

IEEE Recommended Practice for Radio-Frequency (RF) Absorber Evaluation in the Range of 30 MHz to 5 GHz

Sponsor

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Abstract: Realistic and repeatable criteria, as well as recommended test methods, for characterizing the absorption properties of typical anechoic chamber linings applied to a metallic surface are described. Parameters and test procedures are described for the evaluation of RF absorbers to be used for radiated emissions and radiated susceptibility testing of electronic products, in the absorber manufacturer and/or absorber user environment, over the frequency range of 30 MHz to 5 GHz.

Keywords: anechoic chamber, radiated emissions, RF absorber, semianechoic chamber

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Introduction

[This introduction is not part of IEEE Std 1128-1998, IEEE Recommended Practice for Radio-Frequency (RF) Absorber Evaluation in the Range of 30 MHz to 5 GHz.]

Interest in materials that absorb radio-frequency energy has existed for many years. The recent increased regulation of sources of radio waves and equipment immunity levels has led to the need for a more accurate determination of electromagnetic field intensity. As modern measuring antennas and receivers have increased measurement accuracy, the problem of making accurate measurements in less than optimum open-area test sites has become a more important part of the overall measurement procedures. The practice of placing absorbing materials on the walls and ceilings of measuring sites to reduce reflections from these surfaces has become common. Claims for the efficacy of various absorbing materials, however, have led to conflicting reports in the literature, which confuse many potential purchasers of absorbing material. An effort to end that confusion led to the development of a draft recommended practice in 1986. Following several years of work, the current recommended practice was developed.

At the time this recommended practice was completed, the working group had the following membership:

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The following individuals have contributed reviews and comments and have attended meetings:

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The following persons were on the balloting committee:

Stephen H. Berger	Donald N. Heirman	A. Piper
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Alan H. Cookson

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IEEE Recommended Practice for Radio-Frequency (RF) Absorber Evaluation in the Range of 30 MHz to 5 GHz

1. Introduction

1.1 Scope

The purpose of this recommended practice is to recommend realistic and repeatable criteria, as well as recommended test methods, to characterize the absorption properties of typical anechoic chamber linings applied to a metallic surface. This recommended practice covers the parameters and test procedures for the evaluation of radio-frequency (RF) absorbers to be used for radiated emissions and radiated susceptibility testing of electronic products, in the absorber manufacturer and/or absorber user environment, over the frequency range of 30 MHz to 5 GHz.

The recommended methods approach the RF absorber evaluation at three levels:

- a) RF absorber materials bulk parameters
- b) Arrays of commercially available RF absorbers
- c) RF absorbers in actual applications, as in anechoic or semianechoic chambers and lined open-area test sites.

The evaluation measurements can be performed in frequency and/or time domain. This recommended practice, however, does not address the accuracy and limitations of the different evaluation methods. These issues will be addressed in future revisions of this recommended practice.

1.2 Applications

With the proliferation of RF absorber-lined shielded rooms, and their wide utilization for testing equipment for radiated emissions and radiated susceptibility, there is a need to provide repeatable and realistic performance figures of such RF absorbers in the frequency range of 30 MHz to 5 GHz and higher.

Up until now, the data provided by many manufacturers usually has been for normal incidence only. This data eliminates the effect of polarization on the RF absorber performance. For RF absorbers that use a pyramidal structure, the reflection coefficient is also a function of the alignment of the incident wave with the pyramidal structure, especially for high frequencies and for large angles of incidence. Furthermore, only the magnitude of the reflection coefficient typically is provided by the manufacturers. The lack of these kinds of data precludes the accurate calculation of the performance of RF absorber materials in the great majority of applications.

This recommended practice is intended as an aid to RF absorber users and manufacturers. It indicates the type of data that is required by the design engineer and the various methods that may be used to obtain such data. It is hoped that, in the future, both manufacturers and users will utilize the same or accepted equivalent methods to estimate the performance of RF absorbers.

2. References

This recommended practice should be used in conjunction with the following documents:

ANSI C63.2-1996, American National Standard for Electromagnetic Noise and Field Strength Instrumentation, 10 kHz to 40 GHz Specifications.¹

ANSI C63.4-1992, American National Standard for Methods of Measurement of Radio-Noise Emissions from Low-Voltage Electrical and Electronic Equipment in the Range of 9 kHz to 40 GHz.

IEC/CISPR 16-1 (1993-08), Specifications for radio disturbance and immunity measuring apparatus and methods—Part 1: Radio disturbance and immunity measuring apparatus.²

3. Definitions and acronyms

3.1 Definitions

For the purposes of this recommended practice, the following terms and definitions shall apply.

3.1.1 absorber-lined chamber (ALC): A room or enclosure (either shielded or unshielded) with all of its surfaces lined with radio-frequency (RF) absorber material. Commonly referred to as an anechoic chamber.

3.1.2 absorber-lined open-area test site (ATS): An open-area test site (OATS) in which the ground plane is covered with radio-frequency (RF) absorber to suppress ground reflections. *See also: open-area test site (OATS).*

3.1.3 angle of incidence: At a point on a surface, the acute angle between the normal to this surface and the direction of propagation of an incident wave.

3.1.4 antenna factor: a) Quantity relating the strength of the field in which an antenna is immersed to the output voltage across the load connected to the antenna. b) A factor that, when properly applied to the voltage meter reading of the measuring instrument, yields the electric field strength in volts/meter or the magnetic field strength in amperes/meter.

NOTES

1—This factor includes the effects of antenna effective length and mismatch and may include transmission line loss.

2—The factor for the electric field strength is not necessarily the same as the factor for the magnetic field strength.

3—The antenna factor, as determined in ANSI C63.5-1988, is very nearly equal to the free-space antenna factor.

3.1.5 bistatic reflectivity: The reflectivity when the reflected wave is in any specified direction other than back toward the transmit antenna. The transmit and receive antennas are at different locations.

¹ANSI C63 publications are available from the Institute of Electrical and Electronics Engineers, 445 Hoes Lane, P.O. Box 1331, Piscataway, NJ 08855-1331, USA; or from the Sales Department, American National Standards Institute, 11 West 42nd Street, 13th Floor, New York, NY 10036, USA.

²CISPR documents are available from the International Electrotechnical Commission, 3, rue de Varembe, Case Postale 131, CH 1211, Genève 20, Switzerland/Suisse. They are also available in the United States from the Sales Department, American National Standards Institute, 11 West 42nd Street, 13th Floor, New York, NY 10036, USA.

3.1.6 bulk parameters: Complex permittivity, complex permeability, and conductivity properties of the bulk material used in the radio-frequency (RF) absorber. The conductivity may be included in the imaginary part of the complex permittivity.

3.1.7 diffuse scattering: Scattering of incident electromagnetic energy over a range of angles other than the specular direction.

3.1.8 equipment under test (EUT): The equipment being measured or tested, as opposed to support or ancillary equipment.

3.1.9 equivalent reflection coefficient (ERC): The measure of the reflection coefficient of an actual radio-frequency (RF) absorber-lined reflecting surface. The ERC includes not only the RF absorber reflection, but other effects such as mounting fixtures, adhesive, and any air space between the RF absorber and the reflecting surface.

3.1.10 frequency domain: A function in which frequency is the independent variable.

3.1.11 gigahertz transverse electromagnetic (GTEM) cell: A tapered transverse electromagnetic (TEM) cell with an absorber-lined end wall terminated with an absorber load.

3.1.12 intrinsic impedance: For a monochromatic (time harmonic) electromagnetic wave propagating in a homogenous medium, the ratio of the complex amplitude of the electric field to that of the magnetic field.

NOTE—The intrinsic impedance of a medium is sometimes referred to as the characteristic impedance of the medium.

3.1.13 monostatic reflectivity: Reflectivity in which the reflected and incident waves follow the same path but in opposite directions. The transmit and receive antennas are in the same location, and normal incidence occurs at the reflecting surface.

3.1.14 open-area test site (OATS): A site that meets specified requirements for measuring radio-interference fields radiated by an equipment under test (EUT).

3.1.15 parallel polarization: The polarization of a wave for which the electric field vector lies in the plane of incidence.

NOTE—This is sometimes called vertical or transverse magnetic (TM) polarization.

3.1.16 perpendicular polarization: The polarization of a wave for which the electric field is perpendicular to the plane of incidence.

NOTE—This is sometimes called horizontal or transverse electric (TE) polarization.

3.1.17 plane of incidence: The plane containing the normal to the surface of a boundary and the phase vector (β) of the incident wave.

3.1.18 quiet zone: A described volume within an anechoic chamber where electromagnetic waves reflected from the walls, floor, and ceiling are stated to be below a certain specified minimum. The quiet zone may have a spherical, cylindrical, rectangular, etc., shape depending on the chamber characteristics.

3.1.19 radio-frequency (RF) absorber: A material designed to absorb electromagnetic energy. The material may have a flat face or may be formed into pyramids, wedges, or cones. Radar absorber material is commonly referred to as RAM.

3.1.20 reflectivity (R): For a radio-frequency (RF) absorber, the ratio of the plane wave reflected power density (P_r) to the plane wave incident power density (P_i) at a reference point in space. It is expressed in dB as $R = 10\log_{10}(P_r/P_i)$.

3.1.21 reflectometer, time-domain: An instrument designed to indicate and to measure reflection characteristics of a transmission system connected to the instrument by monitoring the step-formed signals entering the test object and the superimposed reflected transient signals on an oscilloscope that is equipped with a suitable time-based sweep. The measuring system basically consists of a fast-rise function generator, a tee coupler, and an oscilloscope connected to the probing branch of the coupler.

3.1.22 return loss: The ratio, in dB, of the power incident upon the discontinuity to the power reflected from the discontinuity.

NOTE—This ratio is also the square of the reciprocal of the magnitude of the reflection coefficient.

3.1.23 scattering matrix: An $n \times n$ (square) matrix used to relate incident waves and reflected waves for an n -port network.

3.1.24 semianechoic absorber-lined chamber (SALC): A room or enclosure (either shielded or unshielded) with all its surfaces, except the floor, lined with radio-frequency (RF) absorber material. The floor is covered with a good conductor.

3.1.25 site attenuation: a) The ratio of the power input to a matched, balanced, lossless, tuned dipole radiator to that at the output of a similarly matched, balanced, lossless, tuned dipole receiving antenna for specified polarization, separation, and heights above a flat reflecting surface. It is a measure of the transmission path loss between two antennas.

NOTE—The above is the classical definition of site attenuation. In ANSI C63.4-1992, the definition is extended to cover broadband antennas as well as tuned dipole antennas.

b) The ratio of the power input to a matched, balanced, lossless, wideband dipole radiator to that at the output of a similarly balanced, matched, lossless, wideband dipole receiving antenna for specified polarization, separation, and heights above a flat reflecting surface.

3.1.26 site-attenuation deviation (SAD): The difference, in dB, of the site attenuation and the free-space distance attenuation between the transmit and receive antennas in an open-area test site (OATS), absorber-lined open-area test site (ATS), semianechoic chamber, or anechoic chamber. *See also:* **absorber-lined open-area test site (ATS)** and **open-area test site (OATS)**.

3.1.27 specular reflection: The reflection of a wave when incident on an infinite planar surface. The angle of incidence is equal to the angle of reflection.

3.1.28 standing wave: A wave in which, for any component of the field, the ratio of its instantaneous value at one point to that at any other point does not vary with time. A pure standing wave results from the interference of two oppositely directed traveling waves of the same frequency and amplitude.

3.1.29 time domain: A function in which the signals are represented as a function of time.

3.1.30 transverse electromagnetic wave (TEW): An electromagnetic wave in which both the electric and magnetic field vectors are perpendicular everywhere to the direction of propagation.

3.2 Acronyms

AF	antenna factor
ALC	absorber-lined chamber
ATS	absorber-lined open-area test site
BPF	bandpass filter
CRT	cathode ray tube

CW	continuous wave
DFT	Discrete Fourier Transform
DUT	device under test
FM	frequency modulation
EMC	electromagnetic compatibility
EMI	electromagnetic interference
ERC	equivalent reflection coefficient
EUT	equipment under test
FFT	Fast Fourier Transform
IF	intermediate frequency
LCR	low-frequency coaxial reflectometer
OATS	open-area test site
OSAD	open-site-attenuation deviation
RAM	radar absorber material
RBW	resolution bandwidth
RF	radio frequency
SAD	site-attenuation deviation
SALC	semianechoic absorber-lined chamber
TDR	time-domain reflectometer
TE	transverse electric
TEM	transverse electromagnetic
TEW	transverse electromagnetic wave
TM	transverse magnetic
VBW	video bandwidth
VSWR	voltage standing wave ratio

4. Measurement instrumentation

4.1 Spectrum analyzers

4.1.1 Introduction

Superheterodyne spectrum analyzers are widely used in applications that are electromagnetic compatibility (EMC) related. Other spectrum analyzer architectures, such as Fast Fourier Transform (FFT) analyzers, do have some limitations relative to the superheterodyne spectrum analyzers in the areas of frequency range, sensitivity, and dynamic range. The ability of the FFT analyzer, however, to characterize single-shot phenomena and measure both the magnitude and phase of an emission often makes EMC trouble-shooting much easier or even possible. See [B3]³ and [B4].

Figure 1 shows a simplified block diagram of a superheterodyne spectrum analyzer. An input signal fed into the instrument's RF input passes through an RF attenuator to a mixer, where it mixes with a signal from the local oscillator. Because the mixer is a nonlinear device, its output includes not only the two original signals, but also their harmonics and the sums and differences of the original frequencies and their harmonics. If any of the mixed signals falls within the passband of the intermediate-frequency (IF) filter, it is further amplified, detected by an envelope detector, digitized, and displayed either graphically on a digital display or numerically on a numeric readout. A scan generator generates a ramp voltage, which deflects the cathode ray tube (CRT) beam horizontally across an analog screen and tunes the local oscillator so that its frequency changes in proportion to the ramp voltage. The horizontal axis of the spectrum analyzer display is calibrated in frequency. The vertical axis is calibrated in amplitude. Virtually all analyzers offer the choice of a linear scale, calibrated in V, or a logarithmic scale, calibrated in dB (e.g., referenced to V, mV, μ V, or mW). The

³The numbers in brackets preceded by the letter B correspond to those of the bibliography in Clause 10.

spectrum analyzer often has more than one mixer stage. This second mixing has a constant-frequency local oscillator that converts the first IF frequency to a lower IF. The reason for multiple conversions is to improve image rejection.

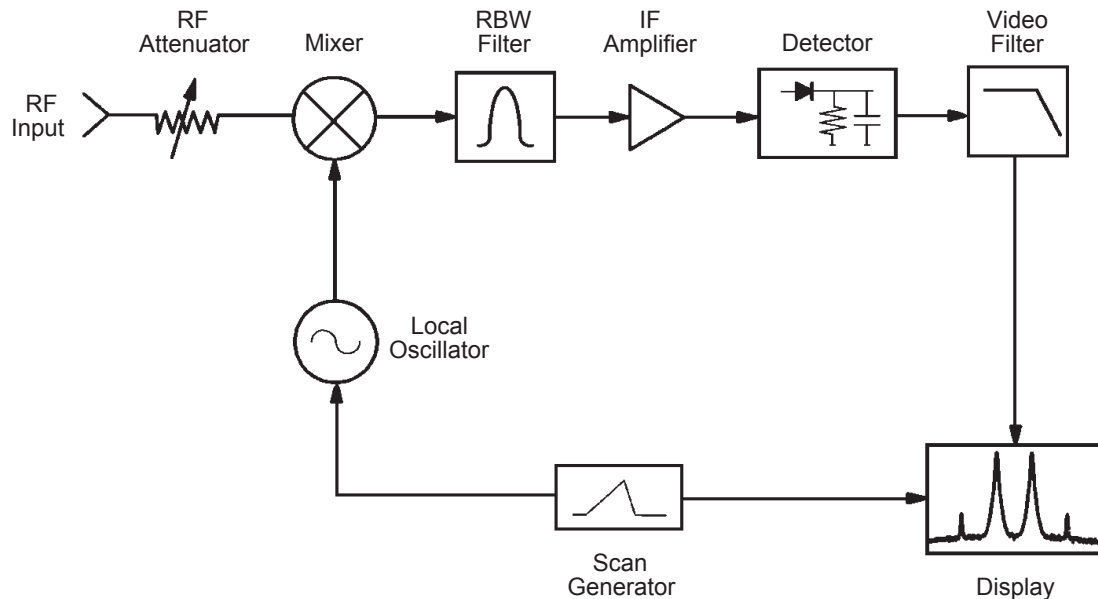


Figure 1—Block diagram of a superheterodyne spectrum analyzer

4.1.2 Input attenuation

The RF attenuator, which can usually be set in 10 dB steps from 0–50 dB (or even higher), is used to condition the amplitude of the input signal in order to avoid mixer saturation and the generation of internal distortion signals, which might also be displayed and mistaken as true signals. The displayed signal-to-noise-floor ratio, the level of the effective noise floor at the input of the analyzer, and the sensitivity of the analyzer are also affected by the RF attenuator setting.

4.1.3 Signal mixing

The mixer is used, along with the local oscillator, to convert the input signal frequency to a fixed IF. Of all the products emerging from the mixer, the two with the greatest amplitudes and, therefore, the two that are most desirable are those created from the sum and the difference of the local oscillator and the input signal. If the frequency of the signal to be examined is either above or below the local oscillator frequency by the IF, one of the desired mixing products will fall within the passband of the IF filter and will appear on the display. The local oscillator frequency is chosen often to produce an IF above the highest tuning frequency. In this case, the tuning equation is

$$f_{\text{sig}} = f_{\text{LO}} - f_{\text{IF}} \quad (1)$$

where

- f_{sig} is the signal frequency (in Hz).
- f_{LO} is the local oscillator frequency (in Hz).
- f_{IF} is the intermediate frequency (in Hz).

The stability of the spectrum analyzer's local oscillator, particularly its first local oscillator, affects the resolution of the instrument. The residual frequency modulation (FM) of the local oscillator is transferred to any mixing product that results from the local oscillator and incoming signals. Therefore, it is not possible to determine whether the signal or the local oscillator is the source of instability. The effects of local oscillator residual FM are not visible in wide resolution bandwidths. Only when the bandwidth approximates the peak-to-peak excursion of the FM does this effect become apparent. For this reason, the minimum bandwidth typically found in a spectrum analyzer is determined at least in part by the local oscillator stability. Another local oscillator attribute, phase noise, also determines the resolution in yet a different manner. Since local oscillators are not perfectly stable, their signals are always frequency or phase modulated to some extent by random noise. As noted before, all instabilities in the local oscillator are transferred to any mixing products; therefore, phase-noise modulation sidebands appear around any spectral component on the display that is far enough above the broadband noise floor of the instrument. The amplitude difference between the displayed spectral component and the phase noise is a function of the stability of the local oscillator as well as the resolution bandwidth. A reduction of the IF filter bandwidth by a factor of 10 will result in a 10 dB decrease of phase-noise level on the display.

4.1.4 IF signal filtering

The IF of a superheterodyne spectrum analyzer is determined by a bandpass filter. This filter selects the desired mixing products (e.g., the sum of the local oscillator and input signal) and rejects all other signals. Because the input signal is fixed and the local oscillator is swept, the products from the mixer are also swept. If a mixing product sweeps past the IF, the characteristics of the bandpass filter are traced on the display. Unless two signals are far enough apart, the traces they cause will fall on top of each other and look like only one response. In this case, individual signals cannot be resolved. Spectrum analyzers, however, have selectable IF filters (e.g., 10 Hz to 3 MHz in a 1-3-10 sequence), so it is usually possible to select one narrow enough to resolve closely spaced signals. Most often, the 3 dB bandwidth of the IF filters are specified, indicating how close together equal amplitude sinusoids can be resolved as two separate signals. The resolution of two signals with unequal amplitude is affected by the filter's shape factor, which is determined by the ratio of the 60 dB bandwidth to the 3 dB bandwidth. Most analog filters in spectrum analyzers are synchronously tuned, have four or five poles, and are nearly Gaussian in shape, which leads to shape factors between 15:1 to 11:1 for narrower filters.

4.1.5 Sweeptime

The selected IF filter also determines the sweeptime, which, in turn, affects the overall measurement time. The bandpass filters are band-limited circuits that require finite times for the filtered signals to reach their maximum amplitude. If the mixing products are swept too quickly, there will be a loss of displayed amplitude and a shift in frequency. The following equation can be used to determine the appropriate sweeptime required to make calibrated measurements (see [B1]):

$$ST = \frac{k \times \text{Span}}{\text{RBW}^2} \quad (2)$$

where

- k is the constant of proportionality (2–3 for synchronously tuned near gaussian filters).
- Span is the displayed frequency range (in Hz).
- RBW is the selected IF bandwidth (in Hz).

4.1.6 IF signal amplification and detection

The IF amplifier boosts the fixed IF signal, according to the current reference level setting (either linearly or logarithmically) and based on the selected display mode, before it is processed further by the detector. Spectrum analyzers typically convert the IF signal to video with an envelope detector. In its simplest form, an envelope detector is a diode followed by a parallel resistor capacitor (RC) combination. The time constants

of the detector are such that the voltage across the capacitor equals the peak value of the IF signal at all times (i.e., the detector can follow the fastest possible changes in the envelope of the IF signal but not the instantaneous value of the IF signal itself). The maximum rate at which the envelope of the IF signal can change is determined by the IF filter bandwidth, as discussed in 4.1.4.

4.1.7 Video filtering

The video filter of the spectrum analyzer is a lowpass filter that follows the detector. As its cutoff frequency is reduced to the point at which it becomes equal to or less than the bandwidth of the IF filter, the high-frequency spectral components of the IF signal's envelope are suppressed, resulting in an averaging or smoothing effect on the displayed signal. This effect is most noticeable in measuring noise, particularly when a wide IF bandwidth is used. The displayed amplitude of broadband coherent signals (e.g., pulses), however, is also affected by the video filter bandwidth.

It should be noted that the video filter has its own response time. When the video bandwidth is equal to or less than the IF bandwidth, the sweep time equation becomes (from [B1])

$$ST = \frac{k \times \text{Span}}{\text{RBW} \times \text{VBW}} \quad (3)$$

where

VBW is the video bandwidth.

4.1.8 Display

Most of the modern spectrum analyzers have a built-in digital display that allows fast data updates at a flicker-free rate without the blooming or fading that is common for analog displays. Once the data has been digitized and stored in display memory, it is updated at the sweep rate of the spectrum analyzer, thus allowing the observation of the actual data update on the display as the instrument is sweeping through the specified frequency range.

Independent of the number of digital points of the display (e.g., 401 or 1001), each point represents a special amplitude value that was detected over some frequency range and some period of time. One display method used is to digitize the instantaneous value of a signal at the end of a digitization interval. This is usually called "sample detector" or "sample mode." Using this mode, the signal variation between individual sweeps is retained, which can be used to identify signal characteristics. Signals, however, might not be displayed at all in the case of wide frequency measurements. In such cases, another method called "peak mode" is used to display the maximum value encountered in each digitization interval to avoid missing signals. A frequency error is usually introduced by this method because the maximum amplitude might be found at some other frequency in the digitization interval. To correctly display both noise and coherent signals, an algorithm is used to determine if the signal rose and fell within the interval represented by the data point. If the signal both rises and falls, then the algorithm classifies the signal as noise (Rosenfell detector). In that case, an odd-numbered data point indicates the maximum value encountered during the digitization interval; whereas an even-numbered point indicates the minimum amplitude.

4.2 Spectrum analyzer and tracking generator

4.2.1 Introduction

A tracking generator is a special type of swept source that is used with a spectrum analyzer to make amplitude and/or precise frequency measurements. The tracking generator has a stable, fixed local oscillator that generates a signal at the IF frequency, f_{IF} , of the spectrum analyzer. Hence, the tracking generator's output

frequency, f_{TG} , can be made to track precisely the spectrum analyzer's input frequency, f_{SA} , since both are effectively tuned by the same local oscillator. This precision tracking exists in a swept-measurement mode and in a fixed-tuned mode, where the output of the tracking generator is simply a continuous wave (CW) signal. Since a flat output level is necessary for swept-frequency applications, the tracking generator generally has an internal leveling loop that ensures a flat output level over the entire frequency range of the spectrum analyzer (see [B4]).

This tracking action allows measurement of the transmitted or reflected amplitude responses of a device (e.g., site attenuation involving two antennas). Using a signal source and spectrum analyzer, the tracking generator allows high-dynamic-range transmission or reflection measurements with fast, continuous sweeps. Furthermore, a spectrum analyzer/tracking generator combination can be used to determine precisely the frequency of unknown signals where the signals to be measured are fed into the spectrum analyzer's input and the tracking generator output is connected to a counter. Most modern spectrum analyzers have a built-in precision counter that provides precision frequency measurement capability.

As shown in Figure 2, the device under test (DUT) can be placed between the tracking generator's output and the spectrum analyzer's input for transmission measurements. Performing reflection measurements, such as to determine the voltage standing wave ratio (VSWR) of a DUT, requires an additional directional coupler or bridge to separate the incident tracking generator signal from the signal reflected by the DUT. Both configurations are used for swept amplitude measurements. No phase information can be obtained with these setups. The spectrum analyzer/tracking generator has a large dynamic range and is capable of measuring very narrowband DUTs (e.g., 20 Hz crystal filters) that are not possible with other swept systems, such as the scalar and vector network analyzers.

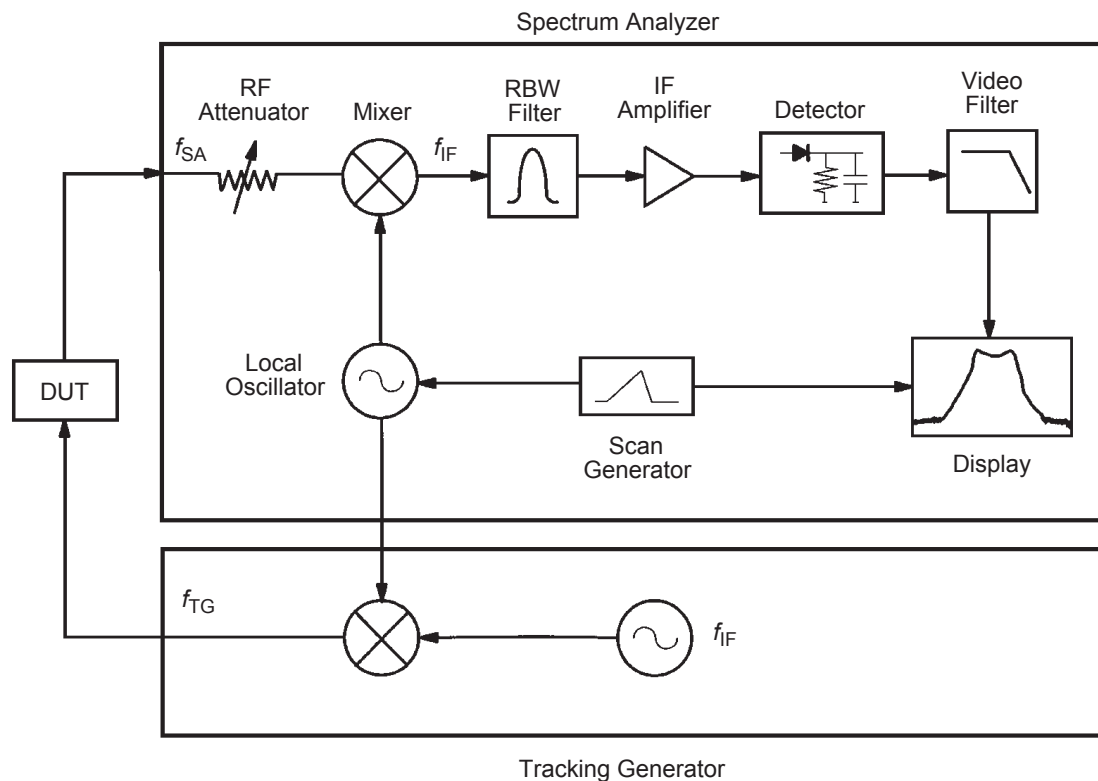


Figure 2—Block diagram of a superheterodyne spectrum analyzer and tracking generator

4.2.2 Dynamic range

The measurement dynamic range of the spectrum analyzer/tracking generator may exceed 120 dB. The key to this large dynamic range is the highly stable and precise tracking of the spectrum analyzer and the tracking generator, which permits overdriving the mixer to the 1 dB compression level while keeping the analyzer distortion products off the display. Furthermore, the spectrum analyzer's lowest IF bandwidth (e.g., 10 Hz or 100 Hz) can be used to reduce the displayed average noise level. Both factors provide a large dynamic range for amplitude measurements.

Precision tracking means that, at every instant of time, the generator fundamental frequency is in the center of the analyzer passband, and all generator harmonics, whether they are produced in the analyzer or generated in the tracking generator itself, are outside the passband. Thus, only the tracking generator fundamental frequency is displayed on the analyzer screen. While these distortion products may exist in the measurement setup, they are completely eliminated from the spectrum analyzer display.

4.2.3 Frequency accuracy

The frequency accuracy of a spectrum analyzer/tracking generator system is dependent upon the frequency tracking between the two instruments. When the tracking is degraded, the tracking generator frequency shifts from the center of the IF bandwidth, and the analyzer response decreases in amplitude and shifts in frequency. This frequency shift is a result of tracking error, and is typically negligible when measuring over wide frequency spans using wide IF filter bandwidths. When measuring narrowband DUTs, however, precise tracking is mandatory to avoid erroneous measurement results.

Measurements of DUTs of very narrow bandwidth (e.g., less than 20 Hz) are potentially limited by the tracking generator's residual FM. The ability to use a very small spectrum analyzer IF bandwidth (e.g., 10 Hz) results in filtering out those components of the residual FM whose fluctuation rates are greater than approximately half of the used IF bandwidth (e.g., 5 Hz), thus leaving only a small fraction of the total residual FM.

4.2.4 Measurement normalization

Normalization is a technique to remove certain components of the absolute system amplitude uncertainty from the measurement result. The procedures for transmission and reflection measurements are different, and the errors being eliminated by normalizing are different as well.

A transmission measurement requires a direct connection that replaces the DUT during the normalization measurement. A sweep is executed over the frequency range in which the actual DUT is to be tested, and the trace is stored in memory. The DUT is then connected in place, and the measurement is made by subtracting the normalization trace from the current trace. This, in effect, removes the system frequency response uncertainty. The measurement errors left are due to display accuracy (e.g., ± 0.5 – 0.8 dB full screen) and the mismatch uncertainties at the input and output of the DUT.

When reflection measurements are made, a different uncertainty calculation, which uses another normalization technique, applies. The reflection measurement amplitude uncertainty can be estimated by the following equation:

$$\Delta\rho = A + B\rho_L + C\rho_L^2 \quad (4)$$

where

- A is the directivity of the coupler.
- B is the frequency response.
- C is the effective source match.
- ρ_L is the reflection coefficient of the DUT (see [B12]).

If the quantities used in this equation are expressed in dB, as they normally are, they should be converted to linear units before using the equation. For example, a frequency response of -20 dB should be 0.1 . The term $B\rho_L$ can be eliminated by normalization. This is done by connecting an open circuit ($\rho = 1$) in place of the DUT and storing the trace with all desired measurement settings. Then a short circuit ($\rho = -1$) is connected in place of the open circuit, and another trace is taken and averaged with the open-circuit trace. Next, the DUT is connected, and the averaged normalization trace is subtracted from the current trace. Since the open circuit and the short circuit both have reflection coefficients of magnitude one (0 dB), ideally, either could have been used as a reference for normalization. However, since the directional coupler is not perfectly directional, and since there exists mismatch between the tracking generator and the coupler, a single normalization trace would include information about errors associated with reflections from internal components of the measurement system, not just from the open-circuit or short-circuit reference, as desired. The open-circuit and short-circuit traces are averaged in order to remove this undesired internal reflection component from the reference data. The effective source match, as well as the coupler directivity, cannot be eliminated with normalization.

4.3 Electromagnetic interference (EMI) receiver

4.3.1 Introduction

Modern EMI receivers are based typically on the superheterodyne principle with one to three frequency conversion stages, just like a spectrum analyzer. Figure 3 is a simplified block diagram of an EMI receiver that shows the commonalities with the spectrum analyzer block diagram in Figure 1. The EMI receiver, however, has additional components, such as a preselector, a preamplifier, specialized IF bandwidths, and detector responses, in order to meet IEC/CISPR 16-1 (1993-08)⁴ requirements. Therefore, a spectrum analyzer alone cannot be used for EMI compliance measurements unless additional equipment is used to satisfy the CISPR requirements. The following paragraphs only describe those blocks in Figure 3 that are specific to the EMI receiver. All others are covered in 4.2.

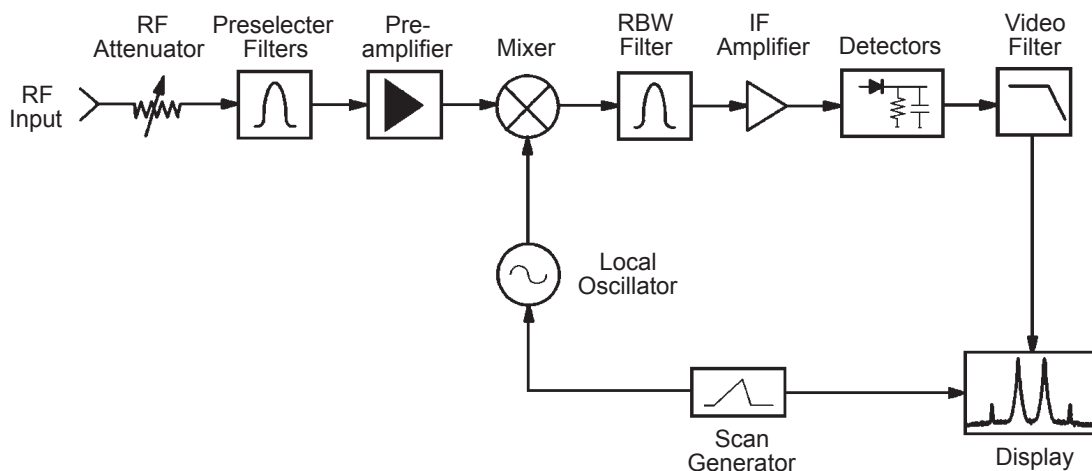


Figure 3—Block diagram of an EMI receiver

4.3.2 RF preselection and preamplification

Both spectrum analyzers and EMI receivers use one or more mixers to convert the unknown signal frequency to the IF frequency. Mixers go into saturation when the input signal exceeds certain levels, as discussed with spectrum analyzers in 4.2. One solution for preventing nonlinear operation is to increase the

⁴Information on references can be found in Clause 2.

input attenuation, which, in turn, reduces the available dynamic range, thus limiting the ability to measure low-level emissions. A better method for measuring broadband signals is to limit the signal spectrum that reaches the mixer to around the desired frequency in order to ensure linear operation. This is accomplished by introducing a preselector filter before the mixer input using either fixed-tuned or variable lowpass and bandpass filters to cover the whole measurement frequency range. The preselection filter passband insertion loss is accounted for by using calibration procedures. Therefore, RF preselection at the input of an EMI receiver enhances the receiver's ability to handle wide dynamic ranges of coherent broadband signals without degrading the dynamic range. Preselection does not prevent the EMI receiver from being overloaded by a signal that happens to be in the passband of a preselector filter. In this case, additional input attenuation needs to be used to avoid mixer saturation.

A low-noise preamplifier, placed after the preselector filter, can be used to improve the EMI receiver's overall system sensitivity. This is particularly useful for radiated EMI measurements in which the test system sensitivity is essentially degraded by the antenna factors of broadband antennas commonly being used.

4.3.3 Specialized IF bandwidths

IEC/CISPR 16-1 (1993-08), specifies IF filter shapes and 6 dB bandwidths that are significantly different from the gaussian shaped filters implemented in most spectrum analyzers. The specialized bandwidths listed in Table 1 should be used during a commercial compliance measurement of DUTs.

Table 1—Specialized IF bandwidths

Characteristic	10–150 kHz	150 kHz to 30 MHz	30 MHz to 1 GHz
6 dB bandwidth	200 Hz	9 kHz	120 kHz
Charge constant (quasi-peak detector)	45 ms	1 ms	1 ms
Discharge constant (quasi-peak detector)	500 ms	160 ms	550 ms
Meter constant (quasi-peak detector)	160 ms	160 ms	100 ms

4.3.4 Specialized detectors

In addition to peak detection, IEC/CISPR 16-1 (1993-08), requires quasi-peak as well as average detection to be used in commercial EMI measurements. The quasi-peak detector was developed to give a reading proportional to the annoyance effect of interference on listeners to broadcast radio. This is accomplished by a circuit that exhibits a time constant and a meter constant, thus simulating the inertia of an analog meter, which provides some smoothing of the readout. These values are listed in Table 1 and are different for the frequency ranges shown. The output of a quasi-peak detector will always decrease with the decrease in the repetition rate of the pulsed emission to be measured. The peak detector output is independent of the repetition rate. Furthermore, the peak detector will always capture the highest signal amplitude, indicating the worst-case value of an emission. The output of a quasi-peak detector might be lower when a broadband signal is measured.

The average detector consists of a peak detector followed by a series-resistor, shunt capacitor averaging network that produces the average of the detected envelope over a certain integration period. An unmodulated sinusoidal signal will produce an average reading proportional to its peak value. For pulsed signals, the average detector produces readings proportional to the duty factor of the signal.

4.4 Vector network analyzers

The minimum network analyzer measurement system consists of a source, a test set, a vector signal processor, and a display. Together, these components comprise a complete excitation/response test system that provides excitation to the DUT and measures the signal transmitted through the device or reflected from its input. The system then processes this data to provide various displays that show the magnitude and phase of these responses. Accuracy enhancement techniques permit measurement calibration at the interface to the DUT, thus minimizing the effect of systematic measurement errors. Figure 4 shows a block diagram of a vector network analyzer system with a swept source providing the excitation to the DUT. A test set provides the points at which the DUT is connected, signal separation devices to measure the four S -parameters, and possibly the first frequency conversion stage. Vector network analyzers usually are heterodyne receivers with multiple frequency conversion. Subsequent filtering allows for reducing the processing bandwidth and, therefore, improving the signal to noise ratio, which ultimately leads to improved sensitivity and dynamic range. Furthermore, the narrowband receiver characteristic suppresses harmonic and spurious signals at other frequencies. The signal processor then provides the amplitude and phase of the desired outputs in amplitude and phase. This information allows the application of sophisticated error-correction techniques that dramatically reduce the measurement ambiguities (see [B13]).

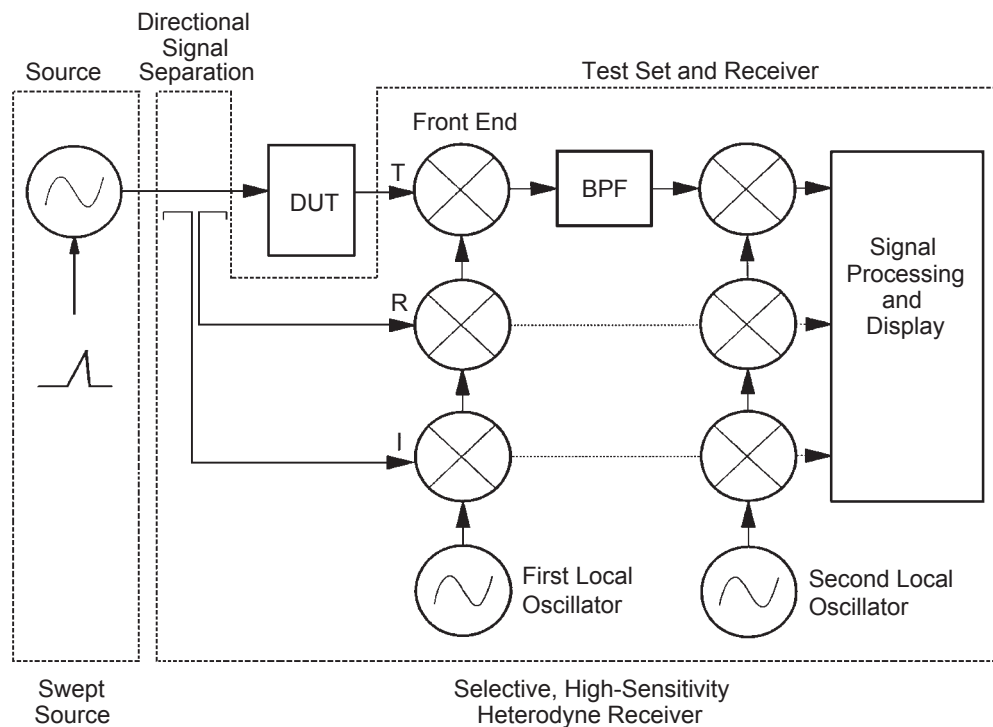


Figure 4—Block diagram of a vector network analyzer

4.5 Scaler network analyzers

The scaler network analyzer, as depicted in Figure 5, measures the magnitude of a microwave test signal by converting it into a low-frequency ac or dc signal by means of a thermoelectric effect or a diode homodyne process. Scaler analyzers with diode detectors as frequency-down converters are inherently broadband receivers. This gives them the capability of measuring any signal in the entire frequency band instantaneously. This feature is mandatory when frequency translating components like mixers have to be characterized, or when high-frequency agility is required. In contrast to a vector network analyzer system, scaler

analyzers offer less sensitivity mainly because of the broadband noise and lower conversion efficiency of the diode converter. Furthermore, there is limited rejection of undesirable signal components, which might lead to erroneous interpretations of measurement results (see [B33]).

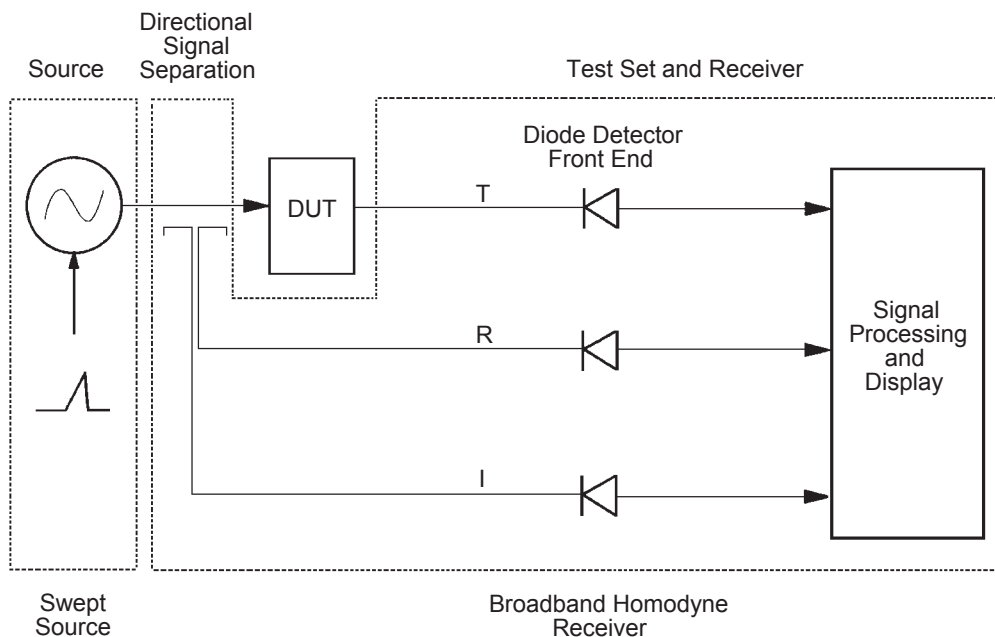


Figure 5—Block diagram of a scalar network analyzer

4.6 Vector voltmeter

4.6.1 Introduction

A vector voltmeter is a dual-channel, autoranging receiver that measures the voltages of two CW signals and the phase difference between them. Any unmodulated RF signal in the frequency range covered by the probes can be used as a source for these measurements, including a source that is part of a DUT. The vector voltmeter's narrowband measuring technique ensures high sensitivity and high dynamic range. The measurements are either of a single channel or of the ratio between the two channels, and the results displayed are as voltage or power, normalized magnitude and phase angle, or real and imaginary components. Measurements can be made either from the input to the output of the DUT or by probing within the device (see [B27]).

4.6.2 Measurement setup

A typical setup for a vector voltmeter measurement is shown in Figure 6. It includes a source (which may be part of the DUT), a reference path, and a measurement path. In order for the source to be suitable for use with a vector voltmeter, it must produce a stable CW signal at the frequency of interest with no modulation. This source signal is fed into a splitter to separate it into a reference signal (channel A) and the excitation for the DUT. The desired DUT response (transmitted or reflected signal) is routed to channel B, which is much more sensitive than channel A. The A input provides the signal to which the instrument is locked, thus serving as a reference to be present at all times for the vector voltmeter to function. Single-channel measurements can be made using the B channel, leaving the A probe in place as a reference. Dual-channel measurements are made by specifying the B/A magnitude ratio or the B – A phase difference as the test parameter. It should be noted that no swept measurements are possible with this setup, since no frequency information is retained along with the actual measurement data by the vector voltmeter itself.

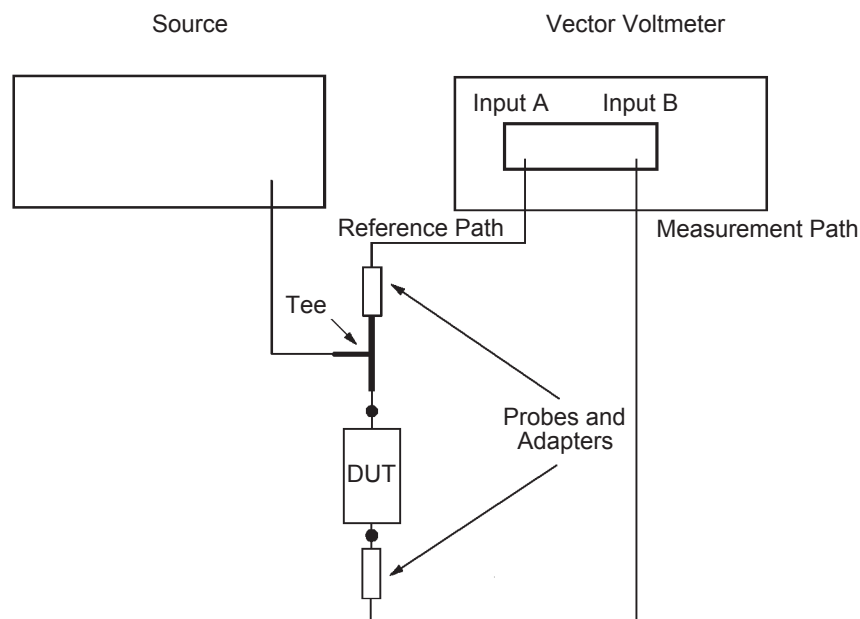


Figure 6—Block diagram of a vector voltmeter test setup

4.7 Time-domain reflectometer

4.7.1 Introduction

Figure 7 shows the basic block diagram of a time-domain reflectometer (TDR). It consists of a very-fast-rise-time pulse generator, an oscilloscope, and a bridging tee with a high-impedance arm toward the oscilloscope. The pulse generator must be well matched at the bridging tee to eliminate reflections back to the signal generator. These reflections would produce a delayed version of the original pulse that would clutter the display if they were to pass again in the forward direction through the bridging tee. There are no directional bridges or couplers to separate incident and reflected waves because they are already separated by time.

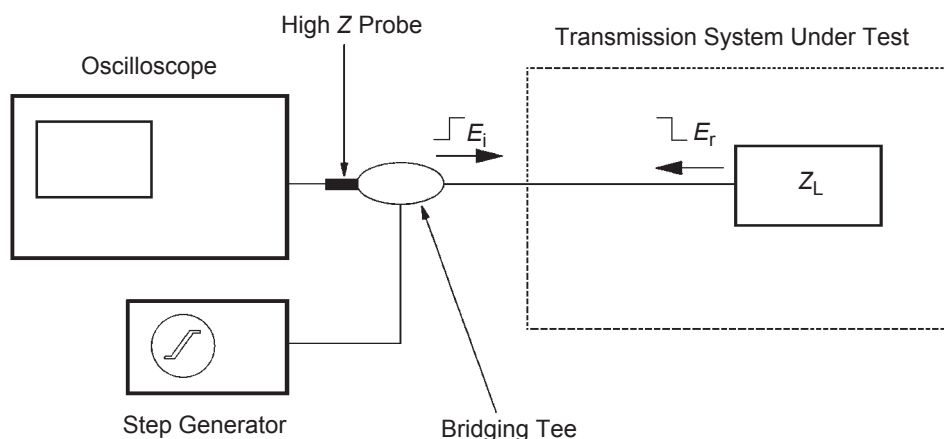


Figure 7—Block diagram of a basic time-domain reflectometer setup