)
5 d.c. voltage meter

Figure 37 – Circuit for the measurement of threshold voltage

6.4.2.3 Circuit description and requirements

For proper measurement it is necessary to have a cavity chamber, the design of which is specified for each type of diode.

The cavity chamber shall meet the following requirements:

– have adjustment elements enabling visual oscillation mode of a diode to be observed within the specified frequency band;
– ensure a specified microwave frequency coupling in the diode supply circuit;
– ensure the heat removal from the diode in the operating conditions. In this case the critical holding temperature should not exceed a specified maximum rating for a particular type of diode. The input resistance of the voltage source should not exceed 10 % of the diode resistance.

6.4.2.4 Measurement procedure

The diode is inserted into the cavity. The supply voltage is gradually increased until oscillation occurs. The moment when the oscillation occurs is defined by the indicator (4). The cavity is adjusted to the maximum indicator reading. When the indicator is overloaded, an attenuator is used. The supply voltage is reduced to zero. These operations are repeated until the minimum voltage value at which oscillation occurs is achieved. The threshold voltage value is read on the d.c. voltage meter at the moment when the oscillation occurs.

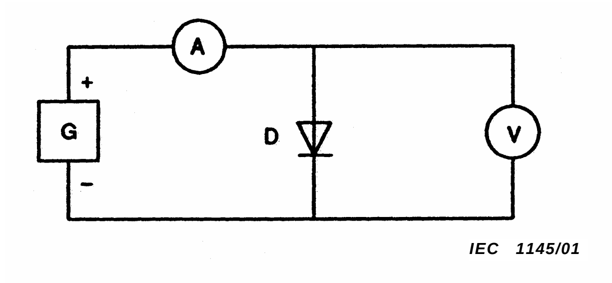
6.4.3 Resistance

6.4.3.1 Voltmeter-ammeter method

6.4.3.1.1 Purpose

To measure the resistance of a Gunn diode under specified conditions.

6.4.3.1.2 Circuit diagram



Key
G d.c. generator
A d.c. ammeter
V d.c. voltmeter
D diode being measured

Figure 38 – Circuit for the measurement of resistance in voltmeter-ammeter method

6.4.3.1.3 Circuit description and requirements

The input resistance of the voltmeter should be high with respect to the maximum resistance of the diode.

6.4.3.1.4 Measurement procedure

The measuring current shall be significantly lower than the threshold current, I_{TO}.

Take into account the voltage drop on wires and terminals connecting the diode to the measurement circuit.

For this purpose the voltage drop on the short-circuited terminals of the diode holder is measured.

The resistance of a Gunn diode is calculated by the formula:

$$R = \frac{U}{I}$$

where

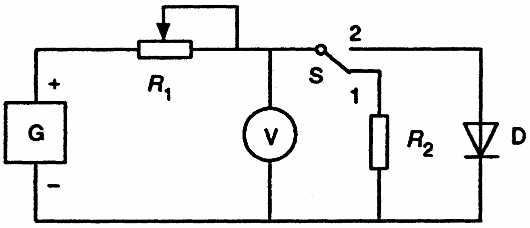
R is the resistance of the diode being measured;

U is the voltage drop on the diode being measured;

I is the current through the diode being measured.

6.4.3.2 Alternative method

6.4.3.2.1 Circuit diagram



Key
G d.c. generator
R_1 variable resistor
V d.c. voltmeter
S switch
R_2 calibrated resistor
D diode being measured

Figure 39 – Circuit for the measurement of resistance in alternative method

6.4.3.2.2 Circuit description and requirements

The internal resistance of the d.c. generator (G and R_1) should be high with respect to the maximum resistance of the diode. The value of the variable resistor, R_1, is chosen so that the current through the diode can be adjusted within ±10 %.

The resistance of the calibrated resistor, R_2, should not be less than the maximum rating of the diode resistance.

6.4.3.2.3 Measurement procedure

The switch S is set in position 1 and the meter scale is calibrated by the resistor, R_1.

Then the switch S is set in position 2 and the resistance value is read on the meter scale.

7 Bipolar transistors

7.1 General

This clause provides terms and definitions, essential ratings and characteristics, measuring methods, and verifying methods for bipolar transistors used in microwave applications. For general items of bipolar transistors, refer to IEC 60747-7.

7.2 Terms and definitions

For the purposes of this clause , the following terms and definitions apply:

7.2.1
output power at 1dB gain compression
$P_{o(1dB)}$
See 8.2.13.

7.2.2
output power
P_o
See 3.3 of IEC 60747-16-2:2002.

7.2.3
power gain at 1dB gain compression
$G_{p(1dB)}$
See 8.2.14.

7.2.4
power-added efficiency
η_{add}
See 8.2.15.

7.2.5
collector efficiency
η_c
ratio of output power to d.c. input power of collector

NOTE This ratio is normally expressed as a percentage.

7.2.6
noise figure
F
See 702-08-57 of IEC 60050-702:1992.

7.2.7
associated gain
G_{as}
See 8.2.23.

7.2.8
minimum noise figure
F_{min}
See 8.2.21.

7.2.9
source reflection factor for minimum noise figure
r_{GFmin}
source reflection factor that gives minimum noise figure

NOTE 1 For source reflection coefficient (factor), see 3.5.3.3 of IEC 60747-7:2000.

NOTE 2 The symbol "Γ_{opt}" is still in common use for the source reflection factor for minimum noise figure.

7.2.10
equivalent input noise resistance
R_n
quotient of the equivalent input noise voltage and the equivalent input noise current (see 3.4.5 and 3.4.6 of IEC 60747-1:2006).

7.2.11
maximum frequency of oscillation
f_{max}
See 3.4.14 of IEC 60747-7:2000.

7.2.12
transition frequency
$\boldsymbol{f_T}$
See 3.4.15 of IEC 60747-7:2000.

7.2.13
frequency of unity current transfer ratio
$\boldsymbol{f_1}$
See 3.4.16 of IEC 60747-7:2000.

7.2.14
maximum available gain
$\boldsymbol{G_{a(max)}}$
See 8.2.8.

NOTE The abbreviation "*MAG*" is in common use for the maximum available gain.

7.2.15
maximum stable gain
$\boldsymbol{G_{ms}}$
magnitude of the ratio of S_{21} in common emitter configuration, to S_{12} in common emitter configuration, the output terminating resistance and the source resistance each having the value R_0

NOTE The abbreviation "*MSG*" is in common use for the maximum available gain.

7.2.16
insertion power gain
$\boldsymbol{|S_{21e}|^2}$
magnitude squared of S_{21} in common emitter configuration, the output terminating resistance and the source resistance each having the value R_0

7.2.17
input power at the intercept point (for intermodulation products)
$\boldsymbol{P_{i,n(IP)}}$
input power at intersection between the extrapolated output powers of the fundamental component and *n*th order intermodulation components, when the extrapolation is carried out in a diagram showing the output power of the components (in decibels) as a function of the input power (in decibels)

NOTE The abbreviation "IIP_n" is in common use for the input power at the intercept point (for intermodulation products).

7.2.18
output power at the intercept point (for intermodulation products)
$\boldsymbol{P_{o,n(IP)}}$
output power at intersection between the extrapolated output powers of the fundamental component and *n*th order intermodulation components, when the extrapolation is carried out in a diagram showing the output power of the components (in decibels) as a function of the input power (in decibels)

NOTE The abbreviation "OIP_n" is in common use for the output power at the intercept point (for intermodulation products).

7.2.19
intermodulation distortion
$\boldsymbol{P_1/P_n}$
See 3.7 of IEC 60747-16-1:2007.

NOTE The abbreviation "IMD_n" is in common use for the *n*th order intermodulation distortion.

7.2.20
load mismatch tolerance
Ψ_L
maximum load VSWR in the range where the device amplifies the input signal with no oscillation and no spurious intensity and/or no discontinuity of the frequency response at all phase angles with specified conditions

7.2.21
source mismatch tolerance
Ψ_S
maximum source VSWR in the range where the device amplifies the input signal with no oscillation and no spurious intensity and/or no discontinuity of the frequency response at all phase angles with specified conditions

7.2.22
load mismatch ruggedness
Ψ_R
maximum load VSWR in the range where the device withstand load mismatch with no degradation at all phase angles with specified conditions

7.3 Essential ratings and characteristics

7.3.1 General

This subclause gives ratings and characteristics required for specifying microwave bipolar transistors.

Microwave bipolar transistors are divided into two categories:

– category A: power devices;
– category B: small power signal devices.

7.3.2 Limiting values (absolute maximum rating system)

7.3.2.1 Electrical limiting values

Limiting values shall be specified as shown in Table 1:

Table 1 – Electrical limiting values

Subclause	Parameters	Min.	Max.
7.3.2.1.1	Collector-base voltage with zero emitter current, V_{CBO}		+
7.3.2.1.2	Collector-emitter voltage with zero base current, V_{CEO}		+
7.3.2.1.3	Emitter-base voltage with zero collector current, V_{EBO}		+
7.3.2.1.4	Collector current, I_C		+
7.3.2.1.5	Storage temperature, T_{stg}	+	+
7.3.2.1.6	Junction temperature, T_j		+
7.3.2.1.7	Either total power dissipation, P_{tot} or		+
	Collector power dissipation, P_C		+

7.3.2.2 Characteristics

Characteristics are to be given at 25 °C except where otherwise stated.

7.3.2.2.1 DC characteristics

The parameters shall be specified corresponding to categories as shown in Table 2 below.

Table 2 – DC characteristics

Subclause	Parameters	Min.	Typ.	Max.	Categories	
					A	B
7.3.2.2.1.1	Collector-base cut-off current, I_{CBO}	+		+	+	+
7.3.2.2.1.2	Emitter-base cut-off current, I_{EBO}	+		+	+	+
7.3.2.2.1.3	Collector-emitter cut-off current, I_{CEO} (where appropriate)	+		+	+	+
7.3.2.2.1.4	Static value of common-emitter forward current transfer ratio, h_{21E}	+		+	+	+
7.3.2.2.1.5	Junction-case thermal resistance, $R_{th(j-c)}$			+	+	
7.3.2.2.1.6	Collector-base breakdown voltage with zero emitter current, $V_{(BR)CBO}$ (where appropriate)	+	+		+	+
7.3.2.2.1.7	Emitter-base breakdown voltage with zero emitter current, $V_{(BR)EBO}$ (where appropriate)	+	+		+	+
7.3.2.2.1.8	Collector-emitter breakdown voltage with zero emitter current, $V_{(BR)CEO}$ (where appropriate)	+	+		+	+

7.3.2.2.2 RF characteristics

The parameters shall be specified corresponding to categories as shown in Table 3 below:

Table 3 – RF characteristics

Subclause	Parameters	Min.	Typ.	Max.	Categories	
					A	B
7.3.2.2.2.1	Common-emitter reverse transfer capacitance, C_{re} (where appropriate)		+	+		+
7.3.2.2.2.2	Common-base output capacitance, C_{obs} (where appropriate)		+	+		+
7.3.2.2.2.3	Base-collector capacitance, C_{cb} (where appropriate)		+	+		+
7.3.2.2.2.4	Either output power at 1dB gain compression, $P_{o(1dB)}$ or Output power at specified input power, P_o	+	+		+	
7.3.2.2.2.5	Power gain at 1dB gain compression, $G_{p(1dB)}$	+	+		+	
7.3.2.2.2.6	Power-added efficiency, η_{add} (where appropriate)	+	+		+	
7.3.2.2.2.7	Collector efficiency, η_c (where appropriate)	+	+		+	
7.3.2.2.2.8	Noise figure, F		+	+		+
7.3.2.2.2.9	Associated gain, G_{as}	+	+			+
7.3.2.2.2.10	Minimum noise figure, F_{min} (where appropriate)		+	+		+
7.3.2.2.2.11	Source reflection factor for minimum noise figure, r_{GFmin} (where appropriate)		+			+
7.3.2.2.2.12	Equivalent input noise resistance, R_n (where appropriate)		+			+
7.3.2.2.2.13	Maximum frequency of oscillation, f_{max} (where appropriate)		+			+
7.3.2.2.2.14	Transition frequency, f_T (where appropriate)		+			+
7.3.2.2.2.15	Frequency of unity current transfer ratio, f_1 (where appropriate)		+			+

Table 3 *(continued)*

Subclause	Parameters	Min.	Typ.	Max.	Categories	
					A	B
7.3.2.2.2.16	**Either**					
	Maximum available gain, $G_{a(max)}$ or maximum stable gain, G_{ms} or Insertion gain, $\|S_{21e}\|^2$	+	+			+
7.3.2.2.2.17	**Either**					
	Input power at the intercept point(for intermodulation products), $P_{i,n(IP)}$ or output power at the intercept point (for intermodulation products), $P_{o,n(IP)}$ or Intermodulation distortion, P_1/P_n (where appropriate)		+		+	
7.3.2.2.2.18	**Load mismatch tolerance, Ψ_L (where appropriate)**			+	+	
7.3.2.2.2.19	**Source mismatch tolerance, Ψ_S (where appropriate)**			+	+	
7.3.2.2.2.20	**Load mismatch ruggedness, Ψ_R (where appropriate)**			+	+	

7.4 Measuring methods

7.4.1 General

The measuring methods of field-effect transistor are applicable, with the terms and symbols bipolar transistor in 7.4.3.3, 7.4.3.4, 7.4.3.5, 7.4.3.6, 7.4.3.7, 7.4.3.8 and 7.4.3.9 beir replaced with the replacing rules shown in Table 4 and 5 or 6. Constant base current, Table 5, or constant base voltage, in Table 6 is used properly.

Table 4 – Replacing rule for terms

Terms to be replaced	Terms to be replaced by
Field-effect transistor	Bipolar transistor
Gate-source voltage, V_{GS}	Base-emitter voltage, V_{BE}
Gate-source cut-off voltage	Base emitter cut-off current
Drain-source voltage, V_{DS}	Collector-emitter voltage, V_{CE}
Drain current, I_D	Collector current, I_C

Table 5 – Replacing rule for symbols in the case of constant base current

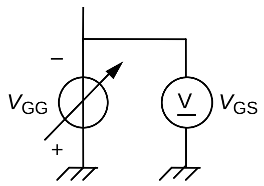
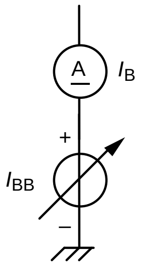
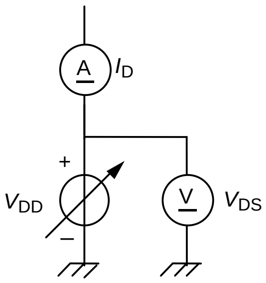
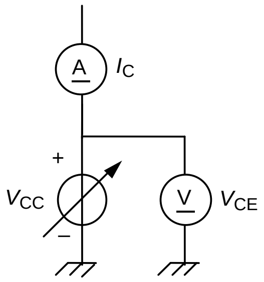
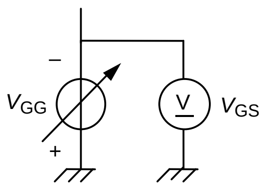
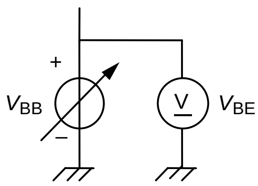
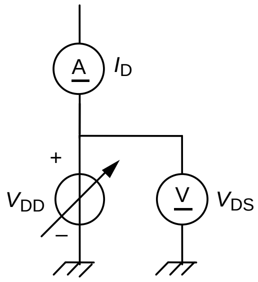
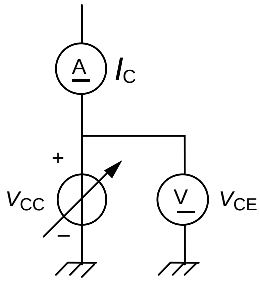
Symbols to be replaced	Symbols to be replaced by
	
	

Table 6 – Replacing rule for symbols in the case of constant base voltage

Symbols to be replaced	Symbols to be replaced by
	
	

7.4.2 DC characteristics

7.4.2.1 Collector-base breakdown voltage with zero emitter current ($V_{(BR)CBO}$), emitter-base breakdown voltage with zero collector current ($V_{(BR)EBO}$) and collector-emitter breakdown voltage with zero base current ($V_{(BR)CEO}$)

The measuring methods of $V_{(BR)CBO}$ and $V_{(BR)EBO}$ given in 6.1.10 of IEC 60747-7:2000, are applicable. The measuring method of $V_{(BR)CEO}$ is applicable by suitably interchanging the base and emitter terminals in 6.1.10.2 of IEC 60747-7:2000.

7.4.2.2 Collector-base cut-off current (I_{CBO}), emitter-base cut-off current (I_{EBO}) and collector-emitter cut-off current (I_{CEO})

The measuring methods given in 6.1.2.1, 6.1.2.2 and 6.1.3 of IEC 60747-7:2000 are applicable.

7.4.2.3 Static value of common-emitter forward current transfer ratio (h_{21E})

The measuring method given in 6.2.7 of IEC 60747-7:2000 is applicable.

7.4.2.4 Junction-case thermal resistance ($R_{th(j-c)}$)

The measuring method given in 6.1.11 of IEC 60747-7:2000 is applicable.

7.4.3 RF characteristics

7.4.3.1 Common-emitter reverse transfer capacitance (C_{re})

The measuring method given in 6.1.8.2 of IEC 60747-7:2000 is applicable, with the term "collector-base capacitance, C_{cb}" being replaced by "common-emitter reverse transfer capacitance, C_{re}".

NOTE The term "collector-base capacitance, C_{cb}" is identical to the common-emitter reverse transfer capacitance.

7.4.3.2 Common-base output capacitance (C_{ob})

The measuring method given in 6.1.8.1 of IEC 60747-7:2000 is applicable.

7.4.3.3 Output power at specified input power (P_o)

The measuring method given in 8.4.3.1 is applicable, with the terms and symbols being replaced with the replacing rules shown in Tables 4 and 5 or 6. “Type of base bias supply” should be added to specified conditions.

7.4.3.4 Output power at 1dB gain compression ($P_{o(1dB)}$)

The measuring method given in 8.4.3.2 is applicable, with the terms and symbols being replaced with the replacing rules shown in Tables 4 and 5 or 6. “Type of base bias supply” should be added to specified conditions.

7.4.3.5 Power gain at 1dB gain compression ($G_{p(1dB)}$)

The measuring method given in 8.4.3.3 is applicable, with the terms and symbols being replaced with the replacing rules shown in Tables 4 and 5 or 6. “Type of base bias supply” should be added to specified conditions.

7.4.3.6 Power-added efficiency (η_{add})

The measuring method given in 8.4.3.4 is applicable, with the terms and symbols being replaced with the replacing rules shown in Tables 4 and 5 or 6. “Type of base bias supply” should be added to specified conditions.

7.4.3.7 Collector efficiency (η_c)

The measuring method given in 7.4.3.6 is applicable, with the term “η_{add}” being replaced by “η_c” which is given by equation (26).

$$\eta_c = \frac{P_o}{V_{CE} \times I_C} \times 100 \qquad (26)$$

where

V_{CE} the collector-emitter voltage in volts;

I_C the collector current in amperes;

P_o the output power in watts.

7.4.3.8 Noise figure (F) and associated gain (G_{as})

The measuring method given in 8.4.3.6 is applicable, with the terms and symbols being replaced with the rules shown in Tables 4 and 5 or 6.

”Type of base bias supply” should be added to specified conditions.

7.4.3.9 Minimum noise figure (F_{min}), equivalent input noise resistance (R_n) and source reflection factor for minimum noise figure (r_{GFmin})

The measuring method given in 8.4.3.7 is applicable, with the terms and symbols being replaced with the rules shown in Tables 4 and 5 or 6.

”Type of base bias supply” should be added to specified conditions.

7.4.3.10 Scattering parameters (S_{ij})

Two measuring methods are given:

– Method 1, using basic instruments and components;

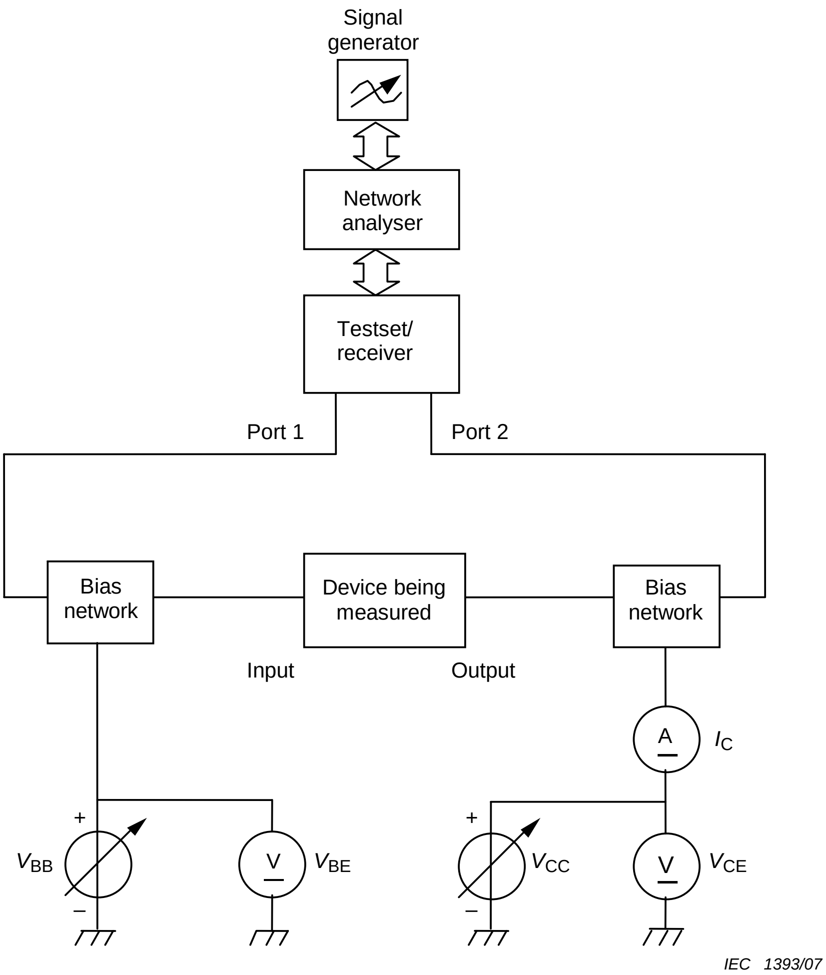
– Method 2, using a network analyser. This method is preferable for microwave frequency.

7.4.3.10.1 Measuring method 1

The measuring method given in 6.1.13.6 of IEC 60747-7:2000 is applicable.

7.4.3.10.2 Measuring method 2

7.4.3.10.2.1 Circuit diagram



NOTE This diagram shows common-emitter configuration as an example. In this case, constant base current, in Table 5, may be used as the base bias supply.

Figure 40 – Circuit for the measurement of scattering parameters

7.4.3.10.2.2 Principle of measurement

The scattering parameters are defined by the following equations:

$$b_1 = S_{11}a_1 + S_{12}a_2$$
$$b_2 = S_{21}a_1 + S_{22}a_2 \quad (27)$$

where a_1 and a_2 are the incident wave quantities, b_1 and b_2 are the reflected wave quantities, all having the dimension (Watt)$^{1/2}$. See Figure 41.

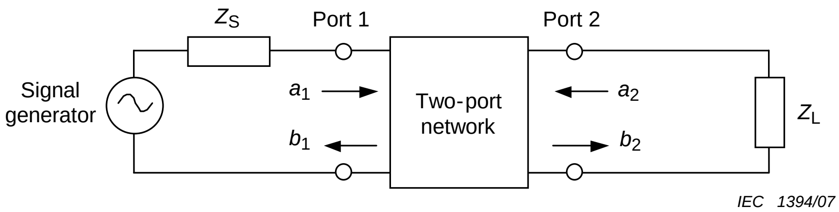


Figure 41 – Incident and reflected waves in a two-port network

Each scattering parameter is defined by the following equations:

$$S_{11} = \frac{b_1}{a_1}\bigg|_{a_2=0} \quad \text{(input reflection coefficient)} \quad (28)$$

$$S_{21} = \frac{b_2}{a_1}\bigg|_{a_2=0} \quad \text{(forward transmission coefficient)} \quad (29)$$

$$S_{12} = \frac{b_1}{a_2}\bigg|_{a_1=0} \quad \text{(reverse transmission coefficient)} \quad (30)$$

$$S_{22} = \frac{b_2}{a_2}\bigg|_{a_1=0} \quad \text{(output reflection coefficient)} \quad (31)$$

Terminating the output port in an impedance equal to the characteristic impedance of the measurement system is equivalent to setting a_2=0. Terminating the input port in an impedance equal to the characteristic impedance of the measurement system is equivalent to setting a_1=0.

7.4.3.10.2.3 Circuit description and requirements

The input port of the device being measured is connected to port 1 of the network analyser, and the output port is connected to port 2.

7.4.3.10.2.4 Precautions to be observed

Oscillation should be eliminated during these measurements.

7.4.3.10.2.5 Measurement procedure

The frequency of the network analyser shall be set to the specified value.

The output power of the network analyser shall be adjusted to the appropriate value that the specified input power is applied to the device being measured.

The network analyser shall be calibrated at the specified reference plane.

The device being measured shall be connected in the specified ground configuration between port 1 and port 2 of the network analyser.

The bias under specified conditions shall be applied to the device being measured.

The specified input power shall be applied from port 1 to the device being measured.

The wave quantities a_1, b_1 and b_2 are measured, and S_{11} and S_{21} are calculated by equations (28) and (29).

The specified input power shall be applied from port 2 to the device being measured.

The wave quantities a_1, b_1 and b_2 are measured, and S_{12} and S_{22} are calculated by equations (30) and (31).

7.4.3.10.2.6 Specified conditions

– Ambient or reference-point temperature
– Measurement frequency
– Input power
– Reference plane
– Ground configuration
– Bias conditions

7.4.3.11 Maximum frequency of oscillation (f_{max})

7.4.3.11.1 Purpose

To measure the maximum frequency of oscillation of a transistor under specified conditions.

7.4.3.11.2 Circuit diagram

See the block diagram of the circuit for the scattering parameter measurement which is shown in 7.4.3.10.

7.4.3.11.3 Circuit description and requirements

See the circuit description and requirements which are shown in 7.4.3.10.

7.4.3.11.4 Precautions to be observed

See the precautions to be observed which are shown in 7.4.3.10.

7.4.3.11.5 Principle of measurement

All four scattering parameters are measured and the maximum unilateral gain $G_{u(max)}$ is calculated as:

$$G_{u(max)} = \frac{|S_{21} - S_{12}|^2}{1 + |S_{11}S_{22} - S_{21}S_{12}|^2 - |S_{11}|^2 - |S_{22}|^2 - (S_{21}S_{12}^* + S_{21}^*S_{12})} \quad (32)$$

Maximum frequency of oscillation (f_{max}) is the frequency where $|G_{u(max)}|$ is equal to 1.

7.4.3.11.6 Measurement procedure

See the measurement procedure for the scattering parameters S_{11}, S_{12}, S_{21}, S_{22} which are shown in 7.4.3.10.

Obtain the maximum frequency of oscillation (f_{max}), according to the following priority.

a) case 1

   $G_{u(max)}$ is calculated by equation (32) using the measured scattering parameters.

   Maximum frequency of oscillation (f_{max}) is the frequency where $|G_{u(max)}|$ is equal to 1.

b) case 2

   If the maximum frequency of oscillation (f_{max}) exceeds the measurement frequency, a small-signal equivalent circuit of the transistor is used to calculate the scattering parameters and $G_{u(max)}$.

   The values of the elements of the small-signal equivalent circuit are extracted from the measured scattering parameters.

   The scattering parameters and $G_{u(max)}$ of the small-signal equivalent circuit are calculated beyond the frequency that meets $|G_{u(max)}| < 1$.

   Maximum frequency of oscillation (f_{max}) is the frequency where $|G_{u(max)}|$ is equal to 1.

c) case 3

   If the maximum frequency of oscillation (f_{max}) exceeds the measurement frequency and a small-signal equivalent circuit is not available, -6dB/octave extrapolation may be used.

   The highest measurement frequency f shall be chosen.

   The maximum frequency of oscillation (f_{max}) is calculated as:

$$f_{max} = f\left|G_{u(max)}\right| \quad (33)$$

7.4.3.11.7 Specified conditions

– Ambient or reference-point temperature
– Bias conditions

7.4.3.12 Transition frequency (f_T)

7.4.3.12.1 Purpose

To measure the transition frequency of a transistor under specified conditions.

7.4.3.12.2 Circuit diagram

See the block diagram of the circuit for the scattering parameter measurement which is shown in 7.4.3.10.

7.4.3.12.3 Circuit description and requirements

See the circuit description and requirements which are shown in 7.4.3.10.

7.4.3.12.4 Precautions to be observed

See the precautions to be observed which are shown in 7.4.3.10.

7.4.3.12.5 Measurement procedure

See the measurement procedure for the scattering parameters S_{11}, S_{12}, S_{21}, S_{22} which are shown in 7.4.3.10.

All four scattering parameters are measured and h_{21} is calculated by the equation of h_{21} which is shown in 3.5.3.2 of IEC 60747-7:2000.

The measurement frequency range shall include the frequency in which $|h_{21}|$ is decreasing at the rate of approximately 6dB/octave.

The appropriate frequency f shall be chosen in the frequency range in which $|h_{21}|$ decreases at the rate of approximately 6dB/octave.

The transition frequency is calculated as:

$$f_T = f|h_{21}| \tag{34}$$

7.4.3.12.6 Specified conditions

– Ambient or reference-point temperature
– Bias conditions

7.4.3.13 Frequency of unity current transfer ratio (f_1)

7.4.3.13.1 Purpose

To measure the frequency of unity current transfer ratio of a transistor under specified conditions.

7.4.3.13.2 Circuit diagram

See the block diagram of the circuit for the scattering parameter measurement which is shown in 6.1.13.6 of IEC 60747-7:2000.

7.4.3.13.3 Circuit description and requirements

See the circuit description and requirements which are shown in 7.4.3.10.

7.4.3.13.4 Precautions to be observed

See the precautions to be observed which are shown in 7.4.3.10.

7.4.3.13.5 Principle of measurement

All four scattering parameters are measured and h_{21} is calculated by the equation of h_{21} which is shown in 3.5.3.2 of IEC 60747-7:2000.

The frequency of unity current transfer ratio (f_1) is the frequency where $|h_{21}|$ is equal to 1.

7.4.3.13.6 Measurement procedure

See the measurement procedure for the scattering parameters S_{11}, S_{12}, S_{21}, S_{22} which are shown in 7.4.3.10.

h_{21} is calculated using the measured scattering parameters.

The frequency of unity current transfer ratio (f_1) is the frequency where $|h_{21}|$ is equal to 1.

If the frequency of unity current transfer ratio (f_1) exceeds the measurement frequency, f_1 is not available.

7.4.3.13.7 Specified conditions

– Ambient or reference-point temperature.
– Bias conditions.

7.4.3.14 Maximum available gain ($G_{a(max)}$)

7.4.3.14.1 Purpose

To measure the maximum available gain of a transistor under specified conditions.

7.4.3.14.2 Circuit diagram

See the block diagram of the circuit for the scattering parameter measurement which is shown in 7.4.3.10.

7.4.3.14.3 Circuit description and requirements

See the circuit description and requirements which are shown in 7.4.3.10.

7.4.3.14.4 Precautions to be observed

See the precautions to be observed which are shown in 7.4.3.10.

7.4.3.14.5 Measurement procedure

See the measurement procedure for the scattering parameters S_{11}, S_{12}, S_{21}, S_{22} which are shown in 7.4.3.10.

The maximum available gain is calculated using the following equations:

$$G_{a(max)} = \left|S_{21} / S_{12}\right| \times \left(K - \sqrt{K^2 - 1}\right) \qquad \text{for } K \geq 1 \tag{35}$$

where K is a stability factor which is given by:

$$K = \frac{1 + \left|S_{11} S_{22} - S_{12} S_{21}\right|^2 - \left|S_{11}\right|^2 - \left|S_{22}\right|^2}{2\left|S_{21} S_{12}\right|} \tag{36}$$

If the stability factor (K) is less than 1, $G_{a(max)}$ is not available.

7.4.3.14.6 Specified conditions

– Ambient or reference-point temperature
– Measurement frequency
– Bias conditions

7.4.3.15 Maximum stable gain (G_{ms})

7.4.3.15.1 Purpose

To measure the maximum stable gain of a transistor under specified conditions.

7.4.3.15.2 Circuit diagram

See the block diagram of the circuit for the scattering parameter measurement which is shown in 7.4.3.10.

7.4.3.15.3 Circuit description and requirements

See the circuit description and requirements which are shown in 7.4.3.10.

7.4.3.15.4 Precautions to be observed

See the precautions to be observed which are shown in 7.4.3.10.

7.4.3.15.5 Measurement procedure

See the measurement procedure for the scattering parameters S_{12}, S_{21} which are shown in 7.4.3.10

The maximum stable gain is calculated using the following equation:

$$G_{ms} = \frac{|S_{21}|}{|S_{12}|} \tag{37}$$

7.4.3.15.6 Specified conditions

- Ambient or reference-point temperature
- Measurement frequency
- Bias conditions

7.4.3.16 Insertion power gain ($|S_{21e}|^2$)

7.4.3.16.1 Purpose

To measure the common-emitter insertion power gain of a transistor under specified conditions.

7.4.3.16.2 Circuit diagram

See the block diagram of the circuit for the scattering parameter measurement which is shown in 7.4.3.10.

7.4.3.16.3 Circuit description and requirements

See the circuit description and requirements which are shown in 7.4.3.10. The transistor under test should be placed in the common-emitter configuration.

7.4.3.16.4 Precautions to be observed

See the precautions to be observed which are shown in 7.4.3.10.

7.4.3.16.5 Measurement procedure

See the measurement procedure for the scattering parameter S_{21} which are shown in 7.4.3.10.

The insertion power gain is calculated using the following equation:

$$|S_{21e}|^2 = |S_{21}|^2 \tag{38}$$

7.4.3.16.6 Specified conditions

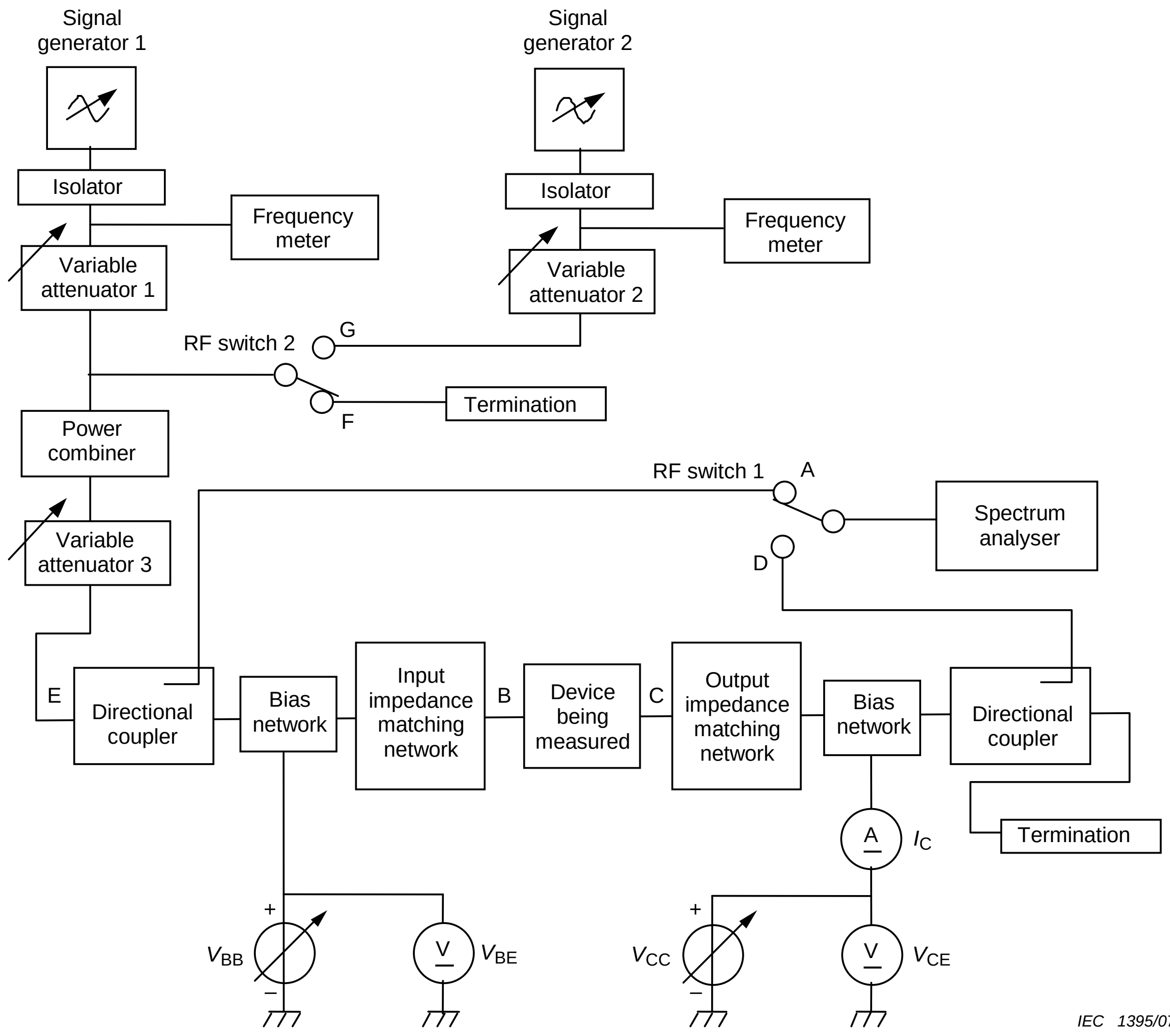
- Ambient or reference-point temperature
- Measurement frequency
- Bias conditions

7.4.3.17 Intermodulation distortion (two-tone) (P_1/P_n)

7.4.3.17.1 Purpose

To measure the intermodulation distortion under specified conditions.

7.4.3.17.2 Circuit diagram



NOTE Constant base current, in Table 5, may be used as the base bias supply.

Figure 42 – Circuit for the measurement of two-tone intermodulation distortion

7.4.3.17.3 Principle of measurement

In the circuit diagram shown in Figure 42 the input power P_i, the output powers P_o and P_n of the device being measured are derived from the following equations:

$$P_i = P_a + L_1 \tag{39}$$

$$P_o = P_b + L_2 \tag{40}$$

$$P_n = P_c + L_2 \tag{41}$$

where

P_o and P_n are the output powers of the input signal and the intermodulation products, respectively;

P_a, P_b and P_c are the values indicated by the spectrum analyser and corresponding to P_i, P_o and P_n, respectively.

L_1 is the difference between the loss L_A and L_B where L_A is the loss from point E to point A and L_B is the loss from point E to point B shown in Figure 42, respectively. L_2 is the circuit loss from point C to point D shown in Figure 42. P_i, P_o, P_n, P_a, P_b and P_c are expressed in dBm. L_1 and L_2 are expressed in decibels.

The intermodulation distortion, P_1/P_n, which is expressed in dBc, is derived from equations (40) and (41) as follows:

$$P_1/P_n = P_o - P_n = P_b - P_c \tag{42}$$

7.4.3.17.4 Circuit description and requirements

The purpose of the isolator is to enable the power level to the device being measured to be kept constant irrespective of impedance mismatches at its input.

The circuit losses L_1 and L_2 should be measured beforehand.

The variable attenuator 3 can be eliminated.

7.4.3.17.5 Precautions to be observed

Oscillation, which is checked by a spectrum analyser, should be eliminated during these measurements. The termination shall be capable of handling the power fed.

Harmonics or spurious responses of the signal generator should be reduced to negligible.

It is better to terminate port D, when the switch is connected to position A, and vice versa.

7.4.3.17.6 Measurement procedure

The bias under specified conditions is applied.

The RF switch 1 is connected to position A.

The RF switch 2 is connected to position F.

The signal generator 1 is turned on and the frequency of the input signal is set to the specified value f_1.

The input signal is applied to the device being measured with the specified level P_i using the spectrum analyser and the variable attenuator 1.

The RF switch 1 is connected to position D.

The input and output impedance matching networks are adjusted using the spectrum analyser so that the fundamental output power shows the maximum value.

The RF switch 1 is connected to position A.

The RF switch 2 is connected to position G.

The signal generator 2 is turned on and the frequency of the input signal is set to the specified value f_2.

The input signal is added to the device being measured with the same level as the fundamental signal using the spectrum analyser and the variable attenuator 2.

The RF switch 1 is connected to position D.

The output powers P_b and P_c in dB of the input signals and the specified intermodulation products, i.e. second order, third order etc., are measured using the spectrum analyser (see Figure 43).

The intermodulation distortion on the specified input power P_i is derived from equations (39) to (42).

If the intermodulation products output powers are different between upper side and lower side, larger value should be used.

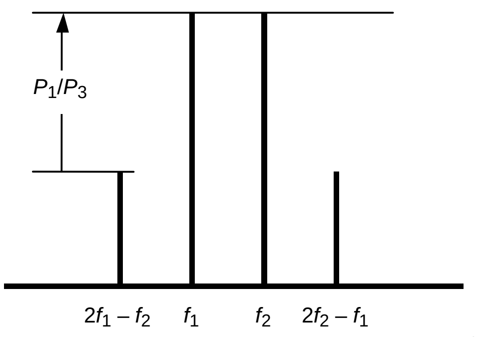


Figure 43 – Example of third order intermodulation products indicated by the spectrum analyser

7.4.3.17.7 Specified conditions

– Ambient or reference-point temperature
– Bias conditions
– Type of base bias supply
– Frequency of input signals, f_1 and f_2
– Input power
– Spurious intensity
– Order of intermodulation distortion

7.4.3.18 Input power at the intercept point (for intermodulation products) ($P_{i,n(IP)}$) and output power at the intercept point (for intermodulation products) ($P_{o,n(IP)}$)

7.4.3.18.1 Purpose

To measure the input power at the intercept point for intermodulation products and the output power at the intercept point for intermodulation products under specified conditions.

7.4.3.18.2 Circuit diagram

See the circuit diagram of 7.4.3.17.2.

7.4.3.18.3 Principle of measurement

Refer the principle of measurements of 7.4.3.17.3.

7.4.3.18.4 Circuit description and requirements

See the circuit description and requirements of 7.4.3.17.4.

7.4.3.18.5 Precautions to be observed

See the precautions to be observed of 7.4.3.17.5.

7.4.3.18.6 Measurement procedure

The bias under specified conditions is applied.

The RF switch 1 is connected to position A.

The RF switch 2 is connected to position F.

The signal generator 1 is turned on and the frequency of the input signal is set to the specified value f_1.

The input signal is applied to the device being measured with the specified level P_i using the spectrum analyser and the variable attenuator 1.

The RF switch 1 is connected to position D.

The input and output impedance matching networks are adjusted using the spectrum analyser so that the fundamental output power shows the maximum value.

The RF switch 1 is connected to position A.

The RF switch 2 is connected to position G.

The signal generator 2 is turned on and the frequency of the input signal is set to the specified value f_2.

The input signal is added to the device being measured with the same level as the fundamental signal using the spectrum analyser and the variable attenuator 2.

The RF switch 1 is connected to position D.

The output powers P_b and P_c in dB of the input signals and the specified intermodulation products, i.e. second order, third order, etc., are measured using the spectrum analyser (see Figure 43).

Changing the power level of the input signals using the variable attenuator 3, the above procedure is repeated within the specified range.

The data obtained are plotted on linear scales.

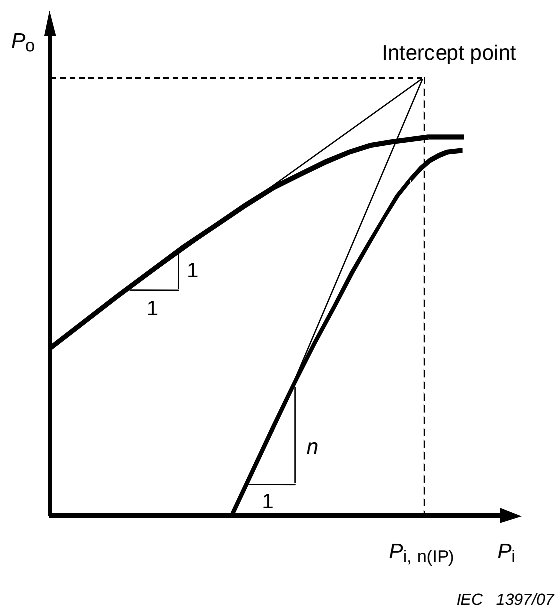
If the intermodulation products output powers are different between upper side and lower side, larger value should be used.

The straight lines of the fundamental signal in the linear region (slope of 1) and the intermodulation products in the linear region (slope of *n*) are extended.

The input power and the output power at the intercept point of the two extended lines are the input power at the intercept point (for the intermodulation products) and the output power at the intercept point (for the intermodulation products) under the specified conditions (see Figure 44).

7.4.3.18.7 Specified conditions

- Ambient or reference-point temperature
- Bias conditions
- Type of base bias supply
- Frequency of input signals, f_1 and f_2
- Input power in linear region
- Spurious intensity
- Order of intermodulation distortion



Key

n order of intermodulation distortion

Figure 44 – Typical intermodulation products output power characteristic

7.5 Verifying methods

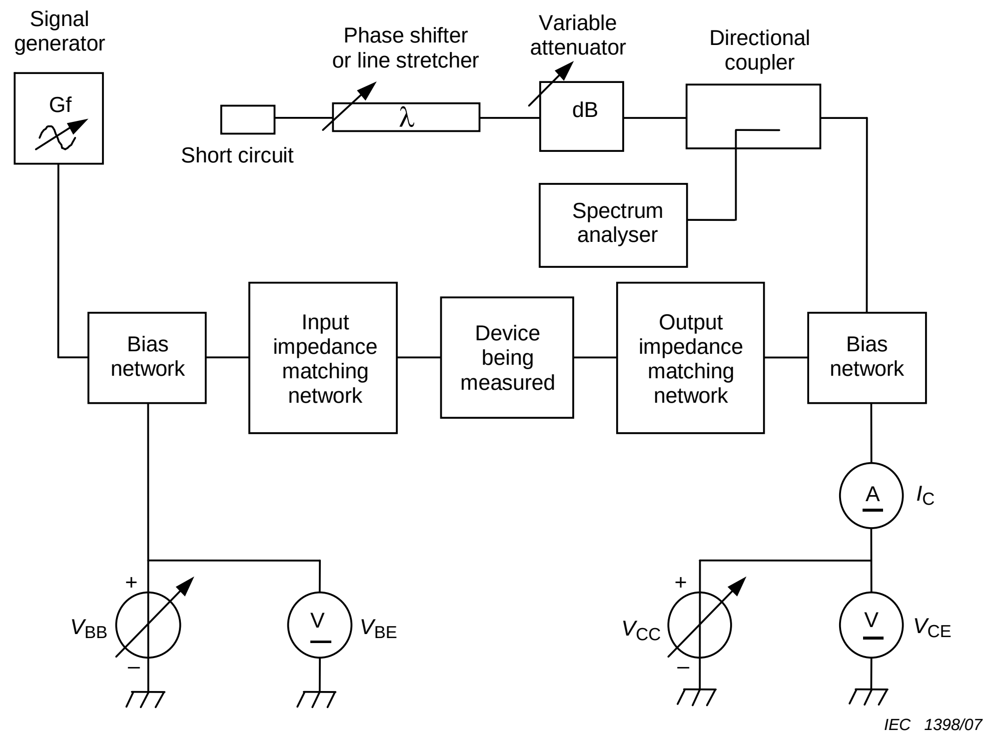
7.5.1 Load mismatch tolerance (Ψ_L)

7.5.1.1 Purpose

To verify the load mismatch tolerance under specified conditions.

7.5.1.2 Verifying method 1 (spurious intensity)

7.5.1.2.1 Circuit diagram



NOTE Constant base current, in Table 5, may be used as the base bias supply.

Figure 45 – Circuit for the verification of load mismatch tolerance in method 1

7.5.1.2.2 Circuit description and requirements

The signal generator shall be capable of operating within specified frequency-band. The signal generator shall have stable characteristics above noise floor with no oscillation and no spurious intensity. The noise floor should be smaller than specified level from the output power. The signal generator must generate specified modulation signal.

The spectrum analyser shall be capable of operating within a specified frequency range for checking that there is no oscillation nor any spurious intensity. The spectrum analyser shall have a specified dynamic range.

The phase shifter shall be capable of keeping the load VSWR or the magnitude of the load reflection coefficient. The line stretcher is suitable for this purpose. The output port of the phase shifter should be shorted.

The purpose of the variable attenuator is to realize the specified VSWR.

7.5.1.2.3 Precautions to be observed

Noise floor or spurious responses of the signal generator should be reduced to be negligible at the VSWR less than the specified one.

The VSWR shall be kept constant during all phase conditions of the phase shifter.

7.5.1.2.4 Test procedure

The load VSWR is set to the specified value by adjusting variable attenuator.

The frequency of the signal generator is set to the specified value.

The modulation of the signal generator is set to the specified condition.

The bias under the specified condition is supplied.

The power level of the signal generator is set to the specified value.

The phase angle is swept continuously by moving the length of the line stretcher.

Ensure that there is no oscillation or spurious intensity less than that specified by using the spectrum analyser at all phase angles.

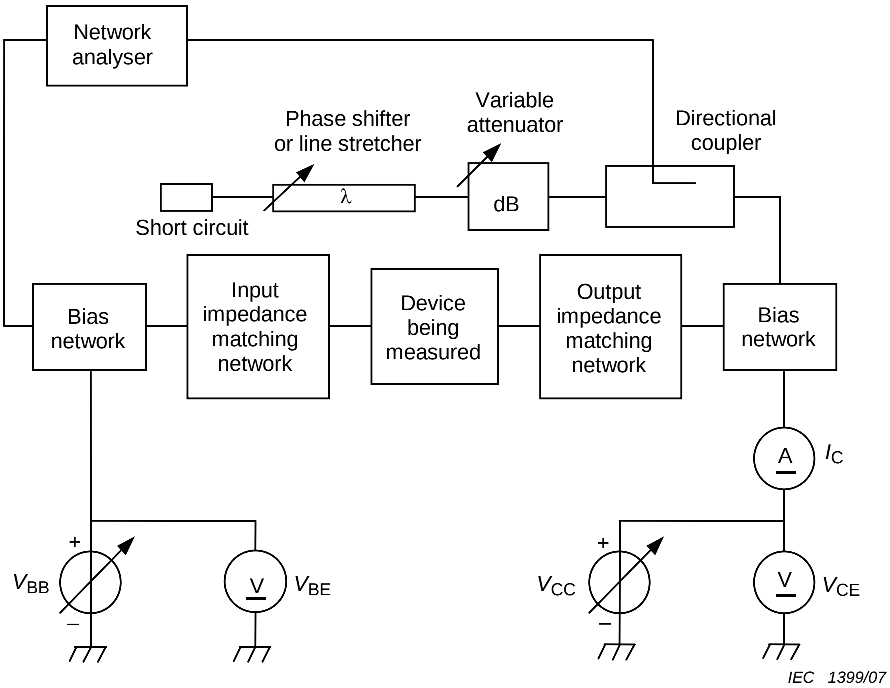
NOTE Instead of the line stretcher, the slide screw tuner can be used. An automatic stub-tuner or an electronic tuner is also used to enable the specified VSWR for convenience. The demerit of the tuners is that phase condition is discrete and cannot be swept continuously.

7.5.1.2.5 Specified conditions

– Ambient or reference-point temperature
– Load VSWR
– Bias conditions
– Frequency of the input signal
– Modulation of the input signal
– Input power
– Spurious

7.5.1.3 Verifying method 2 (no discontinuity of the frequency response)

7.5.1.3.1 Circuit diagram



NOTE Constant base current, in Table 5, may be used as the base bias supply.

Figure 46 – Circuit for the verification of load mismatch tolerance in method 2

7.5.1.3.2 Circuit description and requirements

The network analyser shall be capable of operating within the specified frequency-band.

The phase shifter shall be capable of keeping the load VSWR or the magnitude of the load reflection coefficient. The line stretcher is suitable for this purpose. The output port of the phase shifter should be shorted.

The purpose of the variable attenuator is to realize the specified VSWR.

7.5.1.3.3 Precautions to be observed

The VSWR shall be kept constant at all phase conditions of the phase shifter.

7.5.1.3.4 Test procedure

The load VSWR is set to the specified value by adjusting variable attenuator.

The sweep frequency range of the network analyser is set to the specified value.

The power level of the network analyser is set to the specified value.

The bias under specified condition is supplied.

The phase angle is swept continuously by moving the length of the line stretcher.

No discontinuity of the frequency response is confirmed by using the network analyser at all phase angles.

7.5.1.3.5 Specified conditions

– Ambient or reference-point temperature
– Load VSWR
– Bias conditions
– Frequency range of the input signal
– Input power

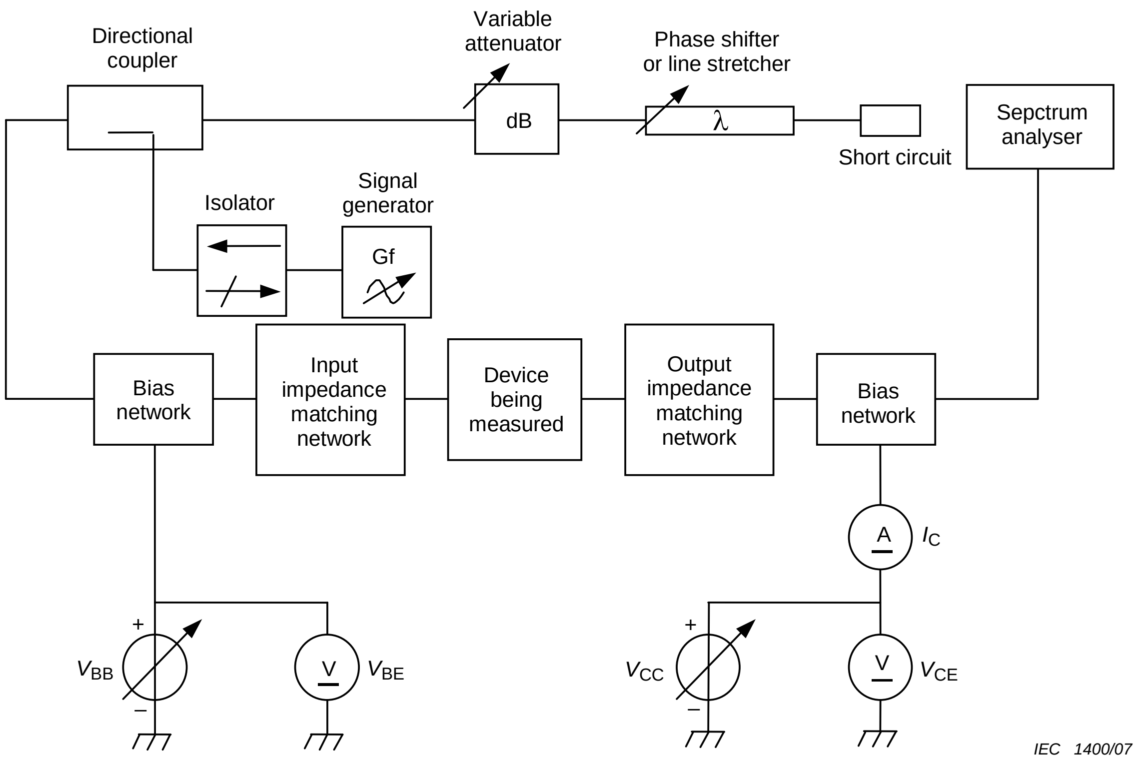
7.5.2 Source mismatch tolerance (Ψ_S)

7.5.2.1 Purpose

To verify the source mismatch tolerance under specified conditions.

7.5.2.2 Verifying method 1 (spurious intensity)

7.5.2.2.1 Circuit diagram



NOTE Constant base current, in Table 5, may be used as the base bias supply.

Figure 47 – Circuit for the verification of source mismatch tolerance in method 1

7.5.2.2.2 Circuit description and requirements

See the circuit description and requirements of 7.5.1.2.2.

The purpose of the isolator is to enable the power level to the device being measured to be kept constant irrespective of impedance mismatches at its input.

7.5.2.2.3 Precautions to be observed

See the precautions to be observed in 7.5.1.2.3.

7.5.2.2.4 Test procedure

The source VSWR is set to the specified value by adjusting variable attenuator.

The frequency of the signal generator is set to the specified value.

The modulation of the signal generator is set to the specified condition.

The bias under specified condition is supplied.

The power level of the signal generator is set to the specified value.

The phase angle is swept continuously by moving the length of the line stretcher.

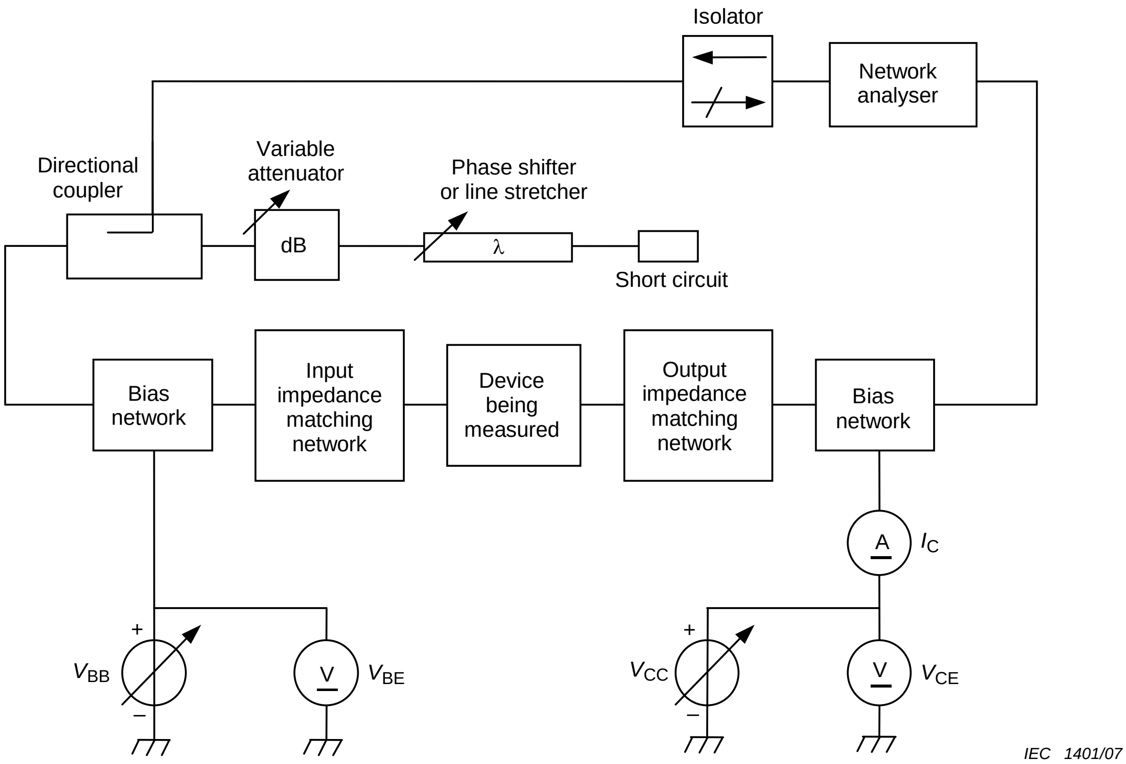
No oscillation or spurious intensity less than the specified condition is confirmed by using the spectrum analyser at all phase angles.

7.5.2.2.5 Specified conditions

– Ambient or reference-point temperature
– Source VSWR
– Bias conditions
– Frequency of the input signal
– Modulation of the input signal
– Input power
– Spurious intensity

7.5.2.3 Verifying method 2 (no discontinuity of the frequency response)

7.5.2.3.1 Circuit diagram



NOTE Constant base current, in Table 5, may be used as the base bias supply.

Figure 48 – Circuit for the verification of source mismatch tolerance in method 2

7.5.2.3.2 Circuit description and requirements

See the circuit description and requirements in 7.5.1.3.2.

The purpose of the isolator is to enable the power level to the device being measured to be kept constant irrespective of impedance mismatches at its input.

7.5.2.3.3 Precautions to be observed

See the precautions to be observed of 7.5.1.3.3.

7.5.2.3.4 Test procedure

The source VSWR is set to the specified value by adjusting variable attenuator.

The sweep frequency range of the network analyser is set to the specified value.

The power level of the network analyser is set to the specified value.

The bias under specified condition is supplied.

The phase angle is swept continuously by moving the length of the line stretcher.

No discontinuity of the frequency response is confirmed by using the network analyser at all phase angles.

7.5.2.3.5 Specified conditions

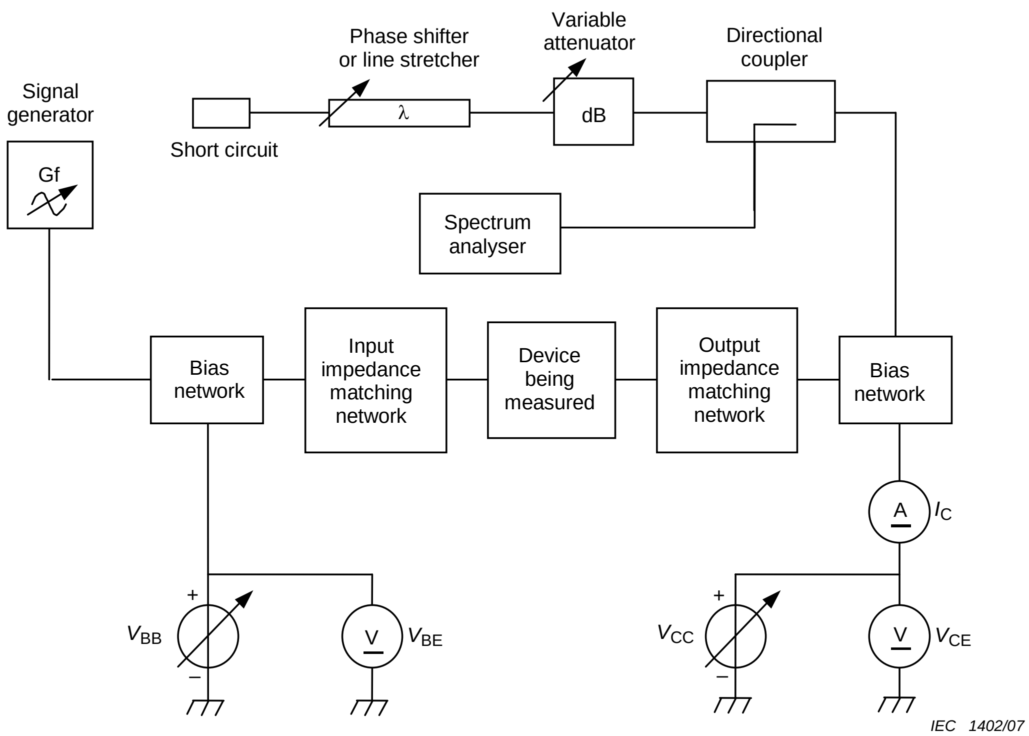
– Ambient or reference-point temperature
– Source VSWR
– Bias conditions
– Frequency range of the input signal
– Input power

7.5.3 Load mismatch ruggedness (Ψ_R)

7.5.3.1 Purpose

To verify the load mismatch ruggedness under specified conditions.

7.5.3.2 Circuit diagram



NOTE Constant base current, in Table 5, may be used as the base bias supply.

Figure 49 – Circuit for the verification of load mismatch ruggedness

7.5.3.3 Circuit description and requirements

The signal generator shall be capable of operating within the specified frequency band.

The spectrum analyser shall be capable of operating within the specified frequency range for checking that there is no oscillation nor any spurious intensity.

The phase shifter shall be capable of keeping the load VSWR or the magnitude of the load reflection coefficient. The line stretcher is suitable for this purpose. The output port of the phase shifter should be shorted.

The purpose of the variable attenuator is to realize the specified VSWR.

7.5.3.4 Precautions to be observed

No oscillation influenced by the verification system should be confirmed for any phase conditions of the phase shifter at the VSWR less than the specified one.

The VSWR shall be kept constant during all phase conditions of the phase shifter.

7.5.3.5 Test procedure

DC and RF characteristics are measured under specified conditions before the following load mismatch test procedure.

The load VSWR is set to the specified value by adjusting variable attenuator.

The frequency of the signal generator is set to the specified value.

The bias under specified condition is supplied.

The power level of the signal generator is set to the specified value.

The phase angle is swept continuously by moving the length of the line stretcher.

The device is kept in operation during the specified operation time at all phase angles.

DC and RF characteristics are measured under specified condition once more.

Load mismatch ruggedness is verified using specified degradation criteria of DC and RF characteristics.

7.5.3.6 Specified conditions

– Ambient or reference-point temperature
– Load VSWR
– Bias conditions
– Frequency of the input signal
– Input power
– Operation time
– Degradation criteria of DC and RF characteristics
– Measurement conditions of DC and RF characteristics

8 Field-effect transistors

8.1 General

This clause provides terms and definitions, essential ratings and characteristics, measuring methods, and verifying methods for field-effect transistors used in microwave applications. For general items of field-effect transistors, refer to IEC 60747-8:2000.

8.2 Terms and definitions

For the purposes of this clause, the following terms and definitions apply.

NOTE This subclause contains the main specific letter symbols in microwave applications.

8.2.1
gate-source cut-off voltage
V_{GSoff}
See 4.3.3 of 60747-8:2000.

8.2.2
gate-source breakdown voltage
$V_{(BR)GSO}$
reverse voltage measured with the drain electrode open at which the gate-source current becomes greater than a specified value

8.2.3
gate-drain breakdown voltage
$V_{(BR)GDO}$
reverse voltage measured with the source electrode open at which the gate-drain current becomes greater than a specified value

8.2.4
channel-case thermal resistance
$R_{th(ch-c)}$
quotient of the difference between the virtual temperature of the channel region and the temperature of the case, and the steady-state power dissipation in the field-effect transistor

8.2.5
maximum frequency of oscillation
f_{max}
See 3.4.14 of IEC 60747-7:2000.

8.2.6
transition frequency
f_T
See 3.4.15 of IEC 60747-7:2000.

8.2.7
frequency of unity current transfer ratio
f_1
See 3.4.16 of IEC 60747-7:2000

8.2.8
maximum available gain
$G_{a(max)}$
ratio of the power delivered to load to the power input to network when the input and output ports are simultaneously conjugately matched to source and load impedances, respectively

NOTE The abbreviation "*MAG*" is still in common use for maximum available gain.

8.2.9
maximum stable gain
G_{ms}
See 7.2.15.

NOTE The abbreviation "*MSG*" is still in common use for maximum stable gain.

8.2.10
insertion power gain
$|S_{21}|^2$
magnitude squared of S_{21} in common source configuration, the output terminating resistance and the source resistance each having the value R_0

8.2.11
output power
P_o
See 3.3 of IEC 60747-16-2:2002.

8.2.12
power-gain compression
ratio of the magnitude of the power gain at a reference signal level to its magnitude at a specified higher signal level

NOTE The power-gain compression is usually expressed in dB.

8.2.13
output power at 1 dB gain compression
$P_{o(1dB)}$
output power where the gain decreases by 1 dB compared with the linear gain

8.2.14
power gain at 1 dB gain compression
$G_{p(1dB)}$
ratio of the power delivered to load to the power input to network at a gain compression of 1 dB

8.2.15
power-added efficiency
η_{add}
ratio of the difference between the output power and the input signal power to the DC input power

NOTE This ratio is normally expressed as a percentage.

8.2.16
drain efficiency
η_d
ratio of the output power to the DC input power of drain

NOTE This ratio is normally expressed as a percentage.

8.2.17
input power at the intercept point (for intermodulation products)
$P_{i,n(IP)}$
See 7.2.17.

8.2.18
output power at the intercept point (for intermodulation products)
$P_{o,n(IP)}$
See 7.2.18.

8.2.19
intermodulation distortion
P_1/P_n
See 3.7 of IEC60747-16-1:2007.

NOTE The abbreviation "IMD_n" is in common use for the *n*th order intermodulation distortion.

8.2.20
noise figure
F
See 702-08-57 of IEC 60050-702:1992.

8.2.21
minimum noise figure
F_{min}
minimum value of the noise figure that can be obtained through adjustment of the source impedance under specified bias condition and a specified frequency

8.2.22
source reflection factor for minimum noise figure
r_{GFmin}
source reflection factor that gives minimum noise figure

NOTE 1 For source reflection coefficient (factor), see 3.5.3.3 of IEC 60747-7:2000.

NOTE 2 The symbol "Γ_{opt}" is still in common use for the source reflection factor for minimum noise figure.

8.2.23
associated gain
G_{as}
power gain when the device is matched (for example, by means of an external network)

NOTE 1 The gain is normally given under conditions matched for minimum noise. In this case the "associated gain for minimum noise" should be used.

NOTE 2 The shorter term "associated gain" may be used if no ambiguity is likely to occur.

8.2.24
equivalent input noise resistance
R_n
quotient of the equivalent input noise voltage and the equivalent input noise current (see 3.4.5 and 3.4.6 of IEC 60747-1:2006)

8.2.25
load mismatch tolerance
Ψ_L
See 7.2.20.

8.2.26
source mismatch tolerance
Ψ_S
See 7.2.21.

8.2.27
load mismatch ruggedness
Ψ_R
See 7.2.22.

8.3 Essential ratings and characteristics

8.3.1 General

This subclause gives ratings and characteristics required for specifying microwave field-effect transistors.

Microwave field-effect transistors are divided into two categories:

– category A: power devices;
– category B: small power signal devices.

8.3.2 Limiting values (absolute maximum rating system)

8.3.2.1 Electrical limiting values

Limiting values shall be specified as follows:

Table 7 – Electrical limiting values

Subclause	Parameters	Min.	Max.
8.3.2.1.1	Drain-source voltage, V_{DS}		+
8.3.2.1.2	Gate-source voltage, V_{GS}		+
8.3.2.1.3	Drain current, I_D		+
8.3.2.1.4	Storage temperature, T_{stg}	+	+
8.3.2.1.5	Channel temperature, T_{ch}		+
8.3.2.1.6	Either total power dissipation, P_{tot} or drain power dissipation, P_d		+

8.3.2.2 Characteristics

Characteristics are to be given at 25 °C except where otherwise stated.

8.3.2.2.1 DC characteristics

The parameters shall be specified corresponding to categories as shown in Table 8 below:

Table 8 – DC characteristics

Subclause	Parameters	Min.	Typ.	Max.	Categories	
					A	B
8.3.2.2.1.1	Drain current with gate short-circuited to source, I_{DSS}	+		+	+	+
8.3.2.2.1.2	Gate current with drain short-circuited to source, I_{GSS}			+	+	+
8.3.2.2.1.3	Gate-source cut-off voltage, V_{GSoff}	+		+	+	+
8.3.2.2.1.4	Gate-source breakdown voltage, $V_{(BR)GSO}$	+			+	+
8.3.2.2.1.5	Gate-drain breakdown voltage, $V_{(BR)GDO}$	+			+	+
8.3.2.2.1.6	Channel-case thermal resistance, $R_{th(ch-c)}$			+	+	

8.3.2.2.2 RF characteristics

The parameters shall be specified corresponding to categories as shown in Table 9 below:

Table 9 – RF characteristics

Subclause	Parameters	Min.	Type	Max.	Categories A	Categories B
8.3.2.2.1	Maximum frequency of oscillation, f_{max} (where appropriate)		+		+	+
8.3.2.2.2	Transition frequency, f_T (where appropriate)		+		+	+
8.3.2.2.3	Frequency of unity current transfer ratio, f_1 (where appropriate)		+		+	+
8.3.2.2.4	Either maximum available gain, $G_{a(max)}$ or maximum stable gain, G_{ms} or insertion power gain, $\|S_{21}\|^2$	+	+		+	
8.3.2.2.5	Either output power at 1 dB gain compression, $P_{o(1dB)}$ or output power at specified input power, P_o	+	+		+	
8.3.2.2.6	Power gain at 1 dB gain compression, $G_{p(1dB)}$	+	+		+	
8.3.2.2.7	Power-added efficiency, η_{add}	+	+		+	
8.3.2.2.8	Drain efficiency, η_d	+	+		+	
8.3.2.2.9	Either input power at the intercept point (for intermodulation products), $P_{i,n(IP)}$ or output power at the intercept point (for intermodulation products), $P_{o,n(IP)}$ or Intermodulation distortion, P_1/P_n (where appropriate)	+	+		+	
8.3.2.2.10	Noise figure, F		+	+		+
8.3.2.2.11	Minimum noise figure, F_{min} (where appropriate)		+	+		+
8.3.2.2.12	Source reflection factor for minimum noise figure, r_{GFmin} (where appropriate)		+			+
8.3.2.2.13	Associated gain, G_{as}	+	+			+
8.3.2.2.14	Equivalent input noise resistance, R_n (where appropriate)		+			+
8.3.2.2.15	Load mismatch tolerance, Ψ_L (where appropriate)			+	+	
8.3.2.2.16	Source mismatch tolerance, Ψ_S (where appropriate)			+	+	
8.3.2.2.17	Load mismatch ruggedness, Ψ_R (where appropriate)			+	+	

8.4 Measuring methods

8.4.1 General

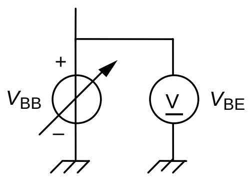
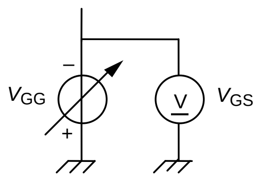
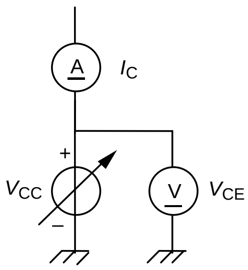
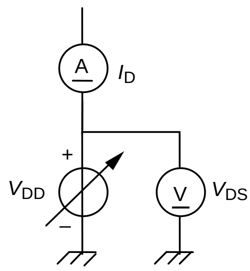
The measuring methods of bipolar transistor are applicable, with the terms and symbols of field-effect transistor in 8.4.3.8, 8.4.3.9, 8.4.3.10, 8.4.3.11, 8.4.3.12, 8.4.3.13, 8.4.3.14, 8.4.3.15 and 8.4.3.16 being replaced with the replacing rules shown in Table 10 or 11.

NOTE The replacing rules shown in Table 10 and 11 are also applicable to the verifying methods of field-effect transistor in 8.5.

Table 10 – Replacing rules for terms

Terms to be replaced	Terms to be replaced by
Bipolar transistor	Field-effect transistor
Base-emitter voltage, V_{BE}	Gate-source voltage, V_{GS}
Base-emitter cut-off current	Gate-source cut-off voltage
Collector-emitter voltage, V_{CE}	Drain-source voltage, V_{DS}
Collector current, I_C	Drain current, I_D

Table 11 – Replacing rules for symbols

Symbols to be replaced	Symbols to be replaced by
	
	

8.4.2 DC characteristics

8.4.2.1 Drain current, with gate short-circuited to source (I_{DSS})

The measuring method given in 6.3 of IEC 60747-8:2000 is applicable.

8.4.2.2 Gate current with drain short-circuited to source, (I_{GSS})

The measuring method given in 6.2 of IEC 60747-8:2000 is applicable, with the drain voltage short-circuited to the source.

8.4.2.3 Gate-source cut-off voltage (V_{GSoff})

The measuring method given in 6.5 of IEC 60747-8:2000 is applicable, with the sentence in 6.5.3 being replaced by "See general precautions".

8.4.2.4 Gate-source breakdown voltage ($V_{(BR)GSO}$)

8.4.2.4.1 Purpose

To measure the gate-source breakdown voltage of a field-effect transistor under specified conditions.

8.4.2.4.2 Circuit diagram

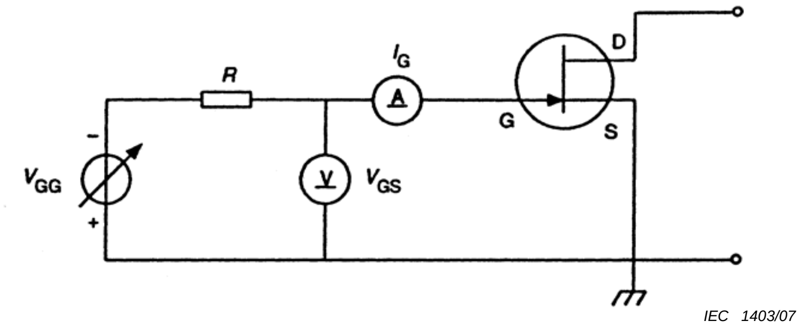


Figure 50 – Circuit for the measurement of gate-source breakdown voltage, $V_{(BR)GSO}$

8.4.2.4.3 Circuit description and requirements

R is a protective resistor.

8.4.2.4.4 Measurement procedure

The gate bias voltage, V_{GG}, is set to the appropriate value.

By varying the gate bias voltage, the gate current, I_G, is set to the specified value, I_{GG}.

The breakdown voltage, $V_{(BR)GSO}$ is the gate-source voltage, V_{GS}, measured at the specified gate current.

8.4.2.4.5 Specified conditions

– Ambient or point temperature, T_{amb} or T_{case}
– Specified gate current, I_{GG}

8.4.2.5 Gate-drain breakdown voltage ($V_{(BR)GDO}$)

8.4.2.5.1 Purpose

To measure the gate-drain breakdown voltage of a field-effect transistor under specified conditions.

8.4.2.5.2 Circuit diagram

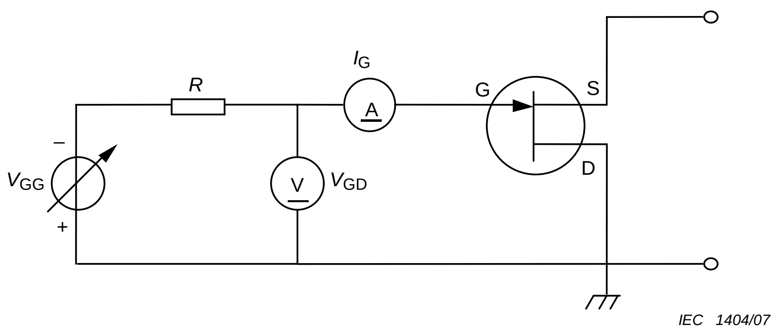


Figure 51 – Circuit for the measurement of gate-drain breakdown voltage, $V_{(BR)GDO}$

8.4.2.5.3 Circuit description and requirements

R is a protective resistor.

8.4.2.5.4 Measurement procedure

The gate bias voltage, V_{GG}, is set to the appropriate value.

By varying the gate bias voltage, the gate current, I_G, is set to the specified value, I_{GG}.

The breakdown voltage, $V_{(BR)GDO}$ is the gate-drain voltage, V_{GD}, measured at the specified gate current.

8.4.2.5.5 Specified conditions

– Ambient or point temperature, T_{amb} or T_{case}
– Specified gate current, I_{GG}

8.4.2.6 Thermal resistance, channel-to-case ($R_{th(ch-c)}$, $R_{th(j-c)}$)

8.4.2.6.1 Purpose

To measure the thermal resistance, channel-to-case of a field-effect transistor under specified conditions.

8.4.2.6.2 Circuit diagram

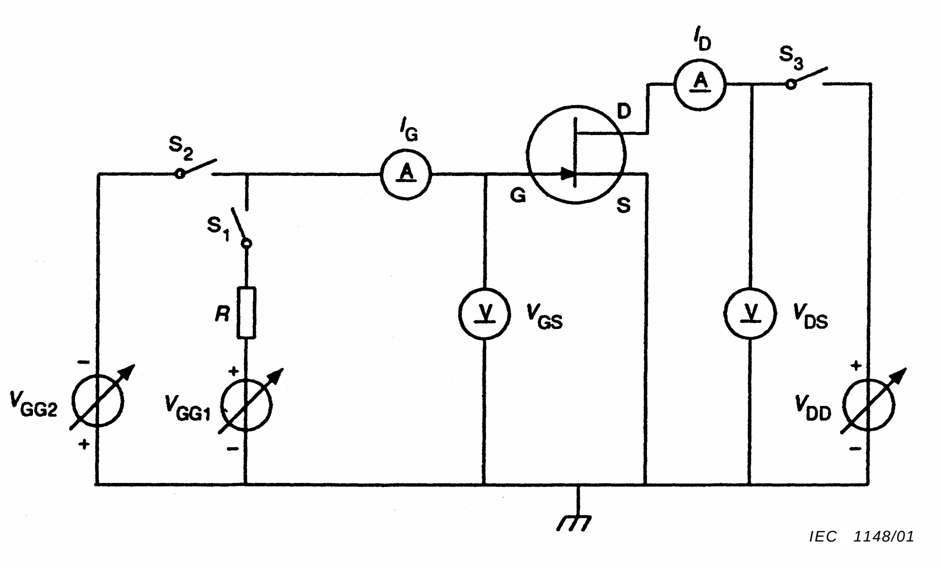


Figure 52 – Circuit for the measurement of thermal resistance, channel-to-case

8.4.2.6.3 Principle of measurement

The method uses the gate-source forward voltage, V_{GSF}, at open-circuit drain for a fixed reference gate forward current, $I_{G(ref)}$, as temperature-sensitive characteristic for the measurement of the (virtual) channel temperature, T_{ch}.

The method consists of two steps:

1) Establishment of the individual calibration curve, $V_{GSF} = f(T_{ch})$ which is then approximated by a straight line. During this calibration, T_{ch} is equivalent to T_{amb}.

   The slope of this line is the temperature coefficient, α, of the gate-source forward voltage:

$$\alpha = \frac{\Delta V_{GSF}}{\Delta T_{ch}} \quad (43) \text{ (see Figure 54)}$$

2) Application of a constant power dissipation, $V_{DS} \times I_D$, until thermal equilibrium is reached, and measurement of the resulting change (ΔV_{GSF}) of the gate-source forward voltage.

   The thermal resistance is then calculated as:

$$R_{th(ch-c)} = \frac{\Delta V_{GSF}}{\alpha} \times \frac{1}{V_{DS} \times I_D} \quad (44)$$

8.4.2.6.4 Circuit description and requirements

Resistor R shall be high enough, compared with the static gate-source input resistance, to ensure sufficient constancy of the reference gate forward current.

V_{GG1} and R may be replaced by a constant current source.

The timing of the DC pulses supplied to the device being measured is shown in Figure 53. The current passing through the voltmeter for V_{GS} shall be negligible compared with the forward gate current of the device being measured.

8.4.2.6.5 Precautions to be observed

See general precautions.

The specified case temperature of the device being measured shall be kept constant during the measurement.

8.4.2.6.6 Measurement procedure

The measurement consists of two parts, the measurement of the thermal coefficient and the measurement of thermal resistance, channel-to-case.

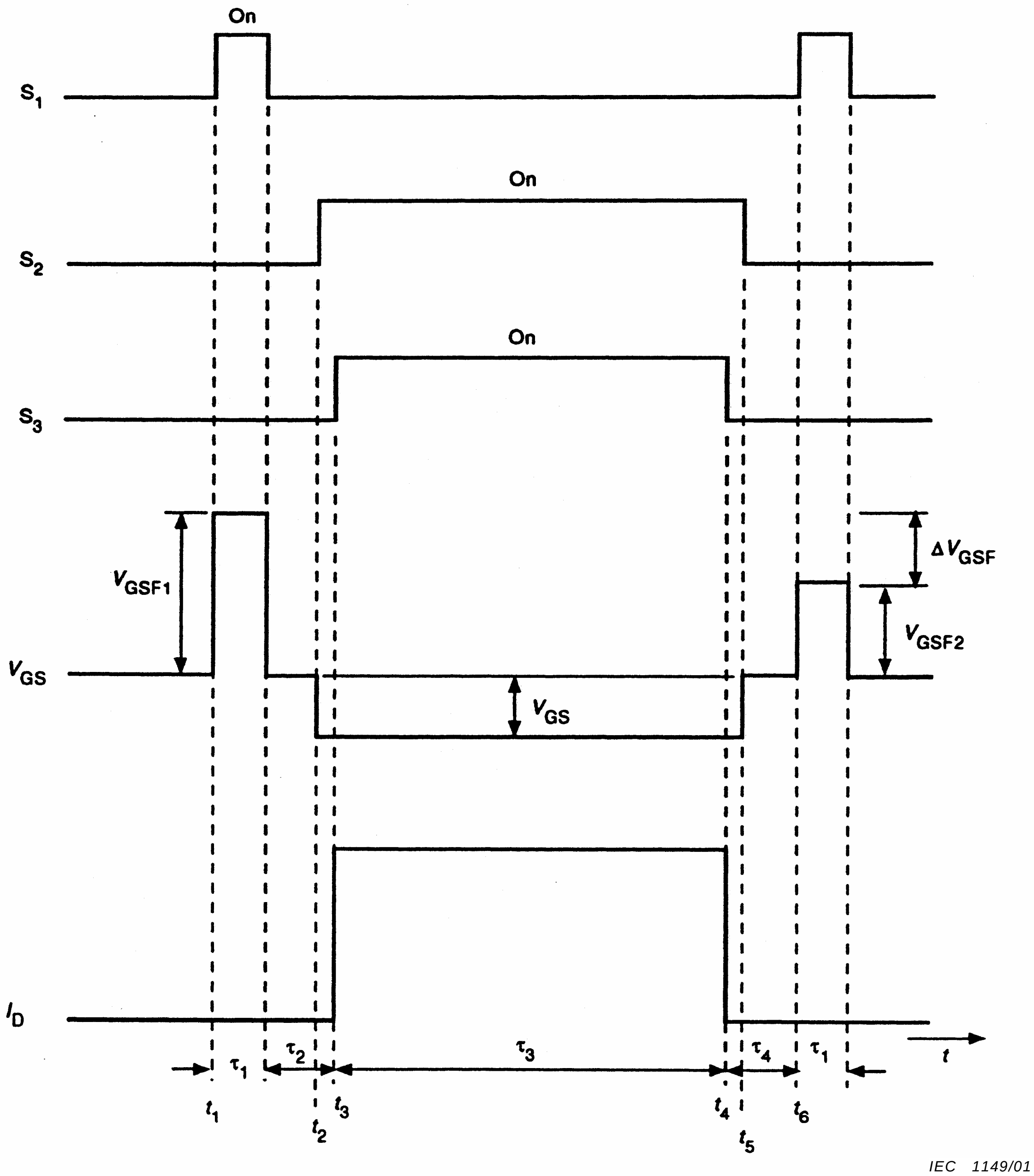


Figure 53 – Timing chart of DC pulse to be supplied to the device being measured

8.4.2.6.7 Measurement of α

The device being measured is mounted on a heating block (or placed in a heating oven) and connected to the circuit as shown in Figure 52. Switches S_2 and S_3 are off, switch S_1 is on.

R is adjusted such that the reference gate forward current, $I_{G(ref)}$, has a suitable value for the purpose of the measurements.

This value is not changed again during the whole measurement procedure for α and $R_{th(ch-c)}$. After this adjustment, switch S_1 is opened.

This calibration curve, $V_{GSF} = f(T_{ch})$, is established by measuring a sufficient number of points of the curve to allow for a sufficiently precise straight-line approximation. For this, in each point of the curve, the temperature of the heating block is set to a measured value, T_{ch}.

After thermal equilibrium is reached, switch S_1 is closed for the same time τ_1 as it will later be closed for at the measurement $R_{th(ch-c)}$. The gate-source forward voltage is recorded.

From the measured points the straight-line approximation is derived, and α is calculated from its slope (see Figure 54).

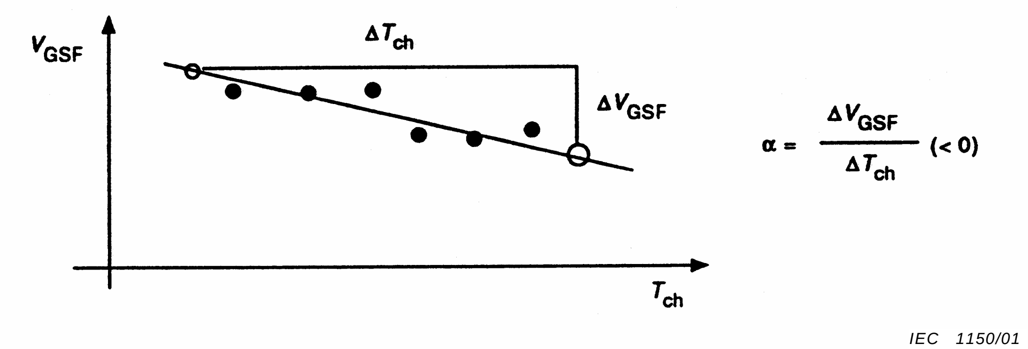


Figure 54 – Calibration curve $V_{GSF} = f(T_{ch})$ for fixed $I_{G(ref)}$, evaluation of α

8.4.2.6.8 Measurement of $R_{th(ch-c)}$

The device being measured is connected to the circuit as shown in Figure 52. The measurement begins after thermal equilibrium is reached. At $t = t_1$, S_1 is switched on for the period τ_1 and the gate-source forward voltage V_{GSF1} is recorded. After τ_1, S_1 is switched off.

Period τ_1 is usually selected to be several tens of microseconds.

At $t = t_2$, S_2 is switched on. At $t = t_3$, S_3 is switched on, and V_{GG2} is adjusted to reach the required value of I_D. V_{DS} and I_D are recorded.

After the period τ_3, when thermal equilibrium is reached, S_3 is switched off at $t = t_4$. Usually, several hundreds of milliseconds for τ_3 will be sufficient. S_2 is switched off at $t = t_5$, just after S_3 is switched off. At $t = t_6$, after the delay time τ_4 has elapsed with respect to delay time τ_4, S_1 is switched on again for the period τ_1 and the gate-source forward voltage, V_{GSF2}, is recorded.

The above-described measurement procedure is repeated for increasing values of τ_4, all other conditions being held constant. The resulting values for V_{GF2} are recorded.

These values are inserted in a graph showing $V_{GSF2} = f(\tau_4)$ (see Figure 55). From the graph, V_{GSF2}* can be extrapolated, which is V_{GSF2} for $\tau_4 = 0$. $R_{th(ch-c)}$ is calculated as:

$$R_{th(ch-c)} = \frac{V_{GSF2}{}^* - V_{GSF1}}{\alpha} \times \frac{1}{V_{DS} \times I_D} \quad (45)$$

Note that both(V_{GSF2}* – V_{GSF1}) and α have a negative value.

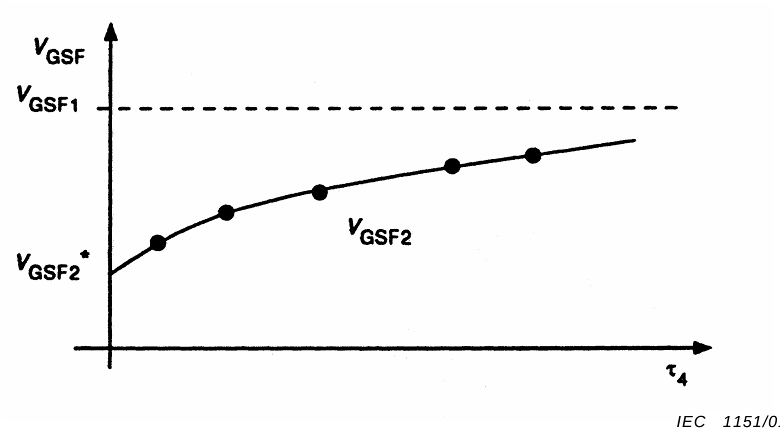


Figure 55 – V_{GSF2} in function of delay time τ_4

8.4.2.6.9 Specified conditions

- Ambient temperature: T_{amb}
- Case temperature: T_{case}
- Reference gate-source forward current: I_G
- Drain current: I_D
- Drain source voltage: V_{DS}

8.4.3 RF characteristics

8.4.3.1 Output power at specified input power (P_o)

8.4.3.1.1 Purpose

To measure the output power at specified input power of a field-effect transistor under specified conditions.

8.4.3.1.2 Circuit diagram

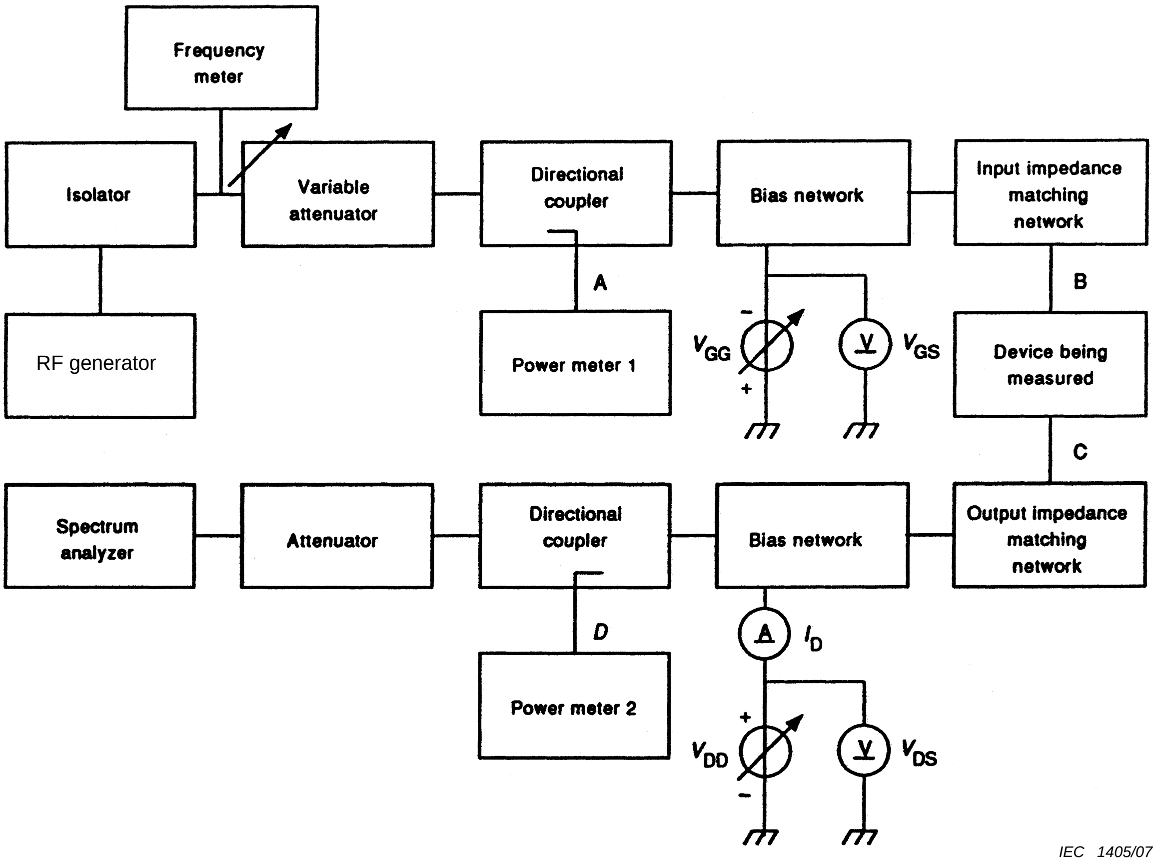


Figure 56 – Circuit for the measurement of output power at specified input power

8.4.3.1.3 Principle of measurements

In the circuit diagram shown in Figure 56, the input power, P_i, and the output power, P_o, of the device being measured are derived from the following equations:

$$P_i = P_1 - L_1 \quad (46)$$

$$P_o = P_2 - L_2 \quad (47)$$

where P_1 and P_2 are the values indicated by the power meters 1 and 2, respectively. L_1 and L_2 are the circuit losses from point A to point B and from point C to point D shown in Figure 56, respectively. P_i, P_o, P_1 and P_2 are expressed in dBm. L_1 and L_2 are expressed in dB.

Output power, P_o, at specified input power is derived from the equation (47).

8.4.3.1.4 Circuit description and requirements

The purpose of the isolator is to enable the power level to the device being measured to be kept constant irrespective of impedance mismatches at its input. The device being measured should be mounted on the test fixture having a good heat flow. The circuit losses L_1 and L_2 should be measured beforehand.

NOTE It is desirable in the measurement of L_1 and L_2 that the input and output-impedance network are tuned beforehand.

8.4.3.1.5 Precautions to be observed

There should not be any abrupt impedance change at the input and output matching circuits. Oscillation, which is checked by a spectrum analyzer, shall be eliminated during these measurements. Termination shall be capable of handling power supply.

8.4.3.1.6 Measurement procedure

The frequency of the RF generator is adjusted to the specified value.

The gate-source voltage, V_{GS}, near gate-source cut-off voltage is applied.

The specified drain-source voltage, V_{DS}, is applied.

The drain current is adjusted to the specified value by varying V_{GG}.

An input power just below the specified value is applied to the device being measured.

The input and output impedance matching networks are adjusted so that the power meter 2 shows the maximum value.

The input power is increased to the specified value and final adjustments are made to the impedance matching networks.

The output power is measured at the specified input power.

8.4.3.1.7 Specified conditions

– Ambient or case-point temperature, T_{amb} or T_{case}
– Drain-source voltage
– Drain current
– Frequency
– Input power

8.4.3.2 Output power at 1 dB gain compression ($P_{o(1dB)}$)

8.4.3.2.1 Purpose

To measure the output power at 1 dB gain compression of a field-effect transistor under specified conditions.

8.4.3.2.2 Circuit diagram

See the block diagram shown in Figure 56.

8.4.3.2.3 Principle of measurement

Power gain, G_p, is derived from the following equation:

$$G_p = P_o - P_i \tag{48}$$

where Po and Pi are derived from equation (46) and (47).

The output power at 1 dB gain compression, $P_{o(1dB)}$, is the value where the gain decreases by 1 dB, compared with the linear gain.

8.4.3.2.4 Circuit description and requirements

See the circuit description and requirements of 8.4.3.1.4.

8.4.3.2.5 Precautions to be observed

See the precautions of 8.4.3.1.5.

8.4.3.2.6 Measurement procedure

The frequency of the RF generator should be adjusted to the specified value.

The gate-source voltage, V_{GS}, near gate-source cut-off voltage is applied.

The specified drain-source voltage, V_{DS}, is applied.

The drain current is adjusted to the specified value by varying V_{GG}.

An input power just below the specified value $P_{i,match}$ is applied to the device being measured.

The input and output impedance matching networks are adjusted so that the power meter 2 shows the maximum value.

The input power is increased to the specified value $P_{i,match}$ and final adjustments are made to the impedance matching networks.

An adequate input power that is the sufficiently low power level to the input power, $P_{i,match}$, is applied to the device being measured.

By varying input power, confirm that the change of the output power in decibels is the same as that of the input power.

The gain, measured in the region where the change of output power in decibels is the same as that of input power, is liner gain G_{lin}.

The input power is increased until the power gain is decreased by 1 dB, compared with linear gain, G_{lin}.

The output power is measured at 1 dB-gain compression point.

8.4.3.2.7 Specified conditions

– Ambient or case-point temperature, T_{amb} or T_{case}
– Drain-source voltage, V_{DS}
– Drain current, I_D
– Frequency
– Input power, $P_{i,match}$

8.4.3.3 Power gain at 1 dB gain compression ($G_{p(1dB)}$)

8.4.3.3.1 Purpose

To measure the power gain at 1 dB gain compression of a field-effect transistor under specified conditions.

8.4.3.3.2 Circuit diagram

See the block diagram shown in Figure 56.

8.4.3.3.3 Principle of measurements

Power gain is derived from the equation (48). Power gain at 1 dB gain compression, $G_{p(1dB)}$, can be calculated by:

$$G_{p(1dB)} = G_{lin} - 1 \tag{49}$$

8.4.3.3.4 Circuit description and requirements

See the circuit description and requirements of 8.4.3.1.4.

8.4.3.3.5 Precautions to be observed

See the precautions of 8.4.3.1.5.

8.4.3.3.6 Measurement procedure

See the measurement procedure of 8.4.3.2.6. See the principle of measurements of 8.4.3.3.3.

8.4.3.3.7 Specified conditions

– Ambient or point temperature, T_{amb} or T_{case}
– Drain-source voltage
– Drain current
– Frequency
– Input power, $P_{i,match}$

8.4.3.4 Power added efficiency (η_{add})

8.4.3.4.1 Purpose

To measure the power added efficiency of a field-effect transistor under specified conditions.

8.4.3.4.2 Circuit diagram

See the block diagram shown in Figure 56.

8.4.3.4.3 Principle of measurements

Power added efficiency (η_{add}) in per cent is given by:

$$\eta_{add} = \frac{P_o - P_i}{V_{DS} \times I_D} \times 100 \qquad (50)$$

where
V_{DS} is the drain-source voltage in volts;
I_D is the drain current in amperes;
P_o and P_i are expressed in watts.

8.4.3.4.4 Circuit description and requirements

See the circuit description and requirements of 8.4.3.1.4.

8.4.3.4.5 Precautions to be observed

See the precautions of 8.4.3.1.5.

8.4.3.4.6 Measurement procedure

The frequency of the RF generator should be adjusted to the specified value.

The gate-source voltage, V_{GS}, near gate-source cut-off voltage is applied.

The specified drain-source voltage, V_{DS}, is applied.

The drain current is adjusted to the specified value by varying V_{GG}.

The specified input power is applied to the device being measured and input and output impedance matching networks are adjusted so that the power meter 2 shows the maximum value.

The output power, P_o, is measured at the specified input power, P_i.

The corresponding drain-source voltage, V_{DS}, and drain current, I_D, are also measured.

8.4.3.4.7 Specified conditions

– Ambient or case temperature, T_{amb} or T_{case}
– Drain-source voltage, V_{DS}
– Drain-current, I_D
– Frequency
– Input power, P_i

8.4.3.5 Drain efficiency (η_d)

The measuring method given in 8.4.3.4 is applicable, with the term "η_{add}" being replaced by "η_d" which is given by equation (51).

$$\eta_d = \frac{P_o}{V_{DS} \times I_D} \times 100 \qquad (51)$$

where

V_{DS} is the drain-source voltage in volts;
I_D is the drain current in amperes;
P_o are expressed in watts.

8.4.3.6 Noise figure (F) and associated gain (G_{as})

8.4.3.6.1 Purpose

To measure the noise figure and the associated gain of a microwave field-effect transistor under specified conditions.

8.4.3.6.2 Circuit diagram

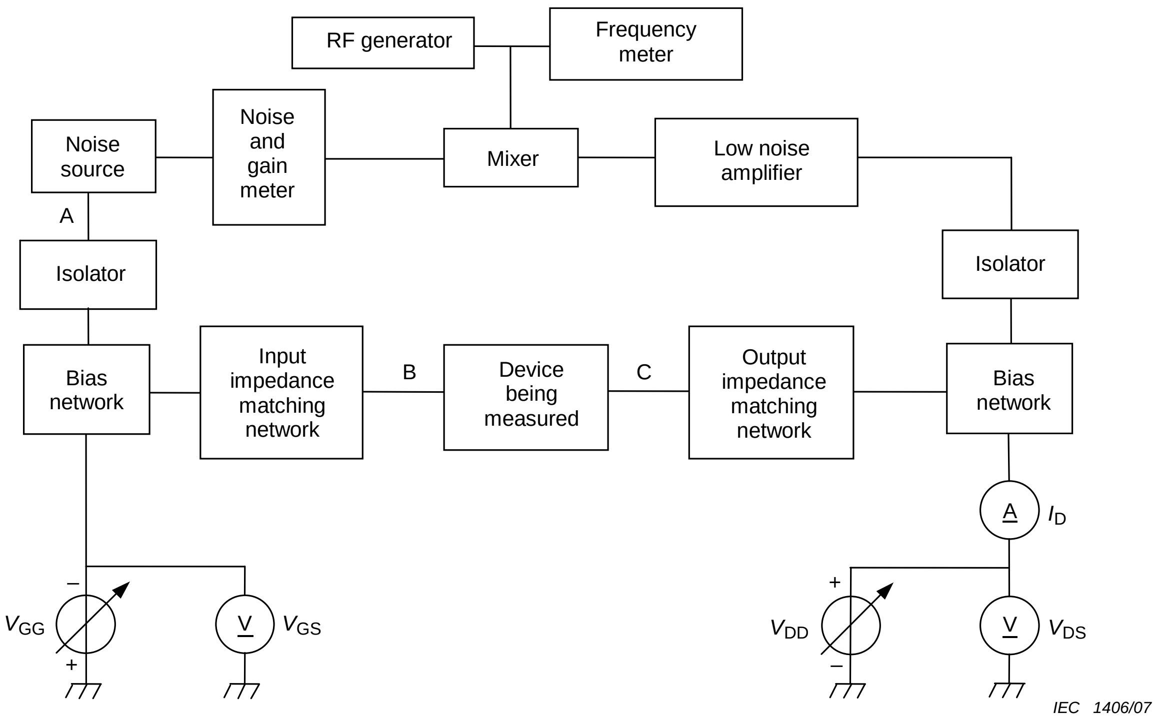


Figure 57 – Circuit for the measurement of the noise figure and associated gain

8.4.3.6.3 Principle of measurement

The noise figure F of the device being measured is derived from the following equation:

$$F = 10\log\left(10^{(F_{12} - L_1)/10} - \frac{10^{F_2/10} - 1}{10^{G_{as}/10}}\right) \quad (52)$$

where

F_{12} is the overall noise figure;

L_1 is the circuit loss from point A to B;

F_2 is the noise figure after point C at the output stage, and

G_{as} is the associated gain of the device being measured.

F, F_{12}, F_2, $L_1$and G_{as} are expressed in decibels. The noise figure measurement is carried out by using the hot and cold measurement method. F_{12}, F_2 and G_{as} are calculated as follows:

$$F_{12} = 10\log\left(\frac{10^{ENR/10}}{(P_{N1}/P_{N2}) - 1}\right) \quad (53)$$

$$F_2 = 10\log\left(\frac{10^{ENR/10}}{(P_{N3}/P_{N4}) - 1}\right) \quad (54)$$

$$G_{as} = 10 \log \left(\frac{P_{N1} - P_{N2}}{P_{N3} - P_{N4}} \right) \qquad (55)$$

where

ENR	is the excess noise ratio of the noise source;
P_{N1} and P_{N2} in W	are the measured noise power under the hot and cold state of the noise source, respectively;
P_{N3} and P_{N4} in W	are the measured noise powers under the hot and cold state of the noise source, respectively, in the case of directly connecting point B to C in Figure 57.

The temperature of the measurement is 290 K.

8.4.3.6.4 Circuit description and requirements

The circuit loss L_1 from point A to B should be measured beforehand.

8.4.3.6.5 Precautions to be observed

The entire circuit shall be shielded and grounded to prevent undesired signals. For noise figure measurement under the single-side-band (SSB) condition, careful attention shall be paid to the image and other spurious responses which are generated by the mixer. These spurious responses should be reduced so as to be negligible.

8.4.3.6.6 Measurement procedure

The frequency of the RF generator is adjusted to the specified condition.

In order to measure the noise contribution of the measurement system, connect point B to C in Figure 57 without the device being measured and set the input and output impedance matching networks to 50 Ω.

The noise power P_{N3} and P_{N4} corresponding to the noise source hot and cold, respectively, are measured.

The noise figure F_2 in decibels is calculated by equation (54).

The device being measured is inserted as shown in Figure 57.

The gate-source voltage V_{GS} (near the gate-source cut-off voltage) is applied.

The specified drain-source voltage V_{DS} is applied.

The drain current I_D is adjusted to the specified value by varying V_{GG}.

During the adjustment of the input and output matching networks, the noise power P_{N1} and P_{N2} corresponding to the noise source hot and cold, respectively, are measured.

The noise figure F_{12} in decibels is calculated by equation (53).

The associated gain G_{as} in decibels is calculated by equation (55).

The noise figure F in decibels is calculated by equation (52).

The input impedance matching network is adjusted to the minimum value of F.

The output impedance matching network is adjusted to the maximum value of G_{as}.

Repeat the above two steps until no further reduction in noise figure F is possible.

8.4.3.6.7 Specified conditions

– Ambient or reference point temperature
– Drain source voltage
– Drain current
– Frequency
– Single-side band or double-side band.

8.4.3.7 Minimum noise figure (F_{min}), equivalent input noise resistance (R_n) and source reflection factor for minimum noise figure (r_{GFmin})

8.4.3.7.1 Purpose

To measure the minimum noise figure, equivalent input noise resistance and source reflection factor for the minimum noise figure of a microwave field-effect transistor under specified conditions.

8.4.3.7.2 Circuit diagram

See the circuit diagram in 8.4.3.6.2.

8.4.3.7.3 Principle of measurement

See the principle of measurement in 8.4.3.6.3.

The noise figure dependence on the source admittance can be expressed as:

$$F = F_{min} + \frac{R_n}{G_s}\left\{(G_s - G_0)^2 + (B_s - B_0)^2\right\} \quad (56)$$

where

F is the noise figure;

F_{min} is the minimum noise figure;

R_n is the equivalent input noise resistance ;

G_s is the source conductance;

B_s is the source susceptance ;

G_0 is the source conductance for $F_{min;}$

B_0 is the source susceptance for $F_{min.}$

To determine the four parameters, F_{min}, R_n, G_0 and B_0, four dimensional simultaneous equations should be solved.

From equation (56)

$$F = F_{min} + \frac{R_n|Y_0|^2}{G_s} - 2R_nG_0 + \frac{R_n|Y_s|^2}{G_s} - 2R_nB_0\left(\frac{B_s}{G_s}\right) \quad (57)$$

where $$Y_0 = G_0 + jB_0 \quad (58)$$

$$Y_s = G_s + jB_s \quad (59)$$

In equation (57), X_1, X_2, X_3 and X_4 are defined as

$$
\begin{aligned}
X_1 &= F_{min} - 2\,R_n G_0 \\
X_2 &= R_n |Y_0|^2 \\
X_3 &= R_n \\
X_4 &= R_n\,B_0
\end{aligned}
\qquad (60)
$$

Then, equation (57) leads to the following equations for n different Y_s:

$$
\begin{aligned}
F_{(1)} &= X_1 + \frac{1}{G_{s(1)}} X_2 + \frac{|Y_{s(1)}|^2}{G_{s(1)}} X_3 - 2\left(\frac{B_{s(1)}}{G_{s(1)}}\right) X_4 \\
&\bullet \\
&\bullet \\
&\bullet \\
F_{(n)} &= X_1 + \frac{1}{G_{s(n)}} X_2 + \frac{|Y_{s(n)}|^2}{G_{s(n)}} X_3 - 2\left(\frac{B_{s(n)}}{G_{s(n)}}\right) X_4
\end{aligned}
\qquad (61)
$$

Substituting X_1, X_2, X_3 and X_4 obtained from equation (61) into equation (60), the four parameters are determined as follows:

$$F_{min} = X_1 + 2\sqrt{X_2 X_3 - X_4{}^2} \qquad (62)$$

$$R_n = X_3 \qquad (63)$$

$$G_0 = \sqrt{X_2 / X_3 - (X_4 / X_3)^2} \qquad (64)$$

$$B_0 = X_4 / X_3 \qquad (65)$$

r_{GFmin}, source reflection factor for F_{min}, is determined from the above G_0 and B_0.

8.4.3.7.4 Circuit description and requirements

See the circuit description and requirements in 8.4.3.6.4.

8.4.3.7.5 Precautions to be observed

See the precaution to be observed in 8.4.3.6.5.

8.4.3.7.6 Measurement procedure

The frequency of the RF generator is adjusted to the specified condition.
The device being measured is inserted as shown in Figure 57.
The gate-source voltage V_{GS} (near gate-source cut-off voltage) is applied.

The specified drain-source voltage V_{DS} is applied.

The drain current I_D is adjusted to the specified value by varying V_{GG}.

The input impedance matching network is adjusted so that the source admittance becomes ($G_{s(10)}$, $B_{s(10)}$).

The output impedance matching network is adjusted so that the maximum power gain is achieved.

The noise figure $F_{(10)}$ is measured in accordance with the procedure described in 8.4.3.6.6.

Repeating the above procedure *n* times, $F_{(10)–(n)}$ are determined for the n source admittance ($G_{s(10)–(n)}$, $B_{s(10)–(n)}$).

The noise parameters: F_{min}, R_n and r_{GFmin} are determined from the equations (61) to (65).

8.4.3.7.7 Specified conditions

– Ambient or reference point temperature
– Drain source voltage
– Drain current
– Frequency
– Single-side band or double-side band.

8.4.3.8 Scattering parameters (S_{ij})

The measuring method given in 7.4.3.10 is applicable, with the terms and symbols being replaced with the replacing rules shown in Tables 10 and 11.

8.4.3.9 Maximum frequency of oscillation (f_{max})

The measuring method given in 7.4.3.11 is applicable, with the terms and symbols being replaced with the replacing rules shown in Tables 10 and 11.

8.4.3.10 Transition frequency (f_T)

The measuring method given in 7.4.3.12 is applicable, with the terms and symbols being replaced with the replacing rules shown in Tables 10 and 11.

8.4.3.11 Frequency of unity current transfer ratio (f_1)

The measuring method given in 7.4.3.13 is applicable, with the terms and symbols being replaced with the replacing rules shown in Tables 10 and 11.

8.4.3.12 Maximum available gain ($G_{a(max)}$)

The measuring method given in 7.4.3.14 is applicable, with the terms and symbols being replaced with the replacing rules shown in Tables 10 and 11.

8.4.3.13 Maximum stable gain (G_{ms})

The measuring method given in 7.4.3.15 is applicable, with the terms and symbols being replaced with the replacing rules shown in Tables 10 and 11.

8.4.3.14 Insertion power gain ($|S_{21}|^2$)

The measuring method given in 7.4.3.16 is applicable, with the terms and symbols being replaced with the replacing rules shown in Tables 10 and 11.

8.4.3.15 Intermodulation distortion (two-tone) (P_1/P_n)

The measuring method given in 7.4.3.17 is applicable, with the terms and symbols being replaced with the replacing rules shown in Tables 10 and 11.

8.4.3.16 Input power at the intercept point (for intermodulation products) ($P_{i,n(IP)}$) and output power at the intercept point (for intermodulation products) ($P_{o,n(IP)}$)

The measuring method given in 7.4.3.18 is applicable, with the terms and symbols being replaced with the replacing rules shown in Tables 10 and 11.

8.5 Verifying methods

8.5.1 Load mismatch tolerance (Ψ_L)

The verifying method given in 7.5.1 is applicable, with the terms and symbols being replaced with the replacing rules shown in Tables 10 and 11.

8.5.2 Source mismatch tolerance (Ψ_S)

The verifying method given in 7.5.2 is applicable, with the terms and symbols being replaced with the replacing rules shown in Tables 10 and 11.

8.5.3 Load mismatch ruggedness (Ψ_R)

The verifying method given in 7.5.3 is applicable, with the terms and symbols being replaced with the replacing rules shown in Table 10 and 11.

9 Assessment and reliability – specific requirements

9.1 Electrical test conditions

Test conditions and test circuits, for each device category, are listed in Table 12. The relevant specification will state which test(s) will apply.

9.2 Failure criteria and failure-defining characteristics for acceptance tests

Failure-defining characteristics, their failure criteria and measurement conditions for each device category are listed in Table 13.

NOTE Characteristics should be measured in the sequence in which they are listed in this table because the changes of characteristics caused by some failure mechanisms may be wholly or partially masked by the influence of other measurements.

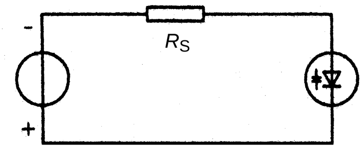
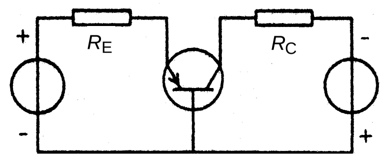
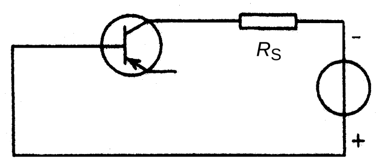
9.3 Failure criteria and failure-defining characteristics for reliability tests

(Under consideration).

9.4 Procedure in case of a testing error

When a device has failed as a result of a testing error (such as a test equipment fault or measurement equipment fault, or an operator error), the failure shall be noted in the data record with an explanation of the cause.

Table 12 – Operating conditions and test circuits

Device categories	Tests	Operating conditions			Test circuits	Remarks
		Current	Voltage	Temperature		
Variable capacitance diodes for tuning application	High temperature reverse bias		$V_R = V_R$ max.	Highest operating temperature $t_{amb(max)}$ or $t_{case(max)}$ as specified		R_s = current limiting resistor
Mixer diodes	Under consideration					
Impatt diodes	Under consideration					
Gunn diodes	Under consideration					
Bipolar transistors	Operating life	$I_C = \frac{P_{tot}\text{max.}}{V_{CE}}$ (see 7.2.6 of IEC 60747-1:2006)	V_{CE} = 0,7 V_{CEO} max. (NOTE 1)	(See 7.2.4 of IEC 60747-1:2006)	 (NOTE 2)	$R_E \geq \frac{10V_{EB}}{I_E}$ $R_C \approx \frac{V_{CB}}{I_C}$
	High temperature reverse bias		$V_{CB} = V_{CBO}$ max.	Highest operating temperature $t_{amb(max)}$ or $t_{case(max)}$ as specified	 (NOTE 2)	R_s = current limiting resistor

NOTE 1 Test conditions should be within the safe operating area if one is specified. The voltage is to be lowered below 0,7 V_{CEO} max. with only as much as necessary to remain within the safe operating area.

NOTE 2 Change circuit appropriately for NPN transistor.

Table 12 *(continued)*

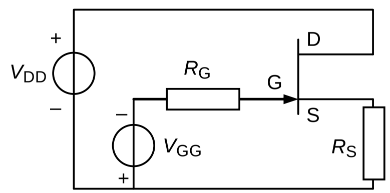
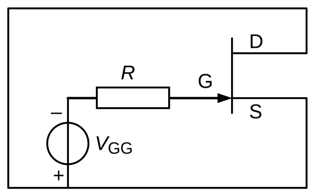
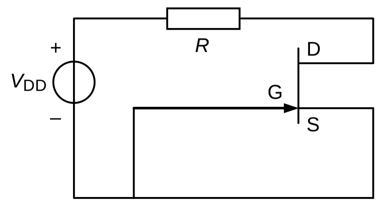
Device categories	Tests	Operating conditions			Test circuits	Remarks
		Current	Voltage	Temperature		
Field effect transistors	Operating life	$I_D = \frac{P_{tot}\text{max.}}{V_{DS}}$ $I_D < I_{DSS}$ min.	V_{GG} = set to obtain the required I_D Preferably, for best regulation; V_{DS} = specified (preferably 0,8 V_{DS} max.) V_{DD} = 2 V_{DS}	t_{amb} or t_{case} as specified See 7.2.4 of IEC 60747-1:2006		R_G = 10 kΩ or as specified $R_S \approx \frac{V_{DS}}{I_D}$
	High temperature reverse bias: Depletion Types		V_{DS} = 0 V_{GS} = 0,7 to 0,8 V_{GSO} max. (preferably 0,8)	Highest operating temperature $t_{amb(max)}$ or $t_{case(max)}$ as specified		R = current limiting resistor
	High temperature reverse bias: Enhancement types		V_{GS} = 0 V_{DS} = 0,7 to 0,8 V_{DSS} max. (preferably 0,8)	Highest operating temperature $t_{amb(max)}$ or $t_{case(max)}$ as specified		R = current limiting resistor

Table 13 – Failure criteria and measurement conditions

Device categories	Failure-defining characteristics	Failure criteria (NOTE 1)	Measurement conditions
Variable capacitance diodes for tuning applications	I_R	>2 × USL	Highest V_R specified for I_R
	V_F	>1,1 × USL	Highest I_F specified for V_F
	Q or r_s	<0,5 × LSL >2 × USL	Lowest V_R specified for Q
Mixer diodes	Under consideration		
Impatt diodes	Under consideration		
Gunn diodes	Under consideration		
Bipolar transistors	I_{CBO}	>2 × USL	Highest V_{CB} specified for I_{CBO}
	h_{21E} (h_{21e} (NOTE 2))	<0,8 × LSL >1,2 × USL	A value of I_C for which a h_{21E} (h_{21e}) tolerance (lower and upper limits) is specified
	V_{CEsat}	>1,2 × USL	Highest I_C specified for V_{CEsat}
	F (NOTE 3)	>USL + 3 dB	Lowest I_C specified for F
Field-effect transistors	$V_{(BR)GS}$ or I_{GSS}	<0,8 × LSL >10 × USL	A value of I_G for which a $V_{(BR)GS}$ tolerance (lower limit) is specified A value of V_{GS} for which a I_{GSS} tolerance (upper limit) is specified
	V_{GSoff}	<0,8 × LSL >1,2 × USL	A value of I_D for which a V_{GSoff} tolerance (lower and upper limits) is specified
	I_{DSS}	<0,9 × LSL >1,1 × USL	A value of V_{DS} for which a I_D tolerance (lower and upper limits) is specified

NOTE 1 USL = upper specification limit.
LSL = lower specification limit.
IVD = initial value of individual device.

NOTE 2 Only where no h_{21E} tolerances are specified or where h_{21E} is unspecified.

NOTE 3 Where applicable.