SCTE | STANDARDS

Network Operations Subcommittee

SCTE OPERATIONAL PRACTICE

SCTE 183 2023

SCTE Measurement Recommended Practices For Cable Systems

Fifth Edition

NOTICE

Table of Contents

List of Figures

Title **Page Number**

List of Tables

Title **Page Number**

Acknowledgements

Fifth Edition Revision (2022)

Building on the extraordinary work performed by the cable engineers that participated in the development of the Fourth Edition, the SCTE's Network Operations Subcommittee Working Group 1 (NOS WG1) took on the task of updating the SCTE Measurement Recommended Practices for Cable Systems in order to create its Fifth Edition.

In doing so, numerous references to the Federal Communications Commission's (FCC) Code of Federal Regulations were updated to reflect changes that the FCC has made to its rules in the intervening years since the Fourth Edition, including digital signal quality (i.e., "proof of performance") and signal leakage rules.

In section 12.3 of this Fifth Edition, Signal Leakage, All-Digital System, information describing the availability and operation of digital-compatible, multi-frequency leakage detectors was added, as well as information regarding a promising new technique to measure signal leakage in situations where the upstream bandwidth has been expanded to encompass the 108 MHz to 137 MHz aeronautical band.

In section 14.1, Bandwidth, significant revisions were made to clarify the term "occupied bandwidth" and its meaning in the context of orthogonal frequency division multiplexing (OFDM) and orthogonal frequency division multiple access (OFDMA) signals, as well as legacy DOCSIS single carrier quadrature amplitude modulation (SC-QAM) channels. Other terms related to bandwidth also received revisions in order to provide further clarity.

In chapter 16, a new tutorial was added, providing a fulsome discussion on the DOCSIS 3.1 specification's downstream physical layer technology: OFDM.

Finally, numerous minor editorial changes were made throughout the document in an effort to improve clarity and readability.

SCTE wishes to thank the subject matter experts of NOS Working Group 1 for their hard work and efforts.

Fourth Edition Revision (2011)

As this vibrant industry prepares for the day that classical delivery of the NTSC transport standard is supplanted by analog transmission of digitally modulated signals, the need again exists to provide a revision of this document. It has been 10 years since the release of the Third Edition, which delved somewhat into the digital domain. It has served well the introduction of digital technology, but now must move to yet higher ground.

This revision tackles the challenges facing current and future technical staff with the awesome responsibility, which has been and will continue to be, to keep signals flowing to our customer-base while maintaining an outstanding quality-of-service product. It's a pleasure to report that a number of top cable engineers that participated in the development of the earlier revisions again volunteered their time to review the Third Edition of the *NCTA Recommended Practices for Measurements on Cable Television Systems*. And with the help of several newly recruited, but well recognized engineering talents, a number of procedures and tutorial papers are added to create this, the Fourth Edition of what is now titled the ***SCTE Measurement Recommended Practices for Cable Systems***.

The Fourth Edition exhibits the familiar structure found in the Third Edition. None of the included links is broken and, where relevant, those links point to one or more of the newly developed procedures. Also, this edition continues a practice begun by those performing this task before us. A brief developmental history lesson of earlier revisions follows this page along with those contributing to its success.

As with former versions of this publication, procedures produced continue the legacy of being written for field maintenance personnel with the intent to help them perform their daily jobs. As the technology becomes increasingly complex, this is not an easy task. I am fortunate to have had the opportunity to repeat as chair of the group completing the revision and take untold pleasure in personally thanking the following subcommittee members for their exemplary untiring efforts. Ray Thomas; Authors: Ron Hranac, Bill Morgan, Rich Prodan, Gregg Rogers, and Gerard Terreault; reality checkers: Alan Baumgartner, Rex Bullinger, Jim Farmer, Steve Johnson, Jerry Parkins, Al Silva, Dean Stoneback, and Steve Windle. A very special thanks to Dean Stoneback for preparing the newly minted drafts and to SCTE's Daniel Howard for assembling the well-structured publication.

Dick Shimp

Third Edition (2001)

Repeating past performance, our exciting industry once again reaches a point in technological advance that supports - no forces - rewriting this document. The last major revision effort took place in the many months before the release of the Supplement on Upstream Transportation Issues in October of 1999. Including that contribution, the Second Edition Revised version solidified testing procedures that are quite useful to present day for quantifying analog channel performance.

But, time marches on. The group of cable television engineers that produced the Supplement on Upstream Transportation Issues exposed a small sampling of the digital challenges that would become a major factor in the decision to release a Third Edition. Most of that original group accompanied by some new faces, with interest peaked by the ensuing digital revolution, cautiously tackled the rewrite of the *NCTA Recommended Practices for Measurements on Cable Television Systems*, all the while wrestling with, "How much is too much…where do we stop?" Quite candidly, things move ahead so rapidly, that we might be fully justified never finishing. But, then, what use is the publication? So, we forced ourselves to pick a target and worked feverishly to move to press.

Obviously, many issues faced the Subcommittee; but none so important as ensuring that the Third Edition fully supports the efforts expended to achieve political unity between august parties whose identities can be found in the historical description under Second Edition Revised (1993), immediately following. In fact, we created the entire structure of this Third Edition in that service; providing cross-references where appropriate to achieve total and unambiguous continuity with the Second Edition Revised.

This Third Edition of the *NCTA Recommended Practices for Measurements on Cable Television Systems* continues a legacy begun many years ago by cable engineering Pioneers intent on serving their industry. I am proud to again mention a number of names appearing elsewhere as previous contributors to this publication and to add new names to this list of industry engineers who worked tirelessly to this end. My sincere thanks to authors: Rex Bullinger, Bob Dickinson, John Ernandez, Jim Farmer, Dave Large, Jerry Monroe, Bill Morgan, Roger Pence, Dan Pike, Les Read, Gregg

Rodgers, Thomas Sloane, Dean Stoneback, Gerard Terreault, Dan Van Winkle, Joe Waltrich, and Fred Wilkenloh; to reality checkers: Francis Edgington, Dave Franklin, Ulysses Green, Kent Lack, Oleh Sniezko, Ray Thomas, and Dan Van Winkle; to final reviewers: Syd Fluck and Walt Ciciora. Special thanks to Editor Dean Stoneback who conquered - mastered - used the word processing software to add personality and structure to the publication.

Second Edition Revised (1993)

In the Spring of 1988, delegates of the NCTA, CATA and NATOA (the National Association of Telecommunications Officers and Advisors, which also represented the United States Conference of Mayors, the National League of Cities and the national Association of Counties) met in Washington to explore the possibility of jointly updating the technical requirements governing cable TV.

Over the following three years, a core group of engineers (six representing franchising authorities and the rest representing the cable industry) struggled to identify core issues and develop standards which met the needs of both sides. With support (and an occasional nudge) from John Wong at the Federal Communications Commission, the group hammered out the issues and reached a historic consensus.

That consensus was presented in the form of a document to the FCC in October 1991. Shortly thereafter, the Commission adopted the bulk of our recommendations, and we concluded the first part of our project. With the rules completed, we turned our attention to developing clear and useful testing recommendations designed to meet the proof-of-performance requirements contained in the new rules. An ad hoc subcommittee of cable industry engineers, chaired by Syd Fluck, produced this document after months of careful and thoughtful deliberation.

This revision is designed to clearly describe testing methods while simultaneously teaching the underlying technology issues. The *NCTA Recommended Practices for Measurements on Cable Television Systems* has long been a standard reference on chief technicians' bookshelves. It is, however, by design a living document, subject to future revisions as technology improves.

Jonathan L. Kramer and Wendell Bailey, Co-chairmen - Joint Task Force on Technical Standards would like to acknowledge the engineers who devoted endless hours, over many years, to complete this task. The committee members served, tirelessly, without complaint and frequently without pay or corporate support to help establish the new technical standards. For the franchising authorities: William Pohts, Thomas Robinson, William Kohutanycz, George Grills, and Lee Afflerbach. For the cable television industry: Mike Angi, Alex Best, Jim Hannon, Ted Hartson, Dick Hickman, Steve Johnson, Tom Jokerst, Dan Pike, Frank Ragone, Dave Willis and Nick Worth.

Syd Fluck, chairman of the subcommittee that produced the revision is grateful to all contributors: Rich Annibaldi, Rex Bullinger, James O. Farmer, Joe Glaab, Tom Hill, Steve Johnson, Chris Krehmeyer, Jack Kouzoujian, David Large, Mike Long, Jerry Marnell, Bill Morgan, Dan Pike, Les Read, Gregg Rodgers, and Ray Rohrer.

Second Edition (1989)

Mike Jeffers, the NCTA Engineering Committee Standards Subcommittee chairman, is grateful to all the contributors who made the second edition of this volume possible, especially; Alex Best, Tim Dugan, Randy Crenshaw, Jack Arbuthnott, Brian James, Larry Dolan, Tim Homiller, Richard Kreeger, Bob Sturm, Sally Kinsman and Marshall Hudson. Thanks, too, to the first edition authors Frank Bias, Robert Schoenbeck, and Nicholas Worth and to every member of these committees: the NCTA ad hoc Subcommittee on Multichannel Television Sound, the SCTE Symbols Committee, and

the Network Transmission Committee of the Video Transmission Engineering Advisory Committee (Joint Committee of Television Network Broadcasters and the Bell System).

Format

Part I of this publication appears exclusively to make unambiguous the transition between Edition Two Revised and Edition Three of the *NCTA Recommended Practices for Measurements on Cable Television System*. Chapter 1 reproduces all Technical Standards Tests prescribed by the Federal Communications Commission (FCC) within Title 47 of the Code of Federal Regulations, Part 76 (47 CFR 76.6xx) applicable to the required performance of selected cable television system parameters.

Part II of the publication compiles a comprehensive set of procedures to measure all commonly needed performance parameters within a contemporary hybrid fiber coaxial cable television network. At least one measurement procedure for each performance requirement specified in Part I appears at some location within Part II. To quickly locate the recommended measurement procedure for any given FCC Technical Standards Test Requirement, simply identify the desired measurement parameter in Part I and page to the referenced Part II procedure provided within the Part I Section.

Part III is a place-holder for historical documents of significance to the cable industry and assembles tutorial material that provides detailed background engineering information about selected topics. While material covered within the Part III papers may be discussed in other publications, those appearing here are prepared uniquely to be supplementary to contributed measurement procedures.

Feedback

If you have questions or suggestions, please send them to standards@scte.org

Preamble

Since the first publication in 1983, the various editions of *NCTA Recommended Practices for Measurement on Cable Television Systems* and *SCTE Measurement Recommended Practices* have served as a *defacto* standard for measurements on operational cable systems, whether made by operators, regulators or independent consultants. The Second Edition was created in 1989 with the assistance of engineers from within the cable industry and from the National Association of Telecommunications Officers and Administrators (NATOA). They were subsequently reviewed by FCC staff members and are, in fact, specifically referenced in the FCC's technical rules. The Second Edition, Revision 1, issued in 1993, was reformatted to directly tie some of the procedures to the applicable FCC performance standards as contained in CFR 47 §76.605.

As technology changes and evolves, additional measurement techniques are required. In 1997, the Upstream Supplement to *Recommended Practices* was added. This supplement contained both tutorial material and measurement practices related to upstream (that is, from subscribers towards the headend) signals. The Third Edition of the *Recommended Practices* was issued in 2002 and featured the addition of a number of much needed basic digital parameter measurement procedures. The Third Edition also included a complete content makeover and Table of Contents links to make the document's CD edition navigation more efficient.

With most services now being based on digital signals, such as digital video, high-speed data (HSD) and voice over Internet protocol (VoIP) telephony, and with the number of HSD signals increasing exponentially due to multiple HSD channels being bonded together supporting more than 100 Mbps to DOCSIS 3.0 devices, the industry needs more methodologies to evaluate digital signals, both individually and as a group. Since many operators are building new plants that are all-digital and converting existing plants to all, or mostly-digital signals, a host of new tests and procedures along with revisions to some existing procedures is required. One key purpose of the Fourth Edition of *Recommended Practices* was to further expand on the previous editions to add greater depth and coverage of measurement practices that are appropriate for digital signals of all types. Of particular note to this point is the addition of tests for determining total signal power and all-digital plant leakage. But a potentially game-changing development that has preceded the Fourth Edition and led to new material is the notion that customer premises equipment (CPE) with DOCSIS capabilities could be used to report on plant conditions and be used to proactively maintain the HFC plant. For example, the pre-equalization coefficients that are negotiated between the DOCSIS CPE and the cable modem termination system may be used to identify micro-reflections in the plant and a group of CPE devices can be used to triangulate the location of the source of the micro-reflection. This means that in addition to the usual suite of test equipment, the cable engineer now has virtual probes throughout the network. Given the importance of this development, significant new material was added to the Fourth Edition on the topic of micro-reflections, including a tutorial in support of understanding this process. Other key topics that were added or significantly enhanced include: group delay, hum on digital signals, modulation error ratio (MER), phase noise, and QAM flatness. Again, as cable networks become all, or mostly all-digital in signals, understanding these measurements and topics is critical to the cable professional whose mission is to keep the network running reliably and optimally.

This Fifth Edition includes updates to content in the Fourth Edition, especially references to FCC rules, signal leakage and proof-of-performance testing in digital systems, new tutorials, and more.

Several levels of testing are appropriate for broadband communications systems and the components from which they are constructed. The procedures in this document are primarily geared towards in-service compliance testing, but when an issue is found, it might be due to a faulty component. Out-of-service testing of faulty components is covered by various SCTE standards and operational practices.

Operational testing of in-service cable systems necessarily involves a compromise between absolute accuracy and disruption of services to subscribers. The FCC rules mandate periodic testing to demonstrate compliance with certain technical parameters. The test procedures contained herein have been designed to interrupt normal services as little as possible, consistent with providing an adequate demonstration of technical performance. In some cases, several versions of a procedure are given in order to provide the operator with several choices of degree of service interruption vs. testing flexibility and equipment cost. Any of the procedures which are referenced to specific FCC requirements have been deemed sufficiently accurate for determining compliant performance.

The procedures contained in this edition of *Recommended Practices* are not the only acceptable methods for determining compliance with FCC performance requirements. They do, however, provide a common basis for testing. To the extent that other methods are used, the testing agency or technician has the burden of showing that the method used employs good engineering practice and provides comparable results to those delineated here.

Important Note on FCC Regulations

SCTE's NOS WG1 notes that the FCC's Code of Federal Regulations (CFR) Title 47, section 76.609 paragraphs (e) and (j) reference "*NCTA Recommended Practices for Measurements on Cable Television Systems*, Second edition, November 1989". SCTE's NOS WG1 has made every effort to transfer procedures from the previous edition to this fifth edition, without material change and with editorial change only where realizing improved clarity. The material in this publication best serves those persons experienced in the cable television technology that have knowledge of equipment calibration requirements coupled with principles that follow good engineering practices observed within the cable industry. Definitions and procedures within this publication do not replace other lawful compliance procedures imposed by enforcement agents of the FCC or alternatives of a franchising authority.

For the most current revision of the FCC Technical Standards, see the Code of Federal Regulations, Title 47 at: https://www.fcc.gov/encyclopedia/rules-regulations-title-47 .

Before You Use This Recommended Practice

Throughout this manual you will notice references to "a properly operating headend." While several of the recommended practices will help you understand the operating parameters of the typical cable headend, this manual cannot specify the exact levels and adjustments needed to put *your* individual headend(s) into proper operation -- there are too many possible combinations of equipment, service offerings and even chief engineers' operating philosophies affecting these levels and adjustments. Use available information and expertise [e.g., vendors' manuals, cable operator engineering or your system chief technician's operating procedures text, and your own system logs, etc.] to establish these base parameters.

It is vital that you are aware of the proper alignment procedures and level adjustments for your particular headend **before** you use these recommended practices for measurements.

Abbreviations and Definitions

AC	alternating current
ALC	automatic level control
AGC	automatic gain control
AM	amplitude modulation
AML	amplitude-modulated microwave links
APL	Average picture level
ATIS	automatic transmitter identification system
ATSC	Advanced Television Systems Committee
ATT	attenuator
AWG	American wire gauge
AWGN	additive white Gaussian noise
BER	bit error rate or bit error ratio
BERT	bit error ratio tester or bit error ratio analyzer
BNN	bit error ratio in the noise notch
BPSK	binary phase shift keying
BTSC	Broadcast Television Systems Committee
CNR	carrier-to-noise
cm	centimeter
CATA	Cable Telecommunications Association
CCIR	International Radio Consultative Committee
CCN	carrier-to-composite noise
CIN	carrier-to-intermodulation noise
CMTS	cable modem termination system
CLI	cumulative leakage index
CPD	common-path distortion
CSO	composite second order
CTB	composite triple beat
CTN	carrier-to-thermal noise
CV	cluster variance
CW	continuous wave
dB	decibel
dBm	decibel milliwatt
dBmV	decibel millivolt
dBW	decibel watt
DC	direct current
DCT	digital consumer terminal
DFB	distributed feedback
DFE	decision feedback equalizer
DFI	discrete frequency interference
DISP	discrete interfering signal probability
DIV.	division
DMM	digital multimeter
DSA	digital signal analyzer
DUT	device under test
DVM	digital volt meter
DVS	Digital Video Subcommittee of the SCTE

ECL	emitter coupled logic
EIRP	effective isotropic radiated power
EVM	error vector magnitude
FCC	Federal Communications Commission
FEC	forward error correction
FET	field effect transistor
FFE	feed forward equalizer
FFT	fast Fourier transform
FM	frequency modulation
FP	Fabry-Perot
FSL	free space loss
HBI	horizontal blanking interval
HFC	hybrid fiber-coaxial (or hybrid fiber/coax)
HRC	harmonically related carriers
ICPM	incidental carrier phase modulation
IEC	International Electrotechnical Commission
IF	intermediate frequency
IIN	Interferometric intensity noise
IMN	intermodulation noise
IRC	incrementally related carriers
IRE	Institute of Radio Engineers
ISI	intersymbol interference
ITU	International Telecommunications Union
ksps	kilosymbols per second
LNB	low noise block converter
MER	modulation error ratio
MTS	multichannel television sound
ms	millisecond
µs	microsecond
mV	millivolt
mW	milliwatt
NATOA	National Association of Telecommunications Officers and Advisors
NCTA	Formerly National Cable Television Association and National Cable & Telecommunications Association, now known as NCTA – The Internet & Television Association
NIST	National Institute of Standards and Technology
NPR	noise power ratio
NTC	Network Transmission Committee of the Video Transmission Engineering Advisory Committee
NTSC	National Television System Committee
p-p	peak-to-peak
PHY	physical layer
PRBS	pseudorandom bit stream; also pseudo random binary sequence
PSK	phase shift keying
PSD	power spectral density
QAM	quadrature amplitude modulation
QPSK	quadrature phase shift keying
RF	radio frequency
RFI	radio frequency interference

RIN	relative intensity noise
rms	root-mean-square
RSS	root-sum-of-squares
SA	spectrum analyzer
scope	oscilloscope
SCTE	Society of Cable Telecommunications Engineers
SINAD	signal-to-noise and distortion ratio
SLM	signal level meter
SNR	signal-to-noise ratio
SSO	single second order
SSPA	solid state power amplifiers
STO	single third order
TDR	time domain reflectometry or time domain reflectometer
T.P.	test point
TTL	transistor-transistor logic
TWT	traveling wave tube
TWTA	traveling wave tube amplifier
VBI	vertical blanking interval
VCO	voltage controlled oscillator
VITS	vertical interval test signal
VSB	vestigial sideband
XMOD	cross modulation

References

The following documents might provide valuable information to the reader.

SCTE References

[SCTE 03]	ANSI/SCTE 03 2016, Test Method for Coaxial Cable Structural Return Loss
[SCTE 17]	ANSI/SCTE 17 2018, Test Procedure for Carrier-to-Noise (C/N, CCN, CIN, CTN)
[SCTE 23-1]	ANSI/SCTE 23-1 2017 (R2022), DOCSIS 1.1 Part 1: Radio Frequency Interface
[SCTE 25-1]	ANSI/SCTE 25-1 2017 (R2022), Hybrid Fiber/Coax Outside Plant Status Monitoring - PHY Layer
[SCTE 40]	ANSI/SCTE 40 2016 (R2021) , Digital Cable Network Interface Standard
[SCTE 47]	ANSI/SCTE 47 2007, Test Method for Coaxial Cable Attenuation
[SCTE 55-2]	ANSI/SCTE 55-2 2019, Digital Broadband Delivery System Out-of-Band Transport – Mode B
[SCTE 55-1]	ANSI/SCTE 55-1 2019,Digital Broadband Delivery System Out-of-Band Transport – Mode A
[SCTE 220-1]	ANSI/SCTE 220-1 2022, DOCSIS® 3.1 Physical Layer Specification
[SCTE 270]	SCTE 270 2021, Mathematics of Cable
[SCTE 258]	SCTE 258 2020, DOCSIS 3.1 Downstream OFDM Power Definition, Calculation, and Measurement Techniques

Standards from Other Organizations

[ITU-T J.83]	ITU-T J.83 2007, Digital Multi-Programme Systems for Television Sound and Data Services for Cable Distribution

Other Published Materials

[Prodan]	Prodan, R., L. Montreuil, A. Kliger, BZ Shen, "Demystifying the DOCSIS 3.1 PHY," SCTE Cable-Tec Expo 2014 Proceedings
[Hranac}	Hranac, R., Society for Broadband Professionals video lecture: "DOCSIS 3.1 – An Overview" (https://www.youtube.com/watch?v=ZwKuPZVh19Y)
[OFDM]	https://en.wikipedia.org/wiki/Orthogonal_frequency-division_multiplexing
[Math}	https://www.mathworks.com/discovery/ofdm.html
[Digital Comm]	Digital Communications, Fourth Edition, by John G. Proakis. © 2001, McGraw Hill, ISBN 0-07-232111-3 (see chapter 12)

[Patent]	U.S. Patent Number: 5,815,488 “MULTIPLE USER ACCESS METHOD USING OFDM” https://patentimages.storage.googleapis.com/bf/c6/b6/c1730e77076099/US5815488.pdf
[Transmission]	Transmission Systems for Communications, Fifth Edition, © 1982, Bell Telephone Laboratories, Inc.
[IEEE]	Pancaldi, F., Vitetta, G. M., Kalbasi, R., Al-Dhahir, N., Uysal, M., Mheidat, H., “Single-Carrier Frequency Domain Equalization,” IEEE Signal Processing Magazine (Volume: 25, Issue: 5, September 2008)

PART I FCC TECHNICAL STANDARDS

Chapter 1 FCC Technical Standards Tests

The Code of Federal Regulations (CFR) §76.609 paragraph (a), describes example measurement methods using the text "[S]ome exemplary, but not mandatory, measurement procedures are set forth in this section." The technical standards below specifically spell out the end result, but they do not specify the method. They only offer examples. The FCC expects these specific results to be obtained using good engineering practice and with defensible methods, but not necessarily strictly 'word for word' as written in either §76.609(a) or in the SCTE Recommended Practices.

1.1 FCC §76.601(b)(2)(i) – Testing Conditions

FCC §76.601(b)(2)(i): ***The operator of each cable television system that operates NTSC or similar channels shall conduct semi-annual proof-of-performance tests of that system, to determine the extent to which the system complies with the technical standards set forth in section 76.605 (b) (4) as follows. The visual signal level on each channel shall be measured and recorded, along with the date and time of the measurement, once every six hours (at intervals of not less than five hours or no more than seven hours after the previous measurement), to include the warmest and the coldest times, during a 24-hour period in January or February and in July or August.***

> See Section 2.1: "Analog Signal Power – Visual, Aural Carrier Level, 24 Hour Variation [FCC §76.605(b)(3), (4), (5)]".

1.2 FCC §76.605(b)(2) – Aural Carrier Center Frequency

FCC §76.605(b)(2): ***The aural center frequency of the aural carrier must be 4.5 MHz, ±5 kHz above the frequency of the visual carrier at the output of the modulating or processing equipment of the cable television system, and the subscriber terminal.***

> See Section 7.2: "Aural Carrier Center Frequency [FCC §76.605(b)(2)]".

1.3 FCC §76.605(b)(3) – Visual Carrier Signal Level

FCC §76.605(b)(3): ***The visual signal level, across a terminating impedance which correctly matches the internal impedance of the cable system as viewed from the subscriber terminal, shall not be less than 1 millivolt across an internal impedance of 75 Ω (0 dBmV). Additionally, as measured at the end of a 30 meter (100 foot) cable drop that is connected to the subscriber tap, it shall not be less than 1.41 millivolts across an internal impedance of 75 Ω (+3 dBmV). (At other impedance values, the minimum visual signal level, as viewed from the subscriber terminal, shall be the square root of 0.0133(Z) millivolts and, as measured at the end of a 30 meter (100 foot) cable drop that is connected to the subscriber tap, shall be 2 times the square root of 0.00662(Z) millivolts, where Z is the appropriate impedance value).***

> See Section 2.1: "Analog Signal Power – Visual, Aural Carrier Level, 24 Hour Variation [FCC §76.605(b)(3), (4), (5)]".

1.4 FCC §76.605(b)(4) – 24 Hour Signal Variation

FCC §76.605(b)(4): ***The visual signal level on each channel, as measured at the end of a 30 meter cable drop that is connected to the subscriber tap, shall not vary more than 8 decibels within any 6-month interval which must include four tests performed in six hour increments during a 24-hour period in July or August and during a 24-hour period in January or February, and shall be maintained within:***

(i) 3 decibels (dB) of the visual signal level on any visual carrier within a 6 MHz nominal frequency separation;

(ii) 10 dB of the visual signal level on any other channel on a cable television system of up to 300 MHz of cable distribution system upper frequency limit, with a 1 dB increase for each additional 100 MHz of cable distribution system upper frequency limit (e.g. 11 dB for a system at 301-400 MHz; 12 dB for a system at 401-500 MHz, etc.); and

(iii) A maximum level such that signal degradation due to overload in the subscriber's receiver orterminal does not occur.

See Section 2.1: "Analog Signal Power – Visual, Aural Carrier Level, 24 Hour Variation [FCC §76.605(b)(3), (4), (5)]".

1.5 FCC §76.605(b)(5) – Aural Carrier Signal Level

FCC §76.605(b)(5): ***The rms voltage of the aural signal shall be maintained between 10 and 17 decibels below the associated visual signal level. This requirement must be met at the subscriber terminal and at the output of the modulating and processing equipment (generally the headend). For subscriber terminals that use equipment which modulate and remodulate the signal (e.g. baseband converters), the rms voltage of the aural signal shall be maintained between 6.5 and 17 decibels below the associated visual signal level at the subscriber terminal.***

See Section 2.1: "Analog Signal Power – Visual, Aural Carrier Level, 24 Hour Variation [FCC §76.605(b)(3), (4), (5)]".

1.6 FCC §76.605(b)(6) – Frequency Response

FCC §76.605(b)(6): ***The amplitude characteristic shall be within a range of ± 2 decibels from 0.75 MHz to 5.0 MHz above the lower boundary frequency of the cable television channel, referenced to the average of the highest and lowest amplitudes within these frequency boundaries. The amplitude characteristic shall be measured at the subscriber terminal.***

See Section 6.1: "Analog Television In-Band Response [FCC §76.605(b)(6)]".

1.7 FCC §76.605(b)(7) – Visual Carrier-to-Noise

§76.605(b)(7): ***The ratio of RF visual signal level to system noise shall not be less than 43 decibels. For class I cable television channels, the requirements of this section are applicable only to:***

(i) Each signal which is delivered by a cable television system to subscribers within the predicted Grade B contour or noise-limited service contour, as appropriate, for that signal;

(ii) Each signal which is first picked up within its predicted Grade B contour or noise-limited service contour, as appropriate;

(iii) Each signal that is first received by the cable television system by direct video feed from a TV broadcast station, a low power TV station, or a TV translator station.

See Section 3.2: " Visual Carrier-to-Composite Noise Ratio [FCC §76.605(b)(7)]".

1.8 FCC §76.605(b)(8) – Coherent Disturbances

FCC §76.605(b)(8): ***The ratio of visual signal level to rms amplitude of any coherent disturbances such as intermodulation products, second and third order distortions or discrete-frequency interfering signals not operating on proper offset assignments shall be as follows:***

(i) The ratio of visual signal level to coherent disturbances shall not be less than 51 decibels for noncoherent channel cable television systems, when measured with modulated carriers and time averaged; and

(ii) The ratio of visual signal level to coherent disturbances which are frequency-coincident with the visual carrier shall not be less than 47 decibels for coherent channel cable television systems, when measured with modulated carriers and time averaged.

See Section 4.1: "Coherent Disturbances – CSO, CTB and XMOD [FCC §76.605(b)(8)]".

1.9 FCC §76.605(b)(9) – Subscriber Terminal Isolation

FCC §76.605(b)(9): ***The terminal isolation provided to each subscriber terminal:***

(i) Shall not be less than 18 decibels. In lieu of periodic testing, the cable operator may use specifications provided by the manufacturer for the terminal isolation equipment to meet this standard; and

(ii) Shall be sufficient to prevent reflections caused by open-circuited or short-circuited subscriber terminals from producing visible picture impairments at any other subscriber terminal.

See Section 12.1: "Subscriber Terminal Isolation [FCC §76.605(b)(9)]".

1.10 FCC §76.605(b)(10) – Low Frequency Disturbances

FCC §76.605(b)(10): ***The peak-to-peak variation in visual signal level caused by undesired low frequency disturbances (hum or repetitive transients) generated within the system, or by inadequate low frequency response, shall not exceed 3 percent of the visual signal level. Measurements made on a single channel using a single unmodulated carrier may be used to demonstrate compliance with this parameter at each test location.***

See Section 4.4: “Low Frequency Disturbances and Hum”.

1.11 FCC §76.605(b)(11) – Color Parameters

FCC §76.605(b)(11): ***The following requirements apply to the performance of the cable television system as measured at the output of the modulating or processing equipment (generally the headend) of the system:***

(i) The chrominance-luminance delay inequality (or chroma delay), which is the change in delay time of the chrominance component of the signal relative to the luminance component, shall be within 170 nanoseconds.

See Section 10.2: "Chrominance - Luminance Delay Inequality [FCC §76.605(b)(11)(i)]".

(ii) The differential gain for the color subcarrier of the television signal, which is measured as the difference in amplitude between the largest and smallest segments of the chrominance signal (divided by the largest and expressed in percent), shall not exceed ±20%.

See Section 10.3: "Differential Gain and Phase [FCC §76.605(b)(11)(ii), (iii)]".

(iii) The differential phase for the color subcarrier of the television signal which is measured as the largest phase difference in degrees between each segment of the chrominance signal and reference segment (the segment at the blanking level of 0 IRE), shall not exceed ±10 degrees.

See Section 10.3: "Differential Gain and Phase [FCC §76.605(b)(11)(ii), (iii)]".

1.12 FCC §76.605(c) – Signal Leakage

FCC §76.605(c): ***As an exception to the general provision requiring measurements to be made at subscriber terminals, and without regard to the type of signals carried by the cable television system, signal leakage from a cable television system shall be measured in accordance with the procedures outlined in §76.609(h) and shall be limited as as shown in table 1 to paragraph (c):***

Table 1 to Paragraph (c)

Frequencies	Signal leakage limit	Distance in meters (m)
Analog signals less than and including 54 MHz, and over 216 MHz	**15 µV/m**	**30**
Digital signals less than and including 54 MHz, and over 216 MHz	**13.1 µV/m**	**30**
Analog signals over 54 MHz up to and including 216 MHz	**20 µV/m**	**3**
Digital signals over 54 MHz up to and including 216 MHz	**17.4 µV/m**	**3**

See Section 12.2: "Signal Leakage [FCC §76.605(c)]".

PART II MEASUREMENT PROCEDURES

Chapter 2 Signal Power

2.1 Analog Signal Power – Visual, Aural Carrier Level, 24 Hour Variation [FCC §76.605(b)(3), (4), (5)]

Definition: Visual carrier level is the rms voltage produced by the visual signal during the transmission of synchronizing pulses, measured across the 75 Ω internal impedance of the cable system. The visual carrier level is normally expressed in decibels with respect to one millivolt rms in a 75 Ω system (dBmV). Aural carrier level is the rms voltage of a channel's aural (sound) carrier measured across the 75 Ω internal impedance of the cable system and is generally expressed with reference to the channel's associated visual carrier level.

FCC §76.601(b)(2)(i): ***The operator of each cable television system that operates NTSC or similar channels shall conduct semi-annual proof-of-performance tests of that system, to determine the extent to which the system complies with the technical standards set forth in section 76.605 (b) (4) as follows. The visual signal level on each channel shall be measured and recorded, along with the date and time of the measurement, once every six hours (at intervals of not less than five hours or no more than seven hours after the previous measurement), to include the warmest and the coldest times, during a 24-hour period in January or February and in July or August.***

FCC §76.605(b)(3): ***The visual signal level, across a terminating impedance which correctly matches the internal impedance of the cable system as viewed from the subscriber terminal, shall not be less than 1 millivolt across an internal impedance of 75 Ω (0 dBmV). Additionally, as measured at the end of a 30 meter (100 foot) cable drop that is connected to the subscriber tap, it shall not be less than 1.41 millivolts across an internal impedance of 75 Ω (+3 dBmV). (At other impedance values, the minimum visual signal level, as viewed from the subscriber terminal, shall be the square root of 0.0133(Z) millivolts and, as measured at the end of a 30 meter (100 foot) cable drop that is connected to the subscriber tap, shall be 2 times the square root of 0.00662(Z) millivolts, where Z is the appropriate impedance value).***

FCC §76.605(b)(4): ***The visual signal level on each channel, as measured at the end of a 30 meter cable drop that is connected to the subscriber tap, shall not vary more than 8 decibels within any 6-month interval which must include four tests performed in six hour increments during a 24-hour period in July or August and during a 24-hour period in January or February, and shall be maintained within:***

> ***(i) 3 decibels (dB) of the visual signal level on any visual carrier within a 6 MHz nominal frequency separation;***
>
> ***(ii) 10 dB of the visual signal level on any other channel on a cable television system of up to 300 MHz of cable distribution system upper frequency limit, with a 1 dB increase for each additional 100 MHz of cable distribution system upper frequency limit (e.g., 11 dB for a system at 301-400 MHz; 12 dB for a system at 401-500 MHz, etc.); and***
>
> ***(iii) A maximum level such that signal degradation due to overload in the subscriber's receiver or terminal does not occur.***

FCC §76.605(b)(5): ***The rms voltage of the aural signal shall be maintained between 10 and 17 decibels below the associated visual signal level. This requirement must be met at the subscriber terminal and at the output of the modulating and processing equipment (generally the headend). For subscriber terminals that use equipment which modulate and remodulate the signal (e.g., baseband converters), the rms voltage of the aural signal shall be maintained between 6.5 and 17 decibels below the associated visual signal level at the subscriber terminal.***

Note: The choice of the term dBmV to define cable television signal levels has caused confusion over the years. Since the name ends in "V", it is natural to assume it refers to a voltage, but by definition, the impedance is 75 Ω and 0 dBmV is equal to 1 millivolt across 75 Ω, or 13.33 nW. Therefore, the measure of dBmV is also a power ratio with a reference of 13.33 nW. The reference value of 0 dBmV is equal to -48.75 dBm.

Required Equipment

- A signal level meter (SLM) or spectrum analyzer
- A precision step attenuator, which is normally part of the spectrum analyzer or SLM
- A subscriber drop cable (30 meters)
- Set-top converter typical of units used in test point region

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Figure 2-1 is a block diagram showing proper setup of the equipment used in this procedure.
3. Record the make, model number and most recent date of calibration of each unit of test equipment used for the test.
4. Fine tune the SLM to the visual carrier to be measured and, if applicable, adjust the SLM compensator as shown on its calibration chart. If using a spectrum analyzer, adjust the center frequency to the frequency of the carrier being measured, adjust the span to 1 MHz and adjust the resolution bandwidth (RBW) and video bandwidth (VBW) to 300 kHz.
5. Insert or remove attenuation from the SLM precision attenuator until the SLM reads within its linear region on the dB scale. Again, fine tune the SLM, if necessary, to find the peak amplitude of the channel's visual carrier. If using a spectrum analyzer, adjust the full-scale reference level to place the carrier in the upper two divisions of the display.
6. Record the measured compensated visual carrier level. If using the spectrum analyzer, this will require placing a marker on the peak of the displayed trace, or adjusting the full scale reference to position the peak of the carrier at the display reference line. Record the air temperature, time and date of the measurement.
7. Now fine tune the SLM to the channel's associated aural carrier and remove attenuation from the precision attenuator until the SLM again reads within its linear region on the dB scale. Fine tune again if necessary to find the peak amplitude of the channel's aural carrier. If using a spectrum analyzer, adjust the center frequency to the frequency of the aural carrier being measured. All other settings may remain the same.
8. Compute and record the aural carrier level with respect to its associated visual carrier level.

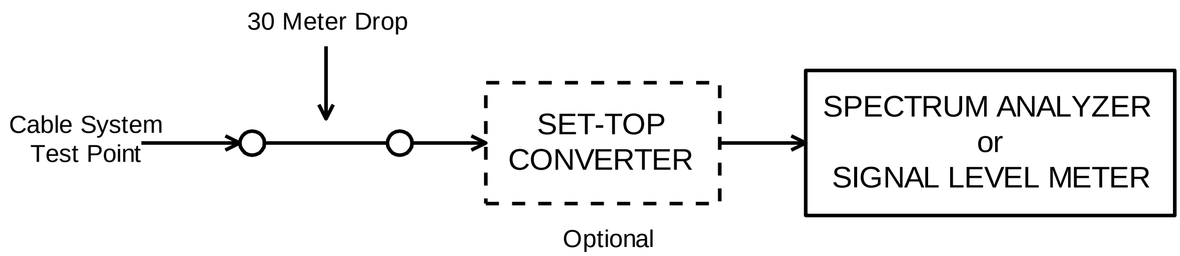


Figure 2-1: Visual, Aural Carrier Level: 24 Hour Variation Test Equipment Setup

Test Methodology

A. At the headend: Repeat the procedure, from step 4 through step 8, for all active cable system channels at the headend. (§76.605(b)(5))

B. At each test point: Repeat the procedure, from step 4 through step 8, for all active cable system channels for each of the system test points at the end of the 30m cable drop. (§76.605(b)(3))

C. At each test point: Repeat the procedure, from step 4 through step 8, for each of the system's test points at the output of the converter terminating the end of the 30m cable drop. The SLM should be tuned to the visual and aural carriers produced by the cable converter and testing performed by tuning the converter to each active cable system channel. (§76.605(b)(3)), (See note 4)

D. At each test point: Re-measure and record all visual carriers (only) at the end of the 30m cable drop using step 4 through step 6 at six-hour intervals three more times (total of four tests in 24 hours). It is only necessary to measure aural carrier levels during the first of the four test intervals. (§76.605 (b)(4), (§76.601(c))

Notes, Hints and Precautions

1. In most cases it is not practical to make signal level measurements on scrambled channels. For this reason, the recommended procedure is to turn off scrambling while making the measurement. When this is not possible, such as during a 24 hour test, it may be possible to measure the peak carrier level by observing several frames. Most analog scrambling systems will maintain a peak video level that is fairly constant, provided that there is at least some black content in the picture. It is necessary to observe the measurement result long enough to capture the peak reading, which may depend on picture content. In addition, the relative level compared to adjacent unscrambled channels may or may not be measured accurately when the picture is scrambled. This can be determined at the headend by measuring signal level with and without scrambling turned on. These issues are explained thoroughly in Section 16.11: “Notes Concerning the Use of Scrambled Channels for Signal Level Measurement”.
2. Signal level meters and spectrum analyzers with multi-channel capability and built-in chart recorders or data output ports for computer interface may automate the signal level measuring process.
3. An appropriate 75 Ω to 50 Ω minimum loss pad must be used with a 50 Ω impedance spectrum analyzer and the measurement result must be adjusted for the loss of the pad.
4. Baseband converters remodulate any tuned channel and generally produce fixed levels for visual and aural carriers independent of tuned channel levels. If a baseband converter or an RF converter incorporating volume control or AGC is used, it is acceptable to record the video carrier level and the aural carrier level on one channel, and note any variation (if any) caused by changing channels. It should be verified that the FCC visual and aural level regulations are met allowing for the converter contribution. If it is established and documented that the converter (of

any type) has at least 0 dBmV of output visual carrier level and the proper aural carrier level with an input signal of the same characteristics, field converter tests for signal level measurements do not have to be performed. Similarly, all types of non-volume control converters should be characterized for their effect on relative aural carrier level and confirmation that the cumulative effect of converters and plant variation meets the FCC requirement.

5. Twenty-four hour variation of a particular visual carrier is the decibel difference between the maximum and minimum levels of the carrier over any 24 hour period.

2.2 Digital Signal Power

Definition: Digital signal power is the power level as measured by a power meter which uses a thermocouple as a transducer. That is, the measurement is the average power in the signal, integrated over the actual occupied bandwidth of that signal. The signal power is normally expressed in decibels with respect to one millivolt rms in a 75 Ω system (dBmV). Thus, the measurement reported is the rms value of the sinusoid that would produce the same heating in a 75 Ω resistor as does the actual signal.

In the case of a burst signal, such as a signal that occupies one assigned time slot in a time division multiple access (TDMA) sequence of time slots, the power reported shall be the equivalent power level as if the signal being measured (any one of the bursts included in the total sequence) was on continuously.

Note: The choice of the term dBmV to define cable television signal levels has caused confusion over the years. Since the name ends in “V”, it is natural to assume it refers to a voltage, but by definition, the impedance is 75 Ω and 0 dBmV is equal to 1 millivolt across 75 Ω, or 13.33 nW. Therefore, the measure of dBmV is also a power ratio with a reference of 13.33 nW. The reference value of 0 dBmV is equal to -48.75 dBm.

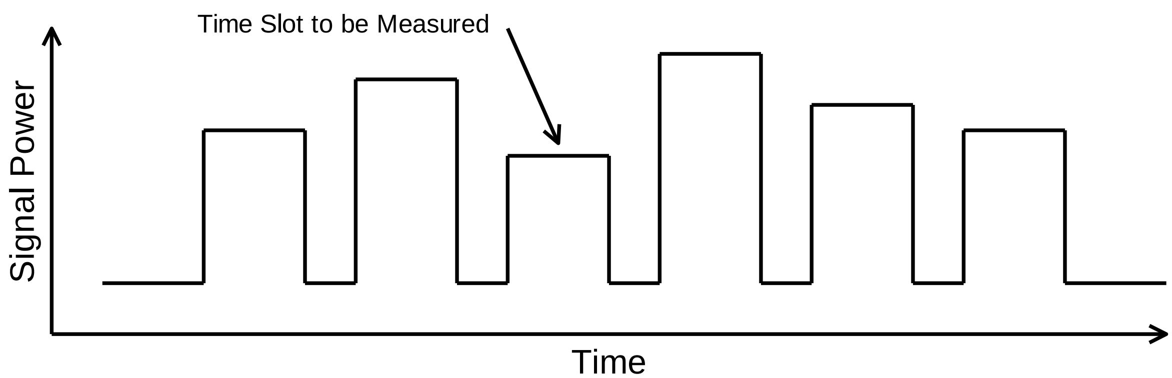


Figure 2-2: Time Slot in a TDMA Signal

Discussion: It is the responsibility of each manufacturer of a signal measuring device to publish a valid procedure that allows a user to translate the power reported by the device to the power according to the above definition. This procedure should include correction factors, if required, to allow the user to measure signals in less than their fully occupied bandwidths and to correct for the occupied bandwidth.

The manufacturer should provide information to allow a user to measure the power of one signal element in a sequence of signal elements in a TDMA signal, regardless of the presence or absence of other elements of the signal sequence. To aid in measuring power levels of burst signals, the manufacturer of a TDMA system should provide at the TDMA receiver, a signal which may be used to synchronize a measuring instrument to any one element of the returning signal sequence. This may be done on a frame basis such that the measuring instrument will setup a delay to the desired signal element. Alternatively, it may be done such that the synchronizing signal appears just prior to the receipt of the signal element to be measured.

This section will describe the proper techniques for measuring digital signal power using spectrum analyzers and digital signal level meters. Measurement of the digital signal power will be the average power within the 3 dB bandwidth (i.e., the symbol bandwidth) of the signal. This section applies to satellite (QPSK), over-the-air-air (VSB) and cable in-band (QAM) and out-of-band (QPSK and FSK) signals. Modulation formats, data rates and symbol bandwidths for these signal formats are shown in Table 2-1. For additional information concerning data rates, Section 14.1: “Bandwidth”.

Modulation Format	Total Data Rate (Mb/s)	Symbol BW (MHz)
ANSI/SCTE 25-1 2017 (R2022) Hybrid Fiber Coax Outside Plant Status Monitoring - Physical (PHY) Layer Specification v1.0	0.04	0.038
L-band QPSK – 24 MHz signal bandwidth	39.02	19.510
L-band QPSK – 36 MHz signal bandwidth	58.53	29.266
QPSK Out-of-Band signal per ANSI/SCTE 55-1 2019, Table 1	2.05	1.024
QPSK Out-of-Band signal per ANSI/SCTE 55-2 2019, Table 2-2	1.54	0.772
QPSK Out-of-Band signal per ANSI/SCTE 55-2 2019, Table 2-2	3.09	1.544
64-QAM	30.34	5.057
256-QAM	42.88	5.361
8-VSB	32.12	5.381

Table 2-1: Digital Signal Formats and Bandwidths

2.2.1 Continuous Signals

Discussion: Three procedures for measuring “noise-like” digitally modulated signals are provided:

1. Manual Single Point Measurement - measures a single point in the middle of the channel and corrects for occupied bandwidth. This measurement does not require any special functions in the analyzer.
2. Power Spectral Density (PSD) - similar to the manual single point measurement, but uses the “noise marker” function within the analyzer which compensates for the IF resolution bandwidth (RBW) and noise measurement errors. Some spectrum analyzers sample data from points around the marker frequency, not just at the marker, increasing the accuracy of the measurement.
3. Automated Measurement - uses automated measurement capability within the analyzer to integrate samples across the entire occupied bandwidth of the channel. This should be the most

accurate measurement approach since it eliminates errors due to frequency flatness problems of the channel being measured.

Many automated digital channel power measurements are now available in lower cost signal level meters. These are all instrument specific and not discussed here. It is up to the user to determine the approach used in these meters and the potential accuracy.

Note: FSK signals are included in this section since they are indeed digitally modulated signals, but they are not "noise-like" since the modulation is two predictable states. For example, the 67 kHz modulation bandwidth of the FSK signal defined in ANSI/SCTE 25-1 2017 (R2022), Hybrid Fiber Coax Outside Plant Status Monitoring – Physical (PHY) Layer Specification v1., is narrow enough to fall within the 300 kHz measurement bandwidth and does not require a correction factor.

PROCEDURE 1 - Manual Single Point Measurement

Required Equipment

- Spectrum Analyzer

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Set the spectrum analyzer as follows:
   - Center Frequency: Center frequency of channel under test
   - Frequency Span: 20 to 30% wider than the channel's occupied bandwidth (see Note 1 below)
   - IF Resolution Bandwidth: 300 kHz
   - Video Bandwidth: ≤ 30 kHz, or until desired display stability is reached
     and/or
   - Number of Video Averages: 100, or until desired display stability is reached
   - Amplitude Scale: 10 dB/div
   - Detector Mode: Sample, if available (see Note 2 below)
   - Reference Level: Position signal trace in upper two divisions of analyzer display
   - Sweep Time: Automatic for calibrated measurement

Note 1: The recommended span settings are as follows:

- 30 MHz or 40 MHz for L-band signals
- 8 MHz for QAM signals transmitted in a 6 MHz channel
- 3 MHz for out-of-band signals such as those defined in Table 2-1

Note 2: The peak of noise-like signals will exceed its power average by an amount that increases (on average) with the length of time the peak is observed. The sample detector is required for making noise measurements with a digital spectrum analyzer to prevent this increase in the measurement result.

3. Using the spectrum analyzer marker, record the signal power at the center of the channel. This signal power is the correct result for the FSK signal defined in ANSI/SCTE 25-1 2017 (R2022). For wider bandwidth signals, proceed to step 4.
4. The power of the digital signal can now be calculated using the following equation:

$$\text{Channel Power} = \text{Marker Value} + K_{RBW} + K_{SA}$$

where:

$$K_{RBW} = 10 * \log\left[\frac{\text{Symbol BW}}{\text{RBW}}\right]$$

The correction factor for the channel's symbol bandwidth is relative to the measurement bandwidth. See Table 2-2 below.

$K_{SA} =$ the correction factor for the analyzer noise measurement which includes:

- under response due to voltage envelope detection
- under response due to log detection
- over response due to the ratio of the equivalent noise bandwidth and the -3 dB bandwidth of the IF resolution bandwidth filter.

K_{SA} is normally a positive number and should be supplied by the spectrum analyzer manufacturer.

Modulation Format	**Symbol BW (MHz)**	**BW Correction (dB)**
ANSI/SCTE 25-1 2017 (R2022) Hybrid Fiber Coax Outside Plant Status Monitoring - Physical (PHY) Layer Specification v1.0	0.038	0.0
L-band QPSK – 24 MHz signal bandwidth	19.510	18.1
L-band QPSK – 36 MHz signal bandwidth	29.266	19.9
QPSK Out-of-Band signal per ANSI/SCTE 55-1 2019, Table 1	1.024	5.3
QPSK Out-of-Band signal per ANSI/SCTE 55-2 2019, Table 2-2	0.772	4.1
QPSK Out-of-Band signal per ANSI/SCTE 55-2 2019, Table 2-2	1.544	7.1
64-QAM	5.057	12.3
256-QAM	5.361	12.5
8-VSB	5.381	12.5

Table 2-2: Correction Factors for 300 kHz RBW Filter

Notes:

The accuracy of this procedure will be affected by the flatness of the signal across its occupied bandwidth. If for example, the signal being measured was 2 dB down at both edges of the modulation envelope, and had relatively flat slope from the center to these edges, the error could be as much as 1 dB. If the signal were 1 dB high at one edge of the channel, and 1 dB low at the other edge (a simple slope down with increasing frequency for example) the errors cancel themselves and the measurement at the middle of the channel gives the correct answer.

The symbol bandwidth of the transmitted signal is the bandwidth which would have been occupied had the signal been filtered by a rectangular filter instead of the smooth filter actually used. Since the filter used is a root-raised-cosine filter, the portion of the signal that is removed by the smooth filter skirts inside the 3 dB points is exactly the same amount as the power that passes through the filter skirts beyond the 3 dB points. The equivalent symbol bandwidth is therefore the 3 dB bandwidth.

PROCEDURE 2 - Power Spectral Density (PSD)

Required Equipment

- Spectrum Analyzer with “Marker Noise” function

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Set the spectrum analyzer as follows:
   - Center Frequency: Center frequency of channel under test
   - Frequency Span: 20 to 30% wider than the channel’s occupied bandwidth (see Note 1 below)
   - IF Resolution Bandwidth: 300 kHz
   - Video Bandwidth: ≤ 30 kHz, or until desired display stability is reached
     and/or
   - Number of Video Averages: 100, or until desired display stability is reached
   - Amplitude Units: 10 dB/div
   - Detector Mode: Sample, if available (see Note 2 below)
   - Reference Level: Position signal trace in upper two divisions of analyzer display
   - Sweep Time: Automatic for calibrated measurement

Note 1: The recommended span settings are as follows:

- 30 MHz or 40 MHz for L-band signals
- 8 MHz for QAM signals transmitted in a 6 MHz channel
- 3 MHz for out-of-band signals

Note 2: The peak of noise-like signals will exceed its power average by an amount that increases (on average) with the length of time the peak is observed. The sample detector is required for making noise measurements with a digital spectrum analyzer to prevent this increase in the measurement result.

3. Enable the Marker Noise function per the analyzer manufacturer's instructions. Place the marker at the center frequency of the channel under test. The Marker Noise function will perform the necessary corrections automatically and provide a power measurement in units of dBmV/Hz. This is a power spectral density (PSD) measurement.
4. Calculate the total power in the signal's symbol bandwidth using the following equation:

$$\text{ChannelPower} = \text{Marker Value} + K_{\text{Symbol BW}}$$

where

$$K_{\text{Symbol BW}} = 10 * \log(\text{SymbolBW})$$

the correction factor for the symbol bandwidth. See Table 2-3 below

Modulation Format	Symbol BW (MHz)	Correction (dB)
L-band QPSK – 24 MHz signal bandwidth	19.510	72.9
L-band QPSK – 36 MHz signal bandwidth	29.266	74.7
QPSK Out-of-Band signal per ANSI/SCTE 55-1 2019, Table 1	1.024	60.1
QPSK Out-of-Band signal per ANSI/SCTE 55-2 2019, Table 2-2	0.772	58.9
QPSK Out-of-Band signal per ANSI/SCTE 55-2 2019, Table 2-2	1.544	61.9
64-QAM	5.057	67.0
256-QAM	5.361	67.3
8-VSB	5.381	67.3

Table 2-3: Correction Factors for PSD Measurements

Example: If the measured PSD of a 64-QAM signal is -61 dBmV/Hz, the total power is equal to:

$$\text{ChannelPower} = -61.0 + 67.0 = +6.0\ \text{dBmV}$$

Notes

The accuracy of this procedure will be affected by the flatness of the signal across the channel's occupied bandwidth. If for example, the signal being measured was 2 dB down at both edges of the modulation sidebands, and had relatively flat slope from the center to these edges, the error could be as much as 1 dB. If the signal were to be 1 dB high at one edge of the channel, and 1 dB low at the other edge (a simple slope down with increasing frequency for example) the errors cancel themselves and the measurement at the middle of the channel gives the correct answer.

The symbol bandwidth of the transmitted signal is the bandwidth which would have been occupied had the signal been filtered by a rectangular filter instead of the smooth filter actually used. Since the filter used is a root-raised-cosine filter, the portion of the signal that is removed by the smooth filter skirts inside the 3 dB points is exactly the same amount as the power that passes through the filter skirts beyond the 3 dB points. The equivalent symbol bandwidth is therefore the 3 dB bandwidth.

PROCEDURE 3 - Automated Measurement

Required Equipment

- Spectrum analyzer or signal level meter with an automated measurement that integrates the power across the full occupied bandwidth.

Test Procedure

Instruments with automated digital channel power measurement capability should make all corrections automatically including any corrections needed for a channel that is not spectrally flat.

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Tune the analyzer or signal level meter to the channel under test
3. Measure the channel power per the instructions provided by the manufacturer of the test equipment.

Note: Some analyzers or signal level meters may display calculated results from a single point measurement instead of integrated power across the band and hence do not correct for flatness. They merely perform a calculation instead of integrating between markers. It is the user's responsibility to understand the measurement algorithm used by the chosen test equipment and the potential measurement errors.

2.2.2 Burst Signals

Discussion: For the discussion of the measurement of burst signals, the signals will be separated into two basic types.

Narrowband Modulation refers to intermittent signals from multiple sources with a modulation bandwidth narrower than the resolution bandwidth of the measuring device. These are typically Out-of-Band FSK type set-top or remote monitoring signals and may be measured using normal peak hold methods.

Wideband Modulation refers to intermittent signals from multiple sources with a modulation bandwidth wider than the resolution bandwidth of the measuring device. Accurate measurement of wideband signals requires the mathematical integration of multiple sample points across the occupied bandwidth of the signal.

The procedures presented here will not distinguish between signals from multiple signal sources in a sequence and will typically measure the level of the highest-powered carrier in the sequence. In order to measure the level of a signal from a specific source in a TDMA signal, a gating trigger is required which is unique to that signal. This gating signal is currently not available from most headend data receivers. Ideally, the specific signal to be measured should be isolated prior to measurement.

Four different measurement approaches are discussed in the following procedures. Procedures 1 and 2 use readily available test equipment without IF or video gating capability. The disadvantage of this approach is that the procedure is more difficult to optimize for minimization of measurement errors and will normally produce a less accurate result. Procedure 3 leverages a characteristic of the DOCSIS specification and measures during the ranging preamble of the upstream signal, providing a more repeatable and stable measurement technique. The last procedure discusses spectrum analyzers with various types of triggered gating capability allowing the use of dedicated average power measurement algorithms internal to the analyzers. This last procedure is the most accurate but also requires referencing the test equipment manufacturer's measurement procedures due to the uniqueness of each approach. This document has attempted to summarize the key steps to the measurement.

This document does not discuss some of the newer test equipment currently available which has specific burst measurement algorithms available. It is up to the user to determine the viability and accuracy of these approaches.

Each procedure discussed below recommends the use of an optional channel selection filter. This filter is used to reduce the total power into the analyzer in order to minimize distortion in the analyzer's front end when measuring low level signals in the presence of higher-level signals. This filter is highly recommended when the upstream path is loaded with many different types of signals at differing levels and the user is attempting to measure one of the lower level signals.

2.2.2.1 Non-Gated Measurements

PROCEDURE 1 - Burst Measurements with a Signal Level Meter:

A signal level meter (with a peak hold function) may be used to measure a single point at the center frequency of the burst signal and the result adjusted for bandwidth and additional corrections associated with the use of the peak detector in the meter. If the signal is nearly flat across its passband and the proper correction factor is used, this method is capable of providing a result within ±2 dB.

Required Equipment

- Signal level meter with a known IF resolution bandwidth

Optional Equipment

- Bandpass filter for channel selection

Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Set the signal level meter as follows:
    - Center Frequency: Channel center frequency
    - IF Resolution Bandwidth: Maximum
    - Detector Mode: Peak Hold
    - Video Bandwidth: Maximum
3. Allow enough time for the peak of the signal to be captured.
4. Record this level which is the measured level.
5. Adjust the measured level using the following equation:

$$\text{Channel Power (dBmV)} = CP_{Meas} + \left[10 * \log\left(\frac{BW_{OCC}}{BW_{Meas}}\right)\right] + K$$

where:

CP_{Meas} = measured level (dBmV)

BW_{OCC} = occupied bandwidth of the burst signal (Hz)

BW_{Meas} = IF resolution bandwidth of the signal level meter (Hz)

K = correction factor provided by signal level meter manufacturer (dB)

Note: This correction factor (K) should compensate for the following discrepancies and will be unique for each type of modulation:

- peak detector used for an average power measurement
- log amplifier response when measuring noise like signals
- IF resolution bandwidth noise equivalent bandwidth
- video filter

This correction factor should be provided by the manufacturer of the test equipment used, or may be generated by the user using the following procedure.

1. Measure a reference continuous digital carrier using the above procedure.
2. Measure the same carrier using a calibrated average reading power meter.
3. The difference between the two measurements is K.

K = reference average power meter result - SLM result

Ideally, this calibration should be done on each meter since the RBW, VBW, and peak detector response will vary from unit to unit.

It should be noted that some signal level meters and spectrum analyzers are currently available with automated burst power measurements and provide an accurate integrated measurement on burst signals. These instruments simplify this procedure and should be used following the manufacturer's recommendations.

PROCEDURE 2 - Burst Measurements with a Spectrum Analyzer:

The spectrum analyzer allows the IF RBW to be set to the optimum bandwidth for the width of the carrier being measured and provides an accurate measurement of the skirts of the carrier. This optimum IF resolution bandwidth is between 1/20 and 1/100 of the signal's occupied bandwidth. By triggering a zero span spectrum analyzer on the detected signal (in sample detection mode), the power spectral density of the pulse can be measured. Using the normalized 1 Hz power density (using the marker noise function available on many analyzers) and the 3 dB bandwidth of the signal as an approximate noise equivalent bandwidth of the signal, fairly accurate approximations of the channel power can be made.

In order to make this measurement, the user must have an approximate knowledge of the burst pulse width being measured, the burst repetition rate (the number of times the pulse occurs per second) and the channel's occupied bandwidth.If the burst repetition rate is not known, it can be measured by observing the burst signal in zero span on the spectrum analyzer while adjusting the sweep speed until at least two burst pulses are displayed in single sweep. If multiple signal sources are present in the same channel, it may be possible to set the analyzer to trigger on the largest signal. For simplicity, this procedure assumes only one signal source is present. Once at least two burst pulses are displayed on the screen, the spectrum analyzer markers can be used to measure the time between pulses. This is the burst repetition rate.

Required Equipment

- Spectrum Analyzer with a variable resolution bandwidth, noise marker functionality, and known frequency response.

Optional Equipment

- Bandpass filter for channel selection

Procedure

1. Setup the equipment as shown in Figure 2-3.

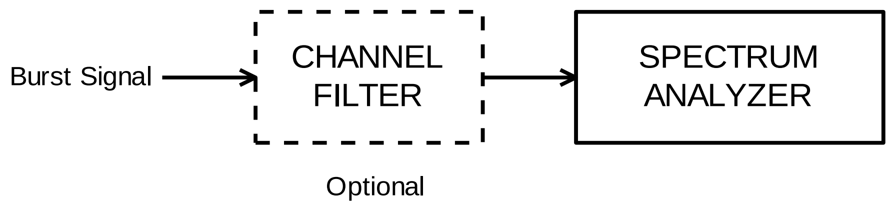


Figure 2-3: Spectrum Analyzer Measurement Setup

2. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

3. Set the analyzer as follows to measure the 1 Hz channel power:

• Center Frequency:	Channel center frequency
• Frequency Span:	0 MHz
• Sweep Time:	$\approx 2 \times \tau_W$
• IF Resolution Bandwidth:	$\approx \frac{4}{\tau_W}$
• Video Bandwidth:	Maximum
• Detector Mode:	Sample, if available (see Note 1 below)
• Trigger Mode:	Video

where

$\tau_W =$ the minimum burst pulse width (second)

Note 1: The peak of noise-like signals will exceed its power average by an amount that increases (on average) with the length of time the peak is observed. The sample detector is required for making noise measurements with a digital spectrum analyzer to prevent this increase in the measurement result.

4. Adjust the trigger level until the pulse can be clearly seen and the screen is continually triggering. The pulse will occupy approximately half of the horizontal display screen when capturing the minimum pulse width. Varying pulse widths will cause the portion of the screen to the right of the minimum pulse to fluctuate more in level as averaging progresses.
5. Set the video averages to 500 (or value sufficient to obtain display stability) and wait for the averaging to complete.
6. Place a marker on top of the displayed pulse and select the marker noise function. Note that the marker value is in dBmV/Hz. This value (CP_{Hz}) should be recorded for use in Step 9.
7. Readjust the analyzer for the following settings to measure the channel's occupied bandwidth:
   - Frequency Span: $\approx 2 \times BW_{OCC}$
   - IF Resolution Bandwidth: $\approx \frac{2}{\tau_W}$ (round down to the next available value)
   - Trigger Mode: Free Run
   - Video Averaging: Off
   - Sweep Time: $\geq \frac{400}{\tau_R}$
   - Detector Mode: Peak Detect

   where:

   $\tau_W =$ burst pulse width (second)

   $\tau_R =$ burst repetition rate (Hz)

   $BW_{OCC} =$ occupied bandwidth (MHz)
8. Turn the video averaging back on.

   (**Note**: some analyzers may need to be reset to peak detect mode).
9. Determine the approximate 3 dB bandwidth of the resultant signal.

   The approximate channel power can be calculated from:

$$\text{Channel Power (dBmV)} = CP_{Hz} + 10 * \log(BW3dB_{OCC})$$

   where:

   $CP_{Hz} =$ the noise power measured with the marker in step 6.

   $BW3dB_{OCC} =$ the 3 dB occupied bandwidth (Hz) measured in step 9.

This method can be used for pulse times as short as 4 µs and is typically accurate to within ±2 dB depending on the spectrum analyzer used.

PROCEDURE 3 - Burst Measurements using the DOCSIS Preamble:

Note: The following procedure was originally created for DOCSIS 1.x upstream signals, but has not been tested on DOCSIS 2.0 and 3.0 signals.

The DOCSIS 1.1 Radio Frequency Interface Specification, ANSI/SCTE 23-1 2017 (R2022) requires a variable-length preamble field which is prepended to the data stream after randomization and Reed-Solomon encoding. At the time the Third Edition was released in 2002, this preamble, although specific to each CMTS receiver, normally began with an alternating pattern mapped onto the QPSK or 16-QAM constellation points. Due to the bit pattern selected, only two constellation points are used and the signal appears as BPSK modulation during this initial portion of the transmission.

This BPSK portion of the preamble provides a predictable signal to use for an accurate and repeatable measurement. In addition, in a fully loaded signal environment a narrow IF resolution bandwidth may be used and the analyzer tuned to one of the BPSK side lobes of the signal to measure the signal quickly and reliably. The length of this BPSK sequence is variable, but the receiver requires at least 20 symbols for reliable acquisition and the chip manufacturers recommend 64 symbols. Typically, this BPSK portion of the preamble will last for about 60 μs.

1. Setup the equipment as shown in Figure 2-3.
2. If you know the occupied channel bandwidth of the signal being measured, proceed to step 6.
3. Set the analyzer as follows to measure the channel's occupied bandwidth:
   - Center Frequency: Channel center frequency
   - Frequency Span: 4 MHz
   - Sweep Time: 20 ms
   - IF Resolution Bandwidth: 100 kHz
   - Video Bandwidth: Maximum
   - Detector Mode: Peak
   - Trigger Mode: Video

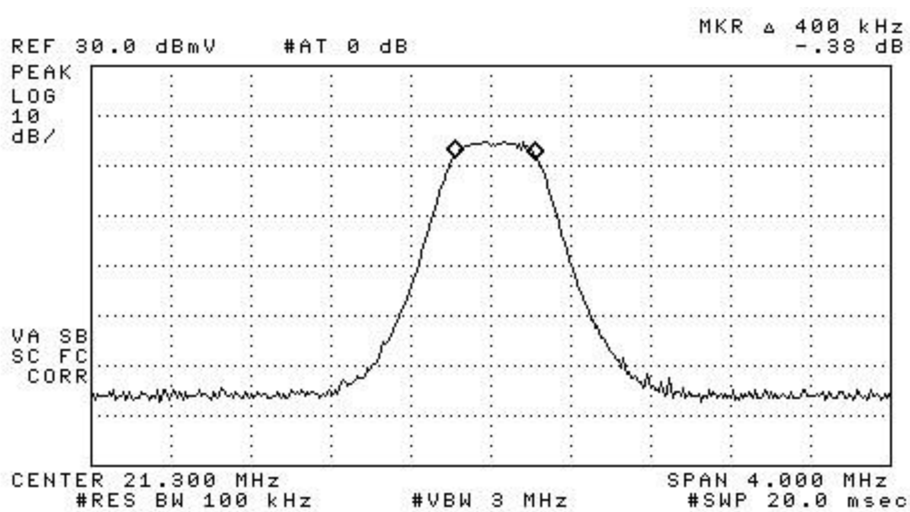


Figure 2-4: Measuring Occupied Channel Bandwidth

4. Measure the 3 dB bandwidth of the detected signal using the spectrum analyzer's markers. See Figure 2-4. Determine the channel's 1/2 symbol rate from Table 2-4.

Occupied Channel Bandwidth	Symbol Rate (ks/s)	1/2 Symbol Rate (ks/s)	300 kHz IF RBW Correction (dB)
200 kHz	160	80	0.0
400 kHz	320	160	+2.0
800 kHz	640	320	+3.0
1.6 MHz	1280	640	+3.0
3.2 MHz	2560	1280	+3.0

Table 2-4: DOCSIS Occupied Channel Bandwidth vs. Symbol Rate and IF Resolution Bandwidth Correction Factor

5. Adjust the analyzer for the following settings to measure the channel power during the preamble:
    - Center Frequency: Channel center frequency - 1/2 Symbol Rate (kHz)
    - Frequency Span: 0 MHz
    - Sweep Time: 100 µs
    - IF Resolution Bandwidth: 300 kHz
    - Video Bandwidth: Maximum
    - Detector Mode: Peak
    - Trigger Mode: Video
6. Adjust the trigger level until the pulse can be clearly seen, and the screen is continually triggering.
7. Place a marker on a signal peak in the preamble portion of the displayed pulse. See Figure 2-5.

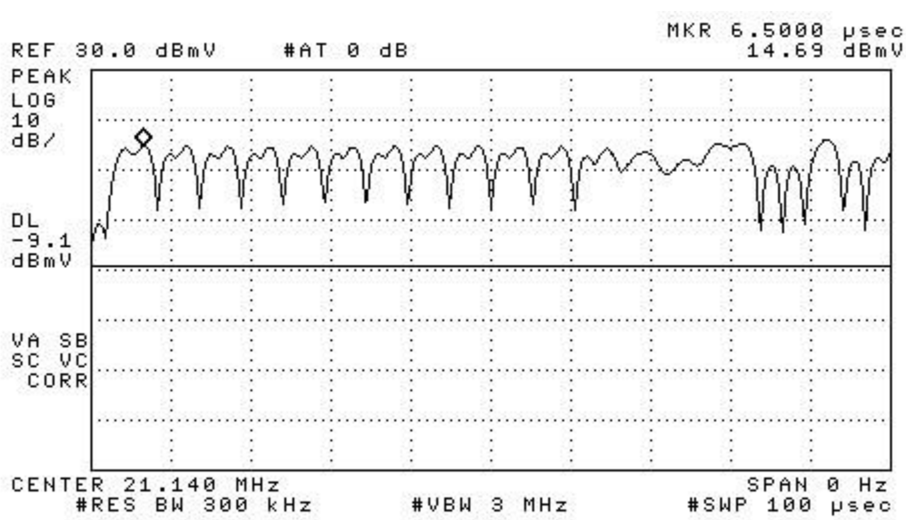


Figure 2-5: Preamble Channel Power Measurement

8. In the example above:
    - Channel center frequency = 21.300 MHz
    - Channel occupied bandwidth = 400 kHz
    - Channel symbol rate = 320 ks/s

- Measurement frequency = 21.300 - 0.160 = 21.140 MHz
- Channel Power = 14.69 + 2.0* = 16.69 dBmV *(see Note below)

Note: At low symbol rates (≤ 320 ks/s) both modulation sidebands will be present within the IF Resolution Bandwidth of the spectrum analyzer. In this procedure, as the symbol rate increases, the upper modulation sideband will move out of the IF Resolution Bandwidth filter. Therefore, the magnitude of the lower sideband being measured will be dependent upon the symbol rate and the rejection of the upper sideband by the IF Resolution Bandwidth filter. Table 1 provides a typical correction factor to the measured result; in this example we have used 2.0 dB for 320 ks/s.

9. The above procedure assumes that the signal being measured has occupied adjacent channels. If this is not the case, a wider resolution bandwidth may be used and the channel power measured at the channel's center frequency. The only change is that the spectrum analyzer is tuned to the center frequency of the channel and a 3 MHz IF resolution bandwidth is used, instead of 300 kHz. In this case, 3 dB does not need to be added to the marker reading and the channel power is measured directly. See Figure 2-6.
10. In Figure 2-6 the channel power is read directly from the marker as 18.03 dBmV.

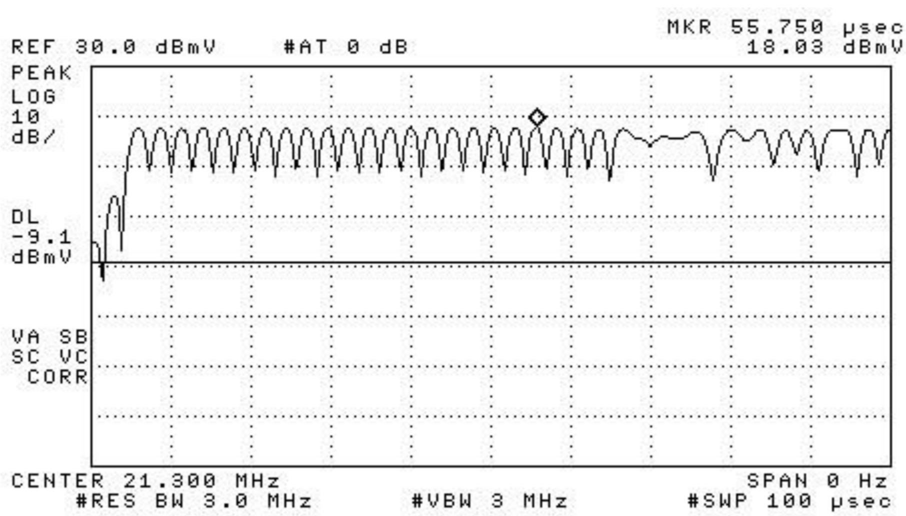


Figure 2-6: Preamble Channel Power Measurement without Adjacent Channels

2.2.2.2 Gated Measurements

There are several different methods that may be used to measure burst signals if a spectrum analyzer is used with video gating capability. Video gating connects the detected RF to the sampling circuitry in the analyzer for a user-controlled period of time (the gate time). The trigger for this gating signal can be generated internally by the analyzer's own video circuitry, or may be generated externally and connected to the analyzer, typically via a rear panel connector. Internal video gating circuitry is relatively new and, therefore, the procedure for making a burst measurement with video gating tends to be analyzer specific. It is our recommendation that the user contact the analyzer's manufacturer for their specific burst carrier measurement procedure. What we will do here is summarize a typical gated procedure to help describe the process.

A narrow video gate can be used to approximate sample detection if the length of the gate (in µs) is short relative to the occurrence of modulation peaks of the signal. In other words, the gate must be short enough that an amplitude peak of the signal is not captured with each sample. With a noise like signal, a gate length of < 20 μs will provide results very close to true sample detection. By carefully adjusting the spectrum analyzer sweep time, resolution bandwidth, video bandwidth and video gating parameters, a very narrow video gate can be used to effectively sample detect a burst RF carrier.

On some analyzers, the video gating will work with the mathematically integrated average power measurement, providing a more accurate result. Currently, video gating is only available on higher performance spectrum analyzers. Once again, it is recommended that the user contact the manufacturer of the user's specific analyzer for the manufacturer's recommended procedure.

Figure 2-7 is an example of a typical measurement setup for a burst measurement using video gating. In some configurations the burst carrier trigger generator is internal to the analyzer. The burst carrier trigger generator detects the envelope of the signal and provides a TTL compatible signal to the gate trigger input of the analyzer when a user adjustable threshold is exceeded.

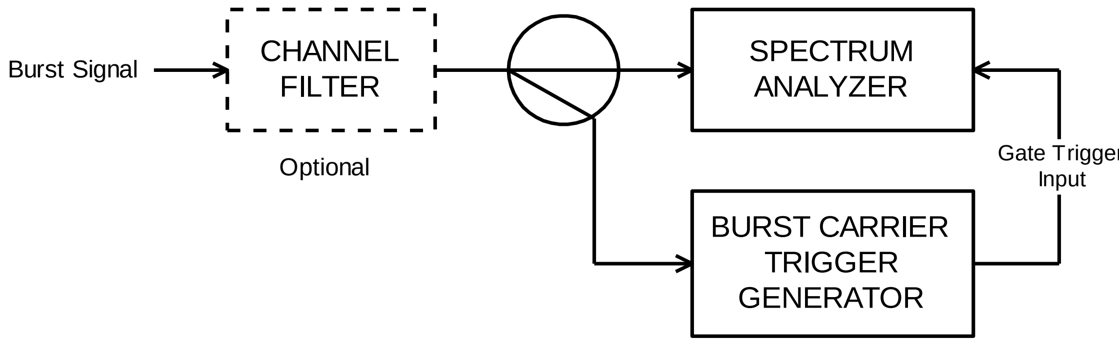


Figure 2-7: Typical Gated Spectrum Analyzer Measurement Setup

The most common burst measurement procedures use the analyzer's average power measurement and the video gating and/or video sweep triggering to enable the detector sampling of the analyzer during the burst pulse on time. Measurement methods which trigger the sweep circuits of the analyzer from the video pulse are more effective when measuring signals with varying pulse repetition rates. Typically, the automated procedures will provide the mathematically integrated average power measurement result directly. The operator's manual of the analyzer should be consulted for specific steps.

2.2.3 Peak-to-Average Ratio

Discussion: The peak-to-average ratio of digitally modulated signals is important to the overall headroom needed to transmit these through the distribution system. The average power needs to be set high enough to provide sufficient margin above the levels of system impairments to assure reliable operation. The peaks of the signal can, if it is set too high, clip the laser or other devices (amplifiers) in the system.

Generally, the peak value reported should be that value at which the signal dwells for a long enough time that the transmission of a single symbol of any other digital signal may be affected (by clipping in any portion of the transmission system). For discussion purposes it is suggested to consider any peak that occurs for a time longer than 20% of a single symbol time for any digital signal which could be clipped.

It is the responsibility of the supplier of a system which employs any type of digital modulation, to publish the peak-to-average power ratio of that signal in a 75 Ω environment. The supplier should make the average power measurement using a meter that responds as defined in Section 2.2: Digital Signal Power.

Accurate measurement of this ratio requires specialized equipment; however, this parameter is not in itself adjustable. This, therefore, is not a parameter that needs to be measured in an operating system but it can be calculated.

The manufacturer should specify the peak-to-average ratio of the digital modulator in use. This parameter is a function of the modulation method used and the filtering applied within the modulator. If the peak-to-average ratio is incorrect, it is an indication of a much more serious distortion such as a modulator problem, severe clipping or severe channel amplitude flatness distortion.

Each of the modulation types may have different symbol rates for a given occupied bandwidth. (For example, QAM and QPSK have a symbol rate approximately equal to the occupied bandwidth, and VSB has a symbol rate approximately twice the occupied bandwidth.)

It is sufficient verification of the peak-to-average ratio to simply check the bit error ratio of the channel, or the constellation measurements (such as Modulation Error Ratio (MER), Error Vector Magnitude (EVM), Cluster Variance (CV), or similar measurements).

These measurements will be quite sensitive to any distortions that can produce an incorrect peak-to-average power ratio.

One may find it useful to specify the percentage of time the signal remains above certain "peak" values to allow for statistical studies of clipping.

Required Equipment

- Spectrum Analyzer with Max Hold and Marker capability.

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Set the spectrum analyzer as follows:
   - Center Frequency: Center of channel under test
   - Frequency Span: Wide enough to see entire signal under test (see Note below)
   - IF Resolution Bandwidth: 300 kHz
   - Video Bandwidth: 30 kHz or lower, or until display stability is reached (30 Hz is preferred)
   - Reference Level: Position top of signal trace approximately 10 dB below the top of the analyzer graticule
   - Number of Video Averages: Until display stability is reached

Note: The recommended span settings are as follows:

- 30 MHz to 40 MHz for L-band signals
- 10 MHz for QAM signals transmitted in a 6 MHz channel

- 3 MHz for out-of-band signals

3. Set the marker to the center frequency of the channel. Record the average value of the digital signal.
4. Turn Video Averaging OFF. Select and activate the analyzer's MAX HOLD function. The amplitude of the digital signal will gradually increase. After about 5 minutes, set the marker to the center frequency and record the value. The difference in dB between this value and the marker reading in step 3 when the signal was averaged is the peak-to-average ratio

2.3 Total Signal Power

Definition: Total signal power or total power is the combined RF power of all signals in a given bandwidth or spectrum, as measured by a power meter which uses a thermocouple as a transducer. That is, the measurement is the average power of all of the signals present, integrated over the actual occupied bandwidth or spectrum of those signals. While total signal power may be expressed in watts, it is normally expressed in decibels with respect to 1 millivolt RMS in a 75 Ω system (dBmV). Thus, the measurement reported is the RMS value of the sinusoid that would produce the same heating in a 75 Ω resistor as does the actual group of signals.

In the case of burst signals, such as signals that occupy assigned time slots in a time division multiple access (TDMA) sequence of time slots, the power reported shall be the equivalent power as if the signals being measured were on continuously and simultaneously.

Note: The choice of the term dBmV to define cable television signal levels has caused confusion over the years. Since the name ends in "V", it is natural to assume it refers to a voltage, but by definition, the impedance is 75 Ω and 0 dBmV is equal to 1 millivolt across 75 Ω, or 13.33 nW. Therefore, the measure of dBmV is also a power ratio with a reference of 13.33 nW. The reference value of 0 dBmV is equal to -48.75 dBm.

PROCEDURE 1 – Manual Measurement

Discussion: When the power of each signal is expressed in watts, total signal power is simply the sum of the individual signal power levels:

$P_{TOTAL\ (watts)} = P1_{(watts)} + P2_{(watts)} + \ldots + Pn_{(watts)}$

Since signal power in cable networks is normally expressed in dBmV, one must convert each signal's power in dBmV to power in watts, add the individual values, then convert the total power in watts to total power in dBmV. While this is a viable method to determine total signal power, it is unnecessarily cumbersome. In a simplified method of calculation, the total signal power in dBmV is equal to the base 10 logarithm of the sum of the anti-logs of the individual RF power levels in dBmV within a given bandwidth or spectrum. Simply measure per-signal RF power levels in dBmV and then combine those measurement results using one of the equations in test procedure step 3 or step 4.

Required Equipment

- Spectrum analyzer, digital signal analyzer, or signal level meter capable of measuring both analog TV channel visual carrier peak envelope power and digital channel power.

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Measure and record the power of each signal in dBmV within the desired bandwidth or spectrum, using the appropriate procedure(s) detailed in Sections 2.1 and 2.2 above.

3. If all of the signals have identical RF power levels, total signal power may be calculated with the following equation:

   $$P_{TOTAL(dBmV)} = P_{SINGLE(dBmV)} + 10\log_{10}(N)$$

   where:

   $P_{TOTAL(dBmV)}$ is the total signal power in dBmV

   $P_{SINGLE(dBmV)}$ is the RF power in dBmV of an individual signal

   N is the total number of signals

4. When the signals being measured have different RF power levels, total signal power may be calculated with the following equation.

   $$P_{TOTAL(dBmV)} = 10\log_{10}[10^{(P1/10)} + 10^{(P2/10)} + \ldots + 10^{(Pn/10)}]$$

   where:

   $P_{TOTAL(dBmV)}$ is the total signal power in dBmV

   P1, P2…Pn represent the RF power in dBmV of each signal being measured

Note: In practical applications where per-signal RF power levels for a large number of signals are unequal, such as would be the case where the power levels are characterized by the sum-of-squares example in Section 4, it will be quite tedious to manually measure each signal's power level and perform a total signal power calculation. Test equipment that supports automated measurement of total signal power across a desired bandwidth or spectrum should be used, as described in Section 4.

PROCEDURE 2 – Automated Measurement

Required Equipment

- Spectrum analyzer, digital signal analyzer, or signal level meter with an automated measurement function that integrates the total signal power across the desired bandwidth or spectrum.

Test Procedure

Instruments with automated total signal power measurement capability should make any required corrections automatically including any corrections needed when signals are of varying types (e.g., analog TV channels versus digitally modulated signals), are not spectrally flat, and when per-signal RF power levels are unequal.

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Set the lower and upper frequency limits of the bandwidth or spectrum in which the signals to be measured are located, and measure the total signal power per the instructions provided by the manufacturer of the test equipment.

Note: Some test equipment performs a total signal power measurement when the instrument is first turned on or reset to ensure that the power present at the RF input connector does not exceed the manufacturer's recommended limits. This functionality also may be accessible though menu selections, avoiding the need to power cycle or reset the instrument. Such a total signal power measurement generally includes all frequencies within the instrument's supported operating bandwidth-for example, 0.1 MHz to 1500 MHz. That bandwidth may exceed the desired total signal power measurement bandwidth or spectrum. Check the test equipment's instructions to determine how to configure a total signal power measurement for the desired bandwidth.

Chapter 3 Noise

3.1 Thermal and Intermodulation Noise

Definition: The definition of noise level shall be the power level of the noise as measured by a power meter that uses a thermocouple as a transducer. That is, the measurement shall be the average power of the noise, integrated over the actual bandwidth of the measurement. The noise level may be presented in decibels with respect to one millivolt rms (dBmV) in a 75 Ω system or decibels with respect to one milliwatt (dBm) along with the measurement bandwidth. Thus, the level reported is the rms value of a sinusoidal voltage that would produce the same heating in a 75 Ω resistor as would the actual noise level in the measurement bandwidth. (Reported noise level is often normalized to a 1 Hz bandwidth.)

There are several types of noise common in cable television systems, which all have properties similar to thermal noise. These types of noise are defined as follows:

Noise: Traditionally, the term used to describe the noise floor of the transmission system. This term, when used generically, may refer to any or all undesired noise and noise-like signals. The terms Thermal Noise, Intermodulation Noise, and Composite Noise are used to identify more clearly the specific components of the noise floor.

Thermal Noise: The thermal noise floor of the transmission system, specifically excluding any contribution from digital intermodulation products. Thermal noise is the random energy inherent in all matter and varies with the thermal agitation of the material. (Note: for fiber optic systems, this term may include noise components such as shot noise and relative intensity noise (RIN), which are not strictly thermal, but are treated as part of the optical system thermal noise contribution.)

Intermodulation Noise: The noise-like signals generated by the nonlinearity of a broadband transmission system carrying digitally modulated signals. These distortion products are analogous to the Composite Second Order (CSO) and Composite Triple Beat (CTB) products generated by analog carriers and appear as a noise-like interference due to the pseudorandom nature of the digital modulation signals. When Intermodulation Noise products fall within the analog portion of the spectrum, their effect on the analog signal is similar to increasing thermal (random) noise. Since Intermodulation Noise is a distortion product, its contribution is dependent on the signal level.

Composite Noise: The combined noise plus noise-like signal sources of non-thermal origin. This includes the total of the thermal noise and intermodulation noise.

Discussion: Carrier-to-composite noise ratio is one of several basic measurements performed on cable television system's downstream paths. For analog channels in the downstream path, noise is commonly stated in a 4 MHz bandwidth. All noise measurements of a downstream path that includes both analog and digital signals are composite noise measurements.

In the upstream path, the baseline noise floor is the result of the noise addition of all the active devices. This is called "noise funneling". This noise is normally measured in a narrow bandwidth to avoid upstream impairments such as ingress and electrical transients. It is generally expressed as an absolute power in dBmV in a 1 Hz bandwidth. Such a noise measurement is intended to characterize the upstream path as though it were devoid of ingressing carriers or impulse type impairments. All noise measurements of an upstream path that include digital signals are composite noise measurements.

Noise-Near-Noise and Beat-Near-Noise Correction

When attempting to measure noise and non-coherent disturbances (often referred to as distortion "beats"), the thermal noise of the test instrument may contribute to the displayed result. Whenever possible it is desirable to subtract the contribution of the test instrument so that only the noise or beat power produced by the device being tested is reported.

Many sections in this document will make reference to noise-near-noise or beat-near-noise correction factors. Most sections include tables for doing such corrections. This information is provided here for general use. The value can be calculated from the equation below or determined from Table 3-1 or Figure 3-1.

$$\text{Correction} = \left| 10 * \log\left(1 - 10^{\frac{-\text{Noise Drop}}{10}}\right) \right| \text{dB}$$

Note 1: When measuring noise, Noise Drop is the difference in level between the noise being measured and the noise floor of the spectrum analyzer when the input signal is removed. When measuring non-coherent distortion beats, Noise Drop is the difference between the beat being measured and the adjacent noise floor. Since the measured beat amplitude is increased by both the spectrum analyzer noise floor and the system noise floor, removing the analyzer input to check noise drop would eliminate the system noise floor contribution and potentially provide an incorrect result.

Note 2: Although the formula will allow one to derive a value for any observed noise drop, any noise drop of less than 2.0 dB is subject to significant potential errors. Therefore, it is recommended that, for noise drop values less than 2.0 dB, 4.3 dB should be added to the calculated distortion/carrier-to-composite noise ratio and the new result is expressed as "greater than" (>) xx.x dB.

Noise Drop	Correction		Noise Drop	Correction		Noise Drop	Correction
2.0	4.3		6.5	1.1		11.0	0.4
2.5	3.6		7.0	1.0		11.5	0.3
3.0	3.0		7.5	0.9		12.0	0.3
3.5	2.6		8.0	0.7		12.5	0.3
4.0	2.2		8.5	0.7		13.0	0.2
4.5	1.9		9.0	0.6		13.5	0.2
5.0	1.7		9.5	0.5		14.0	0.2
5.5	1.4		10.0	0.5		14.5	0.2
6.0	1.3		10.5	0.4		15.0	0.1

Table 3-1: Noise-Near-Noise and Beat-Near-Noise Correction Factors

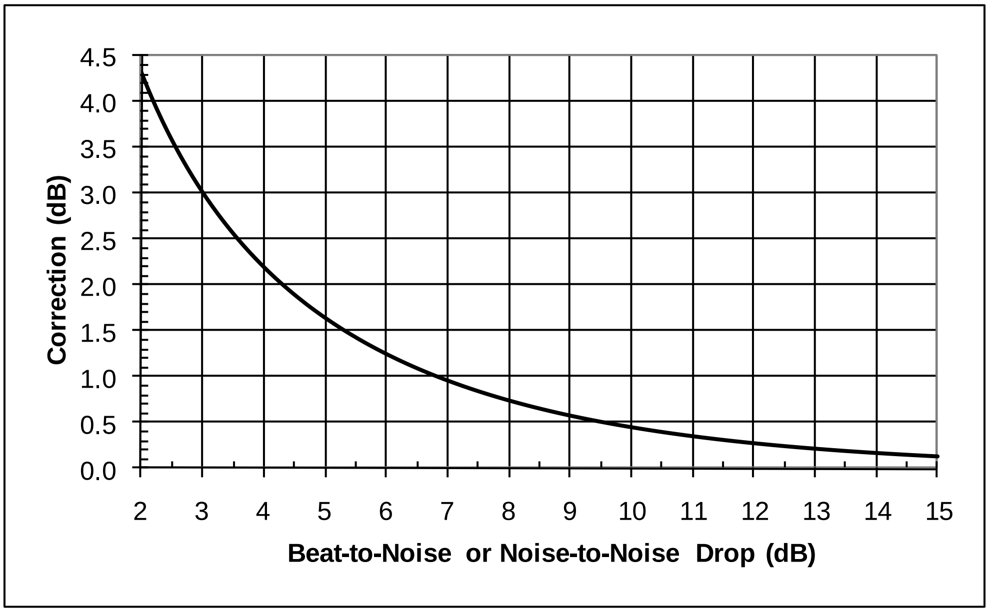


Figure 3-1: Noise-Near-Noise and Beat-Near-Noise Correction Factors

For detailed instructions on measuring noise, refer to the other sections in this chapter. Another useful reference is the SCTE test procedure, ANSI/SCTE 17 2018 Test Procedure for Carrier-to-Noise (C/N, CCN, CIN, CTN).

3.2 Visual Carrier-to-Composite Noise Ratio [FCC §76.605(b)(7)]

Definition: Visual carrier-to-composite noise (CCN) ratio is the rms voltage produced by the visual signal during the transmission of synchronizing pulses, expressed as power in a 75 Ω impedance, divided by the associated system composite noise power in a 4 MHz bandwidth. This ratio is expressed in dB.

The ratio of RF visual signal level to system noise shall not be less than 43 decibels.

Other relevant FCC Regulations: (consult the complete text, CFR 47, Section 76)
§76.5(v), (w), (y)
§76.601(c)
§76.609(a), (b), (e).

Discussion: Carrier-to-composite noise ratio is one of several basic measurements performed on a cable television system. It is useful as a system maintenance tool. For cable television purposes, noise and carrier measurements are referenced to a 75 Ω impedance and a 4 MHz bandwidth.

In cable television systems, the principal sources of this form of noise are active devices, i.e., headend processors, modulators, preamplifiers, cable repeater amplifiers, and set-top converters. The power ratio of the desired carrier to the random noise is an important design parameter that becomes a figure of merit to determine system performance. Noise manifests itself as a picture impairment to the eye and ranges from imperceptible at ratios greater than 53 dB to annoying at ratios less than 40 dB. For ratios in the range of 40 to 53 dB, picture impairment from noise becomes a function of viewer interest in program content, the viewer's critique, and TV receiver characteristics.

The total composite cable system noise at the subscriber terminal is a combination of the individual noise contributions in the channel bandwidth from preamplifiers, processors and/or modulators, distribution system amplifiers, and set-top converters. What the viewer sees is this combined system noise, plus the noise in the original delivered signal, plus the noise added by the TV receiver itself. Cable operators are not required to improve an impaired signal delivered to them nor are they responsible for the TV receiver's noise performance. However, they are charged with minimizing the addition of noise to the signal as it passes through the cable system to the subscriber terminal. The carrier-to-composite noise measurement is intended to measure that degree of further impairment from the first point in the signal path under the control of the cable system operator to the subscriber terminal.

Carrier-to-composite noise is to be characterized at the input to the subscriber terminal. This is interpreted by the FCC to include the contributions of any operator-supplied set-top converters. Furthermore, the converters should reflect the type of converter (in terms of age and manufacturer) supplied to subscribers in the vicinity of the test point.

Unfortunately, direct measurements of the noise contributions of set-top converters is very difficult. Among other problems:

- The low level of the signals at the converter output terminals may limit the measurement range.
- Bandwidth shaping in the converters internally may distort the measurements (in some cases resulting in apparently lower noise at the output than at the input).
- Deleting signals to make measurements, as is common with distribution systems, is generally not possible with converters which use Automatic Gain Control (AGC) as the operating conditions of the converter will shift. Also, Automatic Frequency Control (AFC) circuits may cause a shift in operating frequency without a reference carrier. Finally, some converters require the presence of an aural carrier for proper functioning.
- Substitution of a CW carrier for the normal television signal will not work with some converters as their AGC circuits depend on the presence of a synchronizing pulse.
- In the case of remodulating converters, the output depth-of-modulation may not be the same as at the input.
- Due to the effects of delayed AGC circuits, the effective noise performance of a converter may be a sensitive function of the RF input level.

Given these situations, it may be more practical to make separate measurements of the noise performance of the headend and the distribution system and then compensate for the converter's effect (See Section 3.3: “Measuring Noise of Systems using Converters”).

Test Procedures: The following procedures are intended to serve as reference methods and be usable with generic measurement instruments. Many test equipment manufacturers provide valid and more convenient alternative methods of measuring CCN. For a discussion of the relationship of Video Signal-to-Noise to RF Carrier-to-Noise see Section 16.7: “The Relationship of Cable System Carrier-to-Noise to the Baseband Video System Signal-to-Noise”.

Required Equipment

- A spectrum analyzer capable of IF resolution bandwidth of 300 kHz, with 30 kHz and 100 kHz bandwidths also desirable, or a signal level meter with known noise correction factors for bandwidth and detector noise response.

Optional Equipment

- A tunable bandpass filter or single-channel bandpass filters for the channels to be measured. Noise power bandwidth should be a minimum of 10 times the widest measurement resolution bandwidth used. The maximum filter bandwidth should be narrow enough to prevent instrument overload. The filter passband frequency response at the measurement frequency should be known.
- A broadband low noise preamplifier; 10 dB noise figure maximum with 20 to 30 dB gain. A preamp is not generally required if measuring signal levels greater than +30 dBmV, or if the noise floor drops more than 10 dB when disconnecting the input cable from the instrument input.

PROCEDURE 1 - Using a Spectrum Analyzer

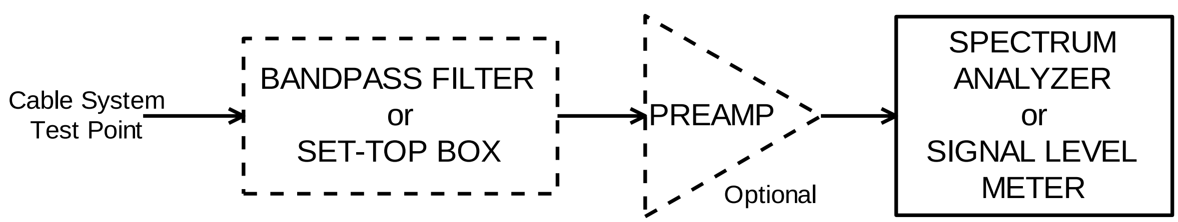


Figure 3-2: Visual Carrier-to-composite noise Ratio Test Equipment Setup

Prepare the Measurement Setup

1. Set the equipment up in the configuration needed as illustrated in Figure 3-2. See text below for configuration information.
2. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing. The accuracy of log scale fidelity is the most critical. Record the appropriate information as called for in §76.601(b)(1).
3. Connect the signal and tune the spectrum analyzer to the carrier to be measured. The carrier should be tuned left of center screen to enable viewing the entire channel bandwidth.

Measure the Carrier Level

4. Adjust the spectrum analyzer as follows to measure the carrier level:
   - IF Resolution Bandwidth: 300 kHz
   - Detector Mode: Peak
   - Video Bandwidth: 300 kHz min.

- Amplitude Scale: 2 dB/div
- Frequency Span: 1 MHz
- Sweep Time: Automatic

5. If using a tunable bandpass filter, set the signal 1 to 2 divisions down from the top of screen using the reference level control and tune the bandpass filter to peak the video carrier.
6. If markers are available, they should be used to measure the carrier level. Otherwise, adjust the reference level to set the peak video carrier at the top graticule line on the spectrum analyzer. **Record this carrier level reference.**

Note: It may be necessary to turn off scrambling if it affects the carrier level measurement. Refer to Section 16.11: “Notes Concerning the Use of Scrambled Channels for Signal Level Measurement” for more information.

Adjust Measurement Setup to Measure Noise

7. To measure the noise, adjust the spectrum analyzer settings as follows:
   - IF Resolution Bandwidth: 30 kHz (wider bandwidths may be used as desired)
   - Detector Mode: Sample, if available (see Note 1 below)
   - Video Bandwidth: 100 Hz, or until display stability is reached
     and/or
   - Number of Video Averages: 25 to 100, or until display stability is reached
   - Amplitude Scale: 10 dB/div
   - Frequency Span: 6 MHz to 10 MHz
   - Sweep Time: Automatic

Note 1: The peak of noise will exceed its power average by an amount that increases (on average) with the length of time the peak is observed. The sample detector is recommended for noise measurements to prevent the peak detector from causing this increase in the measured result.

Note 2: When a bandpass filter is not used, it may be necessary to adjust the spectrum analyzer attenuator to prevent overload. If the attenuator setting is changed at this point, it may be necessary to go back to step (4) above to re-measure the carrier level with the new attenuator setting.

Adjust System to Measure Noise (if necessary)

8. Because video modulation obscures the noise in the defined FCC measurement range, it is necessary to either remove the modulation or measure the noise at the channel edge and use the in-service noise summation method described in Section 3.2: “Notes, Hints and Precautions**”** and Section 3.3: “Measuring Noise of Systems using Converters”.

For in-channel measurements, remove modulation from the channel under test making sure all headend channel noise contributing active devices remain powered up and in the signal path. For over-the-air channels, disconnect the antenna lead to the processor or preamp and terminate the processor or preamp input. For other channels, disconnect the first access to the baseband video signal and terminate the input. Do not just remove power from the processor, preamp, or modulator.

Note: If converters are supplied to the subscribers, see Section 3.3: “Measuring Noise of Systems using Converters” for more information on testing with converters.

Measure the Composite Noise

9. If using a bandpass filter see paragraph 3 of “Notes, Hints and Precautions”. Measure the amplitude of the noise near center screen using either the spectrum analyzer markers if available or the on-screen graticule. Care should be taken to avoid measuring beats which may be present. Record the difference in level between the noise at center frequency and the carrier level reference recorded in step 6. This is the uncorrected carrier-to-composite noise ratio (+dB).

Note 1: Some spectrum analyzers use separate detectors and/or will require reference level adjustment when measuring noise for greater accuracy. Consult the manufacturer's literature.

Note 2: Some modern test equipment automatically calculates and corrects the result for some or all of the correction factors below. Consult the manufacturer's literature.

Note 3: Video filtering and/or averaging may hide "bottoming" of noise at the log detector or analog-digital converter, hence giving a false reading of noise level. To check for this, view the noise with the video filter increased to > 100 kHz and the video averaging turned off. The noise excursion (roughly 15 $dB_{pk\text{-}pk}$) should not frequently reach the bottom of the screen. If this noise bottoming occurs, the noise level into the analyzer’s IF should be increased by one of the following methods:

- wider IF resolution bandwidth
- more IF gain (lower reference level)
- external preamp

This applies to system noise and instrument noise measurements (see noise-near-noise correction, Figure 3-1). Table 3-2 shows the advantages and disadvantages of each method of raising the noise level.

	IF Resolution Bandwidth	**IF Gain**	**External Preamp**
Advantages	No other settings changed, no calculations changes	Allows minimum noise band search, more reliable for channel edge measurements	Improves noise figure of the measurement system
Disadvantages	May hide some discrete products whose power is added to the noise	More elaborate calculations, especially when making noise-near-noise correction	May generate distortion products

Table 3-2: Methods of Raising Noise Level

Adjust Carrier-to-Composite Noise Measurement Result

10. Subtract the following correction factors from the number determined in step (9):

- Convert 30 kHz to 4 MHz: $10 * \log\left[\frac{4\ \text{MHz}}{30\ \text{kHz}}\right] = +21.25\ \text{dB}$
- (other examples: +16.02 dB for 100 kHz and +11.25 dB for 300 kHz)
- Log Detect Rayleigh Noise: +2.5 dB

- IF noise equivalent BW: See manufacturer's specifications (for example: -0.52
- dB for 3 dB filters or +1.10 dB for 6 dB filters)
- Noise-Near-Noise: Disconnect signal from analyzer (or preamp if used) and use Figure 3-1

Example 1

Uncorrected CCN measured in step (9) = 63 dB

Noise drop when disconnecting
cable from analyzer or preamp input = 4 dB (use Figure 3-3 to get 2.2 dB)

CCN = 63 - (+21.25 + 2.5 - 0.52 - 2.2) dB
= 41.97 dB

Note: For the sake of simplicity, the above procedure does not include a correction for bandpass filter frequency response. However, if a fixed filter is being used and its frequency response is known, additional accuracy can be obtained by applying that information to the measurement result.

PROCEDURE 2 - Using a Signal Level Meter

1. Set the equipment up as shown in Figure 3-2 or refer to the manufacturer's manual.
2. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
3. Set the SLM attenuator to get an on-scale meter reading.
4. Record carrier value.
5. Tune the frequency of the SLM 2 MHz to 3 MHz higher than the carrier (if a CW carrier) and remove attenuation from the SLM attenuator until the minimum tunable noise level can be read. If the carrier is modulated, it must be turned off at the headend or modulation must be removed.
6. Record the noise level.
7. Apply appropriate correction factors.

Note: The lowest practical measurable noise level is approximately -40 dBmV. For this reason, measure at a high-level point in the system or use a preamp.

8. The difference between the level in (4) and (7) is the carrier-to-composite noise ratio.

Example 2

- Carrier Level = +31 dBmV
- Noise Level = -20 dBmV
- Carrier-to-Composite Noise (CCN) = 31 - (-20)
- CCN = 31 + 20
- CCN = 51 dB

Notes, Hints and Precautions

1. To allow in-service measurements and/or to deal with the problems inherent in measuring through converters, calculate total system noise by summing the individual in-channel noise contributions of the headend equipment, the plant (as measured at a channel edge), and the noise

of the converter. This allows the actual noise measurement to be made in-service as well as allowing use of automatic equipment that measures system noise at channel edges.

Noise measured at the channel edge will indicate the noise contribution of the distribution system from the output of the headend to the test point. It does not include the noise contributions in the FCC measurement range of the headend equipment or the converter. Adding to the channel edge measurement the in-channel noise contributions from the headend equipment and the converter, as measured individually or calculated from their noise figures, gives the total system noise value. See Section 3.3: "Measuring Noise of Systems using Converters".

Note: The vestigial sideband filter of some modulators does not roll off as quickly as others. This can affect the result when measuring noise at a channel boundary. Try making the measurement using a narrower IF resolution bandwidth.

2. The IF resolution bandwidth used for the spectrum analyzer method is 30 kHz. Theoretically, any bandwidth can be used as long as the correction to 4 MHz is done properly from the chosen bandwidth. An IF resolution bandwidth of 30 kHz was chosen as a compromise between keeping a reasonable measurement speed while minimizing the effects of distortion products and of the carrier (if present) on the lower frequency end of the FCC noise measurement range. An IF resolution bandwidth of 30 kHz also allows noise measurement at the channel edge.
3. Peaking a bandpass filter on noise is difficult, jeopardizing getting repeatable results. Use a bandpass filter only when the preamp and/or measuring instrument are known to overload without it. The following techniques are suggested:
    a) Test the susceptibility of the measuring instrument to overload by measuring CCN with and without the bandpass filter, comparing results. Test with increasing instrument attenuator settings until the results don't improve. Then use the lowest attenuator setting that does not improve results.
    b) Test the preamp for overload by measuring CCN with and without the bandpass filter. If the results are the same, the bandpass filter should not be necessary.
    c) If a bandpass filter is required, it is best to use a fixed filter with a known passband characteristic for the channel being measured. If a tunable bandpass filter is found to be necessary, try not retuning it after it is initially peaked on the carrier by measuring the noise as close to the carrier frequency as possible. This should minimize the effects of any filter roll-off or less than perfect passband flatness.

Note: If testing through a converter with an output bandpass filter, another bandpass filter on its output is not required.

4. For most situations, when measuring a cable plant after several amplifiers with a preamp having a noise figure of less than 10 dB and at levels above about +20 dBmV, the preamp noise correction factor is very small and can be neglected.
5. The carrier-to-composite noise ratio does not change when going through taps or couplers between the high-level output of the last active device (line extender) and the input of a converter in the vicinity. Therefore, assuming signal attenuation is within system specification so that the signal at the converter input is proper, measuring CCN at the high-level output of the last active device is equivalent to measuring at an adjacent converter input driven by that active device.
6. These methods assume the noise floor in the measurement range is flat. If it is not flat, use a method such as described in §76.609(e). However, note that §76.609(e) advises to sum "the power indications to obtain the total power present over a 4 MHz band centered within the cable

television channel." This is slightly different than the definition of system noise in §76.5(w) which is "the 4 MHz bandwidth between 1.25 and 5.25 MHz above the lower channel boundary of a cable television channel". Since §76.609(e) is exemplary, use of the definition in §76.5(w) is advised.

7. When measuring through a converter, set its input level above its minimum specification and below its delayed AGC threshold. Remember to supply carrier(s), CW or with sync, for AGC and/or AFC circuits if needed. Make sure the converter is operating within its specified operating temperature range. See Section 3.3: “Measuring Noise of Systems using Converters”.

3.3 Measuring Noise of Systems using Converters

Carrier-to-composite noise is to be measured at the subscriber terminal, which means through the converter, if converters are provided by the system operator. The converters used should reflect the type of converter (in terms of age and manufacturer) supplied to subscribers in the vicinity of the test point. However, measuring CCN at the output of converters may give misleading and too optimistic results, sometimes even appearing to have better CCN at the output than at the input. This is due to the effects of filters, variations in depth of modulation, input signal level, the effects on video channel noise in the demodulation and re-modulation process and other factors. On the other hand, the effects of delayed AGC could cause worse results.

Consider the case where CCN at the output of a baseband converter is equal to or better than at the input. While this result may meet the letter of the law, it is risky to assume this number truly represents system performance. To obtain a more conservative CCN value, it is advisable to use either the noise summation method described below or the result from a straight RF-to-RF converter with no AGC and having the same or worse noise figure as the baseband converter.

Measuring Noise In-Channel

Noise can be measured at the channel edge or in-channel in the FCC measurement range. To measure in-channel noise, remove the modulation and/or use a substitute carrier. Some converters need a carrier for their automatic gain control (AGC) and automatic frequency control (AFC) circuits to function. For over-the-air channels, disconnect the antenna and substitute a low noise, unmodulated CW signal of the same level and frequency as the received channel. This signal should be known to have phase noise and in-channel noise much lower than the processor under test. In some cases, a CW signal is not sufficient. Some converters may require sync modulation for proper operation of keyed AGC circuits. Additionally, some converters require an aural carrier for proper operation. Be sure of the input requirements of the converter under test. See Section 16.1: “Carrier-to-Noise Measurement through a Set-Top Converter”.

For non-over-the-air channels, terminating the modulator video input will leave the needed carrier in place as long as the other converter input requirements mentioned above are met.

Measuring Noise at Channel Edge

When measuring the noise at the channel edge, it is necessary to add the noise contributions from the headend equipment and the converter. The purpose of the chart below is to find the noise contribution of an active device using its noise figure, its input carrier level, and the CCN of the applied signal at the input. Subtract the Figure 3-3 result from the input CCN to get CCN at the output. Or, use the equation below to calculate the output CCN. Further discussion of this can be found in Section 16.1: “Carrier-to-Noise Measurement through a Set-Top Converter”.

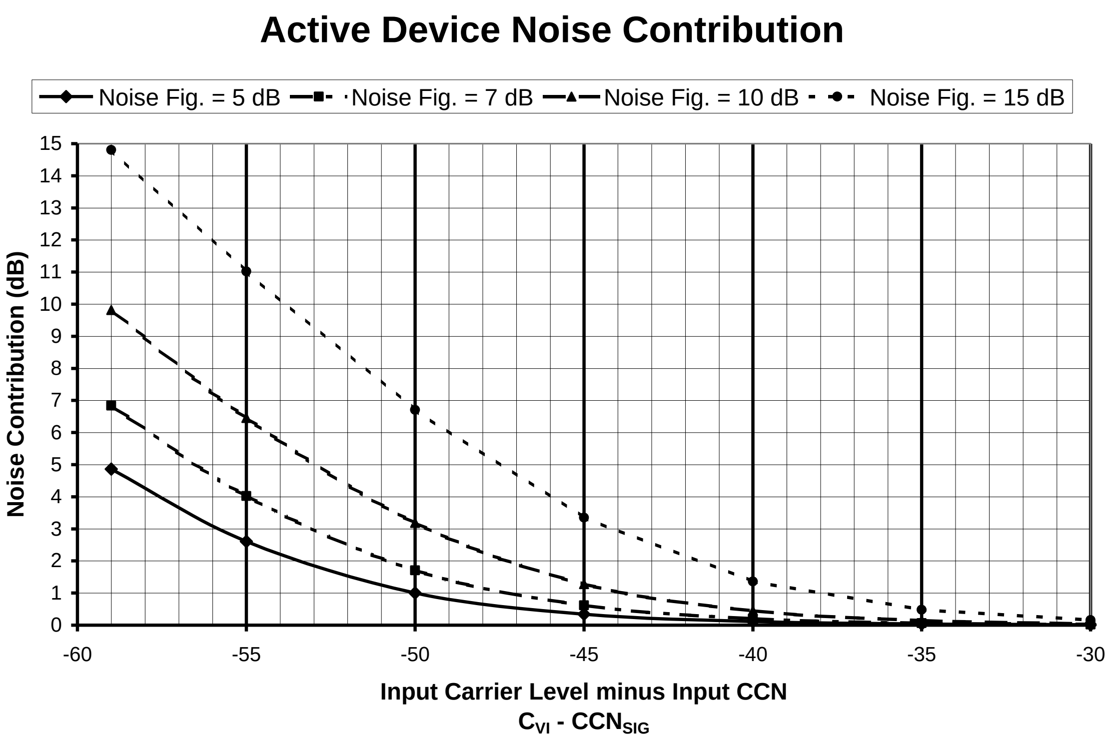


Figure 3-3: Active Device Noise Contribution

The equation below gives the same results as using Figure 3-3 above.

$$CCN_{OUT} = C_{VI} - 10 * \log\left[10^{\left(\frac{CCN_{VI} - CCN_{SIG}}{10}\right)} - 10^{\frac{-59.2}{10}} + 10^{\left(\frac{-59.2+NF}{10}\right)}\right]$$

where:

CCN_{OUT} = Carrier-to-composite noise at device output (dB)

CCN_{SIG} = Carrier-to-composite noise of applied signal at device input (dB)

C_{VI} = Visual carrier level at device input (dBmV)

NF = Device noise figure (dB)

-59.2 = Thermal noise in a 4 MHz noise BW @ 17.5 °C (dBmV)

The noise figure of RF-to-RF converters and amplifiers can be measured using commercially available test equipment or can be obtained from their manufacturer's specifications. The noise figure for baseband converters can be difficult to measure. See Section 16.1: “Carrier-to-Noise Measurement through a Set-Top Converter”.

3.4 Visual Carrier-to-Noise Ratio of a Heterodyne Processor

Definition: Visual carrier-to-noise ratio (CNR) is the rms voltage produced by the visual signal during the transmission of synchronizing pulses, expressed as power in a 75 Ω impedance, divided by the associated processor noise power in a 4 MHz bandwidth. This ratio is expressed in dB.

Discussion: In a headend processor, it is important that carrier-to-noise increase linearly for an increase in input signal level (noise figure remains essentially constant) until it reaches a ratio that assures it has a negligible effect on system picture quality (60 dB or better for the processor).

Required Equipment

- Signal level meter or spectrum analyzer (SLM/SA)
- Variable 75 Ω attenuator 0-60 dB
- RF signal generator or source of television signal

Test Procedure

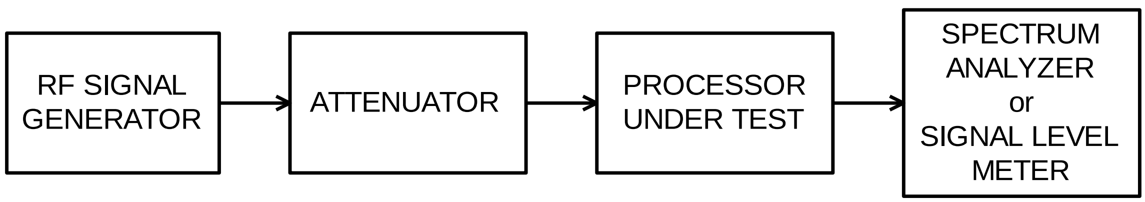


Figure 3-4: Block Diagram of Equipment Setup

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Place 50 dB or more in the attenuator and configure the test equipment as shown in Figure 3-4.
3. Tune RF signal generator to the video carrier frequency of unit under test.
4. Adjust level control on RF signal generator to give a -20 dBmV reading on the SLM/SA (or the manufacturer’s specified minimum input level).
5. Connect SLM/SA to IF test point and tune to IF frequency. Switch the processor to manual gain control and adjust the manual gain control on the Processor to set the IF video carrier level specified by the manufacturer.
6. The aural IF level control should be turned down fully so as to not interfere with the reading accuracy.
7. Connect the SLM/SA to the OUTPUT terminal on the Processor and tune the SLM/SA to the output video carrier frequency. Adjust the level of the output module (not the Video IF level control) as described by the equipment manufacturer to give the specified RF output level. This must also be a manual adjustment.
8. Disconnect the attenuator from the Processor and terminate the Processor INPUT with a 75 Ω load.
9. Adjust the SLM/SA to read the noise level and record this reading. Follow the manufacturer’s recommendation for using the SLM to measure noise. The recommended settings for using a spectrum analyzer are:

- IF Resolution Bandwidth: 30 kHz (wider bandwidths may be used as desired)
- Detector Mode: Sample, if available
- Video Bandwidth: 100 Hz, or until display stability is reached
  and/or
- Number of Video Averages: 25 to 100, or until display stability is reached
- Amplitude Scale: 10 dB/div
- Frequency Span: 6 MHz to 10 MHz
- Sweep Time: Automatic

10. Reconnect the attenuator to the Processor INPUT and remove 10 dB attenuation.
11. If so equipped, reduce the manual IF gain control on the Processor to give the rated output reading on the SLM/SA. Repeat steps 8 through 11 each time removing an additional 10 dB from the attenuator until the attenuation has been reduced by 50 dB from the attenuation used in step 3.
12. Correct all SLM noise readings using the calibration curve or recommended correction factor for signal-to-noise measurements specified by the manufacturer of the SLM. When using a spectrum analyzer, add the following correction factors to the measured noise at each step:

- Convert 30 kHz to 4 MHz: $10 * \log\left[\frac{4\ \text{MHz}}{30\ \text{kHz}}\right] = +21.25\ \text{dB}$
- Log Detect Rayleigh Noise: +2.5 dB
- IF Noise Equivalent BW: See manufacturer's specifications (for example: -0.52 dB for 3 dB filters or +1.10 dB for 6 dB filters)
- Noise-Near-Noise: Disconnect signal from analyzer (or preamp if used) and use Figure 3-1.

13. Plot a curve of the carrier-to-noise ratios vs. input level. The carrier-to-noise ratio is the dB difference between the output level established in step 7 and the various noise level readings recorded in step 9 modified by the correction factor as discussed in step 12.

The first reading, of course, is for a -20 dBmV input and each succeeding reading is for an input signal 10 dB higher.

Performance Objective: A Heterodyne Processor should achieve a 58 dB carrier-to-noise ratio at +10 dBmV input and should achieve a 60 dB or better ratio at maximum rated inputs.

3.5 Digital Carrier-to-Composite Noise (CCN) Ratio

Definition: Digital carrier-to-composite noise ratio is the ratio of the average power of the digital signal in a given bandwidth (less than the channel bandwidth) to the average power of the noise in the same bandwidth. This ratio is expressed in dB.

Discussion: Digital carrier-to-composite noise ratio may be measured using either a spectrum analyzer or a digital signal level meter. Some digital signal level meters will provide a direct readout of digital carrier-to-composite noise ratio. These devices require that an empty channel be selected as the reference channel for noise measurement. The carrier-to-composite noise ratio is then measured by selecting the desired digital channel and then selecting the CCN function on the digital signal level meter. Refer to the manufacturer’s instructions for the proper procedure. This procedure is applicable to continuous digital signals. To measure the level of burst digital signals see Section 2.2.2: "Burst Signals".

Required Equipment

- Spectrum Analyzer

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Set the spectrum analyzer as follows:

• Center Frequency:	Center of channel under test
• Frequency Span:	Wide enough to display entire signal under test (see note below)
• Amplitude Scale:	10 dB/div
• IF Resolution Bandwidth:	300 kHz
• Video Bandwidth:	30 kHz or lower, or until display stability is reached (30 Hz is preferred)
and/or	
• Number of Video Averages:	Until display stability is reached
• Reference Level:	To position top of signal trace approximately 10 - 20 dB below the top analyzer graticule

   Note: The recommended span settings are as follows:
   - 30 MHz to 40 MHz for L-band signals
   - 10 MHz for QAM signals transmitted in a 6 MHz channel
   - 3 MHz for out-of-band signals

3. Place the marker at the center frequency of the channel to be measured. Record the marker reading.
4. Set the center frequency of the analyzer to a clear channel as close as possible to the test channel. Place the marker at this frequency and record the marker reading. This reading will be the composite system noise plus the analyzer noise.
5. Disconnect the signal from the analyzer and terminate the input. Note the decrease in the observed noise. If the noise decreases by 10 dB or more, no correction for the analyzer noise floor is required. If the noise decreases by less than 10 dB, subtract the correction from the noise reading as shown in the graph of Figure 3-1.
6. Subtract the corrected noise reading from the signal power reading. The result is the digital carrier-to-composite noise ratio in dB.

Notes, Hints and Precautions

The "Marker/Noise" function can also be used for measuring digital carrier-to-composite noise ratios. In this case, the CCN ratio is simply the difference between the two readings in dB.

The procedure for measuring digital carrier-to-composite noise is different from the corresponding analog measurement. The reason for this is that digital signals have a uniform power distribution, whereas analog signals have most of their power concentrated in the visual carrier. Since the digital signal is deliberately designed to be noise-like, measurement of digital carrier-to-composite noise ratios does not require a bandwidth correction. This is not the case with analog signals.

The noise floor being measured during this procedure contains intermodulation noise in addition to thermal noise. See Section 3.1: " Thermal and Intermodulation Noise" for a discussion of the different types of noise in a cable system.

3.6 Impulse Noise

Definition: Impulse interference is caused by broadband, fast rise-time signals which may or may not be periodic in nature. These may be the result of electrical faults in the cable system, sheath current effects generally related to power company neutral currents flowing on the outer conductor of the coaxial cable, ingress of signals such as ignition noise or other sources generating interference signals which are not confined to narrow ranges of frequency. Such signals are particularly damaging to data transmissions where short bursts of interference can seriously reduce the data throughput. Quantifying and locating the source(s) of these signals is important for out-of-service setup of the upstream system as well as monitoring and controlling these disturbances while in-service.

Procedure: The following are suggested methods of addressing these interferences in the upstream transmission network and ways of analyzing and presenting the results. Since this is a very complex subject and may utilize a variety of existing and future test equipment the approaches outlined here will need refinement for each specific application.

System Parameters: To be meaningful, the results of testing for impulse interference must be related to the operating levels of the system. In other words, the amplitudes of transients detected need to be quantified in terms of the overall upstream signal power or to the signal levels of the channels in question in order to assess their effects. Refer to Section 5.3: "Discrete Interfering Signal Probability (DISP)" for an applicable reference level procedure.

Required equipment

- For "Method A" below a sampling oscilloscope or other high speed waveform capture device is employed. The data output must be gathered and stored in a computer for analysis.
- For "Methods B & C" below a Spectrum Analyzer or other frequency selective device capable of producing the necessary triggers is required in addition to the equipment of "Method A".

Equipment setup

The equipment configuration of Figure 3-5 is applicable for "Method B" and it is also applicable to "Method A" with the omission of the Spectrum Analyzer, input splitter and associated connections. . Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

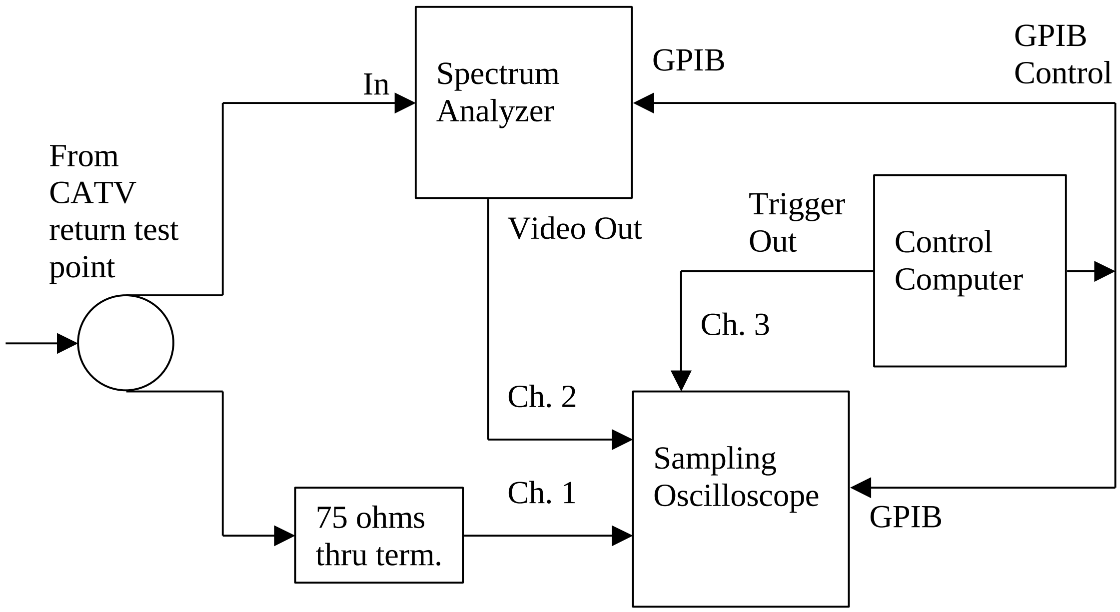


Figure 3-5: Equipment Setup for Impulse Noise

Data Capture

The function of the sampling oscilloscope is to capture real time transient waveforms and transmit them to the computer for storage and later analysis. There are two general conditions under which these captures may be made. In the condition of the plant not being in service (no information signals present), it is possible to simply monitor the upstream spectrum and trigger the sampling oscilloscope on transient events which exceed the peaks of the noise background (Method A). The spectral components and rate of occurrence of such captures give a good indication of the integrity of the upstream system.

In the situation where traffic is being carried on the upstream system capturing transient events is more difficult since there is continuous signal energy present which will trigger the scope and may obscure the data. In order to avoid these problems, it is necessary to examine only parts of the spectrum which are not in use. This is accomplished by using the spectrum analyzer to select a “quiet” portion of the spectrum. In this configuration (Method B) a transient event within the passband of the spectrum analyzer is used to trigger the sampling oscilloscope and the video signal from the spectrum analyzer is captured or an RF capture is initiated (Method C).

Method A

The sampling oscilloscope is set so that it does not trigger on background noise but will trigger at a slightly higher level. This trigger level is available as part of the setup data and should be recorded on the computer with the captured waveforms. A waveform should be captured at every trigger and recorded in the computer. It should be noted that some time is expended in transferring data to the computer and that there may not be the facility to capture subsequent events until the transfer is complete.

Method B

In this method the analyzer is tuned to the selected quiet area of the spectrum using the maximum possible bandwidth while excluding information signals. The analyzer is then set to “zero span” so that it acts as a tuned radio receiver. The amplitude response of the analyzer is set for “linear” in the general case although a logarithmic mode may be selected when the situation warrants. The oscilloscope is fed with spectrum analyzer video output and its trigger is set at a level somewhat above the noise floor as before. Note: Instead of the spectrum analyzer a filter of the desired bandwidth may be substituted and direct RF captures made with the sampling oscilloscope. In either case the transient signals are filtered and therefore only the components of the transient which lie within the preselected band are captured.

Method C

This method is the same as Method B except the sampling oscilloscope takes its input directly from the RF upstream signals and its trigger from the spectrum analyzer. The time waveform of the entire spectrum including the information signals as well as the transients is captured. In order to compensate for the delay of the trigger signal through the spectrum analyzer the sampling oscilloscope must have its buffers configured for continuous sweeping and buffer retention prior to the trigger in order to capture the onset of the transient. The value of this type of capture will become evident in the analysis phase.

Data Acquisition and Storage Requirements

The computer must be programmed to control the instruments as well as retain the captured data. This will require some rather complex software and impose certain limitations on the number and timing of the captures. The same computer may also be employed for data analysis.

Computer Control and Data Analysis

Control

The computer is setup to cause captures either continually upon occurrence or to gather only a preset number of captures in preprogrammed time periods. The buffer size of the captures must be set in the oscilloscope and the computer memory managed accordingly. Commands from the computer to the instruments must be transmitted by some means. The GPIB or IEEE-488 bus, commonly called GPIB (General Purpose Interface Bus), is quite convenient and other configurations such as RS-232 may be employed. It is strongly recommended that the same bus be chosen for both instruments to simplify programming.

Data Analysis

There is a great deal of information contained in captures of transient events as described above. Individual captures of transient events reveal the shapes of the impulse and may allow identification leading to discovery of the source of the impulse interference. Application of Fast Fourier Transform (FFT) processing will separate the spectral components of the impulse and may provide clues to its origin. This FFT data does point out the amount of contamination in each spectral area and hence the information traffic which will be affected by the interference. This processing must be customized to the specific test equipment, system and situation addressed and is beyond the scope of this specification. However, Figure 3-6 illustrates one possible output and display format which illustrates the employment of these techniques. The vertical axis shows the relative frequency of the occurrence of events.

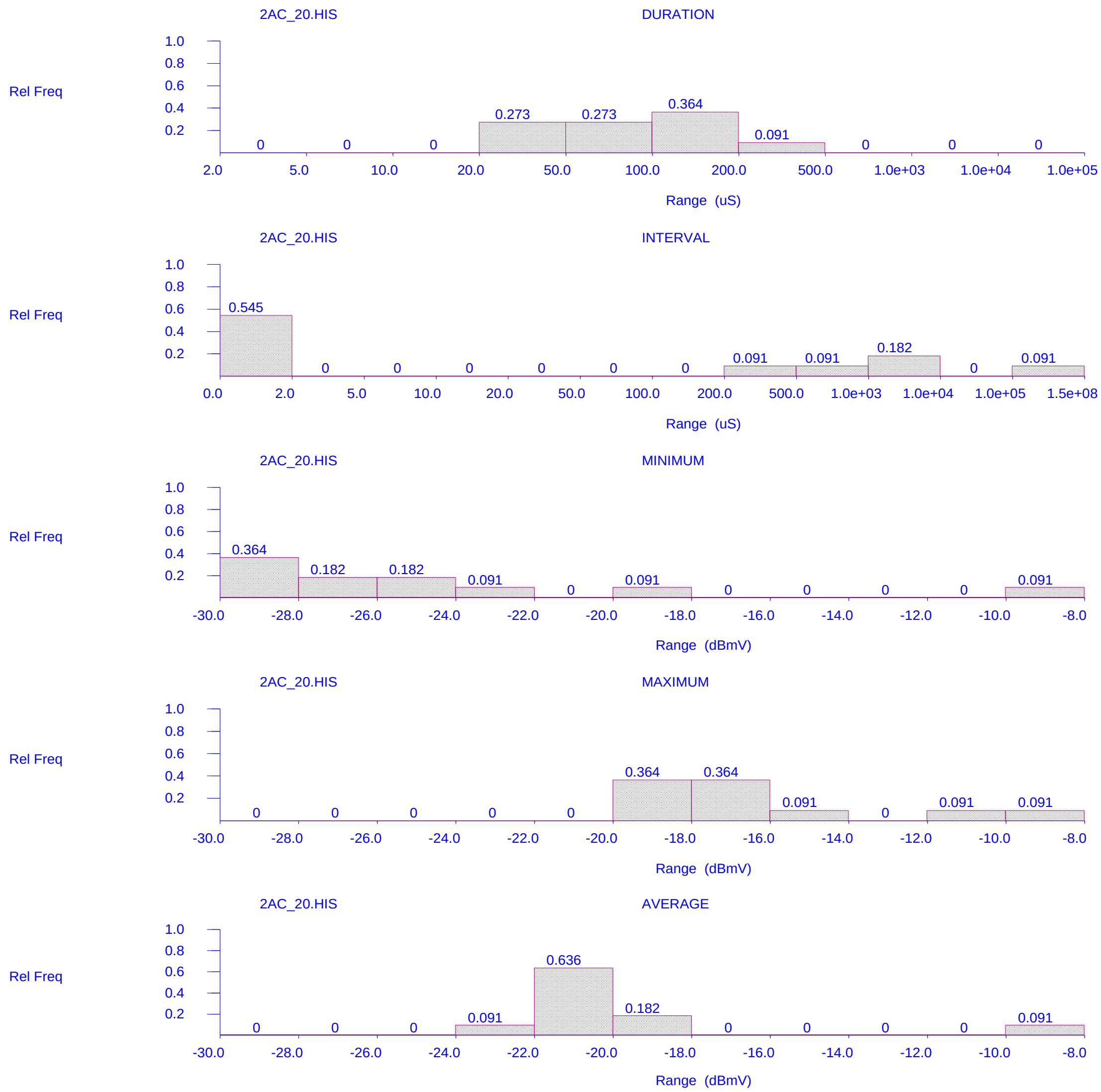


Figure 3-6: Possible Output and Display Format

When data is taken using Method C it is possible to subtract a normal (non-interfered) FFT spectrum from that developed by analysis of the transient capture to generate a first order approximation of the broadband transient spectral components.

3.7 Phase Noise

Definition: Phase noise is the random change in a signal's phase because of the effects of thermal noise in oscillators and the noise injected through external power supplies and control circuitry. Phase noise becomes a problem in cable systems when signals are converted to different frequencies by signal processors, return path band-stacking equipment, microwave links, or other frequency conversion devices. As well, excessive phase noise in standalone signal generation devices such as modulators and cable modems can be problematic.

Discussion: Phase noise on a digital carrier degrades the digital signal and, in the worst case, will cause complete loss of signal demodulation capability. Phase noise effects, when large enough, are visible on the constellation of the demodulated signal (see Section 11.1: "Constellation Analysis"). Modulation error ratio (MER) and bit error ratio (BER) performance will also suffer when excessive phase noise is present. The amount of phase noise that can be tolerated in a given system depends on the modulation type employed and the carrier recovery bandwidth of the demodulator. Phase noise which lies outside the carrier recovery bandwidth is of greatest concern because the demodulator cannot correct for its detrimental effects. To determine what value of phase noise-versus-frequency offset is necessary to ensure acceptable MER and BER performance, consult the equipment manufacturer.

A direct measurement of phase noise on a modulated carrier is available only with select test equipment today. If the digital signal is being phase modulated in the same manner as an analog carrier, an analog phase noise measurement can be used to determine the effective phase noise on the digital carrier.

Phase noise on an unmodulated carrier is typically measured on one side of the carrier at a fixed offset frequency and is specified in dBc/Hz, or dB relative to the carrier normalized to a 1 Hz bandwidth. This result is normally expressed as a negative number. The effective phase noise for a digital signal is the integration of the phase noise at all offset frequencies from the carrier up to one-half of the occupied bandwidth of the carrier on both sides. It is quite common to use a single reference measurement at one offset frequency, usually offset of 10 kHz, assuming the phase noise will roll-off at a 1/f or $1/f^2$ rate. Figure 3-7 illustrates the potential limitation of this single point measurement.

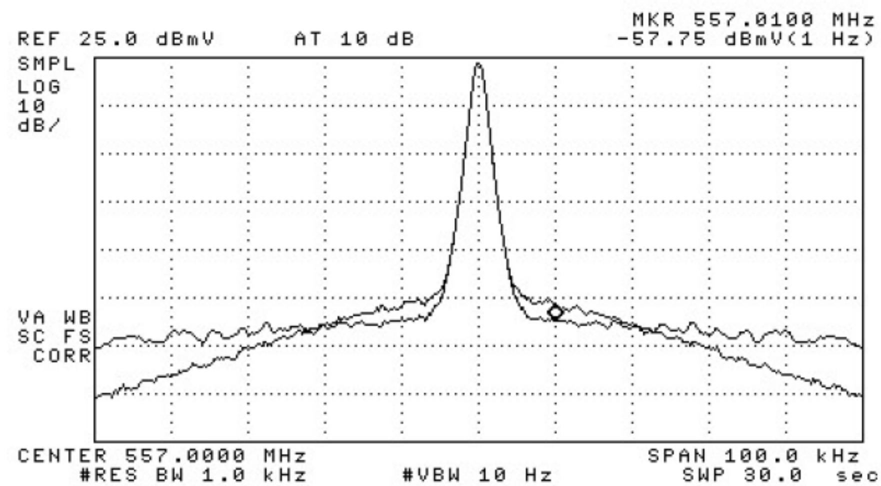


Figure 3-7: Integrated Phase Noise Issues

The diamond marker indicates the measurement of a signal's relatively flat phase noise at offset of 10 kHz offset which measures about 3 dB better than the corresponding signal with the sloped characteristic. In practice, the signal with the flatter phase noise characteristic will have significantly worse MER performance when modulated because of the amount of phase noise outside the QAM demodulator's recovery loop bandwidth, even though it measured 3 dB better at offset of 10 kHz. The integrated phase noise is higher on the signal with the broad characteristic, but the single point measurement at offset of 10 kHz does not indicate this.

Since phase noise is a noise-like signal, it must be measured as such. Noise power levels can be measured using a spectrum analyzer provided that corrections are made for various characteristics of the spectrum analyzer (log detection errors, IF resolution bandwidth errors, etc). Most modern spectrum analyzers have built in noise marker functions which do the corrections automatically. Both the manual method and the noise marker method will be discussed.

Before measuring the phase noise performance of the system, the back-to-back performance of the signal generator and spectrum analyzer should be verified. This should be done with test signal amplitude similar to the levels at the point where phase noise is to be measured. This measurement point will most likely be the forward path headend test point for downstream signals, or the fiber receiver's output at the headend for upstream signals. The procedure "Measuring System Phase Noise Performance" below can be used to verify the test equipment's performance. The back-to-back performance of the signal generator and spectrum analyzer should be at least 10 dB better than the desired system performance level. As an example, if the desired system measurement is -86 dBc/Hz it is desirable to have the signal generator and spectrum analyzer better than -96 dBc/Hz.

Required Equipment

- Spectrum analyzer
  - IF resolution bandwidth (RBW) ≤ 1/10 of offset being measured, typically 1 kHz.
  - Phase noise performance at least 13 dB better than that being measured.
- RF signal generator (may not be necessary if modulator has a CW mode available)
  - Frequency range necessary for required test signals
  - Phase noise performance at least 13 dB better than that being measured.

Forward Path Test Configuration

If testing a modulator or similar standalone signal source, place the modulator in CW mode if this mode is available and skip to step 5.

1. If testing an external upconverter, remove the digital signal from service at the input to the RF upconverter.
2. Set the signal generator as follows:
   - Output frequency to 44 MHz or center frequency required for upconverter under test
   - Output amplitude set for optimum input to the RF upconverter. Refer to the manufacturer's specifications for signal input level.
3. Connect the output of the signal generator to the IF input of the RF upconverter.
4. Connect the spectrum analyzer at the receive site to measure the phase noise of the unmodulated carrier as described in the procedure "Measuring System Phase Noise Performance" below.

Return Path Test Configuration

1. Set the signal generator as follows:
   - Output frequency to the return path frequency to be tested.
   - Output amplitude to the maximum permissible level for input to the return injection point. Refer to the manufacturer's specifications for signal input level.
2. Connect the signal generator to a return path injection point at a return node beyond the last upstream frequency conversion.
3. Connect the spectrum analyzer to the headend test point. If possible, measure directly at the fiber receiver's output to maximize the test signal level in an effort to provide the largest possible signal to spectrum analyzer noise ratio.

4. Use the spectrum analyzer to measure the phase noise of the unmodulated carrier as described in the following procedure - Measuring System Phase Noise Performance.

Measuring System Phase Noise Performance

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Set the spectrum analyzer as follows:
   - Center Frequency: Frequency of signal to be measured
   - Frequency Span: 10 times wider than the offset to be measured (for an offset of 10 kHz, set the span to 100 kHz)
   - IF Resolution Bandwidth: ≤ 1/10 of the measurement offset (for an offset of 10 kHz offset, set the IF RBW to 1 kHz, or less)
   - Video bandwidth: Minimum, typically 10 to 30 Hz
   - Detector Mode: Peak
   - Sweep Time: Auto
3. Adjust the spectrum analyzer’s reference level to place the signal in the upper division of the display.
4. Record the signal’s peak amplitude (C_{pk}).
5. Set the spectrum analyzer detector mode to Sample.
   Note: The peak of noise will exceed its power average by an amount that increases (on average) with the length of time the peak is observed. The sample detector is recommended for noise measurements to prevent the peak detector from causing this increase in the measured result.
6. Measure the phase noise amplitude at the desired offset frequency above and below the signal frequency, using the analyzer noise marker function, if available. Record the highest of the two results (PN). If noise markers are used to make this measurement, skip to step 8.
7. If noise markers are not used, calculate the corrected phase noise (CPN) result by adding the following correction factors to the phase noise result measured in step 6.

Convert IF RBW to 1 Hz:	$10 * \log\left[\frac{1\ \text{Hz}}{\text{IF RBW}}\right]$
for a 1 kHz IF RBW:	$10 * \log\left[\frac{1}{1000}\right] = -30.00\ \text{dB}$
Log Detect Rayleigh noise:	+2.5 dB
IF noise equivalent BW:	See manufacturer's specifications (for example: -0.52 dB for 3 dB filtersor +1.10 dB for 6 dB filters)
Noise-near-noise:	Disconnect signal from analyzer and use Figure 3-1: Noise-Near-Noise and Beat-Near-Noise Correction Factors

Example:

Uncorrected PN measured in step 6		= -43.78 dBmV
Phase noise drop when disconnecting signal from analyzer:		= 4 dB (use Figure 3-10 to get 2.2 dB)
	CPN	= -43.78 + (-30.00 + 2.5 - 0.52 - 2.2) dBmV[1]
		= -74.00 dBmV

8. Compute the phase noise performance of the test system:

   Phase Noise (dBc/Hz) = CPN−C_{pk}

[1] Note: The values in parentheses are in decibels; -43.78 is in dBmV.

Figure 3-8: System Phase Noise illustrates an example of a system phase noise measurement result displayed above the test system's reference signal phase noise. The system phase noise result used in this example is (CPK = +24 dBmV):

$$\text{Phase Noise (dBc/Hz)} = -64.94 - 24 = -88.94 \text{ dBc/Hz @ offset of 10 kHz}$$

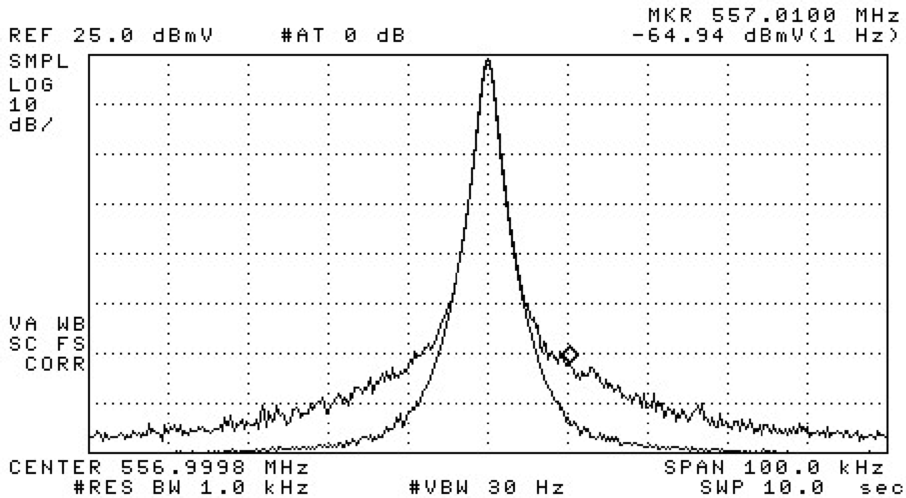


Figure 3-8: System Phase Noise

Chapter 4 Distortion

4.1 Coherent Disturbances – CSO, CTB and XMOD [FCC §76.605(b)(8)]

Definition: Coherent disturbances are undesired signals that occur in cable systems due to factors such as nonlinearities in active devices, ingress, or the addition of unwanted signals to the spectrum by processing equipment.

FCC §76.605(b)(8) The visual signal level to rms amplitude of any coherent disturbances such as intermodulation products, second and third order distortions or discrete-frequency interfering signals not operating on proper offset assignments shall be as follows:

> ***(i) The ratio of visual signal level to coherent disturbances shall not be less than 51 decibels for non-coherent channel cable television systems, when measured with modulated carriers and time averaged; and***
>
> ***(ii) The ratio of visual signal level to coherent disturbances which are frequency-coincident with the visual carrier shall not be less than 47 decibels for coherent channel cable television systems, when measured with modulated carriers and time averaged.***

Discussion: This requirement compares the rms level of in-band interfering signals with the rms voltage produced by the visual signal during the transmission of synchronizing pulses. The interfering signals can have the characteristic of a band of noise or a noise modulated carrier generated by the mixing and combining of distortion products of the system carriers. These are most commonly referred to as Composite Second Order or Composite Triple Beats and are covered in Section 4.1.1: "Composite Second Order (CSO) and Composite Triple Beat (CTB)".

The interfering signals can also be unwanted modulation of the visual carrier caused by modulation components from the other carriers on the system. This is referred to as Cross Modulation and is covered in Section 4.1.2: "Cross Modulation (XMOD)". A third type of coherent disturbance is discrete carriers generated by local oscillators in the headend or hub processing equipment or ingress from over-the-air signals. This type of disturbance is called Discrete Frequency Interference and is discussed in Section 5.2: "Discrete Frequency Interference (DFI)".

The FCC's requirement for coherent disturbances specifies the ratio of visual signal level to the "rms" amplitude of the coherent disturbances. The measurement procedure described in the NCTA recommended practices since 1983 and continued in this SCTE revision uses a spectrum analyzer with logarithmic detection and video filtering to provide the best dynamic range and measurement stability. This method of measuring the coherent disturbance returns the true rms level only if the disturbance approaches a CW carrier. As the coherent disturbance becomes more noise-like, a spectrum analyzer in this mode will return a result up to 2.5 dB lower than the true rms level. In order to maintain continuity with measurement procedures currently in use and to guarantee that the same procedure is used regardless of the type of coherent disturbance, no attempt was made in this revision to adjust the measurement result.

Measure With or Without a Converter?

The FCC specification requires that coherent disturbance measurements be made at the subscriber terminal. This implies that if set-top boxes are used in the system, these measurements must be made at the output of the set-top converter. There are two basic approaches available to meet this requirement: first, is to measure at the output of a representative converter; or second, to measure at the system test points without a converter and adjust the results to compensate for the converter contribution.

Measurements made at the output of the converter have several advantages:

- It may allow making measurements on scrambled channels without disabling the scrambling. This will work for coherent disturbances separated from the visual carrier or CTB using the offset carrier approach with non-baseband converters, discussed in Procedure 1 of Section 4.1.1.
- It provides preselection to the channels being measured to prevent overloading of the measurement device.
- It meets the FCC measurement requirements without compensating the results.
- System frequency response problems are automatically accounted for in the performance measurements.

This approach also has several disadvantages:

- The converter is not designed with the same use requirements as field portable test equipment and is more likely to provide significant uncertainty to the test result if handled improperly.
- AGC or AFC in the converter can cause significant errors when removing the carrier for CTB measurements.
- Measuring CTB at the output of a volume control or baseband converter is complicated by the inability of removing the carrier.

Both approaches have merit. When used properly, either method can be defended as good engineering practice. We have provided procedures which support either approach. If the choice is made to measure at the system test points without a converter, it is recommended that the distortion performance of a sampling of set-top converters be characterized for the compensation of measurement results.

4.1.1 Composite Second Order (CSO) and Composite Triple Beat (CTB)

Definition: Composite Second Order (CSO) is the ratio of the composite beat cluster generated by the direct addition or subtraction of fundamental visual carrier frequencies to the visual carrier level. These products normally fall at ±750 kHz or ±1.25 MHz from the visual carriers. The greatest effect normally occurs at the lower and upper frequencies in use. In IRC systems they will fall exactly 1.25 MHz above and below the carriers, while in HRC systems they will fall exactly on the carriers and be indistinguishable from CTB.

While the use of push-pull and feed forward amplifiers in nearly all cable television coaxial amplifiers minimizes CSO, single-ended equipment, such as fiber-optic and some microwave equipment, exhibits non-symmetrical distortion characteristics, which tend to increase the CSO.

In practice, only the components above the visual carrier are measured. The lower products are usually ignored because they fall between channels. CSO levels, when measured using normally modulated signals, will be about 6 dB better than when measured using CW carriers due to non-synchronization of the modulation.

Definition: Composite Triple Beat (CTB) is the ratio of the composite beat cluster generated by third order distortions resulting from the addition and subtraction of fundamental visual carriers with the second harmonic of other visual carriers or the addition and subtraction of a combination of three visual carriers to the visual carrier level. In a system with standard frequencies (including aeronautical offsets as required), the majority of these distortion products can be measured within ±15 kHz of the visual carriers. If both +12.5 kHz and -12.5 kHz aeronautical offsets are in use, the frequency spread of the distortion is greater and may require a wider measurement bandwidth than the recommended 30 kHz.

Systems are normally designed and specified with performance referenced to the CW carrier test, because this method gives repeatable results. Use of modulated carriers will give results which vary by 2-4 dB over a period of several minutes. CTB measured using normally modulated signals will be about 12 dB better than when measured using CW carriers.

Discussion: Because of the relative nature of this measurement, the absolute accuracy of the test equipment is not as important as the dynamic range and relative accuracy or log scale fidelity. Because of the nature of the distortion products, it is recommended that a 30 kHz resolution bandwidth be used to increase the dynamic range and also that video filtering be used to stabilize the measurement result.

Theoretically, the largest concentration of CSO beats fall at Channels 5 and 6 and at the upper channels. The largest number of CTB products fall at the channels just above the mid frequency of the channel plan. Due to system tilt and amplifier response problems, the worst CTB may occur elsewhere. This will need to be verified in each system and the test channels selected accordingly. For a more in-depth discussion of beat distribution, refer to “Appendix A: Notes on Composite Triple Beat”.

Required Equipment

- Spectrum Analyzer
  - IF Resolution Bandwidth 300 kHz and 30 kHz
  - Video Bandwidth 300 kHz and 10 Hz
- Tunable bandpass filter or fixed bandpass filter for each channel to be measured with 3 dB bandwidth < 15 MHz

Optional Equipment

- Broadband Preamplifier
  - Gain > 15 dB
  - Noise Figure < 10 dB
- Representative Set-top Converter

Test Procedures

Four procedures for measuring CSO and CTB are provided:

1. Measurement of Set-Top Converter Distortion - measures the distortion performance of non-baseband control RF converters
2. Measurement of an Active Channel - measures the distortion performance of the system by removing the video modulation or visual carrier from the channel under test as required
3. Measurement of AM Component on a CW Carrier - measures the composite triple beat performance of the system by measuring the AM modulation component on a CW reference carrier

4. Measurement of Beat Frequency Offset from Carrier - measures the distortion performance of the system by measuring the CSO and CTB in an unused portion of the band and extrapolating the reading to represent the worst case channel

PROCEDURE 1 - Measurement of Set-Top Converter Distortion

The procedure that follows is recommended for measuring the distortion performance of non-volume control RF converters. Because of the wide variety of converters available, any procedure used for this measurement should have the approval of the converter manufacturer.

Note: The distortion performance of modern converters is good enough that their contribution to the subscriber's overall level of distortion is small.

This method may not work on baseband converters because of the signal processing that occurs on the demodulated video signal. Baseband converters require video sync to properly control the video AGC circuits, and therefore complicate the CTB distortion measurement.

For baseband converters, it is recommended that the operator use data provided by the manufacturer. This data shall represent operation of the baseband converter at the particular operating conditions of the system in question. If an accurate measurement of a baseband converter is to be made, the manufacturer's measurement procedure will be required.

To measure the distortion performance of a baseband control RF converter, the visual carrier of the channel being used for measurement must be replaced with a carrier which is the same amplitude as the peak of the original carrier and offset 250 kHz to 500 kHz below the normal visual carrier. This will keep the AGC of the converter at a normal level and allow the CTB product to be seen at the same frequency offset above the substitute visual carrier. The input channel to be measured must be defeated before combining with the signal generator. If the measurement is made at the headend (before any common amplifiers), this can be accomplished by turning off the channel to be tested, otherwise a channel delete filter must be used to remove both the carrier and any system distortion products.

The levels and channel loading used for this measurement must be indicative of the conditions seen by the converter at the subscriber's drop.

Measurement Setup

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Connect the test equipment as indicated in Figure 4-1.

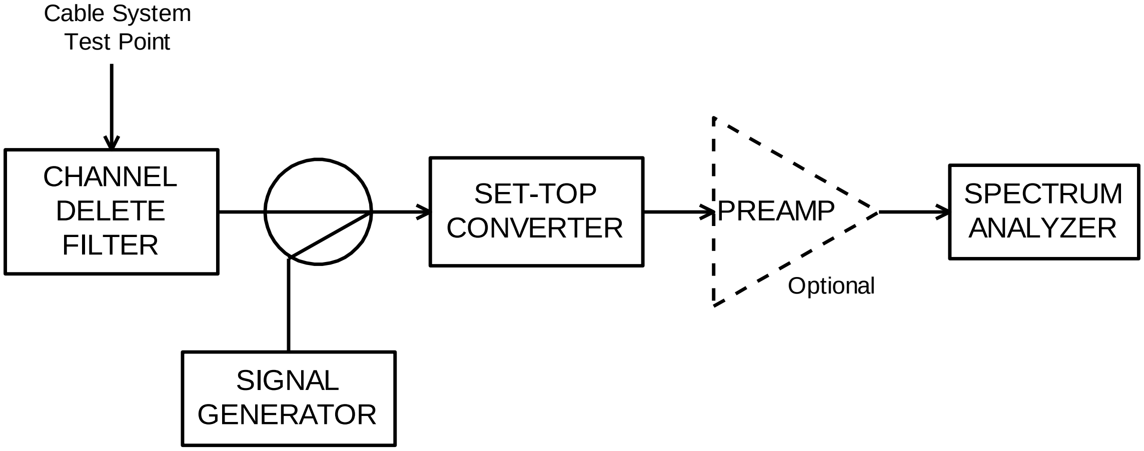


Figure 4-1: Third Order Distortion - Set-Top Converter Test Equipment Setup

3. Set the spectrum analyzer as follows:
   - IF Resolution Bandwidth: 200 kHz or greater
   - Video Bandwidth: Maximum
   - Frequency Span: 5 MHz
   - Amplitude Scale: 10 dB/div
   - Sweep Time: Automatic for calibrated measurement
4. Adjust the spectrum analyzer center frequency to position the converter's visual carrier output to the center of the screen.
5. Adjust the analyzer's full-scale reference to position the carrier in the upper division of the display.
6. Using the analyzer's markers, record the level of the carrier. If the analyzer does not have markers, this may be done by adjusting the full-scale reference to place the carrier exactly on the upper graticule line and recording this full scale reference value.
7. Readjust the analyzer for the following settings:
   - IF Resolution Bandwidth: 30 kHz
   - Video Bandwidth: Minimum (less than 300 Hz)
   - Frequency Span: Unchanged
   - Sweep Time: Automatic for calibrated measurement
   - Number of Video Averages: Until display stability is reached
8. The CTB will be visible above the carrier, offset by the same frequency offset as the visual carrier. The CSO will be visible also offset from its normal ±1.25 MHz and ±0.75 MHz. The marker may be used to measure the peak level of the distortion. If there are no markers, the measurement must be made from the graticules. The beat will move from sweep to sweep due to the low frequency content of the distortion. Video averaging will stabilize the measurement if this feature is available. Otherwise, several traces will need to be observed for an average reading.
9. The distortion magnitude is the difference (in dB) between this level and the carrier level recorded in Step 5. This distortion level (typically near 70 dB) will be combined with the system distortion measured in the next section, to establish the distortion level at the subscriber's terminal.

Note: For the AGC to perform as expected, the frequency response of the converter must be flat from the visual carrier frequency to the frequency of the offset substitute carrier. If, for example, the converter's IF response rolled off by 1 dB at the offset carrier frequency, the CTB would increase by 3 dB.

System Distortion: For the most accurate measurement of distortion products, either the video modulation or the visual carrier must be turned off to measure CSO products. The visual carrier must be turned off to measure CTB products which normally fall at the visual carrier frequency. Some analyzers provide gated measurement capability which samples during quiet lines in the vertical blanking interval, eliminating the need to turn off the video modulation for the CSO measurement. This is instrument specific and is not described in this procedure.

In addition, carrier blanking can be used at the modulator input in the headend to shut off the carrier for selected horizontal scan lines during the vertical blanking interval without causing picture interference. With the appropriate gating capability in the measurement analyzer, measurement samples can be restricted to the time during the carrier blanking and effectively look at the distortion component at the visual carrier frequency. The only customer interference caused by this blanking is a slight buzz in the TV audio which varies depending on the TV receiver used.

This is a much less intrusive approach than shutting off the carrier during the entire measurement. We have not provided a detailed procedure concerning this approach since it is instrument specific, but it is important to be aware that it is available and is an effective way to make the measurement.

PROCEDURE 2 - Measurement of an Active Channel

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Connect the test equipment as indicated in Figure 4-2. The bandpass filter may be necessary to limit the number of carriers (net power) at the input of the analyzer and minimize analyzer created distortions. The preamp should only be used when the signal level available is too low to achieve the required dynamic range.

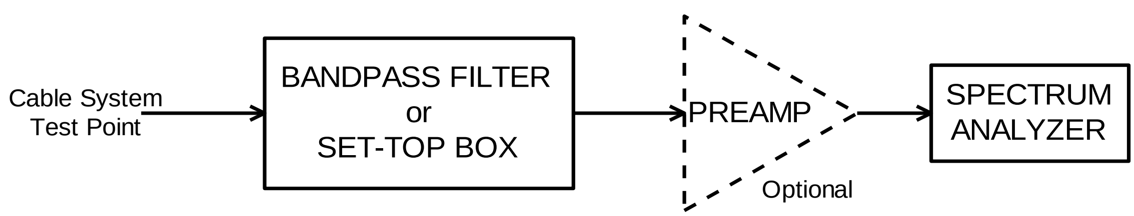


Figure 4-2: Third Order Distortion - Test Equipment Setup

3. Set the analyzer up as follows:
   - IF Resolution Bandwidth: 200 kHz or greater
   - Video Bandwidth: Maximum
   - Frequency Span: 5 MHz
   - Amplitude Scale: 10 dB/div
   - Sweep Time: Automatic for calibrated measurement

4. Adjust the spectrum analyzer center frequency to position the visual carrier of the channel under test to the left of center screen.
5. Adjust the analyzer's full-scale reference to position the peak level in the upper division of the display.
6. Using the analyzer's markers, record the peak level of the carrier. If the analyzer does not have markers, this may be done by adjusting the full-scale reference to place the sync level exactly on the upper graticule line and recording this full-scale reference value.
7. Readjust the analyzer for the following settings:
   - IF Resolution Bandwidth: 30 kHz
   - Video Bandwidth: Minimum (less than 300 Hz)
   - Frequency Span: Unchanged
   - Sweep Time: Automatic for calibrated measurement
   - Number of Video Averages: Until display stability is reached
8. If measuring CTB, the carrier should be turned off at this time. If measuring only CSO, the modulation may be turned off instead of turning off the carrier.
9. If the carrier has been turned off, the display now shows both CTB and CSO. The marker may be used to measure the level of the distortion, or if there are no markers, the measurement must be made from the graticules. The distortion magnitude is the difference (in dB) between the average peak level of the distortion and the carrier. If video averaging is not available, several traces will need to be observed for an average reading.

Notes on Laser Clipping

The CSO measurement is specified with a low video bandwidth setting in the spectrum analyzer in order to find a repeatable average value for the CSO result. This however masks the onset of laser clipping because the high, short duration spikes caused by laser clipping are dramatically reduced by the slow time constant of the video bandwidth. If laser clipping is present, the effect is a higher Bit Error Ratio for the digital carriers. It is possible to check for laser clipping at this point in the procedure by following these steps:

1. Tune the spectrum analyzer to zero span at a CSO beat frequency such as 54 MHz.
2. Set the video bandwidth to 300 kHz or greater.
3. Retain the resolution bandwidth of 30 kHz and other CSO measurement settings.
4. A bandpass filter and external preamp may be needed in some cases to resolve beats at low carrier levels.

Since zero span changes the analyzer display to time domain, laser clipping, if present, will be seen as short duration spikes across the display.

PROCEDURE 3 - Measurement of AM Component on a CW Carrier

This approach eliminates the need to remove the carrier for the CTB measurement and uses a spectrum analyzer as a fixed tuned receiver. This is helpful if communications with the headend are not available or the carrier is not easily controlled. The CTB appears as an AM component on the CW carrier and, by comparing this AM component with the carrier level, the CTB is accurately measured if it is the only AM component present. Hum and Cross Modulation will also appear as AM components and are thus indistinguishable from the CTB with this method.

On an HRC or IRC system, the CTB appears as a CW signal and is not measurable with this method. Caution should be used if this method is applied to set-top converters because of the problems discussed earlier. In addition to the test equipment used in Procedure 2, a true rms reading voltmeter is required.

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Connect the test equipment as indicated in Figure 4-3. The video modulation on the carrier to be measured must be turned off before continuing the measurement.

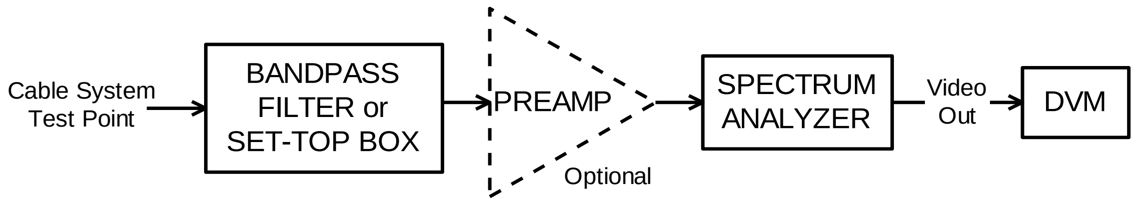


Figure 4-3: Third Order Distortion - Alternate Test Equipment Setup

3. Set the analyzer up as follows:
   - IF Resolution Bandwidth: 200 kHz or greater
   - Video Bandwidth: 30 kHz
   - Frequency Span: 500 kHz
   - Amplitude Scale: Linear (not logarithmic)
   - Sweep Time: Automatic for calibrated measurement
4. Adjust the spectrum analyzer center frequency to move the peak of the visual carrier to center screen.
5. Adjust the analyzer's full-scale reference to position the carrier near the top of the display. If possible, the analyzer's attenuator should be set to 0 dB.
6. Set the analyzer frequency span to 0 MHz.
7. With the voltmeter set to DC, record the voltage of the carrier reference level.
8. Set the voltmeter to AC and record the level of the CTB imposed on the carrier.
9. The distortion magnitude is:

$$CTB = 20 * \log\left[\frac{VOLTS_{AC}}{VOLTS_{DC}}\right] + 0.5dBc \quad (1)$$

Note: Refer to “Appendix B: Notes on the Measurement of CTB with a CW Carrier Present” for more discussion.

PROCEDURE 4 - Measurement of Beat Frequency Offset from Carrier

This procedure eliminates the need to disrupt an active channel by measuring the distortion in an unused portion of the band and extrapolating the reading to represent the worst case channel. The accuracy of this approach is dependent upon the accuracy of the system distortion characterization that is done in Step 1. It is not as accurate as the other procedures, but serves as an acceptable method for guaranteeing compliance.

The first step in this procedure is to characterize the system distortion performance versus channel. This must be done for each portion of the system with unique channel loading, system tilts, AML or fiber links, amplifier spacing, etc. This first step does require interruption of service, but only once as long as the system configuration does not change. The best way to describe this method is by example.

Step 1 - System Characterization

By measuring a representative number of channels throughout the system bandwidth (approximately every 50 MHz), a plot can be generated similar to Figure 4-4. The plot is normalized to the worst-case distortion by subtracting the worst-case value from all others. One plot should be generated for CSO and another for CTB. Either Procedure 1 or Procedure 2 may be used to make this reference measurement. Figure 4-4 is a typical representation of the beat distribution and will not vary significantly from system to system. For a more in-depth analysis of beat distribution, refer to "Appendix A: Notes on Composite Triple Beat".

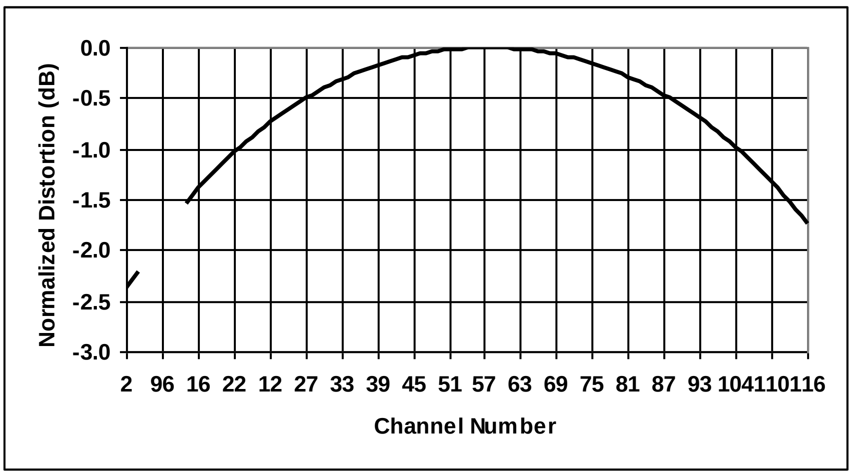


Figure 4-4: Example of Normalized System Distortion vs. Channel

Step 2 - Distortion Measurement

Using Procedure 1 or Procedure 2 with a slight modification, the distortion may be measured at an unused portion of the band. A CW carrier is inserted 6 MHz above the highest visual carrier or at a 6 MHz increment in an unused portion of the band. This carrier must be at the same level as the other carriers in the system. Alternately, an adjacent visual carrier may be used as the reference for this measurement, but the result must be adjusted for the difference in system gain and tilt between the frequency of the measured carrier and the frequency of measuring the distortion. If the system is well behaved and flat, this approach works well. It should be noted that this is quite often NOT the case. Following Procedure 1 or Procedure 2, the distortion is measured at this location.

Step 3 - Extrapolation

The distortion measured in Step 2 (a negative number) is adjusted by the correction factor from Figure 4-4 to represent the worst-case system distortion. For example, if the distortion measured at channel 116 in Step 2 is -58.3 dBc, the worst-case distortion at channel 57 would be:

$$\text{CTB} = \left[-58.3 + 1.7\right] = -56.6 \text{ dBc} \qquad (2)$$

Note Potential Errors

A. Analyzer Noise Floor

If the level of the distortion being measured is not at least 10 dB above the analyzer noise floor, additional attenuation should be removed or a correction factor will need to be used (see Figure 4-5). This can be checked by removing the input cable and looking for at least a 10 dB drop in the level at the frequency of the distortion. If the change is less than 10 dB, the measured level of the distortion should be decreased according to Figure 3-1. As an example, if the beat measured -55 dB and it dropped 7 dB when the input is removed, the corrected beat measurement would be:

$$-55 - 1.0 = -56\,\text{dB}$$

B. Analyzer / Preamp Distortion

If the input level to the analyzer (or preamplifier) is too high, the system distortion will be masked by the measurement system distortion. This can be checked by increasing the input attenuation by a small amount and verifying that the magnitude of the distortion products changes by the same amount. If it is measurement system distortion, the distortion products will drop by a greater amount.

C. Preselector Tuning

If the preselector being used is too narrow, the CSO may be outside the passband of the filter. To prevent this, the preselector should be adjusted to peak the distortion. If a fixed filter is being used, the passband characteristic must be checked beforehand.

Calculation of Distortion at the Subscriber Terminal

Combining the measured distortion of the system with the performance of the set-top converter can be done either mathematically or graphically. Both methods are presented here using the following values as an example:

Set-top converter distortion: $$\text{DIST}_{\text{CONV}} = -75\,\text{dBc} \quad (3)$$

System distortion: $$\text{DIST}_{\text{SYS}} = -56\,\text{dBc} \quad (4)$$

Combined distortion: $$\text{DIST}_{\text{SUB}} = 20 * \log\left[10^{\frac{\text{DIST}_{\text{CONV}}}{20}} + 10^{\frac{\text{DIST}_{\text{SYS}}}{20}}\right] = -55.1\,\text{dBc} \quad (5)$$ [2]

Graphically, using Figure 4-5, the difference between the two readings is 75 - 56 = 19 dB. From the figure, the correction factor is 0.9 dB. The distortion at the subscriber's terminal becomes:

Combined distortion: $$\text{DIST}_{\text{SUB}} = -56 + 0.9 = -55.1\,\text{dBc} \quad (6)$$

[2]This is a worst case calculation. The actual distortion can vary from 10*log to 20*log.

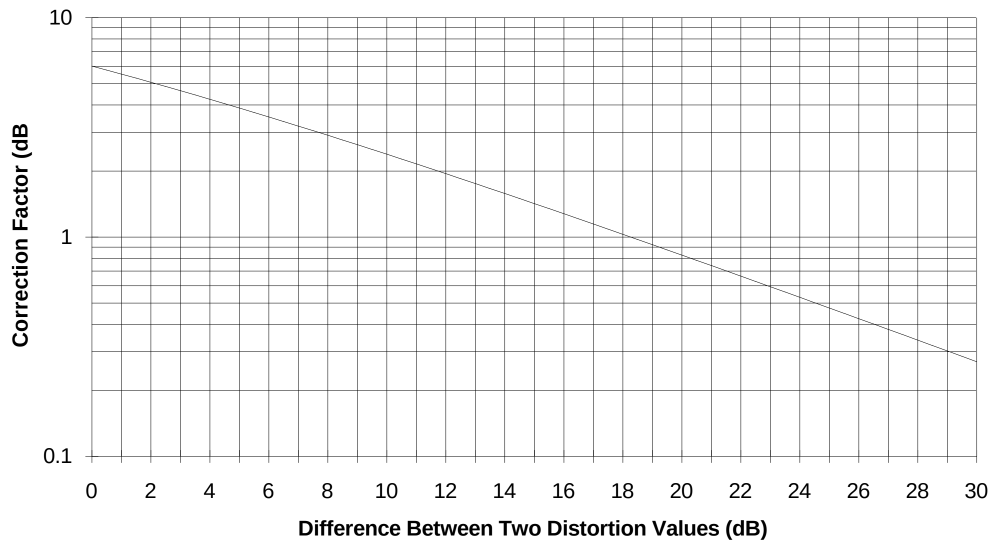


Figure 4-5: Combining Two Distortion Values

Appendix A: Notes on Composite Triple Beat

In a system with standard frequencies, composite triple beat distortion products are measured within ±15 kHz of the visual carriers. It is important to note that although the definition of composite triple beat allows an A+B+C product as well as a 3*A product, these products do not fall on a channel and are usually ignored. There is another third order distortion product that does not conform to the usual definition but whose products do fall on a channel. These are 2*A+B products. Using trigonometry, it can be shown from a trigonometric argument that these products are one-half (-6 dB) of the magnitude of the A+B-C products, and are fewer in number. As a result, in a 20-channel system, their contribution is approximately 0.1 dB and their contribution decreases with an increasing number of channels.

We can define CTB products mathematically as follows:

$$\text{CTB's} = A + B - C$$

where A, B, and C are any of the frequencies on the system, and

$$A \neq B \neq C \quad \text{and} \quad A < B$$

There is a useful expression that can be used to find the number of beats on any channel, with only a very small error. This expression is:

$$\text{Beats} = \frac{N^2}{4} + \frac{(N - M)(M - 1)}{2}$$

where:

Beats = number of beats
N = total number of channels
M = number of the channel being measured

In the middle of the spectrum, the expression simplifies to:

$$\text{Beats} = \frac{3N^2}{8}$$

At the edges of the spectrum, the expression simplifies to:

$$\text{Beats} = \frac{N^2}{4}$$

Also the number of beats at the edges of the band is 2/3 the number of beats in the middle of the band. This ratio is independent of the number of channels provided there are a reasonable number of channels present.

Appendix B: Notes on the Measurement of CTB with a CW Carrier Present

It is sometimes necessary to measure Composite Triple Beat products even though a carrier is present on the channel. The following is an analysis of one method of making these measurements and its results.

Consider a single carrier with a magnitude V_C added to a single distortion product with magnitude V_D. A vector representation is shown in Figure 4-6. The sum of these two signals is now detected in a linear detector and the result is passed through a lowpass filter to eliminate the carrier. This method of analysis yields a result that is easily understood. The output of this first lowpass filter is further split into DC and AC components. Both the DC and the AC voltages can now be measured and the actual distortion ratio calculated.

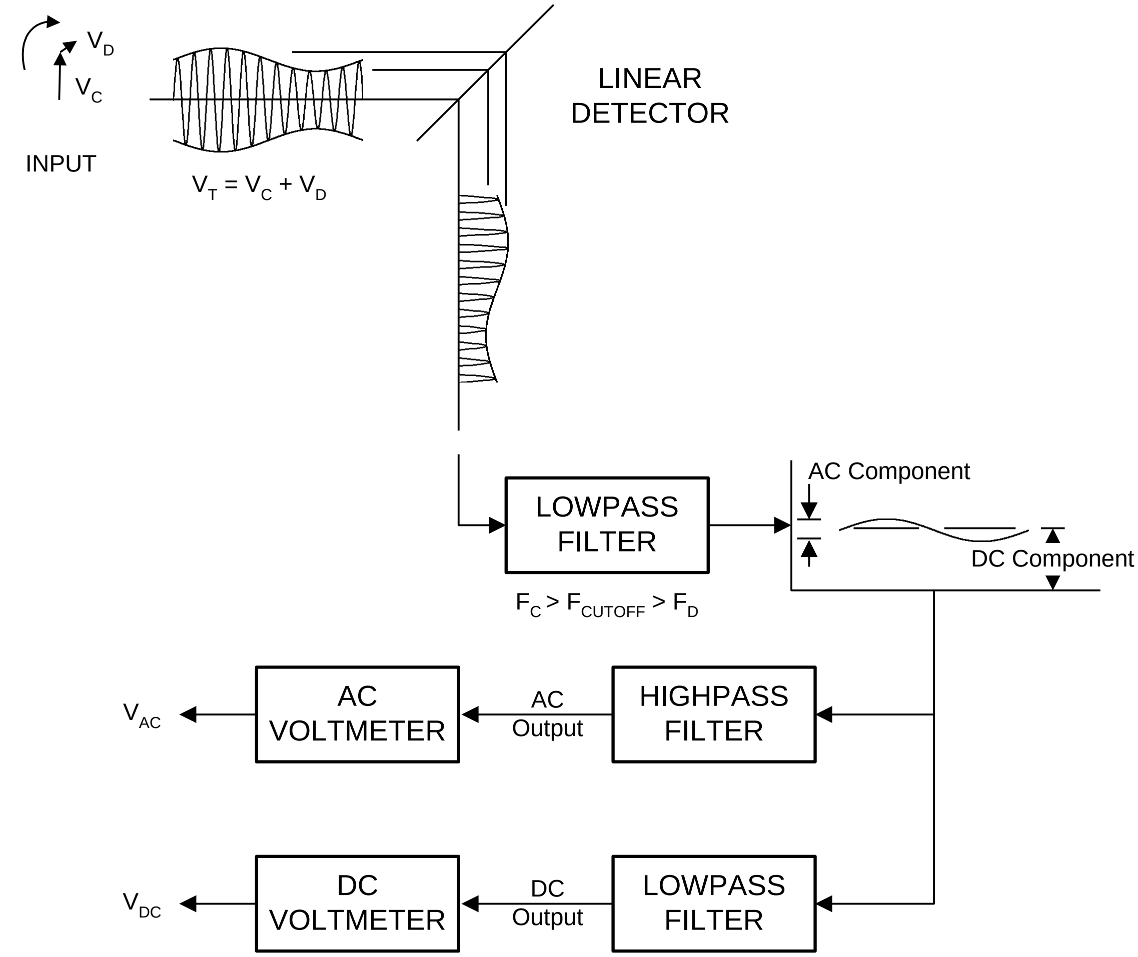
$$V_C = \text{Carrier Voltage}$$

$$V_D = \text{Distortion Voltage}$$

$$V_T = \text{Total Voltage (carrier + distortion)}$$

$$V_T(\text{peak}) = V_C + V_D$$

$$V_T(\text{valley}) = V_C - V_D$$



$$\mathrm{DC\,Output} = V_{DC} = kV_C \quad \text{and} \quad \mathrm{AC\,Output} = V_{AC} = kV_D$$

where k = detector circuit constant

$$\mathrm{CTB\,(dBc)} = 20 * \log\left[\frac{V_{AC}}{V_{DC}}\right] + 0.5\,\mathrm{dB} \quad \text{(for an rms reading DVM)}$$

$$\mathrm{CTB\,(dBc)} = 20 * \log\left[\frac{V_{AC}}{V_{DC}}\right] - 0.5\,\mathrm{dB} \quad \text{(for a rectifying averaging DVM)}$$

Figure 4-6: Measuring CTB with a CW Carrier Present

The cutoff frequency of the highpass filter in AC voltmeters is normally well below 60 Hz, and this measurement will capture the majority of the low frequency components. The output of the high and lowpass filters can be described as follows:

AC output (peak-peak) $= k[(V_C + V_D) - (V_C - V_D)]$

$= 2kV_D$

DC output $= k\left(\frac{(V_C+V_D)+(V_C-V_D)}{2}\right)$

$= kV_C$

In our example, the rms value of the AC output is:

$$\overline{V_D} = \tfrac{1}{2}2kV_D \tfrac{\sqrt{2}}{2}$$

$$= \frac{kV_D}{\sqrt{2}} = \text{ rms value of the distortion}$$

$\overline{V_D}$ is the rms value of the distortion product and is what is read by the rms reading voltmeter.

$$V_D = \frac{\sqrt{2}\ \overline{V_D}}{k}$$

DC output $= kV_C$

And we note that the rms value of the detected carrier is the same as its DC value because it is only a DC voltage, then:

$$\overline{V_C} = kV_C$$

$$V_C = \frac{\overline{V_C}}{k}$$

If we define distortion as:

$$D(dBc) = 20 * \log\left[\frac{V_D}{V_C}\right]$$

$$= 20 * \log\left[\frac{\sqrt{2}\ \overline{V_D}}{k} \frac{k}{\overline{V_C}}\right]$$

$$= 20 * \log\left[\frac{\sqrt{2}\ \overline{V_D}}{\overline{V_C}}\right]$$

$$= 20 * \log\left[\frac{\overline{V_D}}{\overline{V_C}}\right] + 20LOG\left[\sqrt{2}\right]$$

$$= 20 * \log\left[\frac{\overline{V_D}}{\overline{V_C}}\right] + 3\text{ dB}$$

This expression tells us that if the measurements are made with an rms reading voltmeter then we must add 3 dB to the reading. If:

$$20 * \log\left[\frac{V_D}{V_C}\right] = -50\text{ dBc}$$

then the distortion would be:

$$-50 + 3 \text{ or } -47\text{ dBc}$$

So far we have considered only one distortion product. We may have many unrelated products and they can be described as:

$V_D1, V_D2, V_D3, ...$

Since the distortion products are not correlated, we must add their magnitudes as we would add their powers. The total distortion V_t is:

$$\overline{V_T} = \sqrt{\overline{V_D1}^2 + \overline{V_D2}^2 + \overline{V_D3}^2 ...}$$

This does not alter our analysis and the distortion for the case of many products is:

$$D_T(dBc) = 20 * \log\left[\frac{\overline{V_T}}{\overline{V_C}}\right] + 3\,dB$$

where D_T is the total distortion.

We note that the recommended measurement procedure for CTB distortion contained herein uses a spectrum analyzer in the LOG mode to make these measurements. These distortion products, which are made up of many uncorrelated signals, behave as noise and the spectrum analyzer measures these signals about 2.5 dB weaker than the true power in the products. This 2.5 dB correction is used when measuring carrier-to-noise ratio, but is not used for CTB measurements. If we wish to correct our reading to make it agree with the spectrum analyzer, we may do so by adding -2.5 dB to the reading.

$$\begin{aligned} CTB(dBc) &= D_T - 2.5\,dB \\ &= 20 * \log\left[\frac{\overline{V_T}}{\overline{V_C}}\right] + 3\,dB - 2.5dB \\ &= 20 * \log\left[\frac{\overline{V_T}}{\overline{V_C}}\right] + 0.5\,dB \text{ (for an rms reading voltmeter)} \end{aligned}$$

We should cover one more subject which is what happens when a rectifying averaging voltmeter is used in place of the rms voltmeter. Fortunately, this has been analyzed for us and the relationship is well known, but not necessarily easy to understand. The subject is treated well in "Electrical Noise" by Bennett.[3]

According to Bennett, a rectifying averaging voltmeter reads lower than an rms voltmeter, by almost exactly 1 dB, when measuring noise. If we were to use this type of voltmeter to make these measurements, then:

$$\begin{aligned} CTB(dBc) &= 20 * \log\left[\frac{\overline{V_T}}{\overline{V_C}}\right] + 3\,dB - 1dB - 2.5\,dB \\ &= 20 * \log\left[\frac{\overline{V_T}}{\overline{V_C}}\right] - 0.5\,dB \text{ (for a rectifying averaging voltmeter)} \end{aligned}$$

[3] William R. Bennett, *Electrical Noise* (McGraw-Hill, 1960)

4.1.2 Cross Modulation (XMOD)

Definition: Cross Modulation (XMOD), also called *crossmod*, is a distortion phenomenon in which modulation from other carriers are impressed on the test carrier. Cross Modulation is defined as the difference between the detected cross modulation level and the detected level that would correspond to 100% modulation, expressed in dB. It mathematically derives from the third order term in the power series model of distortion in linear amplifiers.

Discussion: In the early days of cable television, when systems carried few channels, cross modulation was often a major limiting factor in system performance. As cable systems added more channels, the impact of cross modulation was reduced as a result of the statistical nature of modulation: some signals are increasing in amplitude while others are decreasing and over many channels the average amplitude will be much lower than if all signals varied synchronously.

Today CSO and CTB are usually taken as better predictors of visible impairment in HFC networks. Many practitioners do not feel there is any need to conduct crossmod tests to show compliance with FCC rules, since composite third order distortion (CTB) measures the same thing and is more closely related to picture impairments, and compliant CTB measurement results indicate that XMOD measurement results will also comply with the rules.

One value in the cross modulation measurement lies in the study of certain narrowband phenomena that look like cross modulation, though in some cases they are not caused by third order distortion. When modulation from a *limited* number of channels is imposed on another channel, the statistics of large channel counts don’t work, and visible cross modulation can once again prove to be a limiting factor.

An example of this is in characterizing the performance of the first intermediate frequency amplifier in a set-top converter which is supposed to amplify only the tuned channel. At the location of the first IF amplifier most non-tuned channel signal power has been rejected, but signals close to the frequency of the tuned signal have not been completely rejected, and can create cross modulation distortion. Since a small number of signals are involved, the cross modulation can be much closer to the worst-case condition of synchronous modulation than it would be if a large number of signals contributed.

Because Cross Modulation distortion is defined as amplitude modulation transferred onto a channel by other carriers in the system, an envelope detector is required to measure it. It has been proven that simultaneous cross modulation measurements with an envelope detector and with a spectrum analyzer measuring the sideband energy, can differ greatly. A spectrum analyzer in the frequency domain measures the energy in the carrier wave and the side bands of a channel under test independently of the phase relationship of these side bands to the carrier. If both amplitude and phase distortion are present, the resultant cross modulation varies from amplitude modulation to phase modulation as a function of frequency. This causes the standard envelope detector method to indicate significantly more favorable results than a spectrum analyzer in the frequency domain.

Several spectrum analyzers in recent years have provided functionality which allows the XMOD AM component on a CW carrier to be measured directly. This includes narrowband 15,734 Hz AM envelope detectors, and time domain functions which sample the detected signal in the time domain, and then perform a Fast Fourier Transform (FFT) on the time domain signal to determine the frequency content at 15,734 kHz. Because these methods are instrument specific, they are not presented here, but they do enable the measurement of the AM component only without generating a 100% modulated reference.

Required Equipment

- Signal generator
- Tunable to the channel to be measured and used to replace the channel in the headend with a CW carrier.
- Capable of 100 % downward amplitude modulation at a frequency of 15,734 Hz with a duty cycle of 50 %.

Note: The 100% modulation reference is not required if the measurement analyzer has time domain FFT capability or narrowband 15,734 Hz AM modulation detection.

- RF receiver or spectrum analyzer
- Bandwidth selectable IF filter, typically 300 kHz or 1 MHz
- Square law or true rms detector for the measurement of AM-XMOD
- Baseband signal output that is directly proportional to the detected modulation signal. The impedance of this output must be known and constant. Typical values are 75 Ω, 50 Ω, and 600 Ω.

Note: The baseband signal output is not required if the analyzer has time domain FFT capability or narrowband 15,734 Hz AM modulation detection.

Optional Equipment

- Baseband analyzer or tuned low frequency (15,734 ± 20 Hz) voltmeter

Note: The tuned low frequency voltmeter is not required if the analyzer has time domain FFT capability or narrowband 15,734 Hz amplitude modulation detection.

- Bandpass filter for each channel to be tested or a tunable bandpass filter.
- Filter should have a bandwidth of between 1 MHz and 6 MHz. This is to prevent distortion from occurring in the measuring equipment and therefore adjacent channel rejection is the critical requirement.
- Broadband Preamplifier (required if measuring with low signal levels)
- Gain > 15 dB
- Noise Figure < 10 dB

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Connect the test equipment as indicated in Figure 4-7.

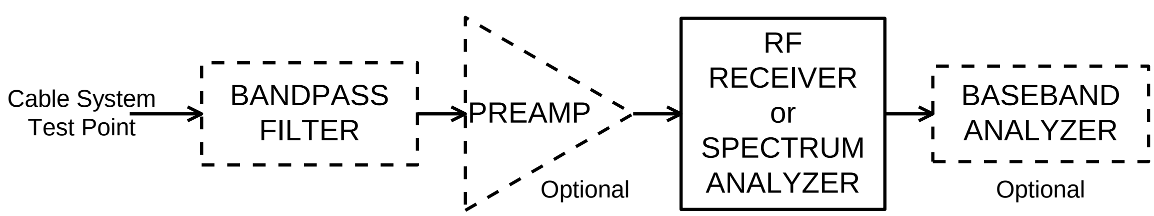


Figure 4-7: Cross Modulation Test Equipment Setup

3. Tune the bandpass filter to the channel under test.

4. If using an analyzer with time domain FFT capability or narrowband 15,734 Hz amplitude modulation detection, the measurement may be made directly at this point, following the manufacturer's procedure. The user needs to verify that the measurement result is calculated

relative to the 15,734 Hz component of 100% modulation, and not relative to the carrier amplitude.

5. If not using an analyzer with direct AM XMOD measurement capability, set the RF Receiver as follows:
   - Center Frequency: Carrier to be measured or set-top visual carrier output frequency
   - Frequency Span: Zero
   - IF Resolution Bandwidth: 1 MHz
   - Video Bandwidth: > 1 MHz
   - Amplitude Scale: Linear (not logarithmic)
   - Sweep Time: 200 μs
6. Turn the 15,734 Hz squarewave, 100 % downward amplitude modulation ON for the carrier under test.
7. Adjust the reference level of the spectrum analyzer so that the signal displayed is at the full scale of the display.
8. Tune the baseband analyzer to 15,734 Hz. Ensure that the signal displayed on the baseband analyzer is at the full scale of the display. Ensure that the resolution bandwidth of the baseband analyzer is ≤ 100 Hz. Note that a narrower baseband measurement bandwidth will produce a lower measurement noise floor and thus a more stable, more repeatable measurement.
9. Record the level of the 15,734 Hz signal measured on the baseband analyzer as *Reference Sideband Level*.
10. Turn the modulation OFF for the carrier under test.
11. Record the level of the 15,734 Hz signal measured by the baseband analyzer as the *RAW XMOD Level*.
12. Tune the baseband analyzer to a flat portion of the spectrum, within the video bandwidth of the RF receiver.
13. Record the level measured by the baseband analyzer as *Noise Floor Level.* Compute the following:

    Noise Drop = RAW XMOD Level – Noise Floor Level.
14. If the *Noise Drop* is less than 2 dB, it is recommended that the optional preamplifier be added to the system. The measurement should then be made again.
15. If the *Noise Drop* is greater than 2 dB, the following *Noise Floor Correction Factor* should be calculated:

    $$\textit{Noise Drop Correction Factor} = 10 * \log\left[1 - 10^{\left(\frac{-\text{Noise Drop}}{10}\right)}\right] \text{ dB}$$
16. Compute Corrected XMOD as:

    $$\textit{XMOD (referenced to sideband)} = \left|\textit{Reference Sideband Level} - \textit{RAW XMOD Sideband Level} + \textit{Noise Drop Correction Factor}\right|$$

Note: This is a positive number, expressed in –dBc.

4.2 Single Second Order (SSO) and Single Third Order (STO) Intermodulation Distortion in the Upstream

Definition: Distortion is an undesired change to the signal as it passes through the cable system. This change may be due to many factors including the nonlinear characteristics of amplifiers, lasers, and photodetectors or the addition of unwanted signals to the spectrum such as conducted or radiated ingress. Single Second Order (SSO) distortion is defined as the ratio in decibels of the level of a second order distortion beat to the level of the carrier(s). Single Third Order (STO) distortion is defined as the ratio in decibels of the level of a third order distortion beat to the level of the carrier(s). Both are expressed with the unit dBc which is decibels relative to carrier.

Other contributors to distortion such as noise, ingress, and hum modulation are discussed in other sections.

Note: This procedure requires free spectrum space on the upstream plant and can cause disruption to any traffic which might exist on the plant during the tests. This test is often performed to verify the alignment and performance of rebuilt or newly built plant. Before performing this test, the plant must be properly setup and aligned. For more information refer to Section 16.4: "Return Plant Setup and Operational Practices".

Discussion: The test is performed by inserting two carriers into the plant (normally at the end of line), and measuring the resulting distortion at the headend. When these two carriers go though nonlinearities (such as amplifiers, lasers, and photodetectors), second and third order distortion products are produced. Assuming that the frequencies of the two carriers are A and B (B>A), this procedure tests the A+B and B-A second order products and the (2*A)-B and (2*B)-A third order products. A, B, and all measured products must fall within the passband of the system.

Required equipment

- Two signal generators or one multi-carrier generator
- Spectrum analyzer
- Optional notch filter

Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Connect the injection equipment in the plant as shown in Figure 4-8.

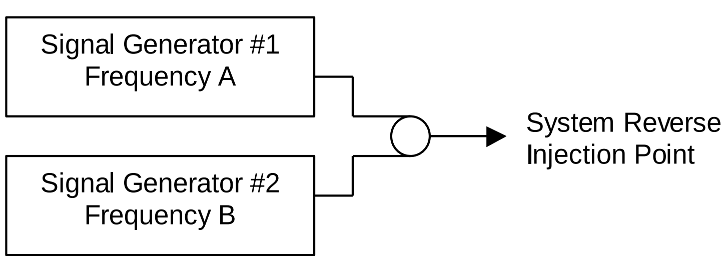


Figure 4-8: Equipment Connection

3. Turn on the signal generators at the desired frequencies and power level. In general, the proper power setting for the signal generators will be the level which hits the amplifiers (and the laser) at the proper total power. Record the lower frequency as "A" and the higher frequency as "B".
4. Connect the spectrum analyzer to the signal to be measured.
5. Set the spectrum analyzer as follows:
   - Center Frequency: "A" MHz
   - Frequency Span: Approx. 1 MHz
   - Amplitude Scale: 10 dB/div
   - IF Resolution Bandwidth: Approx. 30 kHz
   - Video Bandwidth
     and/or

   Number of Video Averages: Until display stability is reached
   - Sweep Time: Automatic for calibrated measurement
   - Input Attn: As required (see text)
6. Adjust the analyzer reference level so that the carrier is close to the top of the screen.
7. Put the marker on the carrier. Record the carrier level.
8. Tune the spectrum analyzer to frequency "B". Put the marker on the carrier and record the carrier level.
9. Compare the two-carrier levels and verify that the difference between the two is within the required plant flatness.
10. Calculate the average carrier level as follows:

$$\text{Average Carrier Level} = 10 * \log\left[\frac{10^{\frac{\text{Carrier A Level}}{10}} + 10^{\frac{\text{Carrier B Level}}{10}}}{2}\right]$$

    Or, use power addition tables to add the powers of the two carriers and then subtract 3 dB.
11. Tune the spectrum analyzer to the A+B beat frequency.
12. Put the marker on the top of the beat. Verify that the spectrum analyzer is not contributing to the distortion reading by varying the spectrum analyzer's input attenuator. Choose the setting which gives the lowest beat level (assure that changing the attenuator by an additional step does not lower the beat further). Calculate the SSO as follows:

    SSO = Average Carrier Level - Beat Level
13. Tune the spectrum analyzer to the A-B beat frequency and measure the SSO as in the previous step.
14. Tune the spectrum analyzer to the (2*A)-B beat frequency.
15. Put the marker on the top of the beat. Verify that the spectrum analyzer is not contributing to the distortion reading by varying the spectrum analyzer's input attenuator. Choose the setting which gives the lowest beat level (assure that changing the attenuator by an additional step does not lower the beat further). Calculate the STO as follows:

    STO = Average Carrier Level - Beat Level

16. Tune the spectrum analyzer to the (2*B)-A beat frequency and measure the STO as in the previous step.

Notes, Hints and Precautions

If spectrum analyzer distortion cannot be avoided, a filter must be used in front of the analyzer. A channel bandpass filter or a notch filter (to remove one of the two carriers) could be used. If a bandpass filter is used, several could be required depending on the location of the carriers and beat frequencies. If a notch filter is used it should have at least 20 to 30 dB of rejection at one of the carrier frequencies and have a constant loss (flat response) at all the measurement frequencies, including the other carrier frequency. When this filter is used, all readings must be referenced to the carrier that is not attenuated.

As an alternative to the above procedure, all beat measurements can be compared to a carrier at that same frequency. To do this, one of the signal generator frequencies needs to be moved from its normal frequency (A or B) to the beat frequency being measured in order to obtain the carrier level reading. Then the signal generator frequency must be moved back to its original frequency (A or B) before the distortion is measured.

Be aware that the distortion products being measured might not be SSO or STO products. For instance, common-path distortion could exist at the same frequency. To check for undesired signals, remove the A and B frequencies and make certain that all SSO and STO beats disappear.

A two-carrier test does not fully stress the capabilities of the return system. In particular, the peak-to-average ratio of two carriers is not as high as the peak-to-average ratio of a fully loaded system. For a further discussion of this issue see Section 16.2: “Peak Voltage Addition”.

4.3 Common-Path Distortion (CPD)

Definition: Common-Path Distortion is the common name for intermodulation products that are observed in sections of the system where both upstream and downstream signals are present. As discussed below, these distortion products are the nonlinear combination of all signals present but have been classically observed in the bi-directional part of the plant.

Discussion: Common-Path Distortion (CPD) is most easily observed in sections of the plant where both the upstream and downstream signals are present. Common-path distortions differ from the ordinary second and third order products only in their sources. Tap plate and seizure connections and end of line terminators are often the locations of the nonlinear elements, which generate the distortion rather than amplifiers, or other active elements. These nonlinear elements are largely local diode-like structures at connection points resulting from dissimilar metal contacts and/or corrosion activity. All signals impinging on these nonlinear elements generate distortion products although the ones usually observed are primarily second and third order products from the combination of downstream visual carriers (other second and third order products exist but are generally of minor importance). Second order products fall at multiples of 6 MHz (in systems with 6 MHz video channel spacing) while third order products appear at 1.25 MHz above and below these frequencies in standard and IRC systems and on-frequency in HRC systems. The contributions of offset channels such as standard channels 5 & 6 are usually not discernible. It should be noted that these products in non-HRC systems seldom exactly overlay one another in frequency since carrier frequency deviations from nominal and intentional offsets (such as those employed for aviation band channels) result in bunches of products in the near vicinity of the nominal locations. While present in both upstream and downstream portions of the system these products are often obscured by carriers and other distortion products in the downstream plant but usually quite apparent in the upstream since there are no downstream signals to obscure them.

Common-path products are elusive, often appearing and disappearing over time. At times none or only one variety is observed (2nd or 3rd order) and sometimes both are seen simultaneously. Amplitudes also vary with conditions. Repairs of contact related rectification elements are often short-lived and may require more drastic actions.

It should also be recognized that the increasing presence of high level return carriers in passive elements such as splitters and taps, opens another avenue for unwanted intermodulation product creation. These devices, when driven with such levels, are prone to transformer or toroid saturation. This phenomenon has effects similar to CPD, by creating unwanted mixing components. Addressing this effect requires out-of-service component evaluation and therefore is not covered in this section.

Common-path and common-path like products are capable of interfering with services as are ingress and other spurious signals. Offending equipment must be located, monitored and corrected before this interference becomes intolerable.

Required Equipment

- Spectrum analyzer (required)

Optional Equipment

- Spectrum capture system capable of unattended periodic sampling (recommended)
- Bandpass or lowpass filter

Test Procedure

Although some common-path distortion may be observed at virtually any drop in the system a meaningful measurement must be made only through passive device(s) whose through path is aligned with the reverse system signal flow. Use of forward system drop ports of directional couplers or taps inserts the passive directional loss in series with CPD products generated further down the forward path and thus distorts the measurement by this attenuation.

The spectrum analyzer must be adjusted so that the system noise exceeds the analyzer noise floor preferably by at least 10 dB. A preamp may be required in some cases. The span and resolution bandwidth settings of the spectrum analyzer are determined by the desired resolution of the data to be taken. In the general case where assessment of the overall contamination of the reverse path is desired a span covering the entire reverse spectrum and employing a 100 kHz resolution bandwidth will indicate the areas of maximum and minimum contamination and thereby assist in selection of the optimum frequency regions for placement of upstream services. Narrower resolution bandwidths and spans may be employed to determine the fine structure of the CPD for more detailed analysis to better understand the generation source(s) of these components or to assist in employment of more sophisticated reduction and/or avoidance techniques.

Caution should be applied to measurements of CPD since it can change slowly or abruptly by significant amounts making it very elusive and requiring a time record of frequency and amplitude to adequately characterize the subject system section under consideration. It is this characteristic which suggests the use of the spectrum capture system recommended above to record the time variations and thereby present a more comprehensive picture of the CPD interference to the system services.

4.4 Low Frequency Disturbances and Hum – Analog Video Carriers

Definition: Low frequency disturbances are the result of unwanted modulation of the desired signal at a rate less than 1 kHz, usually related to the power system frequency or to the video vertical synchronizing frequency. "Hum" is a type of low frequency disturbance and is the amplitude distortion of the desired signals caused by the modulation of these signals with components of the power source.

Low frequency disturbances are expressed as the ratio of the peak-to-peak interference to the peak level of the desired picture carrier in percent.

FCC §76.605(b)(10) ***The peak-to-peak variation in visual signal level caused by undesired low frequency disturbances (hum or repetitive transients) generated within the system, or by inadequate low frequency response, shall not exceed 3 percent of the visual signal level. Measurements made on a single channel using a single unmodulated carrier may be used to demonstrate compliance with this parameter at each test location.***

See “PROCEDURE 1a - Manual Procedure for a Modulator or Processor Using a Spectrum Analyzer” in this section for the procedure used to show compliance with the FCC requirement.

Discussion: This distortion can occur in any piece of active equipment. It is usually due to modulation of the signal by the power line frequency (60 Hz) and its harmonics. Equipment using switching power supply regulation can produce the same form of distortion at higher frequencies. The major sources of power line hum (60 and 120 Hz) are amplifiers with defective module supplies and overloaded system supplies. However, other components, even passives, can introduce hum modulation distortion under certain conditions.

Other low frequency variations may also occur. These may include very low frequency distortion caused by AGC hunting and system intermittents caused by wind and other factors.

A second class of low frequency distortion is disturbances which are caused by, and therefore directly related to, the video waveform. This type of distortion usually manifests itself as field rate transients associated with the vertical sync waveform (Figure 4-9). It is frequency related to the response of AGC systems to the vertical synchronizing information. The time constants of video circuits can have similar effects.

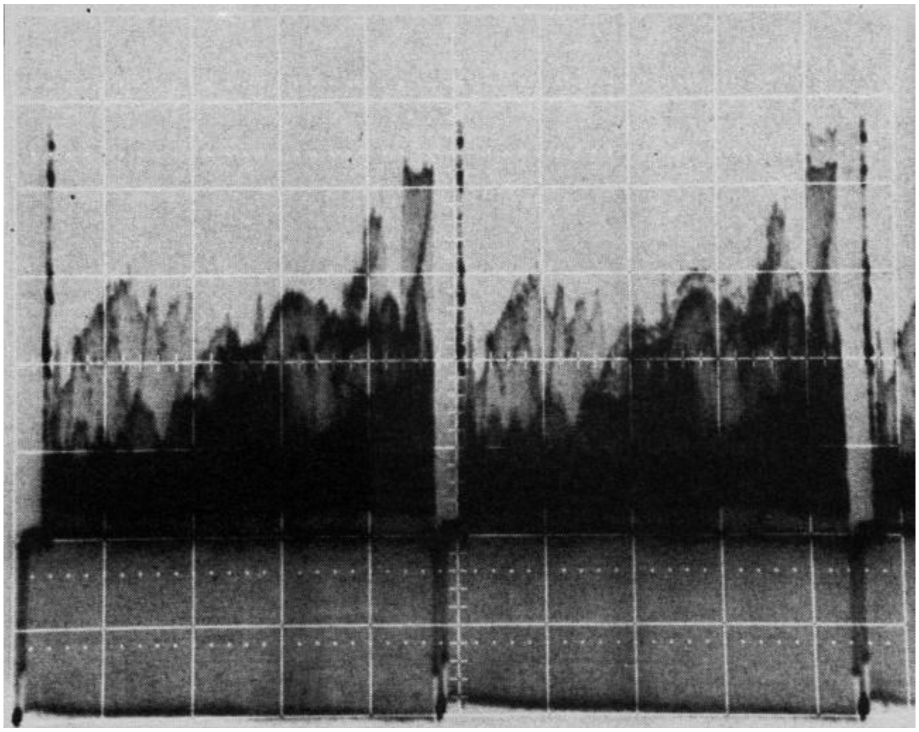


Figure 4-9: Video Waveform with 3% Field Rate Transient Distortion

Measurements of video waveform related low frequency disturbances require a video test waveform generator and modulator with negligible distortion. Power frequency related modulations are measured using only a low hum CW generator.

The above discussion has concerned analog signals only; an equivalent measurement for digital signals is the percentage of the peak-to-peak variation of the signal amplitude compared to the average signal amplitude for a signal which has been averaged over a time period which is very much greater than a symbol time but less than 1 ms.

It is important to recognize that TDMA digital signals with time periods much less than 16.7 ms are not impacted by Hum modulation.

Test Procedures: There are two basic test procedures for low frequency disturbances. They are discussed separately:

1. Manual procedures – using a spectrum analyzer with zero span or demodulator with an oscilloscope or waveform monitor. This will enable the user to visually observe the effect of the hum or other disturbance on the waveform; however, accurate measurement from the display may be difficult, especially for low values of hum. The setup is inherently more cumbersome than the second procedure.

2. Automated procedures – usually a signal level meter with the capabilities of hum measurement. This generates only a number with no display of waveform or picture distortion. But because signal level meters are generally more portable, this may be the preferred method for troubleshooting, especially in the field. Some spectrum analyzers have an automated measurement function for hum - the same discussion for automated signal level meters applies to them. The caveats of peak vs. peak-to-peak and non-power line related disturbances are discussed.

PROCEDURE 1 - Manual Measurements

Required Equipment

To measure the hum with a waveform monitor or spectrum analyzer, the following equipment would be used. Not all of these may be needed, depending on the component under test.

- A waveform monitor or an oscilloscope with 5 millivolts/div sensitivity from DC to at least 500 kHz bandwidth
- A spectrum analyzer with 300 kHz resolution bandwidth, a linear display mode, and zero span capability. It should be able to display a good video waveform.
- A signal generator which will tune to the video carrier frequency of each processor or demodulator channel to be tested and can be adjusted to a proper input level for the unit under test.
- A source of "clean" video signals
- A precision demodulator
- A tunable bandpass filter (preselector)
- A low pass (< 1 kHz) filter

Earlier test procedures included a signal level meter with a video output. The authors are not aware of any signal level meter on the market at this time with a video output.

PROCEDURE 1a - Manual Procedure for a Modulator or Processor Using a Spectrum Analyzer

Note: This procedure is only useable with modulated signals.

Required Equipment

- Spectrum analyzer with linear display and zero span capability.

Optional Equipment

- Bandpass filter (see Discussion)

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Connect equipment as shown in Figure 4-10.

3. Adjust the spectrum analyzer as follows:

   The following settings are <u>required</u>:

   - Center Frequency: Equal to the modulated or unmodulated test carrier
   - Frequency Span: Zero span
   - Amplitude Scale: Linear (not logarithmic)

   The following settings are recommended:

   - IF Resolution Bandwidth: > offset of 10 kHz.
   - Sweep Time: Between 5 and 50 ms/div.

4. Temporarily remove signal and verify the display line is at the bottom under a no-signal condition.
5. Reapply signal and adjust gain to place trace near top of display.
6. Refer to Figure 4-9 (a modulated carrier is shown as an example)

$$\text{Low Frequency Disturbance} = \frac{V_{P-P}(\text{hum})}{V_{RMS}(\text{RFsignal at sync peak})} * (100\%) \qquad (1)$$

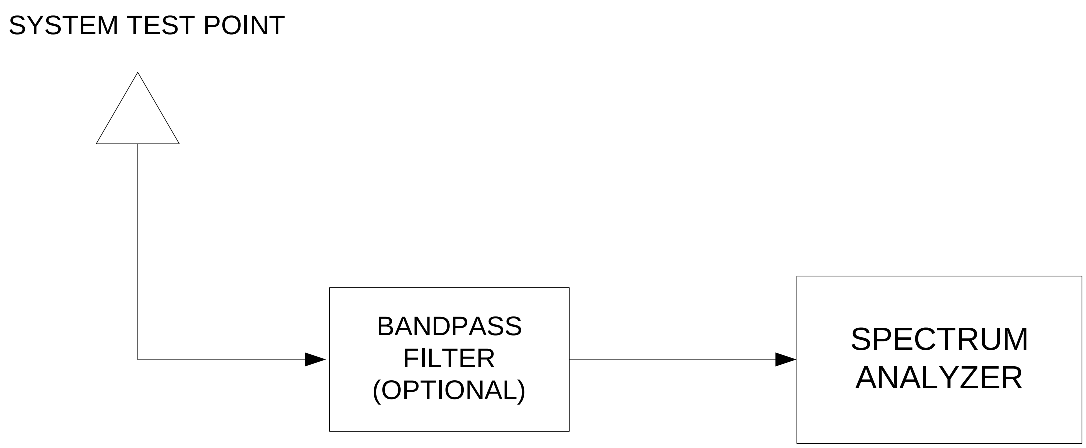


Figure 4-10: Spectrum Analyzer Setup

Note: If the signal being tested is modulated, it will be similar to Figure 4-9.

If the signal is unmodulated, it will appear as nearly a straight line (see Figure 4-11).

$$\text{Low Frequency Disturbance} = \frac{V_{P-P}(\text{hum})}{V_{P}(\text{signal})} * (100\%) \qquad (2)$$

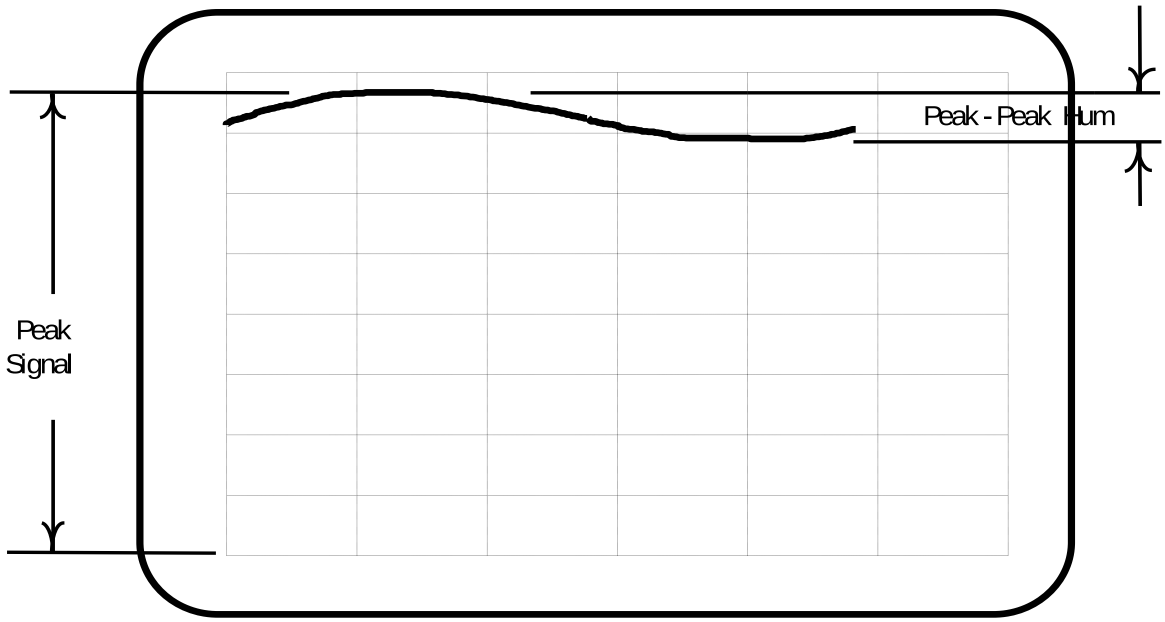


Figure 4-11: Display of Unmodulated or Low-Passed Signal with Hum.

PROCEDURE 1b - Manual Procedure for a Processor Using a Waveform Monitor

A. Signal to hum
   1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
   2. Use the setup of Figure 4-12.
   3. Tune the CW signal generator to the picture carrier of the channel to be tested. Use a variable attenuator, if necessary, to supply this signal to the processor RF input at the normal level. The processor may have internal circuitry to sense video – this should be disabled.
   4. With no input adjust the waveform monitor as follows:
      - Vertical Sensitivity: 0.1 volts/div
      - Sweep Time: 5 ms/div
      - Input: DC coupled
      - Trace position: On a graticule line near the bottom of the screen
   5. Tune the demodulator to the picture carrier output of the processor. As with the processor, any video sensing circuitry must be disabled.
   6. Connect the waveform monitor and adjust it for a five-division deflection on the display from the no signal position. This is the peak RF level reference.
   7. Switch the waveform monitor to AC coupled.
   8. Increase the vertical sensitivity to 5 millivolts/div.
   9. Note the peak-to-peak amplitude. One division corresponds to 1% percent distortion. See Figure 4-11.

B. Video related disturbances
   1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
   2. Connect the normal RF input signal to the processor.
   3. With the RF input to the demodulator disconnected, set the waveform monitor as follows:
      - Vertical Sensitivity: 0.1 volts/div
      - Sweep Time: 5 ms/div
      - Input: DC coupled
      - Trigger: Line Sync
      - Trace position: At the lower edge of the graticule.
   4. Reconnect the demodulator to the combined RF output and tune the video carrier.
   5. Adjust the waveform monitor sensitivity to place the video peaks (sync tips) at the upper edge of the scope graticule. Note the percent variation of the sync envelope.

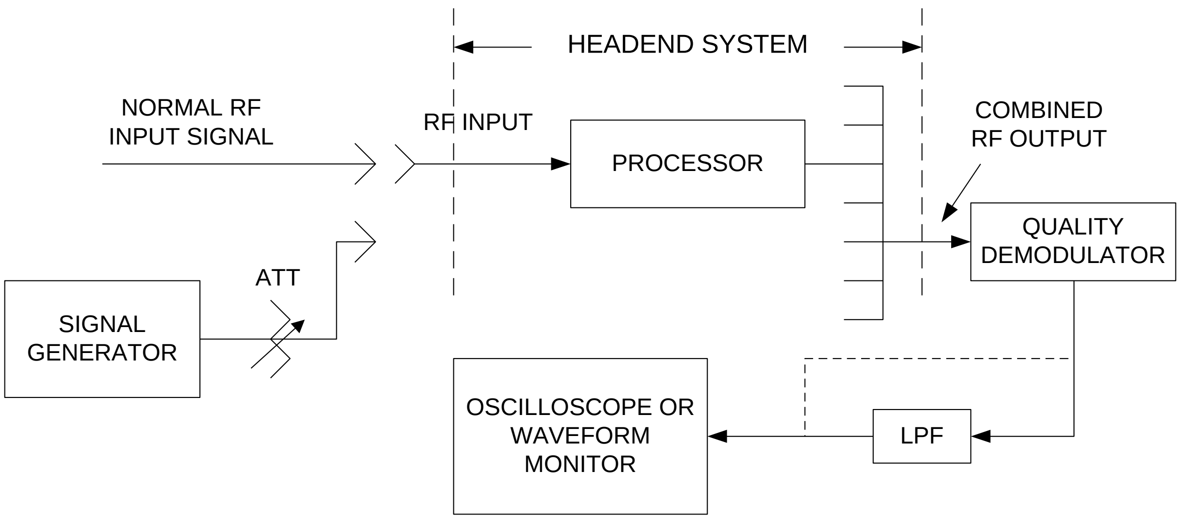


Figure 4-12: Processor Setup for Carrier-to-Hum and Low Frequency Video Disturbance Tests

PROCEDURE 1c - Manual Procedure for a Demodulator Using a Waveform Monitor

A. Video related disturbances
   1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
   2. Connect the equipment as in Figure 4-13 using the normal RF input signal to the demodulator. Provide a proper 75 Ω termination at the waveform monitor.
   3. Set the waveform monitor as follows:
      - Vertical Sensitivity: 0.1 volts/div
      - Sweep Time: 5 ms/div

- Trigger: Line Sync

4. Adjust the video output level of the demodulator such that the maximum peak-to-peak video (at 87.5% depth of modulation) amplitude is 7 divisions. With a standard 8 division graticule, this will allow the direct measurement of the disturbances as *percent of full screen.*

B. Signal to hum

1. Perform the setup as above, except adjust the demodulator for 0.87 volts peak-to-peak video with the scope at 0.2 volts. div. sensitivity.
2. Replace the normal RF input to the demodulation with the CW generator output tuned to the picture carrier. The level should be equal to that of the "normal" RF signal.
3. Insert the low pass filter between the demodulator under test and the waveform monitor. It is assumed that this filter provides proper impedance matching to the demodulator and has negligible loss at low frequencies.
4. Switch the scope to AC coupled and increase the vertical sensitivity to 0.01 volts/div. The peak-to-peak hum can now be read as above with one division equal to 1 percent or 40 dB carrier-to-hum.

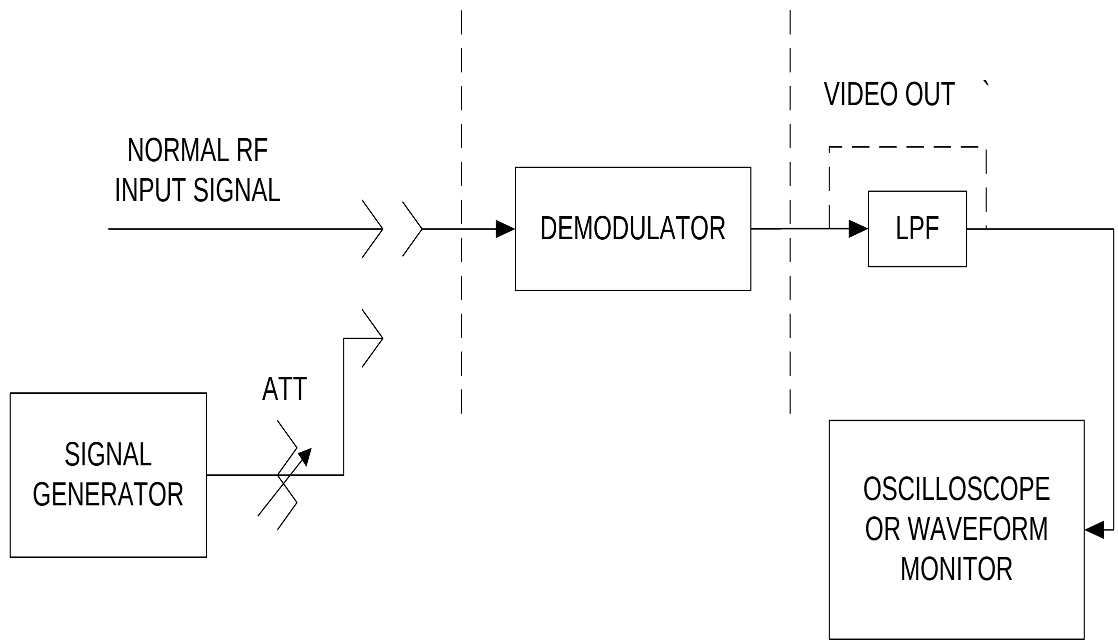


Figure 4-13: Demodulator Setup for Carrier-to-Hum and Low Frequency Video Disturbance Tests

PROCEDURE 1d - Manual Procedure for a Modulator Using a Waveform Monitor

A. Signal to hum

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Connect the equipment as in Figure 4-14 but do not connect the video input to the modulator.
3. Perform steps 3 through 9 in “PROCEDURE 1b - Manual Procedure for a Processor Using a Waveform Monitor”.

B. Video related disturbances
   1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

   2. Connect the normal video input to the modulator and set the peak depth of modulation to 87.5%.

   3. Perform steps 3 through 9 in “PROCEDURE 1b - Manual Procedure for a Processor Using a Waveform Monitor”.

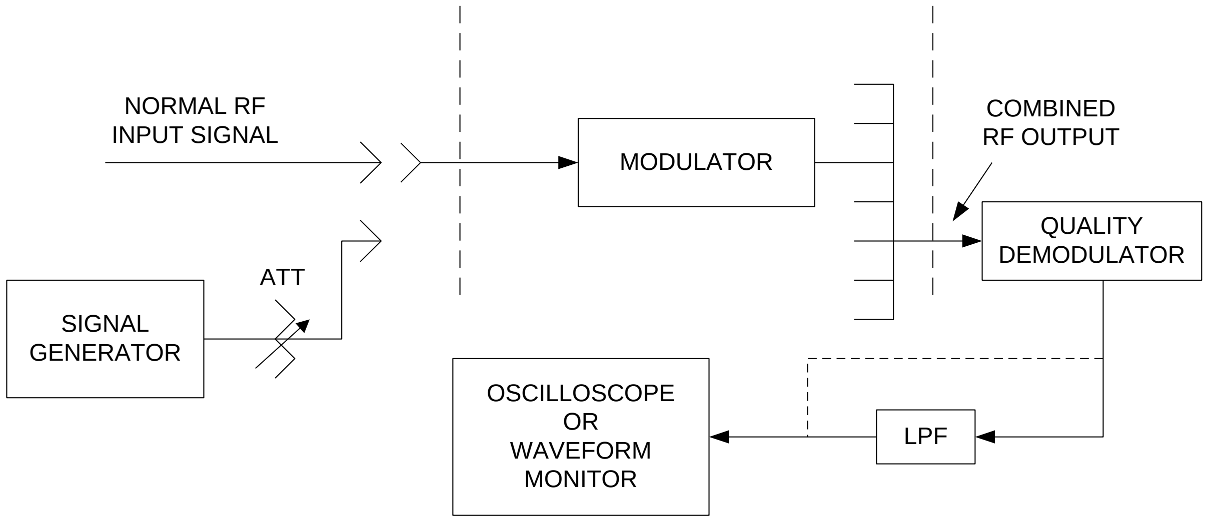


Figure 4-14: Modulator Setup for Carrier-to-Hum and Low Frequency Video Disturbance Tests

Procedure 2 - Automated Measurements

Note: Automated test equipment to measure modulation distortion is available. The "Discussion" section treats some of the possible variation that may occur among test instruments. If it is desired to measure manually, the following procedures are recommended.

Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

A bandpass filter ("preselector") tuned to the frequency under test may be necessary if the measurement equipment generates cross modulation products; i.e., vertical sync information on other channels is transferred to the unmodulated test channel and gives a falsely high hum reading. A possible symptom of this would be a high hum reading but no observed picture degradation.

Certain automated test equipment tests only for 60 and 120 Hz components of a low frequency signal. If there is reason to suspect non-power line related interference, one of the procedures described here should be used. It should be noted that most commercially available automated test equipment reads (peak interfering voltage/peak desired signal) and will display readings approximately one-half those measured by with a waveform monitor (Procedure 1) Persons correlating automated readings with those made with Procedure 1 or with other procedures not described here should be aware of this difference.

Notes, Hints and Precautions

In the demodulator measurement, where the video waveform is used as a level reference for the hum measurement, the accuracy is a direct function of the accuracy with which the modulation depth is known. If a Vertical Interval Test Signal (VITS) white level reference is present, use it as the 87.5% level. If the demodulator is chopper equipped, the actual modulation depth can be determined with it.

4.5 Low Frequency Disturbances and Hum – Digital Modulation

The **Definition** and **Discussion** for low frequency disturbances and hum on analog carriers found in the previous section is still applicable to digital signals, but some additional theoretical background is necessary.

Theoretical background

Digital signals can be defined as continuously variable “analog signals” that go through a given set of discrete values at specific instants of time. The set of discrete values are used to represent digital values known as symbols. The specific interval of time between the instant at which the “analog signal” should be sampled to recover the symbols is equal to the symbol rate.

In cable telecommunications, most digital signals are transmitted using quadrature amplitude modulation (QAM), meaning that the discrete values are given in 2 axes (sine-cosine, real-imaginary or amplitude-phase) in a regular checkerboard or constellation pattern. As an example, in 64-QAM there are 8 values in each axis for a total of 64 possible combinations, allowing the transmission of 64 different symbols. One can also think of the 64 symbols as being represented by 64 combinations of amplitude and phase in the transmitted RF signal. Each symbol represents a 6-bit word. Refer to Figure 4-15 a) below for an example of a 64-QAM constellation.

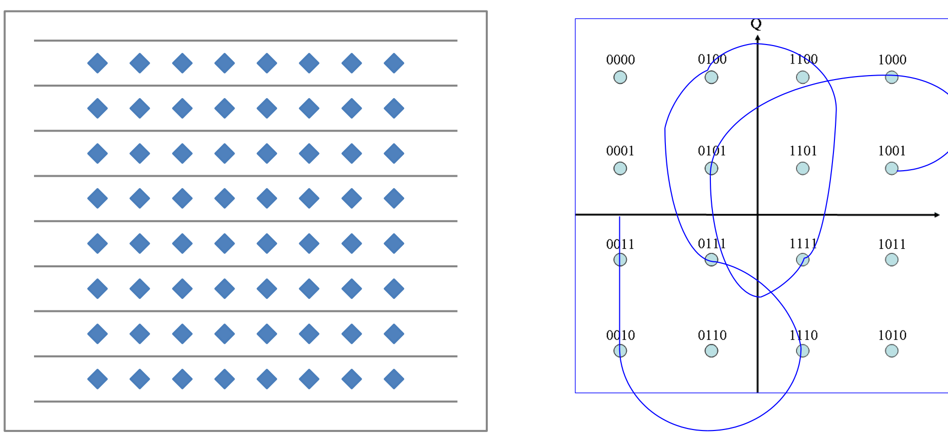


Figure 4-15: a) 64-QAM constellation and b) 16-QAM inter-symbols trajectory

In between the sampling instants, the “analog signal” glides from one symbol value to the next symbol value. Because of the analog nature of the physical signal, it cannot go instantly and directly from one value to the next – to do so would require infinite bandwidth. Instead, it takes a path that can overshoot the next symbol, and the amount of overshoot is a function of how bandwidth-constrained via filtering the symbols are at the modulator, as shown in Figure 4-15 b). On the one hand, as the signal alpha, and thus also the bandwidth is increased by design, the trajectory overshoot and peak-to-average power ratio decrease. On the other hand, as the bandwidth and signal alpha decrease, the trajectory overshoot and the peak-to-average power ratio increase. Keep in mind that symbols, adjacent in time, are spread all over the set of discrete values.

QAM signals have large peak-to-peak variations, symbol to symbol as well as inter-symbol trajectory, up to 30 dB. Because of the substantial instantaneous variations, averaging over a relatively long period of time (e.g., 20 ms) is necessary to extract hum from the signal. In order to distinguish QAM signal variation caused by hum modulation from the information-bearing modulation of the QAM signal, the QAM signal must be filtered using a much smaller bandwidth than the QAM signal modulation (several MHz) but larger than the hum modulation bandwidth (hundreds of Hz).

What is the effect of hum on digital signals? Low to moderate levels of hum can be tracked by the receiver and will have no effect. At higher levels of hum, the gain variation introduced by hum will cause errors in the detected signals. These errors will have no effect on the delivered information until the impairment can no longer be corrected by the FEC and post-FEC errors show up (cliff effect).

A QAM receiver has two mechanisms to track the amplitude variations of the incoming signal. The first is an AGC (Automatic Gain Control) that tracks the average signal power; the response of the AGC is in the tens of milliseconds. The second line of defense is the adaptive equalizer. The equalizer's main tap will also track amplitude variations at a fast rate (same order of magnitude as the symbol rate). The equalizer tends to optimize the MER by reducing the linear distortions (such as hum). With time varying impairments the tracking mechanism is always hunting for the best tradeoff, but it is never perfect. There is a residual error and this robs part of the margin before errors show up because of linear and nonlinear distortions.

A typical QAM receiver can track out a moderate amount of hum before the desired signal is significantly impaired. While the actual amount may vary somewhat among QAM receiver implementations, the design and operation of a QAM receiver is such that it generally can tolerate more hum than, say, the picture of an analog TV channel. That said, ANSI/SCTE 40 states that the worst-case hum in a downstream digital signal is not to exceed 3% peak-to-peak, the same maximum value in [§76.605(b)(10)] for analog TV channels.

Measurement philosophy

This constantly varying signal does not have a repetitive reference level like analog video where sync pulses have a given set point. The amplitude variation of this noise-like QAM signal is more difficult to measure. One searches for amplitude variation trends, using a receiver such as a spectrum analyzer in zero span mode. The narrower the measurement bandwidth the larger the apparent variations, so it is best to use a larger resolution bandwidth (RBW), such as 1 MHz or 3 MHz.

Because the variations being looked for are slow (on the order of 16 ms) captures need to be taken over a long period but also averaged over many capture periods. A 20 ms sweep time is recommended, as it is slightly longer than required to capture 60 Hz, and long enough for 50 Hz. To reduce noise, an averaging mode is suggested. A 100 average means that 100 captures, all 20 ms in length, are taken and averaged. During each sweep, data is collected at each "point" on the display. If there are 1000 points, the instrument samples the signal many times during each sweep (example: 1000 samples at 20 µs over the 20 ms period). When the sweeps are averaged, the sampling is repeated 100 times such that there are 1000 points, each point being a 100 long average. It is important that the sampling sequences be synchronized to the "hum" frequency (power line, or other frequency of interest).

After the data is collected, one more processing step is required because the residual "noise" in the 1000 points has a large spectrum of frequencies. A low pass filter (1000 Hz or 500 Hz cut-off) is applied to recover the 60, 120 and 180 Hz hum, video sync related frequency, or any other desired frequency.

Measurement procedure

1. Set the spectrum analyzer in zero span at the center of the QAM channel.
2. Set in average mode (100 or max), sample detector (usually automatically selected in average mode), adjust the number of data points to 1000 or 1001, set horizontal time to 20 ms (full span), and trigger to power line (or ext source if required).
3. Set the analyzer in log mode. Theoretically the linear mode is preferable, but with the large instantaneous variations (>30 dB), one can easily saturate at the top and bottom of the scale.
4. Set the analyzer to 1 MHz resolution bandwidth (RBW), and 3 kHz video bandwidth (VBW).
5. Start a measurement sequence, and when average has reached 100 the measurement should be halted automatically. If not, halt it manually.
6. Transfer the 1000 averaged data points into a spreadsheet.
7. Apply a 0.5 kHz low pass filter to the 1000 linear data points; this is a sliding average. The 20 µs sampling gives a "bandwidth" of 25 kHz, so a 50 to 1 sliding average gives a 1 kHz BW. According to Nyquist sampling law, 2 samples minimum are required to represent a signal. Two samples equal 2 * 20 µs = 40 µs, the inverse of 40 µs is 25 kHz. If we filter by a ratio of 50 to 1 the effective sampling is reduced to 50 * 20 µs = 1 ms, or a bandwidth of 0.5 kHz. See Figure 4-16.
8. Calculate the peak-to-peak filtered data in dB. Since the filtered data is still "noisy," it is best to calculate the peak-to-peak from the filtered data. For example, using a spreadsheet, the peak-to-peak variation can be extracted from the filtered data using the 0.995 and 0.005 "centile", that is, the fifth largest and fifth smallest data. See Figure 4-17.
9. The peak-to-peak variation for hum in dB is converted to linear scale and then converted to percent:

   $Hum_{\%}$ = 100 * [1 - 10^(Hum_{dB} / 20)].

   For example, for a variation of 0.5 dB, we get $Hum_{\%}$ = 100 *[1 - 10^(0.5 / 20)],

   $Hum_{\%}$ = 100 * [1 – (10^0.025)] = 100 * [1 – 1.059] = 100 * 0.059 = 5.9%

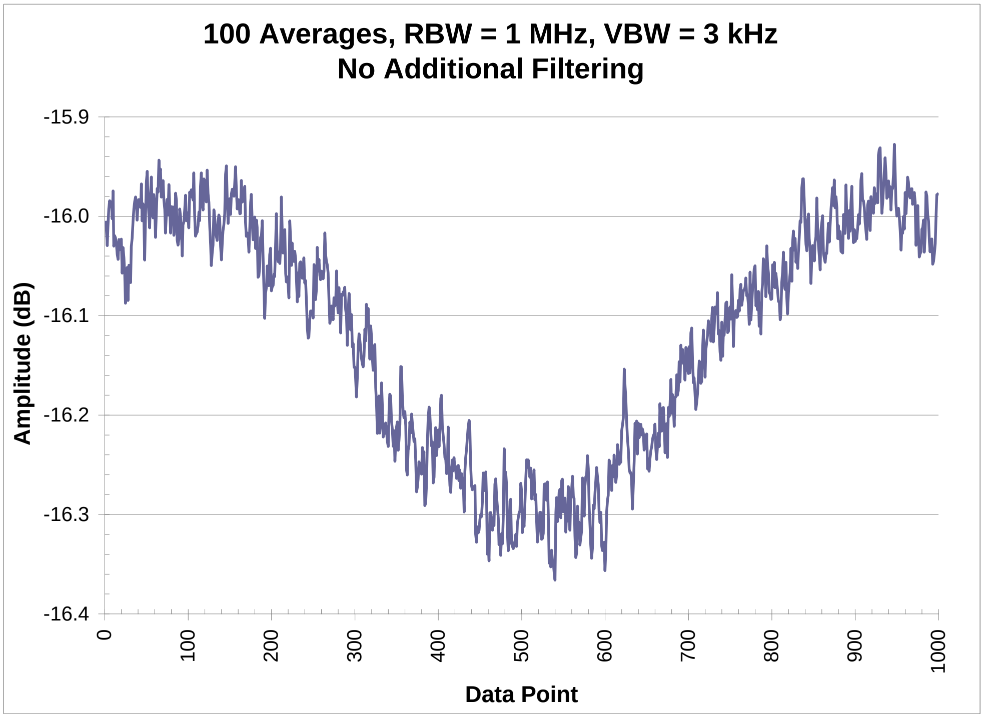


Figure 4-16: Data recovered from the Spectrum Analyzer

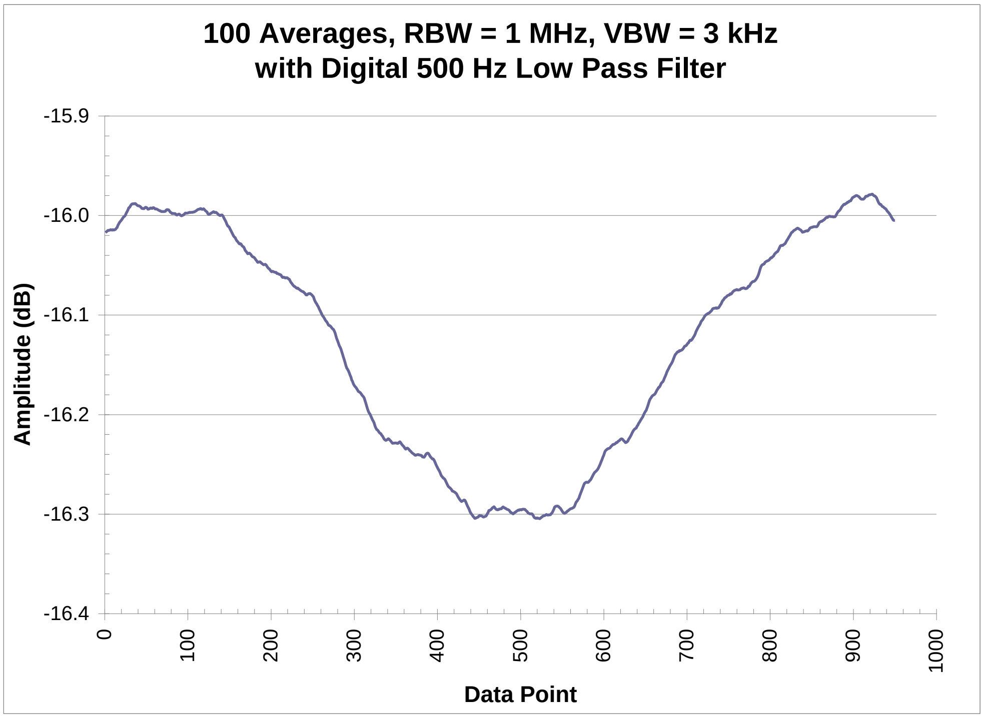


Figure 4-17: HUM from a QAM signal recovered after 500 Hz Filtering, 0.33 dB pk-pk

Note: The spectrum analyzer minimum requirements, specific to measuring the hum on a digital signal, are the following:

- High resolution in logarithmic mode (min 0.01 dB)
- At least 1000 sample points
- Capability to recover the data
- Capability to set RBW and VBW, while in zero span mode
- Trigger capability appropriate for the low frequency disturbance being measured
- At least 100 averages in sample mode

Note: Some test equipment supports automated measurement of hum modulation in digital signals. Refer to the test equipment manufacturer’s documentation for specific instructions on how to perform automated hum modulation measurement (when supported).

Chapter 5 Interfering and Spurious Signals (Plant Measurements)

5.1 Spurious Signals

Definition: The dictionary defines the word "spurious" as illegitimate, false, counterfeit. Similarly, a spurious electrical signal is a false signal whose presence in the output spectrum of an RF processing device is undesired.

Required Equipment

- Spectrum analyzer with variable IF resolution bandwidth and variable RF and IF attenuation. The user should be aware that the IF attenuation is quite often effectively controlled by the reference level adjustment and it may be coupled to the attenuator.
- 75 Ω to 50 Ω minimum loss pad (not needed for spectrum analyzer with 75 Ω input)

5.1.1 Spurious Signal Level of a Signal Processing Device

Test Procedure: The following procedure should be used when measuring the level of spurious signals at the output of a single headend device such as a signal processor or modulator.

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Connect to the output test point of the Device Under Test (DUT) or the test point of the combining network to the Spectrum analyzer through the minimum loss pad. The minimum loss pad is not required if the spectrum analyzer has a 75 Ω input. See Figure 5-2.

   Note: In some cases the levels at the test point will be too low to make an effective measurement. In this case, the output of the DUT rather than a test point can be connected to the spectrum analyzer. However, removing the output of the DUT from the combining network will take the channel out-of-service.

3. Operate the signal processor or modulator at the manufacturer's specified input and output levels.
4. Set the spectrum analyzer RF attenuator so that the video carrier is approximately at full scale, with the IF gain at the nominal value (usually the RF attenuator is coupled to the reference level) and adjust the frequency control to display both the video and audio carrier of the device under test, with frequency span set to 1 MHz/div.
5. Adjust the IF resolution bandwidth to 100 kHz or 30 kHz (narrow-band spurious signals are more easily discriminated against noise with a narrow IF bandwidth), and the amplitude scale to 10 dB/division-logarithmic. Slow sweep time as necessary to maintain a calibrated display.
6. At this point, RF attenuation must be lowered so that the noise floor of the spectrum analyzer is at least 70 dB below the level of the visual carrier displayed. At the same time RF attenuation must be high enough to prevent generation of spurious signals internally within the spectrum analyzer.
7. Adjust the reference level so that the modulation peak of the video carrier coincides with the 0 dB or top graticule line of the spectrum analyzer.
8. Carefully examine the channel of interest for spurious signals and record their levels in dB relative to the modulation peak of the video carrier. Add video filtering (narrow video bandwidth and/or video averaging) to enhance the display and reduce sweep time if necessary to obtain display stability. See Figure 5-1.

9. In addition, slowly examine the full operational bandwidth of your network for any spurious signals. Record their frequencies and relative levels.

Note: The spurious output signal level specification of a headend processing device shall be the highest spurious signal measured in a given frequency band and shall be expressed in dB below the visual carrier level of the device under test. The device shall be tested at least at two output levels and conditions so stated in the manufacturer's specification.

Hint: To determine whether or not a spurious signal is generated within the spectrum analyzer, note the level of the spurious signal and insert a 3 dB pad in the input to the analyzer. If the spurious is reduced by more than 3 dB it is being generated within the spectrum analyzer.

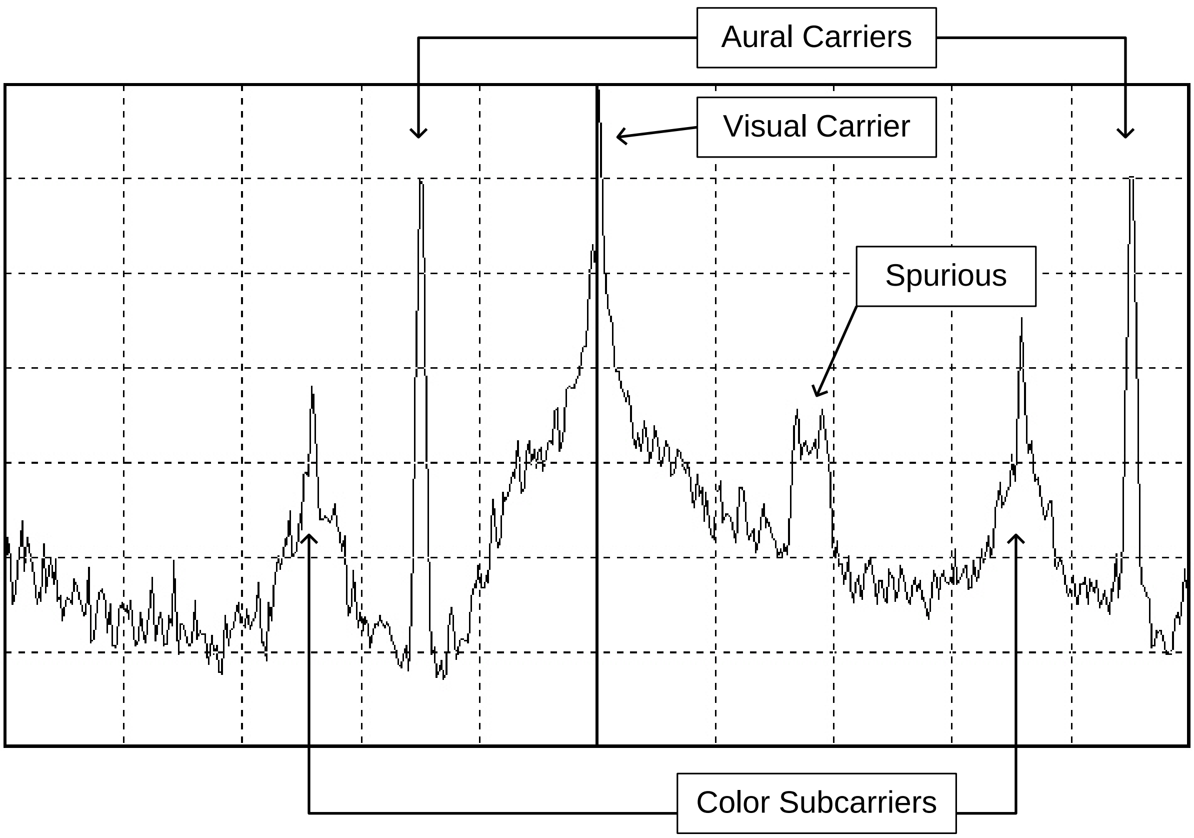


Figure 5-1: Example of Spurious Signals

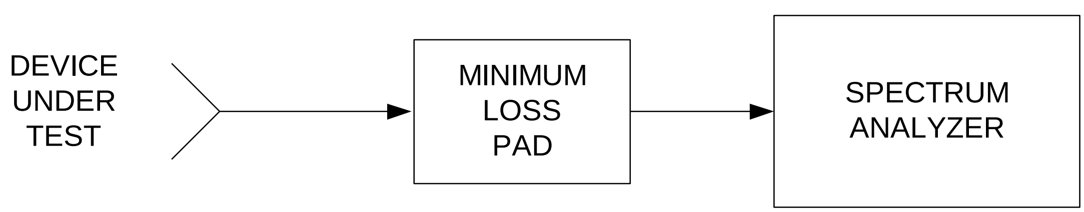


Figure 5-2: Measurement of Spurious Signals

5.1.2 Co-Channel Interference

Definition: Co-channel Interference is the reception of an undesired but legitimate video signal on the same nominal channel as the desired one. The undesired channel is normally offset by ±offset of 10 kHz or ±20 kHz from the desired one.

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Connect the headend output test points to the spectrum analyzer through the minimum loss pad. Adjust the frequency control of the spectrum analyzer to display the visual carrier of interest in the center of the graticule.

3. Adjust the spectrum analyzer RF attenuator so that the video carrier is approximately at full scale, with the IF gain at the nominal value (usually the RF attenuator is coupled to the reference level) so that the noise floor is at least 60 dB below the carrier of interest, and so that no spurious signals are generated within the analyzer. Adjust the reference level so that the modulation peak of the visual carrier coincides with the 0 dB or top graticule line of the analyzer.

4. Perform the following spectrum analyzer adjustments:
   - IF Resolution Bandwidth: = 1 kHz
   - Video Bandwidth: = 100 Hz, or until display stability is reached
     and/or
   - Number of Video Averages: ≥ 25, or until display stability is reached
   - Frequency Span: = 5 kHz or offset of 10 kHz/div.
   - Amplitude Scale: = 10 dB/div
   - Sweep Time: ≅ 50 ms/div or as necessary to obtain calibrated display

5. Carefully examine the regions between the visual carrier and the first two modulation sideband pairs for the presence of co-channel signals at ±offset of 10 kHz or ± 20 kHz from the visual carrier. Record the level of the co-channel signal in dB below the visual carrier and record its frequency relative to the visual carrier.

Note: The level of co-channel interference may vary significantly during changing seasons and weather conditions which temporarily produce abnormal signal propagation conditions.

5.1.3 Examining the Combined Headend Output Spectrum For Spurious Signals

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Connect the headend test point to the spectrum analyzer through the minimum loss pad and adjust the frequency control to display both the visual and aural carrier of interest with a frequency span of 1 MHz/div.

3. Adjust the spectrum analyzer RF attenuator so that the video carrier is approximately at full scale, with the IF gain at the nominal value (usually the RF attenuator is coupled to the reference level) so that the noise floor is at least 70 dB below the visual carrier level, and so that no spurious signals are generated within the analyzer. Adjust the reference level so that the modulation peak of the visual carrier of interest is on the 0 dB or top graticule line.

4. Perform the following spectrum analyzer adjustments:
   - IF Resolution Bandwidth: = 100 kHz or less
   - Video Bandwidth: = 100 Hz, or until display stability is reached
     and/or
   - Number of Video Averages: ≥ 25, or until display stability is reached
   - Amplitude Scale: = 10 dB/div
   - Sweep Time: = As slow as necessary to obtain calibrated display
5. Carefully scan the channel of interest for spurious signals. It may be necessary to replace the modulated visual carrier of the channel under test, with a CW carrier of the same frequency and amplitude in order to examine the region immediately adjacent to the video or chroma carriers.

   Note: Removing the modulated visual carrier takes the channel out-of-service.
6. Repeat this procedure for each channel carried on the cable system.

Performance Objective: It is also good engineering practice to remove any undesired signals present at the antenna input to the signal processing devices. The signals of local FM or television broadcast stations, mobile radio stations or citizen's radio stations may be present with enough strength to overload the input stages of the processing device.

If the spurious signal is apparently generated by the combining process, be certain that your headend combining network has adequate port to port isolation. Utilize directional couplers with high directivity or a packaged headend combining network with high port to port isolation and high return loss specifications.

For interconnection of processing devices with antenna downleads or the combining network; use a cable which has a high degree of shielding effectiveness, together with a connector for best results.

Loose connectors or poorly shielded cable can cause cross talk from high level outputs to low level inputs of signal processors. Deteriorated connectors within the combining network can act as non linear "rectifiers" and generate spurious signals.

5.1.4 Examining the Signal Processor Antenna Inputs for Spurious Signals

Employ the same technique used to measure spurious emissions from a single processing device to examine the antenna downleads for the presence of spurious signals. Make due allowance for the customarily lower levels of the antenna downlead signals.

5.1.5 Solving Spurious Signal Problems

If spurious signals are being generated within a signal processing device, first check to be certain about the device by examining the antenna downleads for the presence of spurious signals. Make due allowance for the customarily lower levels of the antenna downlead signals.

Next, check overall swept frequency alignment and carrier frequency accuracy. Be certain to check for deteriorated RF connectors at the module/housing interface.

If the processing device is operating at the correct levels and the swept frequency alignment is proper, the spurious signals may be generated in one of the active stages of the processing device.

Examine the output spectrum of the individual modules in the processing device to isolate the faulty stage. Frequent culprits in a heterodyne processor or modulator are the RF output transistor and the upconverter mixer or oscillator.

Many manufacturers recommend the installation of a passband filter at the output of each modulator or heterodyne signal processor to suppress out-of-band spurious signals which might fall on other channels. Consult the maintenance manual for your headend processing equipment to see if passband filters are recommended.

Spurious signals may be the result of the nonlinear combination of two or more desired carriers within a headend processing device or of harmonic generation within the device. Spurious signals generated in this fashion are referred to as intermodulation products.

Spurious signals may also originate outside the head end. Because of excessive ambient signal level, insufficient antenna directivity or inadequate filtering of external signals, undesired interference may enter the head end, pass through the various processing devices and be present in the output spectrum of the head end. If a spurious television signal is produced in this manner and it shares the same nominal channel assignment with a desired television carrier the result is called co-channel interference.

5.1.6 Spurious Signal Level of a Digital Processing Device

Required Equipment

- Spectrum analyzer with 75 Ω input

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Disconnect the output of the device under test from other headend equipment and connect it to the spectrum analyzer.
3. Operate the device at the manufacturer’s specified input and output levels
4. Set the spectrum analyzer as follows:
   - Center Frequency: Center of channel under test
   - Frequency Span: Wide enough to see entire signal under test plus adjacent channels. (See Note below)
   - Input Impedance: 75 Ω
   - IF Resolution Bandwidth: 300 kHz
   - Video Bandwidth: 30 kHz or lower, or until display stability is reached (30 Hz is preferred)
   - Reference Level: To position top of signal trace approximately 10 dB below the top of the analyzer graticule
   - Amplitude units: dBmV
   - Number of Video Averages: Until display stability is reached

   Note: The recommended span settings are as follows:

- 50 MHz to100 MHz for L-band signals
- 20 MHz for QAM signals transmitted in a 6 MHz channel
- 6 MHz for out-of-band signals

5. Set spectrum analyzer RF attenuation to a value such that the noise floor of the analyzer is at least 50 dB below the top of the digital signal.
6. Measure the power of the digital signal using the procedures described in Section 2.2.1: "Continuous Signals". Record this information.
7. Slowly examine the entire cable spectrum for spurious output signals. Record their levels and calculate the difference in dB relative to the level measured in step 6.

   Note: Spurious signals are most likely to be found in the upper and lower adjacent channels.

Note: The spurious output signal level of a digital device shall be the highest spurious signal measured in a given frequency band and shall be expressed in dB relative to the digital signal power at the device's output.

Hint: To determine whether or not a spurious signal is generated within the spectrum analyzer, note the level of the spurious signal and insert a 3 dB pad at the input to the analyzer. If the spurious signal is reduced by more than 3 dB, it is being generated within the analyzer.

5.2 Discrete Frequency Interference (DFI)

Definition: Discrete Frequency Interference is a disturbance to the signal composed of discrete signals which may be generated by local oscillators in processing equipment, status monitoring carriers, AGC pilots, encoder jamming carriers, or ingress to the system from over-the-air signals.

FCC §76.605(b)(8): ***The ratio of visual signal level to rms amplitude of any coherent disturbances such as intermodulation products, second and third order distortions or discrete-frequency interfering signals not operating on proper offset assignments shall be as follows:***

> ***(i) The ratio of visual signal level to coherent disturbances shall not be less than 51 decibels for noncoherent channel cable television systems, when measured with modulated carriers and time averaged; and***
>
> ***(ii) The ratio of visual signal level to coherent disturbances which are frequency-coincident with the visual carrier shall not be less than 47 decibels for coherent channel cable television systems, when measured with modulated carriers and time averaged.***

Discussion: This requirement compares the rms level of the in-band interfering signals with the rms voltage produced by the visual signal during the transmission of synchronizing pulses. This FCC requirement includes not only discrete frequency interfering signals, but also intermodulation products and second and third order distortions. These other distortion products are discussed in Section 4.1.

Measure With or Without a Converter?

The FCC specification requires that coherent disturbance measurements be made at the subscriber terminal. This implies that if set-top boxes are used in your system, then these measurements must be made at the output of the set-top. The likelihood that a set-top converter would cause DFI is very remote, and if this were to occur it would be a faulty set-top. Therefore, this procedure has been written without using the set-top converter. It may still be the operator's choice to use a (non-baseband) set-top converter as a pre-selection device to increase the dynamic range of the measurement or enable DFI measurements on scrambled channels.

Because of the relative nature of the measurement, the absolute accuracy of the test equipment is not as important as the dynamic range and relative accuracy or log scale fidelity.

Measuring DFI generally requires removing the video modulation to see interfering signals in the channel. Some spectrum analyzers provide video gating which allows sampling the video during quiet lines in the vertical interval. This eliminates the need to remove the video modulation during the test and will provide the same accuracy. Video gating is typically instrument specific and therefore not described in this procedure. Another approach used to identify DFI is a negative peak detector or minimum peak hold. These approaches will suppress the video modulation, but will also suppress the DFI, unless it is a CW carrier.

Required Equipment

- Spectrum analyzer
    - IF Resolution Bandwidth 300 kHz and 30 kHz
- Video Bandwidth 300 kHz and 10 Hz

Optional Equipment

- Broadband Preamplifier
    - Gain > 15 dB
    - Noise Figure < 10 dB
- Tunable bandpass filter or fixed bandpass filter for each channel to be measured with 3 dB bandwidth < 15 MHz
- Set-top Converter

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Connect the test equipment as indicated in Figure 5-3. The bandpass filter or set-top converter may be necessary to limit the number of carriers (net power) at the input of the analyzer and improve the dynamic range. The preamp should only be used when the signal level available is too low to achieve the required dynamic range.

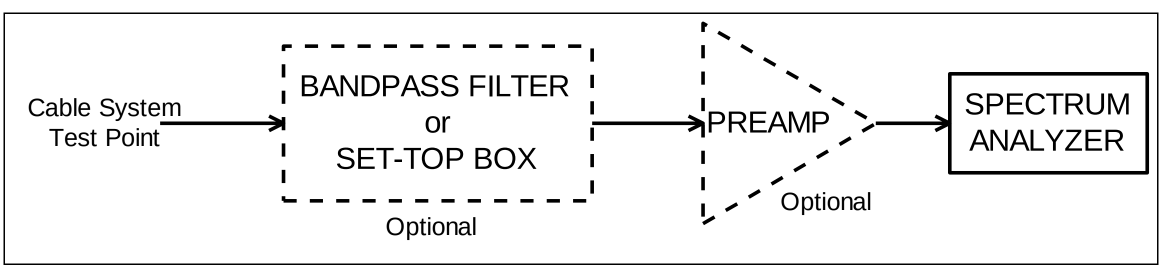


Figure 5-3: Discrete Frequency Interference - Test Equipment Setup

3. Set the analyzer up as follows:
    - IF Resolution Bandwidth: 200 kHz or greater

- Video Bandwidth: Maximum
- Frequency Span: 5 MHz
- Amplitude Scale: 10 dB/div
- Sweep Time: Automatic for calibrated measurement

4. Adjust the spectrum analyzer center frequency to position the visual carrier of the channel under test to the left of center screen.
5. Adjust the analyzer's full-scale reference to position the sync peak level in the upper division of the display.
6. Using the analyzer's markers, record the sync peak level of the carrier. If the analyzer does not have markers, this may be done by adjusting the full-scale reference to place the sync level exactly on the upper graticule line and recording this full scale reference value.
7. Turn the channel under test video modulation off.
8. Readjust the analyzer for the following settings:
   - IF Resolution Bandwidth: 30 kHz
   - Video Bandwidth: Minimum (less than 300 Hz)
   - Frequency Span: Unchanged
   - Sweep Time: Automatic for calibrated measurement
   - Number of Video Averages: Until display stability is reached
9. The display now shows both DFI and CSO. The marker may be used to measure the level of the interfering signal, or if there are no markers the measurement must be made from the graticules. The distortion magnitude is the difference (in dB) between the average peak level of the interfering signal and the carrier.

Note potential errors

A. Analyzer Noise Floor

If the level of the distortion being measured is not at least 10 dB above the analyzer noise floor, additional attenuation should be removed or a correction factor will need to be used. This can be checked by removing the input cable and looking for at least a 10 dB drop in the level at the frequency of the distortion. If the change is less than 10 dB, the measured level of the distortion should be decreased according to Figure 3-1. As an example, if the DFI measured -55 dB and it dropped 7 dB when the input is removed, the corrected beat measurement would be:

$$-55 - 1.0 = -56\text{dB}$$

B. Preselector Tuning

If the preselector being used is too narrow, the DFI may be outside the passband of the filter. To prevent this, the preselector should be adjusted to peak the distortion. If a fixed filter is being used, the passband characteristic must be checked beforehand.

5.3 Discrete Interfering Signal Probability (DISP)

Definition: Discrete Interfering Signal Probability (DISP) is a measure of the percentage of the time that undesired discrete signals which appear at the upstream headend output port exceed certain thresholds which are defined relative to the total signal handling capability of the system. Using the data obtained, it is possible to evaluate the overall interfering signal performance of a system independently of intended signal carriage and also to predict levels of destructive interference to specific signal types as a function of frequency, operating levels and time.

Discrete signals are defined as those signals whose energy is primarily concentrated within a relatively narrow bandwidth (approximately 100 kHz) and which are present for time periods of at least several milliseconds (see discussion at end of procedure). Sources of interfering signals include ingress from over-air signals, antenna-conducted egress from subscribers' directly-attached receivers, spurious and harmonic signals from operator-attached terminal equipment and common-path intermodulation distortion.

The procedure is specifically intended to exclude ingress effects from electrical transients (impulse noise), which affect much wider frequency ranges for shorter time periods.

PROCEDURE

- **Determine System Parameters**

Reference Level

Since measurements are to be made relative to the total power handling limit of the network and will, presumably, be made at a headend test point, the first step is to determine the reference level at that test point. Referring to Section 16.4: "Return Plant Setup and Operational Practices", the Reference Level for this procedure is the level appearing at the headend upstream test point when an upstream amplifier Reference Point is driven with a signal whose level is equal to the total power handling capability of the system (P_1 in section 16.4). That level will be equal to (P_8 - A) if the same test point is used for this measurement as was used for system alignment.

Threshold Levels

The procedure will determine how often ingressing signals exceed defined thresholds relative to the Reference Level. In order to characterize the general performance of the system, a sufficient number of thresholds should be defined to allow reasonable extrapolation to intermediate levels. If specific signal types and operating levels have been determined, it may be useful to set one or more thresholds that are meaningful to those signals.

For example, if the headend test point level resulting from driving the system at its maximum recommended input level is +10 dBmV and the upstream level of a particular QPSK signal is 13 dB below this input level, then the level of the QPSK signal at the headend test point will be -3 dBmV. Since QPSK signals have a discrete signal interference tolerance of approximately -15 dBc, then a logical threshold for testing would be -18 dBmV which is -28 dB relative to the Reference Level.

Limited testing has shown that the probability of discrete interfering signals exceeding a given threshold may vary about 5:1 for a 5 dB change in threshold. Thus, it is suggested that threshold level increments not exceed that value.

- **Equipment Setup**

Measurements are taken using a spectrum analyzer connected to the headend upstream test point determined above with the measurement data captured and downloaded to a connected personal computer (PC) for analysis.

The spectrum analyzer must, at a minimum, be capable of being programmed to take repeated level measurements over a defined frequency range and must be capable of downloading those measurements to the attached PC. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

- **Data Gathering and First Level Analysis**

Method A: External Data Reduction

Program the analyzer to make repeated measurements of peak power level over the entire upstream band. The number of measurement points must be such that the frequency increment between points does not exceed 100 kHz. Set the resolution bandwidth to 100 kHz and the video bandwidth to 300 Hz. If the sweep rate does not automatically set itself to compensate for the bandwidth, manually set it so that the dwell time at each frequency is at least 2 ms.

Set the analyzer to take data over a window of no more than one hour (a faster sampling rate will give greater time resolution to the results, but will multiply the total database size by the same ratio). The data from each sweep during the measurement window must be downloaded to the PC for analysis. If the data from all sweeps during each time window can be stored within the analyzer, then downloaded at the end of the window, the results will be better than if the data from each sweep must be downloaded before another sweep can begin; however, the memory requirement may exceed that available in the analyzer. For instance, with 401 measurement points and 2 ms time per point, the sweep time will be about 3 seconds and therefore 1200 sweeps will be completed in one hour. If the data is stored within the analyzer and each data point requires 2 bytes of storage, then about 1 MB of storage will be required.

The PC should have two 3-dimensional matrices defined to hold the data. One axis of each matrix should be measurement frequencies, the second should correspond to the defined threshold levels, while the third corresponds to the successive time windows. The first matrix will contain “raw data” counters, while the other is the “DISP” matrix. In addition to the matrices, a sweep counter variable needs to be defined for each time window to hold the number of analyzer sweeps during that window.

Initialize the counter variables and the raw data matrix values (all integers) to zero. Non-integer values in each cell in the DISP matrix should be equal to the value in the corresponding cell in the raw data matrix divided by the value in the sweep counter variable for the corresponding time window.

The downloaded data from each sweep will typically appear as a serial listing of peak levels at successive frequencies. The peak level information for <u>each frequency</u> should be used to modify the values in one column of the raw data matrix as follows:

1. Compare the received level at the first frequency with the lowest defined threshold. If it is greater, then increment by one the value in the raw data matrix corresponding to that threshold, frequency and time window.
2. Move to the adjacent cell in the raw data matrix that corresponds to the next higher threshold (at the same frequency and time window) and repeat step 1.

When the processing of data from the lowest frequency is complete, then increment the frequency and repeat.

When all the data from one sweep is processed, the sweep counter variable for that time window should be incremented by one.

When the time window has expired, repeat for additional measurement windows. Data should be taken over at least 24 hours and additional sets of 24 hour data should be taken to determine the consistency of the results.

The result of this process will be a raw data matrix with cells representing specific frequencies, threshold levels and time windows. The integer values in those cells will be the number of sweeps during the time measurement window for which the measured levels exceeded the defined threshold.

The DISP matrix will also have dimensions representing frequencies, time windows and threshold levels, as shown in Figure 5-4, but each value in U1, will represent the probability that a signal is present at frequency F1 and in time window T1 whose level exceeds threshold level L1.

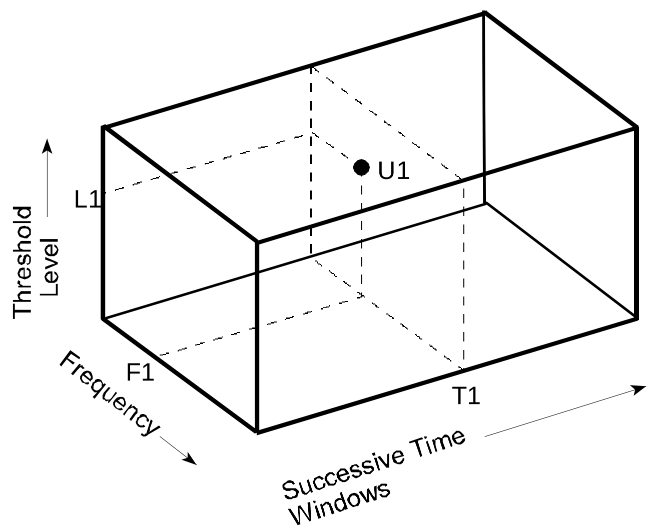


Figure 5-4: Discrete Interfering Signal Probability (DISP) Data Base

Method B: Internal Data Reduction

Some spectrum analyzers include the capability to create the two-dimensional raw data matrices corresponding to each time window internally using a "PDF" function, whereby the various threshold levels are downloaded to the analyzer and "virtual traces" are defined, each of which contains counters, as a function of frequency, of the number of sweeps for which the level exceeded one threshold level. A master sweep counter is also included.

Using the internal PDF function, the data for all the sweeps in a time window are summarized into one set of values for each defined threshold and can be directly downloaded into probability matrices in the PC at the end of each time measurement window.

- **Data Post Analysis**

The data from the DISP matrices can be analyzed in various ways to quantify the discrete interfering signal performance of the system. In the case of cable systems which are already transporting upstream signals, a preliminary step to data analysis is to exclude measurements of already-occupied frequencies. The following are illustrative of useful interpretations of the remaining unavailability data.

Average Discrete Interfering Signal Probability (ADISP) vs. Threshold Level

First, for each threshold level, average the DISP numbers across frequency and time (see Figure 5-5). This value is the Average Discrete Interfering Signal Probability (ADISP) for a given threshold value. Plot this against threshold level (relative to the Reference Level). The result is a service-independent plot of ADISP vs. relative threshold level. This is the most fundamental measure of system performance and the primary result of this procedure.

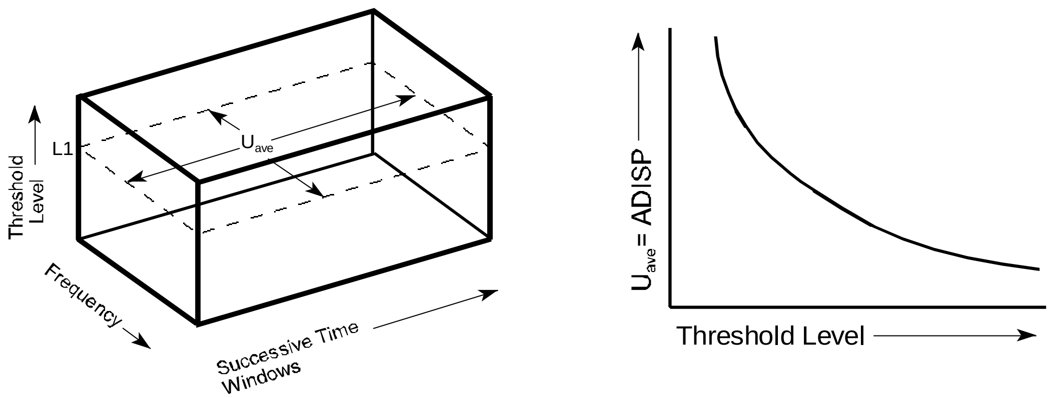


Figure 5-5: Average Discrete Interfering Signal Probability vs. Threshold Level

DISP vs. Frequency and Threshold Level

First, at a given threshold level, plot the time average of each frequency's DISP vs. frequency (see Figure 5-6). Then, using a multi-line chart, add lines for all the threshold values of interest. This chart allows prediction of performance vs. frequency and operating level.

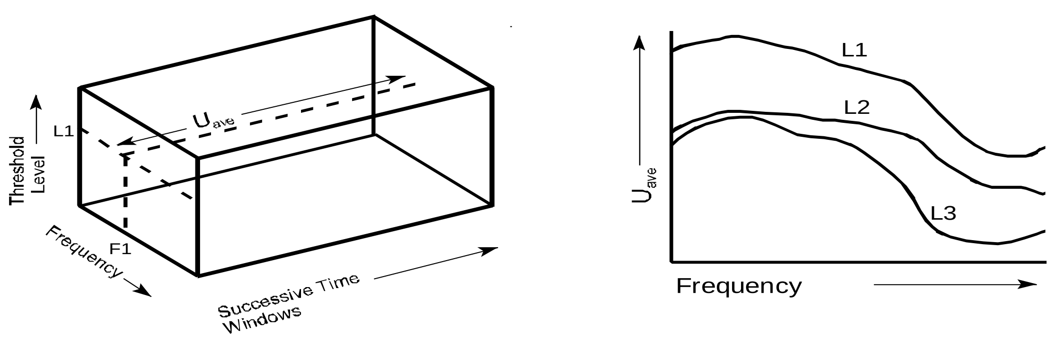


Figure 5-6: DISP vs. Frequency and Threshold Level

Specific Frequency DISP vs. Time and Threshold Level

Plot the DISP of a specific frequency at a threshold of interest vs. time (see Figure 5-7). Repeat, if desired, for different thresholds. This allows an analysis of the time variance of the DISP at a specific frequency and how it varies with operating level.

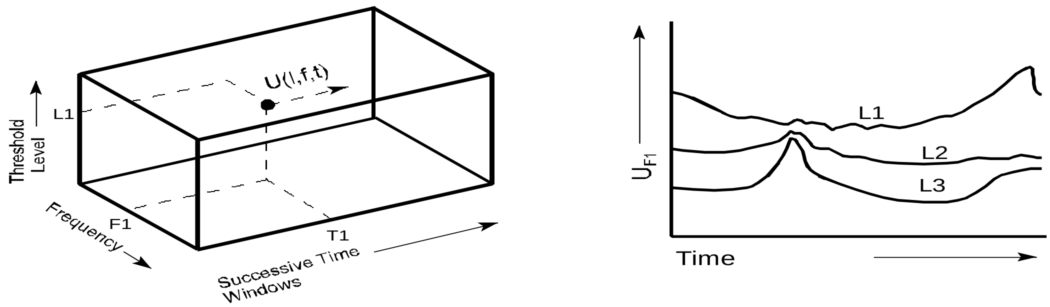


Figure 5-7: Specific Frequency DISP vs. Time and Threshold Level

Specific Channel Unavailability (SCU)

Where specific services have channel bandwidths that exceed the frequency resolution of the raw data and operate at specified signal levels, channel unavailability can be determined similarly to Specific Frequency DISP vs. time, except that instead of plotting the DISP of a particular frequency and defining the thresholds relative to the total power handling capability of the network, it is necessary to define the levels relative to the onset of destructive interference for the service in question and to plot the highest DISP of all the frequencies within F1±BW/2, where F1 is the channel center frequency and BW is the susceptibility bandwidth of the service (see Figure 5-8). Note that this is an approximate process, as two simultaneously occurring interfering signals within the susceptibility bandwidth could individually be less than the threshold, but their instantaneous voltage sum could cause interference.

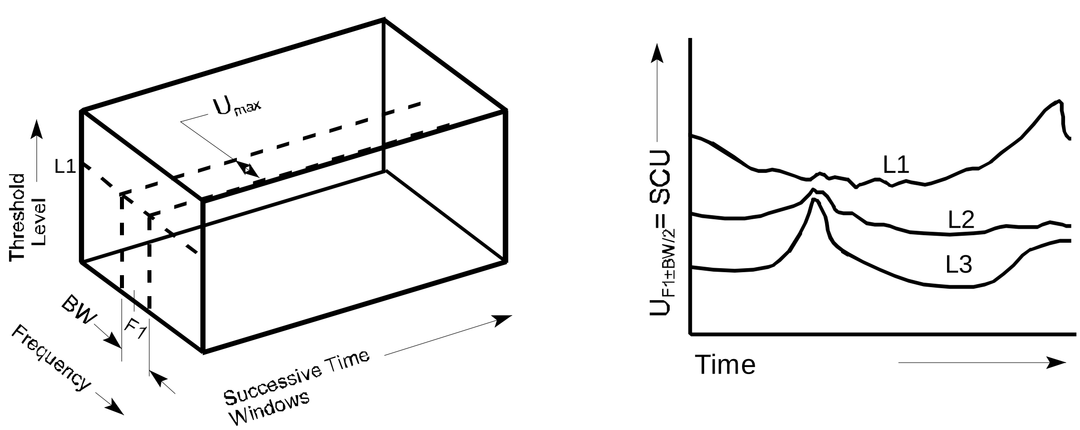


Figure 5-8: Specific Channel Unavailability (SCU) vs. Operating Level

DISCUSSION

The measurement of ingressing signals of all types (plus common-path intermodulation products) is, necessarily, a statistical process. Unlike downstream measurements, the signals measured are generally not under the control of the operator and occur semi-randomly in level, frequency, time and location. This procedure is a method whereby a system can be evaluated with respect to the average presence of upstream potentially-interfering carriers. Since absolute levels are difficult to analyze and summarize in a meaningful way and are of secondary interest relative to their effects on communications, the procedure simply determines the percentage of time they exceed one of several defined thresholds. Since the thresholds are defined only relative to the total signal handling capability of the network, the DISP is totally independent of specific system operating parameters or operating signal levels.

Where systems being evaluated are partially loaded with existing intentional carriers, the frequencies occupied by those carriers should be excluded from the analysis. It is understood that analysis of systems with a significant percentage of frequencies occupied will be less accurate than systems with no upstream traffic.

The measurements taken using this procedure will be no better than the raw data. For instance, if analyzers are used which must download the data from each sweep before the next sweep is started, then the majority of time may be taken up with data transfer and many interfering signals may not be detected. Similarly, if only a single day's data is taken and interfering signal levels vary widely from day to day, the data may be statistically meaningless. The statistical accuracy of the measurement process in either case can be improved by taking data over longer periods of time and averaging the results in each time/frequency/threshold cell of the probability matrix.

The choice of analyzer bandwidths is a compromise. 100 kHz was chosen as it was wide enough to contain all common-path intermodulation components in any one group and also wide enough to contain all the sideband energy from expected ingressing signals, yet narrow enough to be useful in identifying usable frequency ranges for upstream signals without being so narrow that the quantity of data to be gathered becomes unwieldy. A wider resolution bandwidth would also reduce the sensitivity because the effect of integrated wideband noise.

Choose a 300 Hz video bandwidth in order to exclude electrical transients: the rise time at that bandwidth is about 1.1 ms, while most transients are no more than a few µs long. Making the bandwidth even smaller would have unnecessarily lengthened the required dwell time at each frequency, while making it longer would admit some transient energy.

Chapter 6 Frequency Response

6.1 Analog Television In-Band Response [FCC §76.605(b)(6)]

Definition: In-band frequency response is the variation in the amplitude response of a single channel. Depending on the source of the information, it might be measured as a function of modulating frequency or, as with downstream frequency response, as a function of RF frequency within the channel.

FCC §76.605(b)(6): ***The amplitude characteristic shall be within a range of ± 2 decibels from 0.75 MHz to 5.0 MHz above the lower boundary frequency of the cable television channel, referenced to the average of the highest and lowest amplitudes within these frequency boundaries. The amplitude characteristic shall be measured at the subscriber terminal.***

Discussion: The FCC only requires testing of analog NTSC television signals and requires testing over the frequency range extending from 0.5 MHz below the visual carrier to 3.75 MHz above it, but optionally allows testing as a function of the visual carrier modulating frequency (from 0.5 MHz to 3.58 MHz). While it may be measured through the entire system from modulator or processor input to each field test point, many of the procedures may involve interruption of programming. This both causes subscriber complaints and increases the labor cost involved as personnel are required at both headend and field test point in that case. It is equally valid to measure the response to the point at which distribution sweep signals are inserted (generally the headend test point) and add to that the distribution system in-band response as determined above plus the response variation of a typical set-top converter used in the system. Only if this worst-case procedure results in an out-of-specification result will it be necessary to actually measure the total system response on each channel.

Several specific procedures follow, roughly in order of increasing complexity. If available test signals are sufficient to allow testing of the required number of channels using simple procedures and those channels are representative of the headend as a whole, then more complex procedures can be avoided (though operators are obligated to meet the specifications on all channels, whether tested or not).

PROCEDURE 1 - Overall Response from Programmer Input to Subscriber Test Point Using Programmer VITS

This procedure measures the total of programmer response variations, transmission variations, cable system variations, set-top converter response, and test equipment variations. It is simple to perform and is a worst-case test.

Required Equipment

- A precision demodulator with verified good internal frequency response.
- A set-top converter for measuring total response at the subscriber's termination point.
- A video waveform monitor with line and field select capability.
- A subscriber drop cable (30 meters), if testing is to be done at field test locations.
- A variable attenuator, if required, to set input level to the demodulator.

Measurement Steps

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Connect the equipment as shown in Figure 6-1.

3. Tune to the first channel to be tested. Using the waveform monitor line and field select controls, examine the first 20 lines of both fields to see if a multiburst waveform is available as part of the programmer-supplied vertical interval test signals (VITS). If multiburst signals are available and usable (not too noisy or so rolled off as to be meaningless), record the peak-to-peak amplitude (in IRE units or volts) of each of the bursts up to and including 3.58 MHz. The variation in levels will include the effects of the original transmitter, antennas, and everything else in the path between the original VITS generator and the waveform monitor.

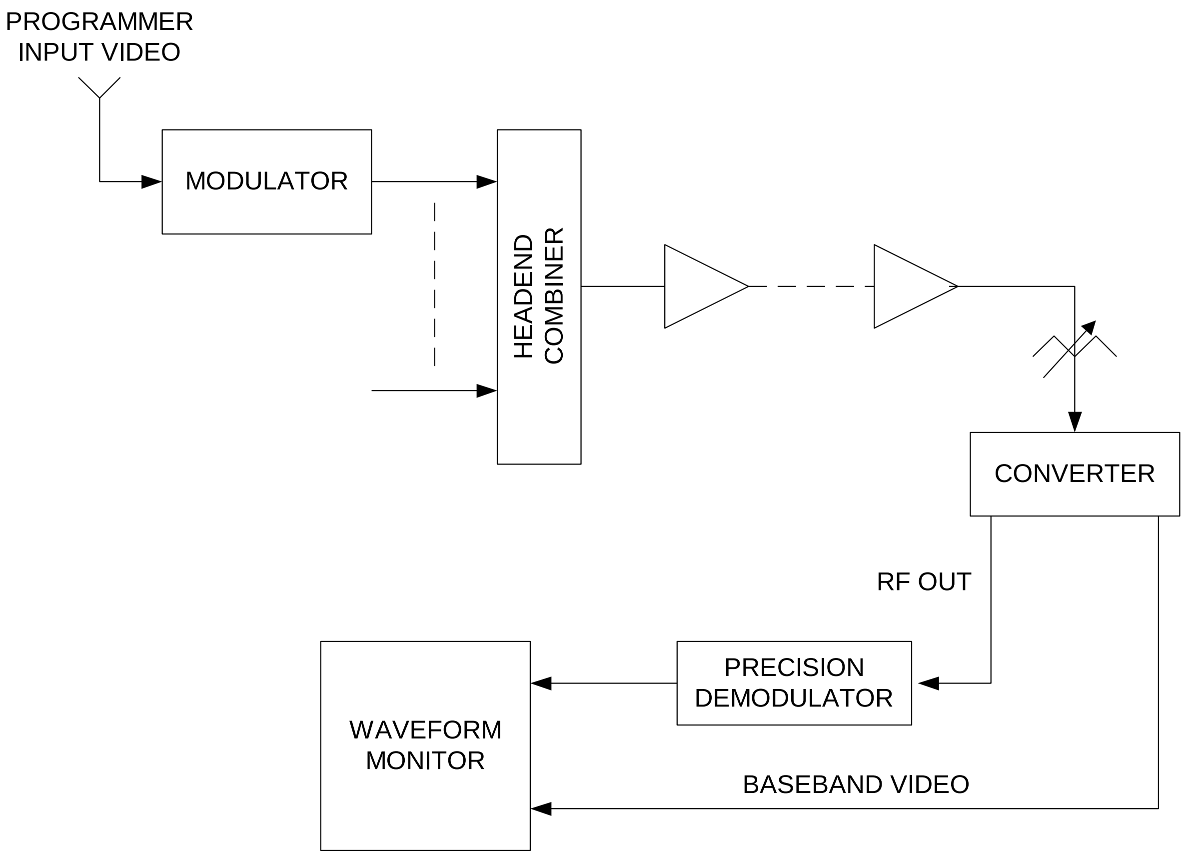


Figure 6-1: Equipment Setup

4. Calculate the response variation using:

$$\text{Response}(\pm\text{dB}) = \frac{1}{2}\left[20 * \log\left(\frac{\text{Maximum Burst Amplitude}}{\text{Minimum Burst Amplitude}}\right)\right]$$

5. If the response of the headend, added to the distribution system and converter in-channel response variation (assuming headend and distribution system are separately measured), is no greater than ± 2 dB, no further measurements are required on that channel.

6. Tune to the next channel to be measured and repeat steps 3-5.

PROCEDURE 2 - Normalizing for As-Received Response Variations: Part A – Channels Received as, or Demodulated to, Baseband

Although cable operators are responsible only for response variations that occur after reception, the above procedure includes all variations. In the case of channels with usable programmer-supplied VITS whose overall response, as measured above, exceeds the allowable tolerance, the following additional procedure will remove the effect of all variations occurring before reception. This procedure applies to all channels received at, or demodulated to, baseband, including satellite channels and those received by FM, digital link, or similar means.

Measurement Steps

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Remove the internal or external termination from the modulator input for the channel under test. Connect the loop-through video signal directly to the video waveform monitor as shown in Figure 6-2.
3. Select the multiburst waveform for viewing and record the peak-to-peak amplitude (in IRE units or volts) of each burst up to and including 3.58 MHz. The variations will include all factors prior to reception by the cable system.

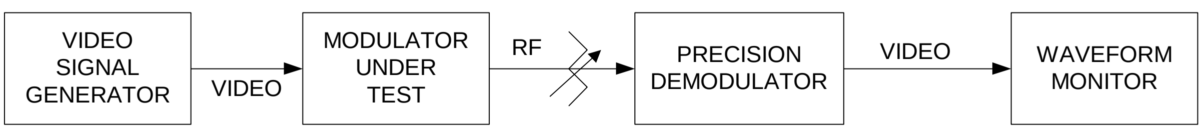


Figure 6-2: Measuring Input Flatness

4. Now calculate the difference between each of the previously recorded amplitudes (for that channel at the test point) and the new amplitudes in dB using see Table 6-1.

$$\text{Difference}_n\ (\text{dB}) = 20 * \log\left[\frac{\text{Test Burst Amplitude}_n}{\text{Reference Burst Amplitude}_n}\right]$$

5. The individual variations represent the relative response, at each multiburst frequency, of the portion of the cable system being tested plus the agile test demodulator. Half the difference between the largest and smallest of these individual variations is the total response variation, as specified by the FCC. The total variation can be calculated from:

$$\text{Total Variation } (\pm\text{dB}) = \left[\frac{(\text{Maximum Difference}) - (\text{Minimum Difference})}{2}\right]$$

Burst Frequency (MHz)	Reference Amplitude (IRE)	Test Amplitude (IRE)	Difference (dB)	Total Variation (dB_{pk-pk})
0.5	80	78	-0.22	±1.35
1.0	78	76	-0.23	
2.0	81	70	-1.27	
3.0	70	63	-0.92	
3.58	56	40	-2.92	

Table 6-1: Sample calculation for one channel.

6. Repeat steps 2-5 for other video signals which are received as, or demodulated to, baseband and which have usable programmer VITS.

PROCEDURE 3 - Normalizing for As-Received Response Variations: Part B-Processed (not demodulated) Over-the-air Channels

This procedure is the equivalent of the above for processor channels. Unlike the above, it also corrects for demodulator response variations (assuming the demodulator has a similar response on every channel).

Measurement Steps

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Connect the test equipment of Figure 6-1 to the input test port of the processor for the channel under test. If that signal level is insufficient, it will be necessary to use the processor input signal directly. Tune the precision demodulator to the input channel frequency.
3. Make the measurements and calculations in steps 3-5 of Procedure 2.
4. Repeat for each processor channel to be tested.

Measuring Modulator Channels using Alternate VITS Source

In cases where the channel is locally generated or where the programmer has not supplied a usable multiburst signal, several alternate procedures are possible:

1. Substitute, for the normal programming, an available signal from another programmer which includes a usable multiburst signal.

> ***Caution: you must either have the right to transmit the alternate programming used or disconnect the modulator from the distribution system while the measurement is being made!***

2. Put a VITS inserter in series with the normal programming. This instrument will "clean" one or more lines of the vertical interval and insert VITS into the video signal.

3. Replace the normal programming with a full-field video test generator. This will interrupt programming but allow response to be measured with waveform monitors that do not include line select capability or with spectrum analyzers.

> ***Caution: Only some versions of multiburst work properly when using spectrum analyzers for response analysis. In particular, signals which are combined with other signals, such as the NTC-7 Combination, or signals with unequal width burst packets will give false flatness readings.***

PROCEDURE 4 - Measuring Processor Channels Using RF Sweep

When broadcasters do not provide usable multiburst signals, it is possible to directly sweep the processor.

Required Equipment

- A bench sweep generator with a frequency range that covers the input frequency of the processor to be tested.
- A directional coupler (approximately 20 dB) with good isolation and flatness.
- A spectrum analyzer with a frequency range that covers the output frequency of the processor to be tested.

Measurement Steps

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.Connect the equipment as shown in Figure 6-3.
2. Set the sweep generator to sweep the frequency range from at least 0.5 MHz below the visual carrier to 3.75 MHz above the visual carrier. Set the level to approximately 20 dB below the visual carrier.
3. Tune the spectrum analyzer so that the entire output channel is displayed. Set the amplitude scale to 1 or 2 dB/division. Adjust the input attenuator and gain so that the sweep signal is within the display window.
4. Adjust the generator sweep rate to sweep the channel in approximately 1/2 second. Adjust the analyzer to approximately 300 kHz resolution and adjust the sweep rate so that the response is "painted" on the screen. If the analyzer has a storage or peak-hold mode, enable it. The exact adjustments will depend on the capabilities of the analyzer. Note that under some conditions (for instance some character-generator patterns) there may be visual sidebands that will exceed the amplitude of the sweep trace and give false flatness readings.

Hints and Precautions

While the above procedure can be performed with the processor "on-line", there will be a high degree of subscriber interference while the sweep generator is in use. When using analyzers with trace storage or peak hold capabilities, however, only a single sweep will be required, limiting interference to a few seconds, at most.

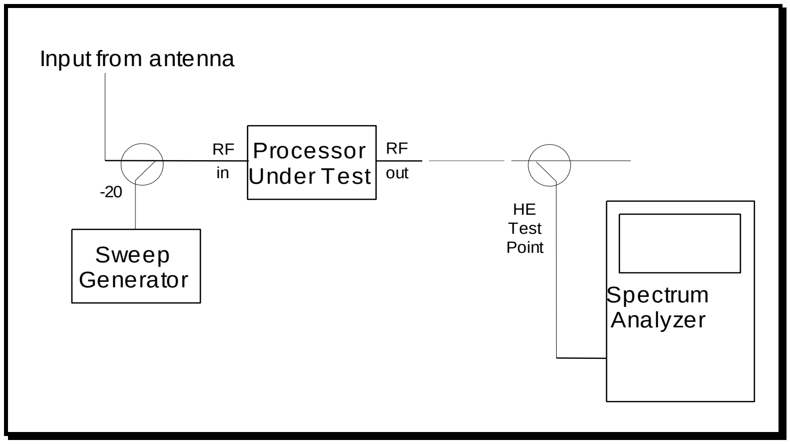


Figure 6-3: RF Processor Sweep Setup

An alternative to RF sweeping of processors where there are no usable programmer VITS is to use a VITS generator and test modulator to create an input signal. While this is an acceptable alternative, the response of the test modulator adds to the uncertainty of the measurement.

PROCEDURE 5 - Measuring In-Band Response Using a Video Sweep Signal and Spectrum Analyzer

This procedure measures the response variations from modulator or processor input to either headend output or, as with the above procedures, to the field test points. It requires interruption of normal video programming. It differs from multiburst testing in that it provides a continuous response curve and allows the use of a spectrum analyzer rather than a waveform monitor for display.

Required Equipment

- Spectrum analyzer with peak-hold capability. A non-peak-hold unit may be used if the display is photographed with a time exposure to record the trace maximum.
- Video sweep signal generator. This unit generates a conventional video signal in which the visual information is replaced by a frequency varying monotonically from 0.5 MHz to 5.0 MHz or higher. This signal may vary at the line rate or frame rate.
- Test modulator (for processor channels only).

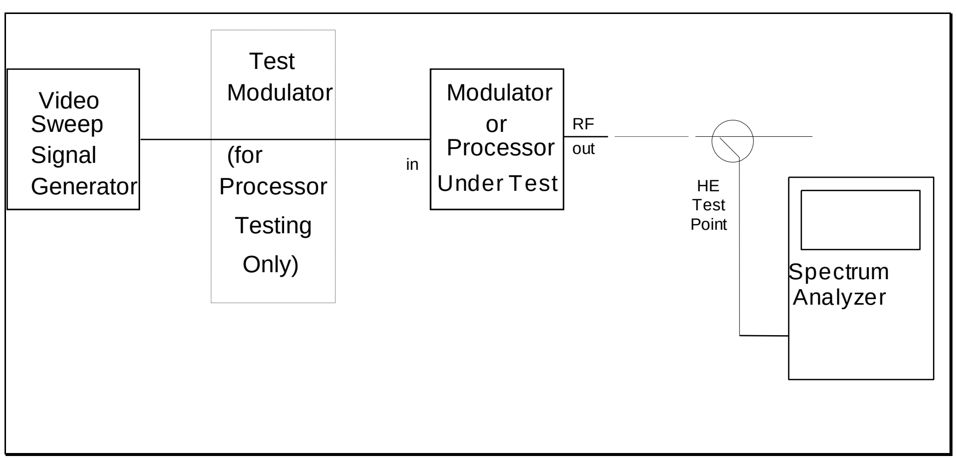


Figure 6-4: Video Sweep of Modulators or Processors

Measurement Steps

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Connect the equipment as shown in Figure 6-4. Use the test modulator only when measuring processors. Unless the modulator has been previously characterized for response, it should first be measured using this procedure (and a back-to-back connection between modulator and spectrum analyzer) and its measured variation subtracted from the data obtained when measuring the processor under test.
3. Tune the spectrum analyzer to the channel under test. The spectrum analyzer should be in log mode. Set the sweep rate so the trace can "paint" the response when the peak hold mode is enabled, as shown in Figure 6-5.
4. If frequency markers are available on the analyzer, set them to 0.5 MHz below the visual carrier and 3.75 MHz above the visual carrier (F_1 and F_2 in the figure).
5. The variation, in dB_{pk-pk}, is half the difference between the maximum and minimum amplitudes (L_1 and L_2 in the figure) between the frequency markers, excluding the visual carrier itself. In the example shown, half the peak-to-peak variation is approximately 2.75 dB which exceeds the allowable variation.
6. Photograph or print the display, if desired, for a permanent record.

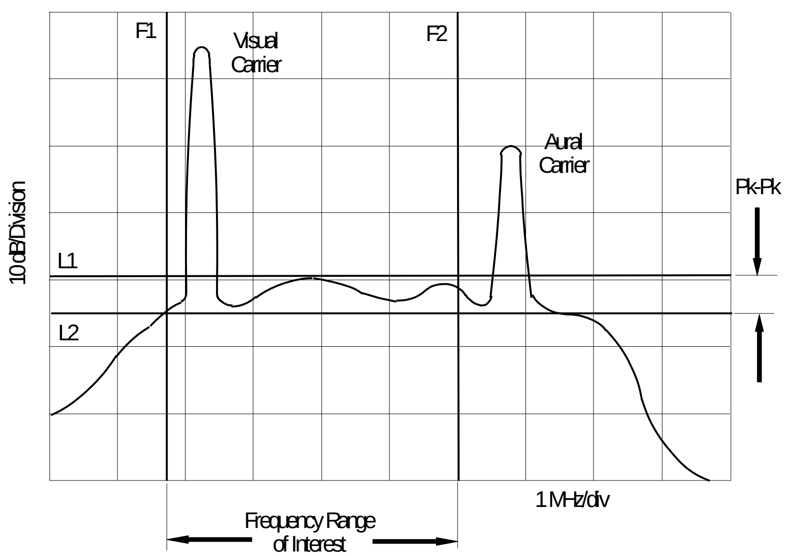


Figure 6-5: Typical Video Sweep Response

6.2 Frequency Response - Forward

Definition: Frequency response is a measure of the overall gain variation of a cable system as a function of frequency. It is normally measured in dB peak-to-peak (sometimes called peak-to-valley) or as ± dB (half the peak-to-peak value). In some cases, peak-to-valley is defined relative to a specific frequency characteristic such as the uniformly-sloped output of system amplifiers.

Discussion: Frequency response variations affect the performance of cable systems in a variety of ways. The variation over wide frequency ranges, a function of the response of the distribution system, primarily affects the relative level of various channels. While there is no specific FCC regulation of system response, these variations degrade noise and distortion performance and also make it harder to meet the FCC's specifications with regard to delivered signal level variation (§76.605(b)(4)). Secondarily, variations in the distribution system can affect the response within each channel.

Absent major “fine-grain” system response variations (caused, for instance, by resonant conditions or serious impedance mismatches), variations in the response of headend equipment, converters, channelized microwave transmitters and any other single-channel processing equipment generally have a greater effect on in-channel response than the broadband distribution system. Fine-grain variations can affect not only the relative amplitude of different frequency components of the visual signal, but relative aural/visual carrier level and chroma delay.

Overall system frequency response is measured over the broadband portion of the plant. In-band response can either be measured through the entire network or separately in the headend/signal processing equipment and the distribution system. While each has its advantages, separate measurement of headend and distribution plant generally involves less signal interruption for subscribers.

Test Procedures: Distribution system response over the full bandwidth is generally specified and measured between any two points in the system for which the nominal gain is unity (0 dB), in such a way as to exclude the effect of channel tilt (full, half or block). Typically, response of all-coaxial distribution systems is measured at an amplifier output with reference to the output of the headend channel multiplexer and its associated amplifier where the operating signal levels are nominally the same as for the point under test. In the case of hybrid fiber-coaxial (HFC) systems, levels are usually nominally flat through the optical portion of the network and a uniform gain slope is created within the first coaxial amplifier. In that case, either the optical portion of the network must be measured separately (with the reference for the coaxial portion taken at the output of the first coaxial amplifier or the measurement equipment must include means for normalizing to a given optimal gain slope value.

For measurements of transmission systems in which some portions of the network have different levels and/or slope (for instance, bridger or line extender levels vs. trunk levels in a trunk-and-feeder coaxial network), the effects of the nominal reference changes must be offset either by adjusting the test signal input levels or by applying appropriate corrections.

Although it is possible to test a distribution system using bench sweep methods (continuous running sweep generator at the headend and detector and triggered oscilloscope for display in the field), that method requires removal of normal signals and is of use only in testing plant that is not in service. Virtually all in-service testing is done using one of several proprietary sweep systems designed to either minimize or eliminate interference to subscribers. A generic description of the various methods used follows.

PROCEDURE 1 – High-Level Simultaneous Sweeping

High-level simultaneous sweeping is accomplished by inserting a swept reference signal into the headend output spectrum at a level 15-20 dB above the levels of video carriers. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing. The sweep generator covers the entire spectrum in a few milliseconds and has a delay between sweeps of several seconds. The effect on subscribers viewing analog television programs is that a few lines of each channel turn momentarily black every few seconds. If the sweep signal happens to be present in a channel during the vertical blanking interval (VBI), the picture may be momentarily unstable on some television sets. If the sweep signal passes through channels carrying digital signals, data errors will occur. Many systems using high-level sweeping use timers to shut the system down except when measurements are being taken to minimize complaints.

The receiver consists of a wideband detector and a triggered oscilloscope. The oscilloscope trigger circuit detects the onset of sweep by the increase in total RF power level in the system and initiates a linear sweep across the screen whose timing is the same as that of the transmitter. The vertical axis displays the variation in total system delivered power as the transmitter sweeps across the spectrum. Since the variations in the many television carriers will tend to average out and the energy in the sweep signal is 30-100 times that of any single carrier, the total delivered RF energy will tend to follow the variation in the amplitude of the sweep signal.

High-level sweep systems must be used carefully in order to achieve accurate results. If the reference level is too low, the trace will be noisy as picture carrier variations will tend to mask system response. If the level is set too high, the total amount of energy will tend to saturate the amplifiers and given an unrealistically flat response readout. If the interval between sweeps is too long, system alignment is very inefficient because much time is spent waiting for new response data after each adjustment, while if it is set short, subscriber complaints increase.

Aside from direct subscriber picture interference, high level sweep systems interfere with AGC carriers used by amplifiers. This may be eliminated by notch filters at the transmitter, but at the expense of a portion of the response curve. Finally, high-level sweep systems interfere with most types of digital transmissions, whether contained within the VBI of video signals or modulated on independent carriers.

Because of its many limitations, high-level sweeping is seldom used in modern cable television systems.

PROCEDURE 2 – Low-Level Synchronous Sweep

Low-level synchronous sweep systems inject their headend reference carrier at a level that is approximately 30 dB below that of the analog visual carriers. They also inject a pilot carrier in an unused portion of the bandwidth that is utilized by the receiver to recover the sweep signal. Current systems sweep the entire spectrum continuously. Since the generally accepted threshold of visibility of an interfering carrier in a television channel is about 60 dB below the visual carrier, the Low-level sweep signal is not truly non-interfering, however the relatively low dwell time per channel and a carefully chosen sweep rate minimize the subjective effect. As with high-level systems, operators typically turn such systems off except when measurements are being made.

Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

The receiver is essentially a tracking spectrum analyzer. It recovers the pilot carrier and uses the information contained in it to tune synchronously with the transmitter. By use of a relatively narrow bandwidth (several hundred kHz), it is possible to see the level of the swept carrier except where the levels of normal system carriers are higher (near the visual and aural carriers).

As with high-level systems, low-level synchronous systems must be adjusted carefully to optimize operation. If the reference sweep is set too high, subscriber interference becomes highly visible. If it is set too low, the trace becomes noisy and the percentage of bandwidth for which usable sweep information can be recovered decreases considerably. If block tilt is used at the headend, it might be difficult to find an acceptable compromise level. An advantage of the system is that it is capable of displaying system response and carrier levels on the same instrument very easily.

PROCEDURE 3 – Visual Carrier Reference Systems

A third method does away with the swept reference carrier altogether and uses the normal system analog visual carriers as the reference. In this method, the instrument calibrates itself at the headend by recording the amplitude of each carrier in turn. When a reading is taken in the field, it again reads each carrier and constructs a curve based on the difference in levels between each carrier as measured at the headend and the level when measured at the field test point.

Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

While the method is totally non-interfering, it has definite limitations. First, although an apparently continuous curve is displayed, in fact only a single data point, at best, per 6 MHz is recovered, possibly masking "fine grain" response characteristics. Second, no data is available in portions of the spectrum where there are no analog visual carriers. Third, the data is only as good as the stability of the headend carriers, so that any headend carrier variation is indistinguishable from a variation in system response.

PROCEDURE 4 – Hybrid Systems

One system provides a hybrid between a high-level and visual carrier reference system. With this system, the operator can choose to use analog visual carrier reference signals to avoid interference where there are active video channels while inserting a transmitted high-level reference carrier where there is none. The transmitted signal is synthesized, as opposed to a continuous sweep, so that data is taken only at discreet frequencies. If greater resolution is desired within channels, the operator can choose to tolerate some degree of interference in exchange for greater frequency resolution. The interference is less than a conventional high-level sweep, however, as the sampling time is shorter. As with Visual Carrier Reference Systems, where headend carriers are used for reference, this system is only as accurate as the stability of the headend. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

PROCEDURE 5 – Vertical Interval Sweep Systems

In this method, the reference signal's presence in each analog television channel is timed to coincide with an unused portion of a line in the vertical interval, and is thus undetectable by subscribers. The reference carrier level is set to approximately the same level as the visual carrier. That, plus its presence only when there is no active video present, assures a noise-free presentation. As with the low level system, a pilot signal is also transmitted to synchronize the receiver with the transmitter.

Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

As with the Hybrid System, the sweep generator is a frequency synthesizer, so that data is only taken at discreet frequencies. The operator can select frequency resolution, with the data refresh rate varying inversely with the number of data points per sweep.

The methodology of vertical interval timing is specific to analog television channels. Thus, this method has the same limitations as the previous method when measuring response in channels used for other than analog video signals.

Notes, Hints and Precautions

Overall Response Variation: Figure 6-6 shows a typical system response variation. Assuming a response nominalized for intentional system gain/slope variation, the overall peak-to-valley is the difference between the highest and lowest level *within the frequency range of interest*. In the example given, the system specified bandwidth is from 54 MHz to 270 MHz (the unshaded portion of the curve). Within that range, the response varies by about 2.6 dB.

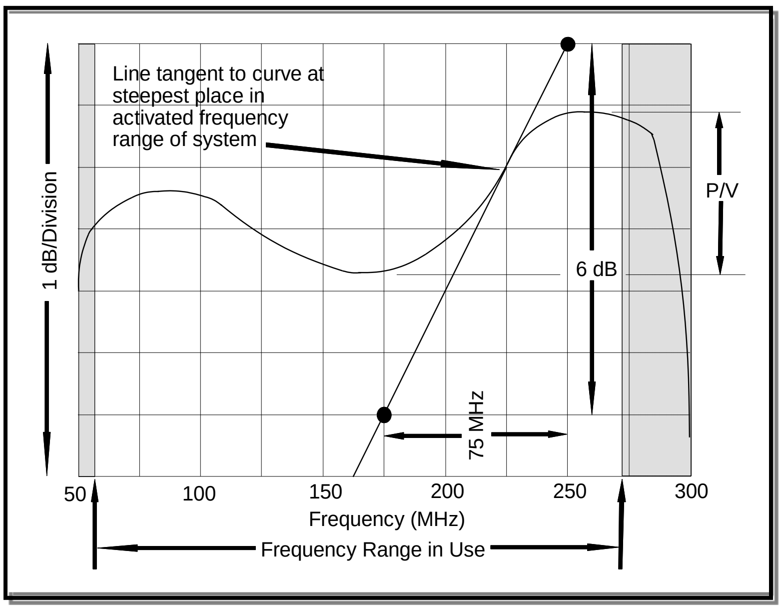


Figure 6-6: Hypothetical System Response and Maximum Rate of Change

While there is no specific regulatory requirement on overall system response, a generally accepted guideline is that the frequency response of coaxial distribution systems (measured in dB peak-to-valley) should be no greater than N/10 + 2 for the trunk line where N is equal to the number of trunk amplifiers in cascade. When the feeder system is included in the measurement, it should be no greater than N/10 + 3 where N is equal to the total number of trunk amplifiers plus the distribution and line extender amplifiers in cascade. For HFC systems with relatively small node serving areas, one possible guide is N/10 + 3 where N is the number of cascaded amplifiers after the node. Unless otherwise stated, overall system response is specified with reference to the nominal reference response and excludes intentional gain slope variation.

Contribution to In-Band Response Variation: While it may be difficult to identify the frequency boundaries of individual video channels on a wide-spectrum response plot, a good approximation of the greatest effect per channel can be made as follows: First, examine the response curve to find the greatest rate of change in the activated portion of the system bandwidth. Draw a line through the curve that has the same slope as this steepest portion. Now determine the slope of the line by marking off a segment of the line and dividing the amplitude change of that segment by the frequency change across the same portion. This will give the slope in dB/MHz. Multiplying that by 4.25 will give the amount of peak-to-peak change in the FCC-specified bandwidth. Half that is the maximum change (as ± dB) over the specified portion of any single television channel.

In the example pictured (Figure 6-6), the steepest part of the relevant curve occurs at approximately 225 MHz at this test point. A straight line is drawn so that it is tangent to the curve at this point. Considering the portion of the line that lies between the two heavy dots, it can be seen that the slope of the line is 6 dB per 75 MHz or 0.08 dB/MHz.Multiplying that by 2.125, we get ± 0.17 dB of maximum response variation in the specified portion of any channel due to distribution plant response variation.

Hints and Precautions: Regardless of the methodology used for sweep testing, the headend setup and field setup should always be checked for flatness accuracy before proceeding to the system test points. This can be accomplished by connecting the field setup drop cable to the headend final test point to get the reference trace. Match problems in the various splitters, couplers, attenuators, *etc*. used in the setups (particularly the couplers used for the headend transmitter insertion and the headend final test point) can affect the accuracy of the final measurements. These problems can usually be corrected by the use of better devices or the strategic placement of attenuator pads to improve marginal return loss.

Sometimes, the headend final test point can be configured to give exactly the same levels as a typical trunk station, so that each succeeding amplifier can be configured to match the headend output.

Block tilt (such as the difference in level between analog video and digital signals) established in the headend should carry through the system unaffected; therefore, systems with block tilt should be aligned flat (relative to the reference slope, if any), even though block tilt exists between various portions of the spectrum.

In some instances, it may be necessary to either pre-equalize the sweep signal, or equalize the input of the receiver. These situations involve non-standard spacing between amplifiers of which a system operator with special designs will be aware.

Systems using channelized microwave will generally need to have the sweep reference transmitter located at the microwave receiver as there is no single insertion point for such systems. Also, some microwave transmitters may not tolerate intermittently-present additional high-level signal without causing picture disruption. In such cases, operators must be aware of the stability of the sweep transmitter and provide an appropriate operating environment for it in the field.

Where broadband optical links form a portion of the tested system, care must be exercised, especially with high-level sweep systems, to avoid downstream laser clipping due to the additional RF drive power from the sweep transmitter.

In the case of systems which split the spectrum and use two or more broadband transmission paths between the headend and remote hubs (split fiber or split broadband microwave systems), the transmitted signal must be inserted In both paths and special circuitry used to create a single headend reference port. In these systems, care must be taken to match the insertion and reference couplers and to avoid coupling between the transmission paths.

Operators should be aware of two possible limitations when using system overall sweep response to measure variations within a single channel: first, that sweep systems with limited frequency resolution (one point per channel or less) may not properly register fine-grain response variation due to impedance mismatch in some conditions and second, that any error in the graphical procedure of determining the maximum slope of the overall response variation will cause a proportional error in the final result.

6.3 Frequency Response - Return

Definition: The flatness of the reverse path is the relative gain difference at various frequencies throughout the range of the reverse path. For U.S. cable systems this is usually from 5 MHz to 42 MHz (but can be greater). The entire path from a subscriber to the headend can be tested or any segment can be tested separately. (For example, the fiber portion and the coaxial portion could be tested separately if desired.)

When testing just a portion of the system, the signal source will be connected at the end of the tested segment that is furthest from the headend (or node) and the signal measurement device will be at the end of the tested segment that is closest to the headend. See Section 16.4: “Return Plant Setup and Operational Practices” for a discussion of the appropriate test points.

Flatness is measured when setting up the system or when verifying performance.

PROCEDURE 1 - Sweep System (or Equivalent)

Required Equipment

- A sweep system capable of reverse path testing

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Using an automated reverse sweep system inject the signal at the source point to be tested (usually either a subscriber tap or at an amplifier test point). This system will provide a measurement of the flatness of the reverse channel.
3. When making this measurement use caution to not allow the test system to inject a test signal that might interfere with existing signals in the transmission system. This includes signals that are not seen at the injection point, but which are being sent to the headend by another portion of the system.

PROCEDURE 2 - Out-of-service

Discussion: This method uses generic test equipment, but requires that the system be tested out-of-service.

Required Equipment

- A sweeping or stepping signal generator
- Spectrum analyzer

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Using the signal generator, inject the signal at the source point to be tested (usually either a subscriber tap or at an amplifier test point).
3. At the upstream point to be tested, use the spectrum analyzer in the max-hold mode to build up a display of the flatness of the reverse path.

Notes, Hints and Precautions

Any non-flatness of the signal generator itself must be subtracted from the display to provide the correct answer.

A spectrum analyzer plot should be made at the upstream point to be tested before the signal generator is connected. This will show the ingress and any spurious signals present. These signals should be noted as they are not to be used as a part of the flatness measurement.

When testing by inserting any signal into an operating system, there is the possibility of causing nonlinear operation due to compression or laser clipping. This is the result of adding enough power to the total already arriving at the laser such that the maximum allowable (without distortion) is now exceeded.

Use caution to ensure that when adding a sweep or other testing signal that the system remains linear. Consult the equipment manufacturer if necessary.

6.4 QAM Signal In-Channel Flatness

PROCEDURE 1 – In-service (tilt only with spectrum analyzer)

Discussion: Normally, a QAM signal should have a flat spectrum over most of the occupied channel. Deviation from flatness can be the result of reflections, in-channel response of RF upconverters or other causes. This procedure will deal with deviations from flatness which produces a positive or negative tilt in the spectrum over the width of the channel.

Required Equipment

- Spectrum analyzer with 75 Ω input and marker delta measurement capability

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Set the spectrum analyzer as follows:
   - Center Frequency: Center of channel under test
   - Frequency Span: Wide enough to see entire signal under test (See Note below).
   - Input Impedance: 75Ω
   - IF Resolution Bandwidth: 300 kHz
   - Video Bandwidth: 30 kHz or lower, or until display stability is reached (30 Hz is preferred)
   - Reference Level: Position top of signal trace approximately 10 dB below the top of the analyzer display.

- Number of Video Averages: Until display stability is reached

Note: The recommended span settings are as follows:

- 30 MHz to 40 MHz for L-band signals
- 10 MHz for QAM signals transmitted in a 6 MHz channel
- 3 MHz for out-of-band signals

3. Move the marker to the center frequency of the channel under test. Change the marker position to the frequency shown in Table 6-2 as Marker 1. Select the Marker Delta function. Move the second marker to frequency shown in Table 6-2 as Marker 2. Read and record the dB difference in amplitude between the markers. This number is the approximate in-channel tilt of the signal.

Note: This procedure measures the tilt of a QAM signal in less than the channel bandwidth. One can divide the measured tilt by the measurement bandwidth, expressing tilt in dB/MHz. Then multiply this tilt rate by 6 to express the channel tilt over a 6 MHz bandwidth by linear extrapolation.

Modulation Format	Marker 1 Position	Marker 2 Position
L-band QPSK – 24 MHz signal bandwidth	CF: 9.755 MHz	Marker 1 + 19.51 MHz
L-band QPSK – 36 MHz signal bandwidth	CF: 14.635 MHz	Marker 1 + 29.27 MHz
QPSK Out-of-Band signal per SCTE DVS-178, Table 1	CF: 512 kHz	Marker 1 + 1.024 MHz
QPSK Out-of-Band signal per SCTE DVS-167, Table 2-2	CF: 386 kHz	Marker 1 + 772 kHz
QPSK Out-of-Band signal per SCTE DVS-167, Table 2-2	CF: 772 kHz	Marker 1 + 1.544 MHz
64-QAM	CF: 2.53 MHz	Marker 1 + 5.06 MHz
256-QAM	CF: 2.68 MHz	Marker 1 + 5.36 MHz
8-VSB	CF: 2.69 MHz	Marker 1 + 5.38 MHz

Table 6-2: Marker Positions for Tilt Measurement

Figure 6-7 shows an example of a tilt measurement.

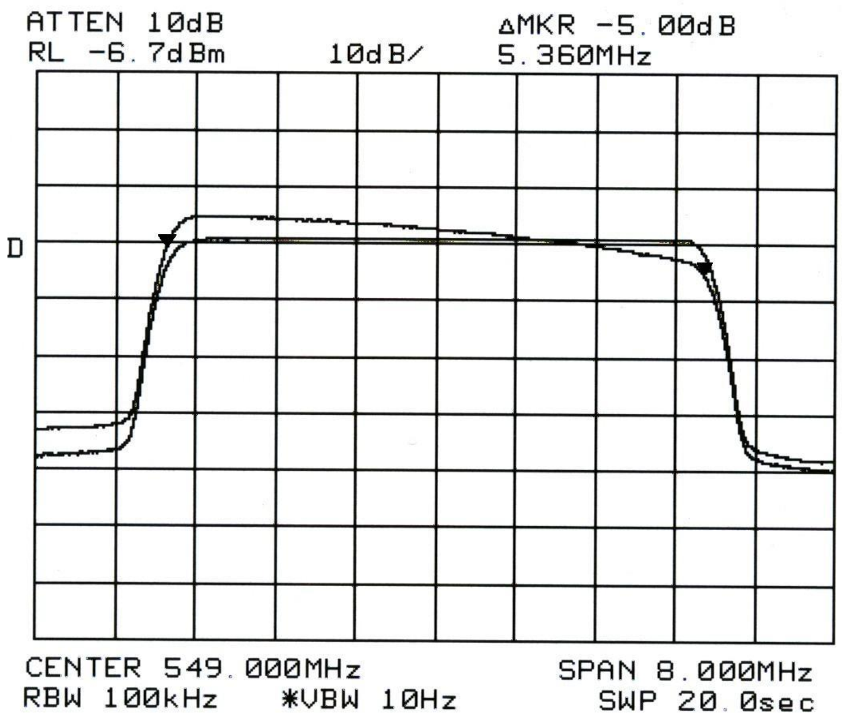


Figure 6-7: Tilt Measurement Example

PROCEDURE 2 – In-service (more generalized in-channel flatness with spectrum analyzer)

Discussion: The following method focuses on downstream QAM signals, and is essentially the same as what is described in Procedure 1. This variation of Procedure 1 uses the test equipment display's on-screen vertical graticule as an amplitude-versus-frequency response limit guideline rather than using markers.

Required Equipment

- Spectrum analyzer with 75 Ω input

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Set the spectrum analyzer as follows:
   - Center Frequency: Center of channel under test
   - Frequency Span: Wide enough to see entire signal under test (See Note below).
   - Vertical Scale: Value that corresponds to desired frequency response flatness measurement limit—for example, 3 dB/division for 3 dB p-p amplitude-versus-frequency response limit
   - Input Impedance: 75 Ω

- IF Resolution Bandwidth: 30 kHz
- Video Bandwidth: offset of 10 kHz or lower, or until display stability is reached (1 kHz is preferred)
- Reference Level: Position top of signal trace approximately in the center of the top division of the analyzer's graticule.
- Number of Video Averages: Two or more, until display stability is reached

Note: The recommended span settings are as follows:

- 6 MHz (600 kHz/division) for Annex B or similar QAM signals transmitted in a 6 MHz channel
- 8 MHz (800 kHz/division) for Annex A or similar QAM signals transmitted in an 8 MHz channel

3. If the top of the signal trace—that portion occupying slightly less than the QAM signal's symbol rate bandwidth—fits within the top division of the analyzer's graticule, the signal's approximate flatness within that bandwidth meets the desired frequency response flatness specification. Figure 6-8 illustrates an example of this measurement.

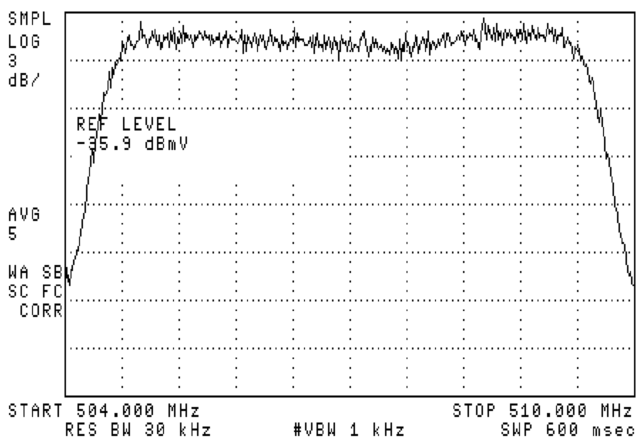


Figure 6-8: Flatness Measurement with Spectrum Analyzer

Notes, Hints, and Precautions

This procedure provides a coarse go, no-go measure of the approximate flatness of most of the QAM signal, but not the flatness of the entire channel bandwidth. It includes amplitude response contributions of the signal source, the signal itself, and the channel through which the signal is transmitted. The results require some interpretation because of the inherent shoulders of the QAM signal (-3 dB at the symbol rate bandwidth). As such, the approximate bandwidth that can be measured is somewhat less than the symbol rate bandwidth. For example, when checking the flatness of an Annex B 64-QAM signal, the approximate bandwidth for this type of measurement is 5.05 - (6 - 5.05) / 2 = 5.05 - 0.475 = 4.575 MHz. Another way of calculating the approximate measurement bandwidth is to use the Nyquist excess bandwidth factor. For Annex B 64-QAM, the excess bandwidth is 18%, so the approximate bandwidth that can be measured using this procedure is 5.05 MHz * (1 - 18%/2) = 5.05 * (0.91) = 4.59 MHz. For Annex B 256-QAM, the approximate bandwidth is 5.36 * (1 - 12%/2) = 5.36 * 0.94 = 5.04 MHz.

PROCEDURE 3 – In-service (in-channel flatness with digital signal analyzer)

Discussion: Many digital signal analyzers support measurement of in-channel flatness of QAM signals. Some digital signal analyzers support downstream-only measurements and some also support upstream measurements. The latter usually requires that a separate continuous QAM test signal be transmitted in the upstream for the duration of the measurement, while the former uses existing downstream QAM signals.

Required Equipment

- Digital signal analyzer with in-channel flatness measurement function
- Continuous QAM test signal source for upstream measurement, if a continuous signal is required

Test Procedure

1. Follow the digital signal analyzer manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Follow the digital signal analyzer manufacturer’s instructions for in-channel flatness measurement.
3. Read and record the in-channel flatness of the QAM signal. The figure below shows an example of this measurement.

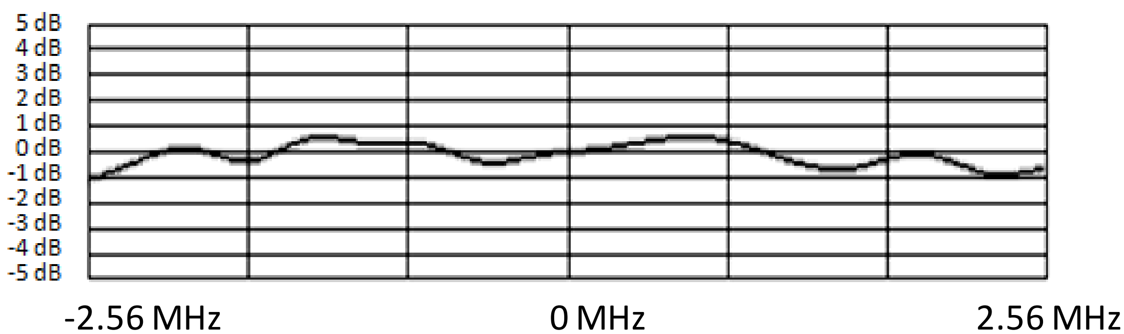


Figure 6-9: Flatness Measurement with Digital Signal Analyzer

Notes, Hints, and Precautions

This procedure derives its indicated in-channel flatness from the digital signal analyzer’s adaptive equalizer coefficients. Resolution and accuracy are limited in part by the number of taps in the adaptive equalizer and the equalizer tap spacing. The procedure provides a measure of the approximate flatness within a bandwidth that is equal to the QAM signal’s symbol rate bandwidth. The measurement results include amplitude response contributions of the signal source, the signal itself, and the channel through which the signal is transmitted. On instruments that support in-channel flatness measurements, the same bandwidth considerations apply that were discussed in Procedure 2. Check with the test equipment manufacturer to see if the displayed results compensate for the inherent shoulders in the QAM signal (allowing for flatness measurement over the entire symbol rate bandwidth).

PROCEDURE 4 – In-service

Discussion: First generation low-level synchronous sweep systems inject their headend reference sweep carrier at a level that is approximately 30 dB below that of the analog visual carrier peak envelope power (PEP). Newer versions of this technology inject a sweep signal that is approximately 40 dB down, and use digital signal processing techniques in the receiver to recover the much lower level sweep signal. Low-level sweep transmitters also inject a pilot carrier in an unused portion of the bandwidth that is utilized by the receiver to recover the sweep signal. Low-level systems sweep the entire spectrum continuously, but the receiver’s settings can be adjusted to facilitate per-channel flatness measurements. While the risk of interference to analog TV channels is low, unless newer “40 dB down” versions of this technology are used, there is increased likelihood of interference to digital channels. Operators typically turn such systems off except when measurements are being made.

Required Equipment

- Synchronous low-level broadband sweep transmitter and receiver

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Follow the sweep equipment manufacturer’s instructions for amplitude-versus-frequency response measurements.
3. Read and record the indicated amplitude-versus-frequency response flatness within the channel under test.

Notes, Hints, and Precautions

The sweep receiver is essentially a tracking spectrum analyzer. It recovers the pilot carrier and uses the information contained in it to tune synchronously with the transmitter. Low-level synchronous sweep systems must be adjusted carefully to optimize operation. If the reference sweep is set too high, bit errors may occur in the digital channels. If it is set too low, the trace becomes noisy and the percentage of bandwidth for which usable sweep information can be recovered decreases considerably. One advantage to the use of low-level sweep is that it allows measurement of the full channel bandwidth’s amplitude-versus-frequency response without the need to remove the QAM signal from the channel being measured.

PROCEDURE 5 - Out-of-service

Discussion: This method uses generic test equipment, but requires that the channel be tested out-of-service. It facilitates measurement of the entire channel amplitude-versus-frequency response.

Required Equipment

- A sweeping or stepping signal generator
- Spectrum analyzer

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Temporarily disable or disconnect the signal source from the channel being tested.
3. Using the signal generator, inject the test signal at the source feeding the point to be tested—for example, a spare port in the headend or hub site combining network.
4. At the downstream point to be tested, use the spectrum analyzer in the max-hold mode to build up a display of the flatness of the channel under test while the signal generator’s frequency is varied to cover the entire channel bandwidth.
5. Read and record the results, and enable or reconnect the channel’s original signal source.

Notes, Hints and Precautions

Any non-flatness of the signal generator itself must be subtracted from the display or otherwise normalized to provide the correct indicated amplitude-versus-frequency response.

When testing by inserting any signal into an operating system, there is the possibility of causing nonlinear operation because of compression or laser clipping. This is the result of adding enough power to the total already arriving at the laser such that the maximum allowable (without distortion) is now exceeded.

Use caution to ensure that when adding a sweep or other test signal that the system remains linear. Consult the equipment manufacturer if necessary.

Be aware that insertion of a manual sweep or similar signal in an empty channel slot may interfere with adjacent analog TV or QAM signals during the test.

6.5 Nominal Relative Carrier Power Levels and Carrier Level Variations

Definitions

The nominal relative carrier power level is defined as the difference between the power level of each digitally modulated carrier type relative to the reference analog carrier power level, expressed in dB. Each nominal relative carrier power level is uniquely specified within a range and applies to all carriers of a given type for a cable system at the demarcation point of entry to the premises.

The variation in signal power of each individual carrier is defined as the deviation from its nominal relative carrier power level.

Analog signal power is defined in Section 2.1. Digital signal power is defined in Section 2.2.

Discussion

A new measurement and analysis technique for carrier power levels is described. This technique can be used to estimate the variation from the ideal carrier levels in a coaxial distribution system for both analog and digital signals. The nominal relative carrier power level offsets between analog and digital signals are accounted for and the actual carrier level variations from their ideal values are estimated.

It can be shown that the signal level attenuation solely due to frequency dependent passive cable attenuation (or "tilt") can be reliably calculated from the measurement of actual absolute carrier power levels, even in the presence of large random channel-to-channel signal level amplitude variations. Thus, the ideal signal levels of each channel are represented by these calculated absolute carrier power levels where amplitude variations are removed such that the minimum mean squared error of the channel-to-channel amplitude variations from the ideal level for the total set of actual carrier levels is achieved.

The ideal level of each analog and digital carrier is defined as the carrier level obtained as the Least Squares fit using Linear Regression (i.e., slope and intercept) on the set of analog carrier levels and digital carrier levels with nominal relative carrier power level offsets applied by carrier type (for example, a QAM signal set some number of decibels below what an analog TV channel's visual carrier on the same frequency would be) versus the square root of carrier frequency.The details of this method are given in Section 6.6 Method for Estimation of Nominal Relative Carrier Power Levels.

Test Procedure:

1. Measure the visual carrier level of each analog signal using the procedure described in Section 2.1.
2. Measure the digital signal power of each digital signal using the procedure described in Section 2.2.
3. Create a spreadsheet with all measured power vs. visual carrier frequency or digital signal center frequency. A 12-channel sample is Table 6-3.

Channel	Type	Frequency (MHz)	Measured (dBmV)
2	Analog	55.25	10.56
15	Analog	127.26	8.95
11	Analog	199.25	7.76
33	Analog	277.26	6.98
45	Analog	349.26	7.14
57	Analog	421.25	5.67
69	Analog	493.25	4.46
81	256QAM	567.00	-1.15
94	256QAM	645.00	-3.12
111	256QAM	717.00	-2.80
123	64QAM	789.00	-8.57
135	64QAM	861.00	-9.24

Table 6-3: Example Power Measurements for 12 Channels before Modification

4. Increase the measured digital signal powers in the table by the value of the system design's nominal relative carrier power level offset. In Table 6-4, we have increased the 64-QAM channels by 10 dB and the 256-QAM channels by 6 dB since the example system's design parameters

stipulate that 64-QAM channels should be set 10 dB below analog video carriers and 256-QAM channels should be set 6 dB below analog video carriers.

			Power	
		Frequency	Measured	Adjusted
Channel	Type	(MHz)	(dBmV)	(dBmV)
2	Analog	55.25	10.56	10.56
15	Analog	127.26	8.95	8.95
11	Analog	199.25	7.76	7.76
33	Analog	277.26	6.98	6.98
45	Analog	349.26	7.14	7.14
57	Analog	421.25	5.67	5.67
69	Analog	493.25	4.46	4.46
81	256-QAM	567.00	-1.15	4.85
94	256-QAM	645.00	-3.12	2.88
111	256-QAM	717.00	-2.80	3.20
123	64-QAM	789.00	-8.57	1.43
135	64-QAM	861.00	-9.24	0.76

Table 6-4: Example Power Measurements for 12 Channels after Modification

5. Calculate the least squares fit trend line to fit this adjusted data (**y** parameter) using the square root of the frequency as the **x** parameter in the equation: **y = mx + b**. This may be calculated many different ways, using linear regression functions in readily available math or spreadsheet software or custom algorithms included in test equipment measurements. The trend line in Figure 6-10 is the calculated estimate of ideal carrier power level without carrier level variations at the point these measurements are being made. The trend line provides a reference that can be used to calculate carrier level variation from the ideal levels. These carrier level variations are plotted in Figure 6-11.

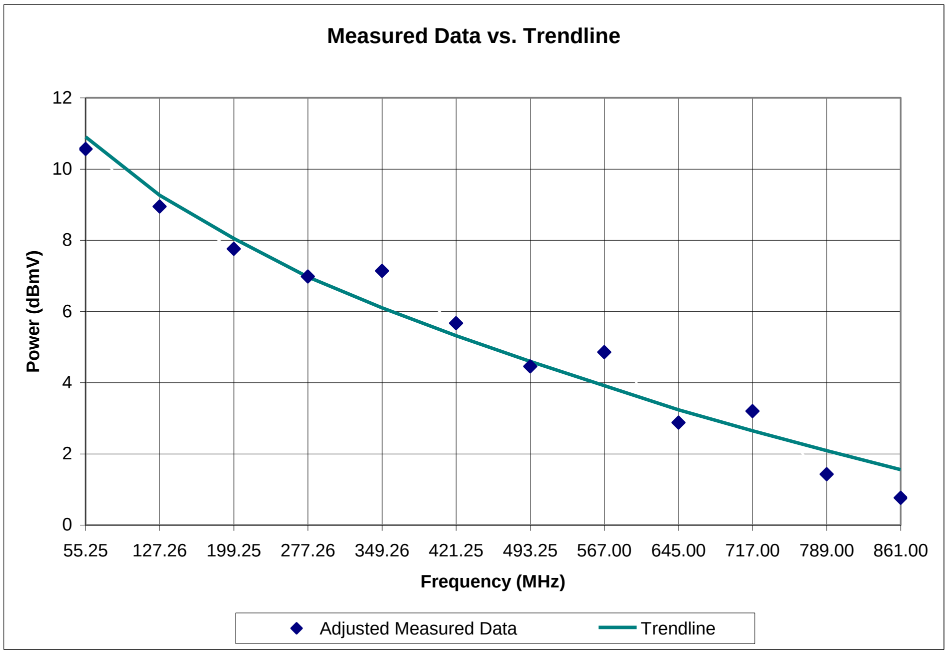


Figure 6-10: Adjusted Measured Data vs. Trend Line.

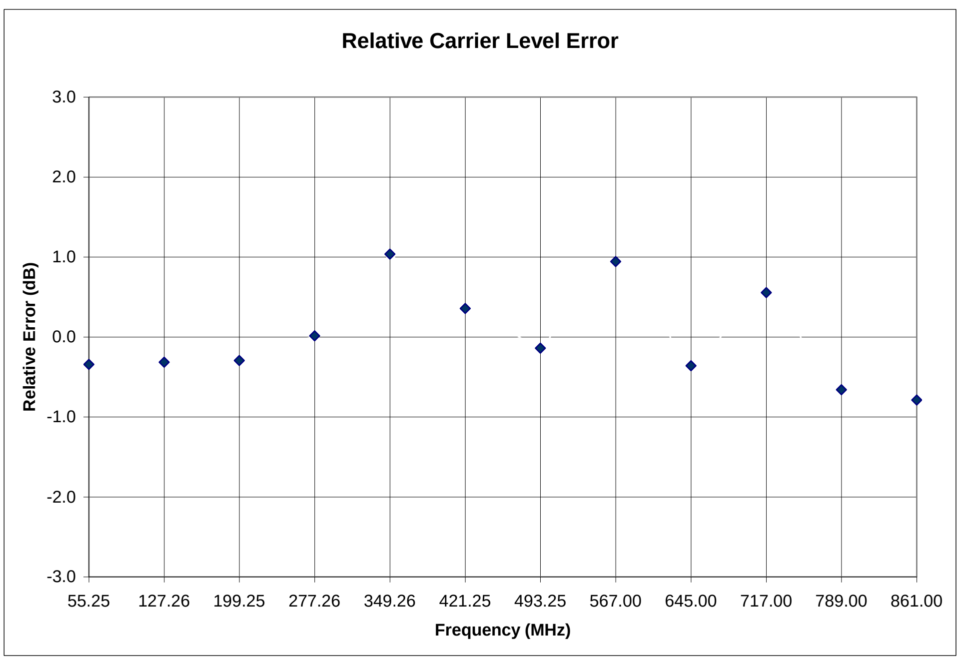


Figure 6-11: Carrier Level Variations.

6.6 Method for Estimation of Nominal Relative Carrier Power Levels

Introduction

Consider the ensemble of analog and digital signals at the receiver input. These signals have traversed the entire transmission path from the originating source in the headend through electro-optical conversion and fiber optic transmission to the node through optical-electrical conversion and transmission through coaxial cable distribution including multiple amplifiers and equalizers.

It will be shown that the signal level attenuation solely due to frequency dependent passive cable attenuation (or "tilt") can be reliably calculated from the measurement of actual absolute carrier power levels, even in the presence of large random channel-to-channel signal level amplitude variations. Thus the "ideal levels" of each channel would be represented by these calculated absolute carrier power levels where amplitude variations are removed such that the minimum mean squared error of the channel-to-channel amplitude variations from the ideal level for each carrier is achieved.

The coaxial cable produces signal attenuation per unit length that is frequency dependent whereby the higher frequency signals are subjected to greater attenuation than the lower frequency signals over the same length of cable. The attenuation in dB is proportional to the square root of frequency.

Estimation of Frequency Dependent Passive Cable Attenuation

Specifically, for the cable attenuation A_1 dB at frequency f_1 and A_2 dB at frequency f_2, the ratio of attenuations A_1/A_2 is equal to the Cable Loss Ratio $\sqrt{f_1/f_2}$.

The tilt can be calculated by

$$Tilt = A_H - A_L = A_H - A_H\sqrt{f_L/f_H} \text{ (dB)}$$

where:

A_H = signal attenuation at the highest carrier frequency f_H

A_L = signal attenuation at the lowest carrier frequency f_L

Tilt = the difference in signal attenuation between f_L and f_H.

Therefore the signal attenuation at the highest carrier frequency is given by

$$A_H = \frac{Tilt}{1-\sqrt{f_L/f_H}} \text{ (dB)}$$

The signal attenuation at the lowest carrier frequency is given by

$$A_L = A_H\sqrt{f_L/f_H} \text{ (dB)}$$

Therefore, the *relative* signal attenuation at a frequency f where $f_L \le f \le f_H$ is given by

$$\mathrm{A}(f) = \mathrm{A}_H \sqrt{f/f_H} - \mathrm{A}_L \text{ (dB)}, \text{ where } 0 \le \mathrm{A}(f) \le Tilt.$$

Substituting $x = \sqrt{f}$ yields

$$\mathrm{A}(x) = \frac{\mathrm{A}_H}{\sqrt{f_H}} x - \mathrm{A}_L$$

which is a linear function of attenuation versus the square root of frequency.

If the signal power at the lowest frequency denoted by P_L dBmV is known, then the signal level P(*f*) at all other frequencies f due to tilt absent any other variation is given by

$P(f) = P_L - A(f)$ (dBmV)

or by the linear function

$$P(x) = P_L - A(x) = P_L - \frac{\mathrm{A}_H}{\sqrt{f_H}} x + \mathrm{A}_L \quad \text{(dBmV).}$$

Note that P(*f*) is the absolute carrier power levels as a function of frequency which is precisely the definition of the ideal level of every carrier.

Estimation of Carrier Level Variation

Suppose one wants to estimate the ideal levels of all carriers measured at the receiver input that have random variations ΔP(*f*) of up to ±6 dB added to each carrier measured. Then tabulating the measured carrier powers (ideal levels plus variations) P(*x*) + ΔP(*x*) as a function of the square root of frequency yields a scatter plot of carrier power ideal carrier levels plus variations about a straight line P(*x*).

The ideal level of each carrier could be found from the carrier levels obtained as the Least Squares fit using Linear Regression (i.e. the line F(*x*) = m*x* + b with slope m and intercept b) on the set of analog carrier levels and digital carrier levels (with nominal relative carrier power level offsets applied by carrier type) versus the square root of carrier frequency.

Using the method of Least Squares for determining the best linear fit for the data yields

$$m = \frac{\sum_{i=1}^{N}(x_i - \bar{x})(y_i - \bar{y})}{\sum_{i=1}^{N}(x_i - \bar{x})^2}$$

$$b = \bar{y} - m\bar{x}$$

where $\bar{x} = \frac{1}{N}\sum_{i=1}^{N} x_i$ and $\bar{y} = \frac{1}{N}\sum_{i=1}^{N} y_i$ are the means of x and y respectively.

An example of this method for determining the ideal values from measured carrier power levels with uniformly distributed random variations between -6 to +6 dB added to each ideal carrier level is shown in Figure 6-12 and Figure 6-13 for an ideal carrier power of 9 dBmV at the lowest frequency carrier centered at 57 MHz with a tilt of 11 dB resulting in an ideal carrier level of -2 dBmV at 1 GHz. Both the linear fit as a function of the square root of the frequency and the carrier power as a function of frequency by straightforward independent variable substitution are shown in the figures. The actual ideal levels are shown in blue, the ideal levels plus uniformly distributed random variations are shown in magenta, and the estimated ideal levels calculated with a Least Squares Linear Regression on these magenta carrier powers plus variations are shown in yellow. Note the close correlation between the actual ideal levels and the estimated ideal levels (typically within 1 to 2 dB) obtained with the extreme 12 dB peak-to-peak maximum variations of the measured carrier levels.

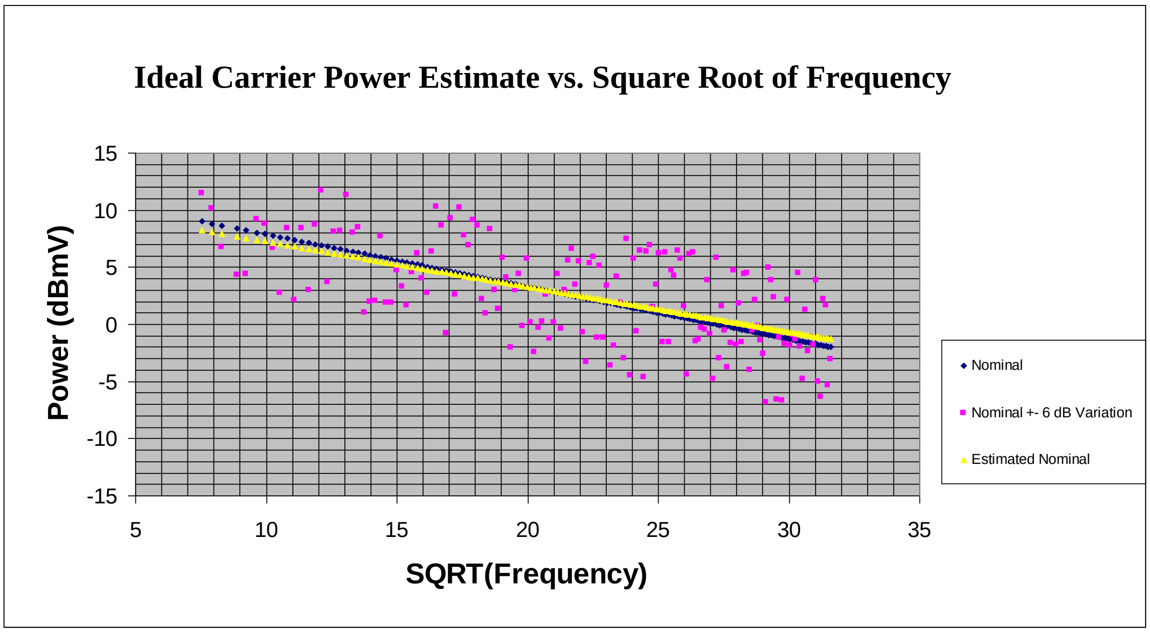


Figure 6-12: Ideal Carrier Power Estimate vs. Square Root of Frequency

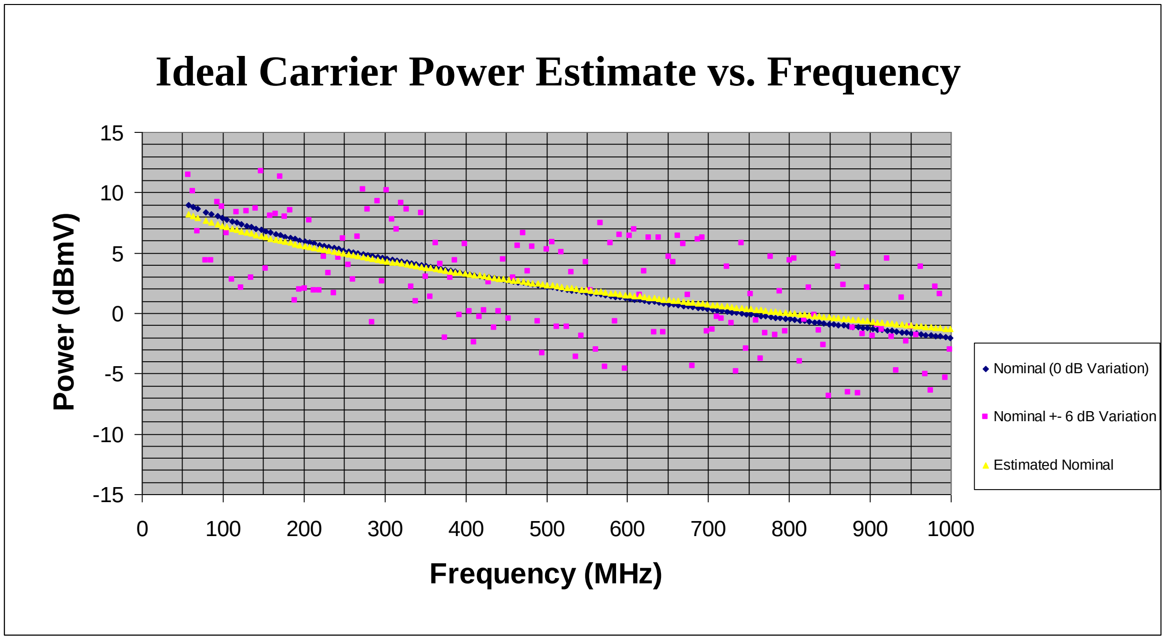


Figure 6-13: Ideal Carrier Power Estimate vs. Frequency

Chapter 7 Frequency

7.1 Visual Carrier Center Frequency

Discussion: Analog video carriers are usually characterized by the carrier frequency. A frequency counter cannot in general be used to measure the carrier frequency except where video modulation and the sound carrier can be removed during the measurement. Some specialized equipment exists that can be used if available. If a spectrum analyzer has a marker with an accurate frequency counter, it can be used to measure the carrier frequency as shown in Figure 7-1. The accuracy required is much greater than the accuracy to which the carrier frequency must be held. It is necessary to hold frequency accuracy within ±5 kHz to satisfy current FCC Rules, so the measuring instrument must have an accuracy on the order of ±0.5 kHz. Note that accuracy is not the same as resolution. An instrument that measures to the nearest 0.1 kHz may or may not be accurate enough. It has sufficient resolution, a necessary but not a sufficient condition for making an accurate measurement. Accuracy must be confirmed by independent measurement traceable to the National Institute of Standards and Technology (NIST), or by the manufacturer’s specification. This is true of any instrument used to make measurements.

Note: Not only must the frequency counter that reads the marker frequency be accurate enough, the marker must be set with enough resolution. This will require the frequency span on the analyzer to be sufficiently low that the marker can be set accurately on the carrier.

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Connect the picture carrier under test to the spectrum analyzer. With the frequency span set to about 10 MHz the presentation will be as shown on the left side of Figure 7-1. Center the picture carrier in the screen and reduce the span until it is 100 kHz or less. Keep the picture carrier in the center. Be sure that sweep speed and IF (resolution) bandwidth settings are such that the analyzer remains calibrated - usually this can be assured by keeping the two parameters in the automatic mode. Note that it is normal for the amplitude of the carrier to drop as the IF (resolution) bandwidth is reduced. With modulation the carrier amplitude will appear to change. However, it will always be greater than the sidebands surrounding it.
3. Set the analyzer’s marker on the picture carrier and measure the frequency.

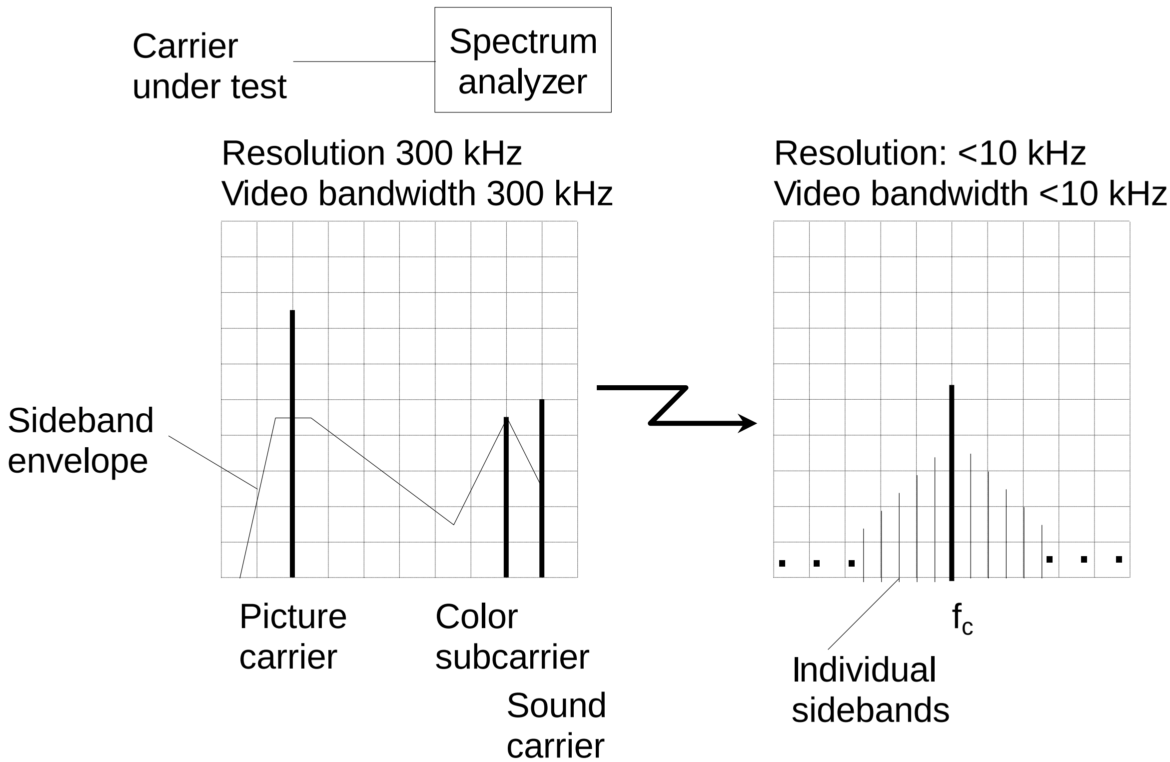


Figure 7-1: Measurement of Visual Carrier Frequency Using a Spectrum Analyzer

Alternate Procedure

If the spectrum analyzer does not have an accurate (±0.5 kHz) marker, the frequency substitution method shown in Figure 7-2 may be used. The output of a CW signal generator is split if necessary to supply signal to a frequency counter. If the generator has an accurate (±0.5 kHz) internal counter, it may be used. The remaining signal is coupled to the carrier under test through a directional coupler. The signal generator amplitude is adjusted for a convenient level on the spectrum analyzer screen and the frequency is adjusted until the carriers overlap. The frequency can then be read from the counter. Be sure to follow the same spectrum analyzer setting guidelines as above.

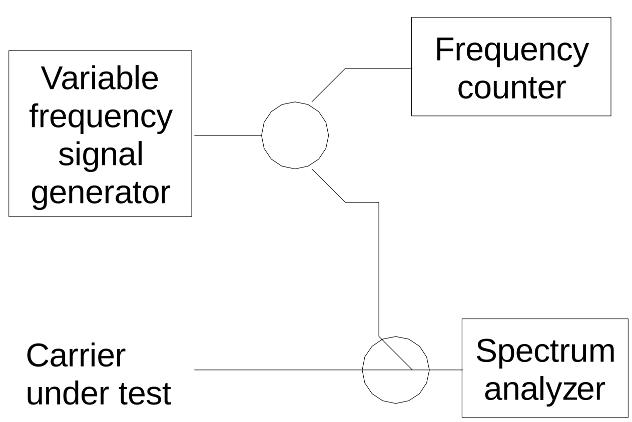


Figure 7-2: Measurement of Visual Carrier Frequency Using the Signal Substitution Technique

7.2 Aural Carrier Center Frequency [FCC §76.605(b)(2)]

Definition: The aural center frequency measurement is the difference in frequency of the aural carrier and the associated visual carrier.

FCC §76.605(b)(2): ***The aural center frequency of the aural carrier must be 4.5 MHz, ± 5 kHz above the frequency of the visual carrier at the output of the modulating or processing equipment of the cable television system, and the subscriber terminal.***

Required Equipment

- A frequency counter covering the direct frequency range to be measured
- A demodulator with 4.5 MHz audio subcarrier output
- A channel selector (if a converter is used, it must be an RF non-volume control type.)

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Connect the equipment as shown in Figure 7-3.
3. Read the frequency of the 4.5 MHz subcarrier output of the demodulator directly on the frequency counter.
4. Best results are obtained with no modulation on the sound carrier. If this is not possible, choose a long gate time on the frequency counter.

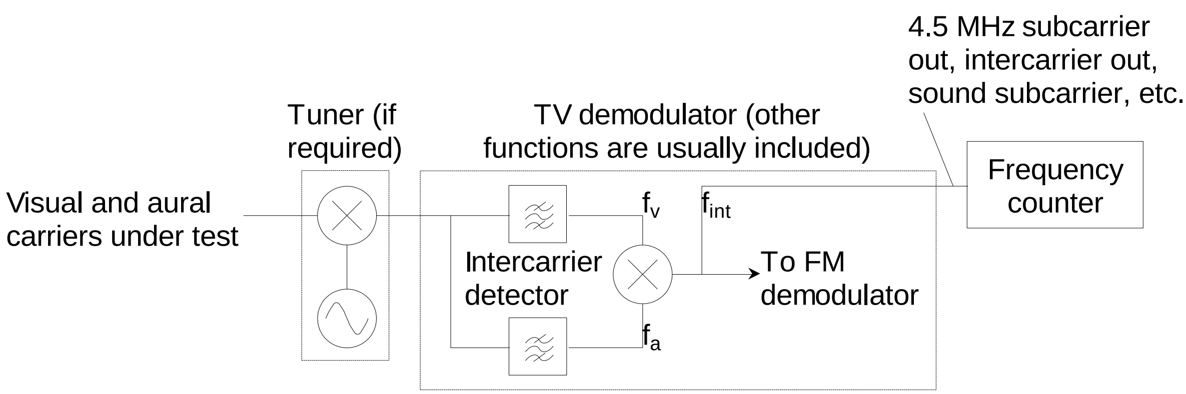


Figure 7-3: Aural Carrier Frequency Separation Test Setup

Discussion: The FCC rules require that the aural center frequency measurements be conducted at the headend or last point of modulation **AND** at the converter output at each test point. However, the aural frequency delivered to the subscriber can be completely characterized by measuring at the last point of modulation **AND** at the output of a baseband converter. Recommended practice is to measure an adequate sample of converters. Keep in mind that if several types of converters are in place in the system, this process must be done to sufficiently characterize all types of converters in use.

The FCC has eliminated standards for video carrier frequency tolerance outside of the FAA frequency bands (108.0 MHz to 137.0 MHz and 225.0 MHz to 400.0 MHz). Within those ranges the video carrier frequency must be properly offset and held to a +/- 5 kHz tolerance per FCC rule 76.612.

7.3 Digital Carrier Center Frequency

Discussion: The center frequency accuracy of a digital signal is required to be ±30 kHz by the DOCSIS RF specification SP-RFIv1.1-106-001215, Table 4.13. If a later FCC Rule establishes a tighter frequency tolerance, then it must be followed. If a measurement to this accuracy is needed, then a specialized instrument is needed. Consult the manufacturer's literature for instructions on using it. The procedure shown below can be used to make an estimate of the center frequency, but it will not yield an accurate measurement. An accurate measurement in this context must be made to ±3 kHz.

Digital signals have little or no transmitted carrier. There is a carrier but it is suppressed and is not visible on the spectrum analyzer (8- and 16-VSB transmissions do have a small carrier component transmitted). Lacking a specialized instrument, the center frequency of a digital carrier may be estimated by measuring the frequency of the point on either side of the signal, at which the signal power is below that of the flat portion of the spectrum by 3dB (see Figure 7-4). The measurement may be made using an analyzer's internal marker or by the substitution method, as discussed in the above section on measuring analog frequencies. Reading the point where the signal amplitude is -3 dB is suggested. Measure the same point on each side, calling them f_1 and f_2.

The center frequency is then obtained by averaging the two:

$$\text{Center Frequency} = \frac{f1 + f2}{2}.$$

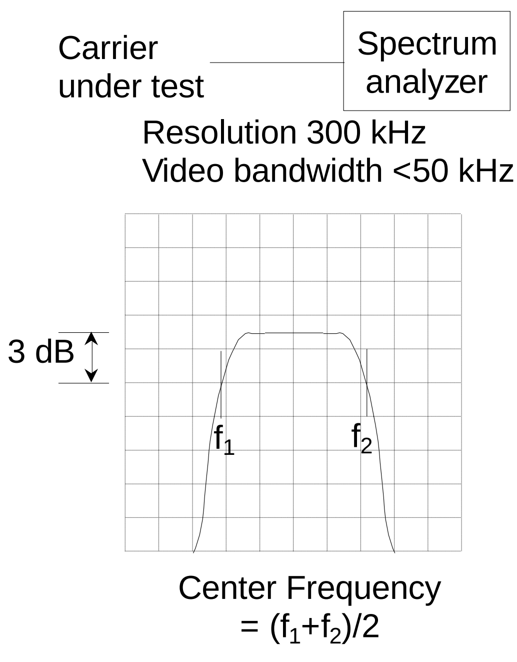


Figure 7-4: Estimating the Center Frequency of a Digital Signal Using the Occupied Bandwidth

For QAM signals (including BPSK and QPSK), the suppressed carrier is at this center frequency. For VSB signals it is 2.69 MHz below the center frequency. For OFDM signals there is no one carrier, but the most appropriate frequency to report is the center frequency.

A number of factors limit the accuracy of this approach, so that it is considered to provide an estimate of center frequency, but it is not accurate enough to provide a true measurement of center frequency.

Chapter 8 Delay

8.1 Propagation Delay

Any signal transmitted from one point to another is delayed because it cannot travel faster than light. In fiber optic cable, the propagation velocity is around 70% that of light in free space. In coaxial cable, the propagation speed is somewhat greater but still lower than the speed of light. This propagation delay is due to the medium in which the signal travels. If the signal travels through several media, the propagation time through each media must be added to obtain the total propagation delay.

Besides propagation delay due to the finite speed at which the signal travels, if a signal is digitized and sent as part of a multiplexed data stream, it will be delayed due to having to be queued into the multiplex of data. Compression and decompression algorithms also contribute to propagation delay.

For many purposes propagation delay is of no consequence; however, for some purposes, propagation delay can be immensely important. The cable engineer should be aware of those circumstances in which propagation delay is important and should be aware of how close the system is to a problem.

Propagation delay due to the medium may be computed from

$$\text{Propagation Delay} = \frac{L}{cV_R}\ \text{seconds,}$$

where

$$L = \text{Length of path (meters or miles)}$$

$$c = \text{speed of light } (3X10^8 \text{ meters per second or } 186{,}000 \text{ miles per second})$$

$$V_R = \text{relative propagation velocity in the medium (decimal fraction).}$$

In certain cases an unusually long path between the headend and a subscriber could be troublesome. For example, the DOCSIS specification calls for a maximum propagation delay of 0.8 ms in each direction, or a total round trip delay of 1.6 ms. which is equivalent to about 168 km of fiber, one way. Few systems are built to be this long, but if you attempt to operate an unusually large loop feeding a node, for example, you might violate this maximum delay. Or if you attempt to connect modems in one city with a CMTS in another, you could have a problem. A similar issue applies to set-top communications circuits. A CMTS manufacturer may choose to implement his system to tolerate longer communications paths, but he is not required to do so.

Another example is that of a voice signal that has been carried as a packetized signal (VoIP). The packetized signal can be delayed significantly in transit. If the delay exceeds 5 ms (as it almost certainly will) echo suppression must be used. This is not a problem to implement and most manufacturers will likely know when they have to implement it. But it is something of which to be aware. Other examples are likely to develop as cable systems add new digital signal processing and signaling processes.

8.2 Group Delay

Definition: Group Delay is the rate of change of phase with respect to frequency and is a measure of the time delay experienced by each frequency component of a signal as it passes through a circuit. Delay inequalities are the result of variations in group delay which occur over the occupied bandwidth of the signal.

Discussion: In well matched or non reactive circuits, phase will vary linearly with frequency resulting in nearly constant group delay and therefore little delay inequality. If delay inequality is very small (significantly less than the symbol length of a digitally modulated signal) over the frequency range spanned by a channel, then all frequency components of the signal will arrive at the same time and the time domain signal will experience little distortion. However, if reactive mismatches occur, such as near the cutoff of a diplex filter or due to the mismatch of a defective component, significant delay inequalities can occur causing some frequency elements of the signal to be delayed relative to others resulting in signal distortion.

The procedure in this section is applicable to return path measurements. For measurements in the forward path, refer to Section 10.2: "Chrominance - Luminance Delay Inequality [FCC §76.605(b)(11)(i)]" or Section 9.3: "Digital Adaptive Equalizer Impulse Response".

Test Procedure

Delay inequalities can be approximated using one of the multipulse 12.5T pulses. The 12.5T pulse that falls closest to the digital signal frequency should be selected. The bandwidth of the 12.5T pulse is approximately 320 kHz. For channel bandwidths greater than 320 kHz it will be necessary to step the modulator frequency over the bandwidth of interest or make estimates from the observations of more than one of the 12.5T pulse frequencies.

Required Equipment

- NTSC multipulse video signal generator
- Reverse modulator
- Reverse demodulator
- Waveform monitor

Test Procedure

1. Connect the equipment as shown in Figure 8-1.
2. Set the modulator's output to the closest T channel for the frequency band of interest.
3. Set the demodulator to the selected channel.
4. Observe the 12.5T pulse on the waveform monitor.
5. Consult Section 10.2: "Chrominance - Luminance Delay Inequality [FCC §76.605(b)(11)(i)]" and/or Section 11.4: "Digital Group Delay" for measurement procedures

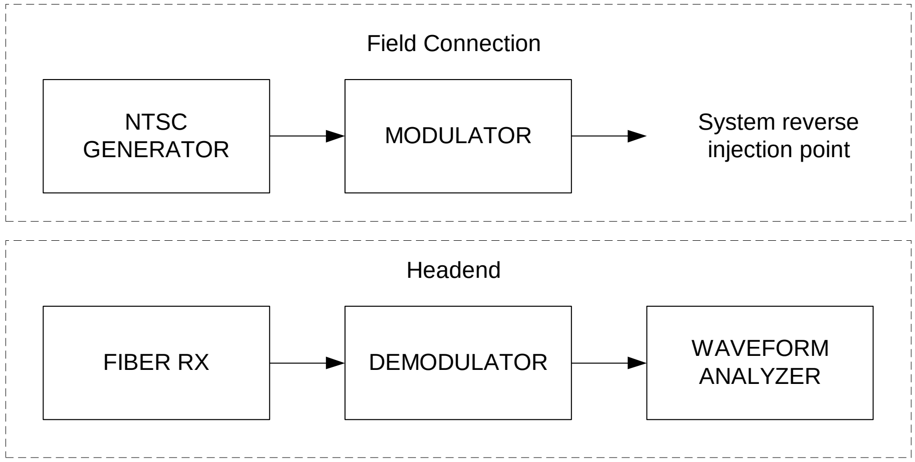


Figure 8-1: Equipment Connection

Chapter 9 Reflections

9.1 Micro-Reflection Overview

The study of transmission lines often begins with the assumption that a signal source, a lossless transmission line, and load have equal impedances. Under that scenario, all of an incident wave's power transmitted from the source is absorbed by the load. In the real world, the impedances of the source, transmission line, and load are seldom, if ever, exactly the same. Furthermore, transmission lines are not lossless, and will attenuate the signals passing through them.

The following definitions may be helpful in understanding impedance mismatches and reflections in cable networks. A signal source can be thought of as a headend modulator's RF output, the downstream output of an amplifier, the upstream output of a cable modem, the output of a tap or splitter, or any other physical device or location in the network we want to define as the origination point of a signal (the signal may actually originate in a different source somewhere else—for instance, a frequency-agile signal generator or analog TV modulator in the headend—but for the purpose of evaluating impedance mismatches and reflections in a network the source can be any convenient device or location). A transmission line in a typical cable system refers to the coaxial cable used transport signals from one point to another. A load can be the input of an amplifier, tap, connector, cable modem or set-top, or even a 75-ohm resistive terminator on a tap spigot.

Consider a situation in which a transmission line and load do not have equal impedances. When this happens we say that an impedance mismatch exists, and some or all of the incident wave will be reflected by the impedance mismatch back toward the source. In a cable network, the nominal impedance is said to be 75 ohms. The key word here is "nominal"—every connector, splice, passive, active, and even the cable itself represents an impedance mismatch of some sort. The question is just how severe is each impedance mismatch?

One common way to characterize the severity of an impedance mismatch is return loss, which is the difference in decibels between the amplitude of the incident and reflected waves. This is described mathematically as $R_{dB} = 10\log(PI/PR)$, where R_{dB} is return loss in decibels, PI is incident power and PR is reflected power, both in watts. If all of the incident power is absorbed by the load, the return loss is infinite and there is no impedance mismatch. If the load is an open circuit, short circuit, or pure reactance, the return loss is 0 dB—that is, 100 percent of the incident wave's power is reflected back toward the source. Typical return loss values for components used in coaxial cable networks vary from a few dB to 30 dB or more.

For more information about the mechanics behind the creation of a reflection at an impedance mismatch—specifically an open or short circuit—see "Impedance Mismatches and Reflections," in the December 2005 issue of Communications Technology magazine. The article also is available on-line at https://www.scte.org/documents/3393/05-12-0120impedance20mismatches20and20reflections.pdf

Where does the term micro-reflection come from and what does it have to do with reflections caused by impedance mismatches? The simple answer is that reflections and micro-reflections are the same thing. The terminology difference is largely related to the reflection's time delay relative to the incident signal. When discussing data transmission over cable networks, we are especially interested in reflections that have very short time delays, on the order of less than a symbol period to perhaps several symbol periods—in other words, close-in or "micro" reflections. Why are micro-reflections important? Simple: A micro-reflection falls into a class of impairments known as linear distortions, and can cause amplitude ripple (standing waves), group delay ripple, inter-symbol interference, and degraded modulation error ratio (MER).

What, then, is meant by a reflection's time delay? One example that may help explain time delay involves ghosting in the picture of an analog TV channel. A trailing-edge ghost, which appears on the right side of the main image, is generally caused by a reflection somewhere in the signal path between the transmitter or modulator and TV set. The ghost is the same image as the main image, but it has been delayed slightly time-wise relative to the main image. Let's say we're looking at a 19-inch diagonal display TV set (Figure 9-1), which, with a 4:3 aspect ratio has a screen height of 11.4 inches and a screen width of 15.2 inches.

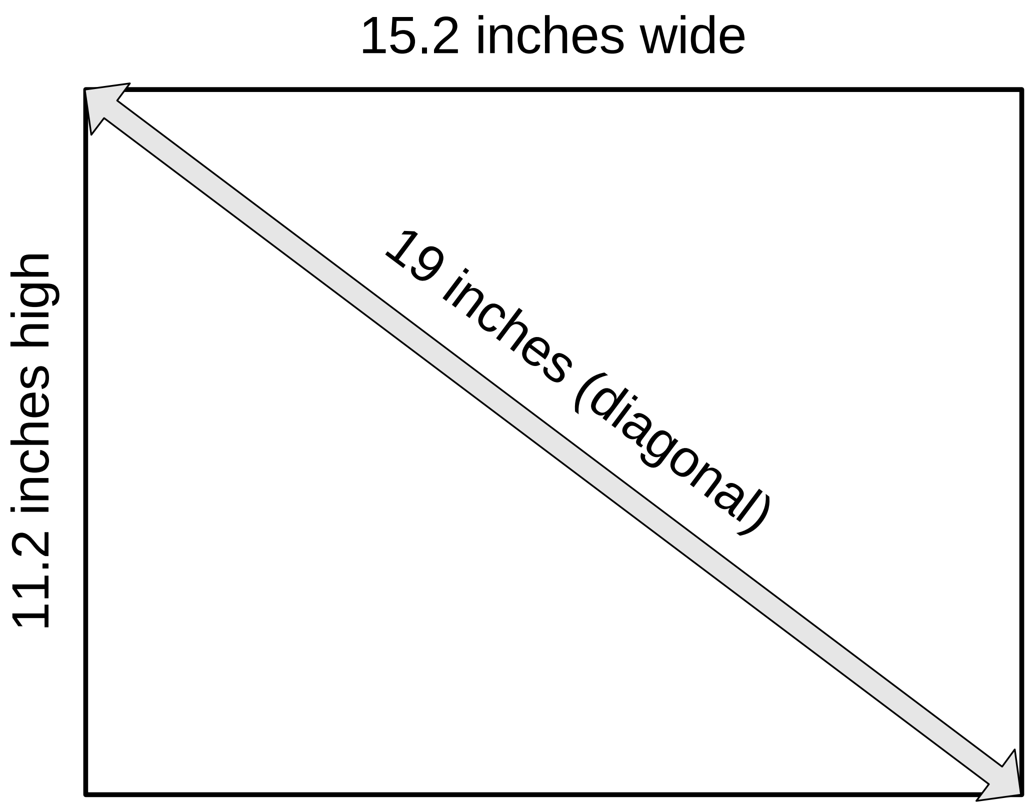


Figure 9-1: Determining the Height And Width Of A 19 Inch Diagonal Television Display With 4:3 Aspect Ratio

Assume a trailing edge ghost is offset to the right of the main image by about three-quarters of an inch, or 5 percent of the TV screen's width. We can estimate the ghost's time delay using that information. In an NTSC picture, each horizontal line takes 63.5556 microseconds (µs) to scan across the TV screen. The horizontal blanking portion of the line is 10.9 µs, leaving 52.66 µs for the active video portion of the line. Assuming the TV set's display shows all of the active portion of each line, 5 percent of 52.66 µs is 2.63 µs. That is, the ghost is delayed slightly relative to the main image; in this example the ghost's time delay is 2.63 µs.

This basic concept of reflections applies to both analog TV channels and digitally modulated signals.

Let’s look at another example, this one with two water-damaged taps separated by 100 feet of coax. **Note:** To simplify the discussion, this example is for a single-frequency. Refer to the Figure 9-2.

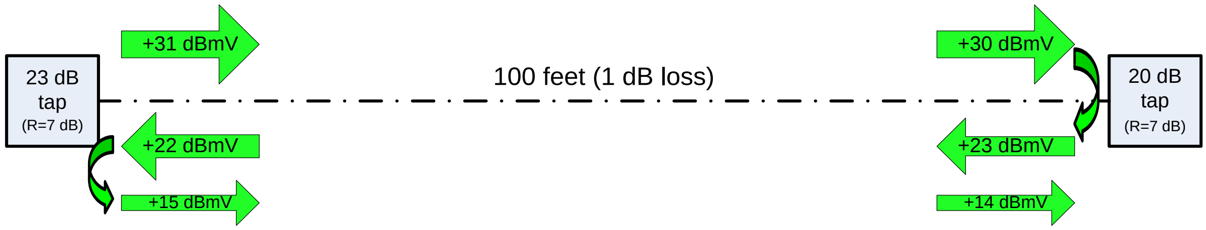


Figure 9-2: Micro-reflection Creation Between Two Water-Damaged Subscriber Taps

Assume that the span of coax has 1 dB of loss, and the water damage has caused both taps’ return loss to degrade from an out-of-the box 15 to 18 dB to just 7 dB. For the sake of discussion, the first tap is a 23 dB 4-port, and the second tap is a 20 dB 4-port. Further assume that an incident signal whose amplitude is +31 dBmV leaves the output connector of the 23 dB tap. When that incident signal reaches the 20 dB tap’s input connector, its amplitude will be +31 dBmV – 1 dB cable loss = +30 dBmV. Most of the incident signal will continue downstream beyond the 20 dB tap, but some of that signal will be reflected by the 7 dB return loss impedance mismatch of the water-damaged 20 dB tap. The amplitude of the reflection will be +30 dBmV – 7 dB return loss = +23 dBmV. That first reflection will travel back towards the 23 dB tap, where its amplitude will be +23 dBmV – 1 dB = +22 dBmV. Here most of the reflection will continue upstream beyond the 23 dB tap, but the 7 dB return loss of the also-water-damaged 23 dB tap will result in the first reflection being re-reflected at an amplitude of +22 dBmV – 7 dB return loss = +15 dBmV. This second reflection will travel toward the 20 dB tap, reaching it at +14 dBmV. And so on.

What will we see at the input connector of the 20 dB tap? The signals of primary concern are the +30 dBmV incident signal and the +14 dBmV reflection. The amplitude difference between the two is +30 dBmV – +14 dBmV = 16 dB. We can also say that the reflection is –16 dBc relative to the incident signal. The –16 dBc reflection will have a time delay relative to the incident signal, defined by the roundtrip propagation time between the 23 dB and 20 dB taps. The cable span between the two taps is the previously mentioned 100 feet. Since it takes RF signals 1.17 nanosecond (ns) to travel through 1 foot of coax that has 87 percent velocity of propagation, the reflection’s roundtrip time is (100 feet + 100 feet) x 1.17 ns = 234 ns. This tells us that the –16 dBc reflection’s time delay relative to the +30 dBmV incident signal at the input to the 20 dB tap is 234 ns. For what it’s worth, a –16 dBc 234 ns reflection is sufficient to be a problem for 256-QAM (quadrature amplitude modulation) operation.

How does the term micro-reflection fit in all of this? As mentioned previously, reflection time delays on the order of less than a symbol period to several symbol periods are of interest with digitally modulated signals. A downstream DOCSIS® 256-QAM signal has a symbol rate of 5.360537 megasymbols per second (Msym/s), so the period of each symbol is the reciprocal of the symbol rate, or 1/5,360,537 = 0.000000186548 second. In this case, reflections with a time delay of somewhat less than 187 ns to a few microseconds can be described as micro-reflections.

9.2 Reflections in the Return Path – Analog Measurement

Definition: Reflections are the result of impedance mismatches between elements in the cable plant such as amplifiers, couplers, splitters and the coax that connects them. When an upstream signal encounters a mismatch, a portion of the signal is reflected back towards the signal's source. The magnitude of the reflected signal depends on the degree of mismatch. This reflected signal is then re-reflected by the mismatch at the source. The re-reflected signal is then traveling upstream where it arrives at the headend as a reduced version of the desired signal that has been delayed in time. Depending on the size of a reflection, it can be a significant source of interference resulting in a reduction in the bit error ratio performance of the system.

Note: Modems with adaptive equalizers can cope with substantially greater reflections with short delay times (usually several symbol times) than they can with delays beyond the equalizer’s limits. Because of this time sensitivity, there is value in knowing reflections in both amplitude and delay time. Modems with adaptive equalizers that can be interrogated as to the equalizer setting can be valuable indicators of reflection problems.

Test Procedure: The following procedures describe two methods for measuring reflections in the return path. Both methods involve injecting an RF pulse into the return path and comparing the amplitudes of the incident pulse to the reflected pulse. Unused spectrum is required in the return path to perform these tests.

PROCEDURE 1 - 2T Pulse

The 2T pulse in the NTSC multipulse test signal may be used to look at reflections with time delays in the range of about .125 to 1.0 µs. The actual maximum delay that can be observed is limited by the time spacing of the multipulses. Some NTSC waveform generators can be set to have only the 2T pulse, which extends the maximum time delay that can be observed to as much as 30 µs.

The method described in this procedure uses simple amplitude detection producing results that are a function of the test frequency and the reflection delay. In order to determine the maximum value of a reflection, the test frequency must be varied until the maximum reflection is observed.

Required Equipment

- NTSC multipurpose video signal generator
- Reverse modulator with 250 kHz or less frequency steps
- Reverse demodulator with 250 kHz or less frequency steps
- Waveform monitor

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Connect the equipment as shown in Figure 9-3.
3. Set the modulator's output to the closest T channel for the frequency band of interest.
4. Set the demodulator to the selected channel.
5. Observe the 2T pulse on the waveform monitor. See Figure 9-4.
6. Set the 2T pulse to 100 IRE.
7. Observe and note the amplitude of the reflections.
8. Change the modulator and demodulator’s frequency by 250 kHz. Note: Finer steps may be used to improve accuracy.
9. Observe and note the amplitude of the reflections.

10. Repeat steps 8 and 9 until the maximum reflection is noted. The frequency of the test signal may need to be moved over several MHz depending on the length of coax involved in the reflection.
11. Note the maximum reflection.

$$\text{Reflections} = 20 * \log\left(\frac{\text{Reflected (IRE)}}{\text{100 (IRE)}}\right) \quad (1)$$

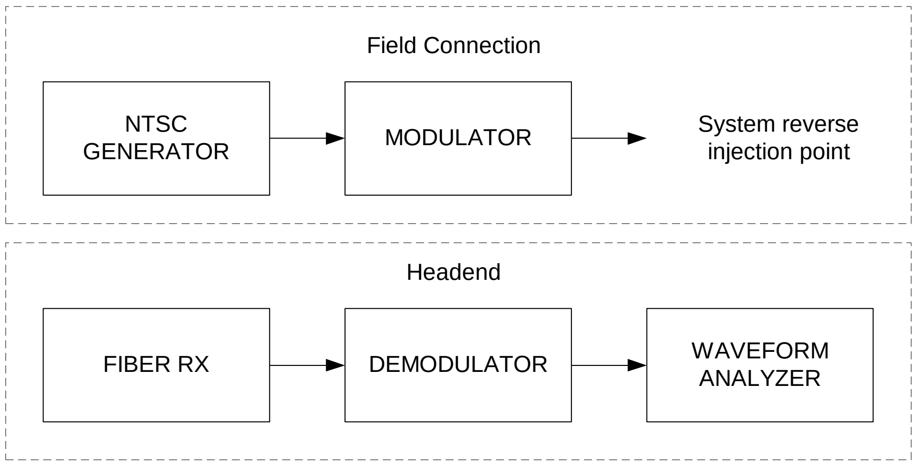


Figure 9-3: Equipment Connection

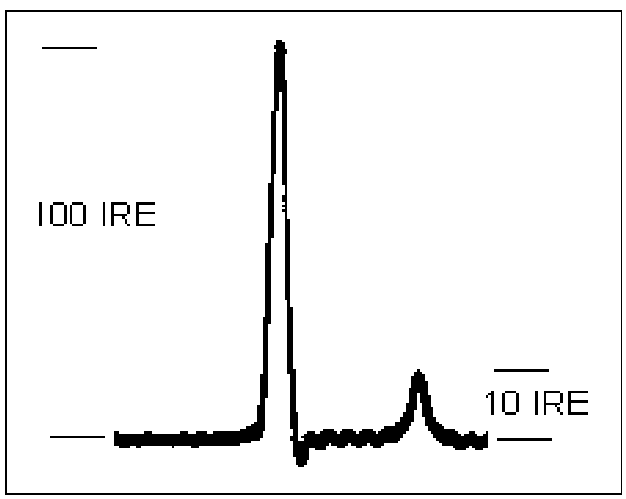


Figure 9-4: 2T Pulse With - 20 dB Reflection

PROCEDURE 2 - Signal Generator Pulse

A signal generator with pulse or square wave modulation capability is injected into the system at the desired point. The pulse is then displayed on a digital storage oscilloscope located at the headend.

The resolution of this test is limited by the fall time of the signal generator pulse. A fall time on the order of 75 ns will resolve reflections involving approximately 25 feet of coax. Longer fall times will result in proportionally less resolution (limited to reflections involving longer runs of coax). In any case, it is desirable to have the fall time less than 1 symbol time for the highest speed data to be carried.

Required Equipment

- RF signal generator with pulse modulation capability having an ON/OFF ratio at least 10 dB greater than the reflection to be measured
- Pulse generator or function generator
- 100 MHz (or greater) digital oscilloscope

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Connect the equipment as shown in Figure 9-5.
3. Set the signal generator's output for the center of the channel of interest.
4. Set the pulse/function generator to produce a 100 kHz square wave.
5. Observe the pulse on the digital oscilloscope.
6. Set the leading edge of the pulse to full scale. Note that reflections delayed by less than the pulse width will cause distortions in the pulse amplitude. It is important to establish the reference amplitude very near to the leading edge of the pulse and observe the variations from this reference. Also note that pulse distortions may be an indication of group delay problems.
7. Use single trigger to capture the leading edge of a pulse near the left edge of the display and adjust the sweep time to place the reflections near the right edge. See Figure 9-6.
8. Observe and note the amplitude of the reflections with respect to the full scale amplitude of the pulse.

$$\text{Reflections} = 20 * \log\left(\frac{\text{Reflection (div)}}{\text{Leading Edge (div)}}\right) \qquad (2)$$

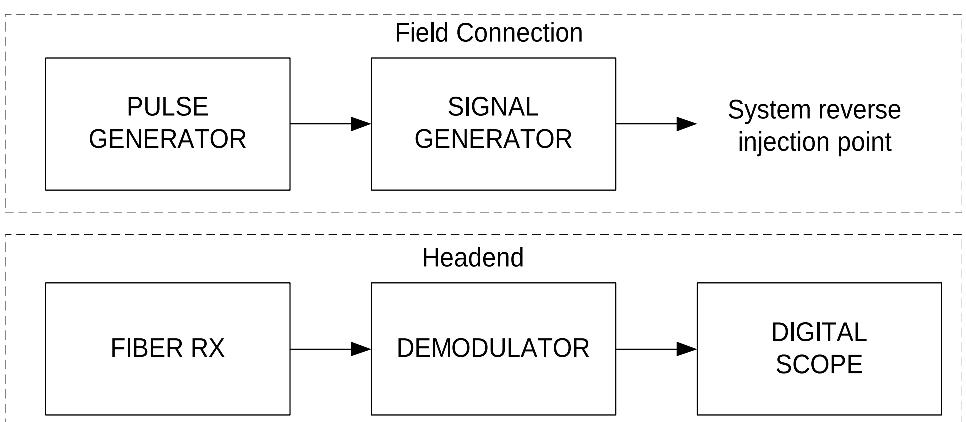


Figure 9-5: Equipment Connection

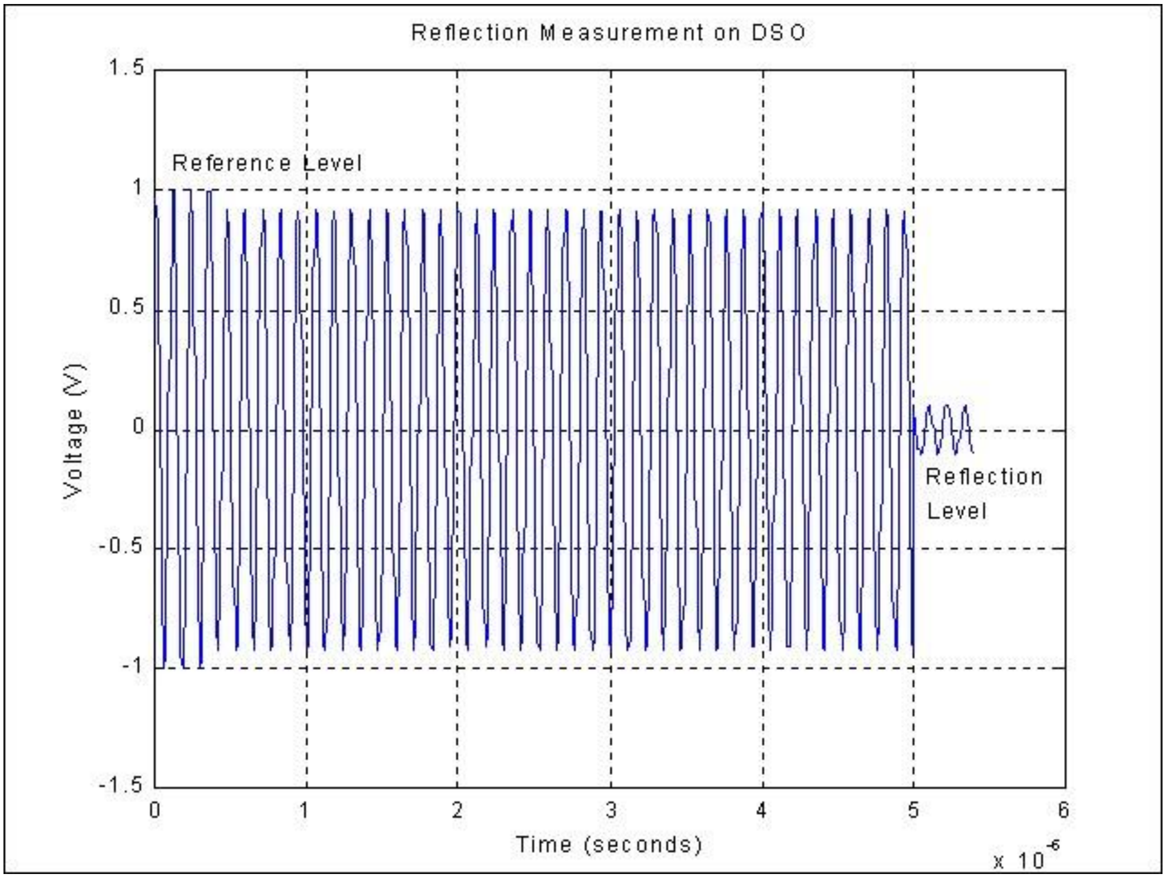


Figure 9-6: Example of -18 dB Reflection

9.3 Digital Adaptive Equalizer Impulse Response

Definition: The impulse response of an adaptive equalizer is defined as the value of the taps and is usually expressed in dB relative to the Main tap.

Discussion: The tap values of the adaptive equalizer, once it has reached convergence, can be used to characterize echoes on the system. Note that the equalizer taps only indicate the presence of reflections that are within the equalizer's range. That is, the equalizer can only report echoes that it is capable of correcting.

If a reflection is within the equalizer's range, the values of the decision feedback equilizer taps will correctly indicate the time delay and magnitude of the post echo. For example, a 64-QAM signal has a symbol period of about 0.2 µs. Therefore, each tap in a T spaced equalizer contributes 0.2 µs. of delay range. With a 16 tap T-spaced equalizer, the maximum echo that could be corrected would be 16 x 0.2 = 3.2 µs. Similarly, a 24 tap equalizer is capable of correcting echoes of up to 4.8 µs. For 256-QAM, the symbol period is 0.187 µs and, therefore, the corresponding echo handling capabilities would be 2.98 µs and 4.46 µs, respectively.

If a reflection is within the equalizer's range, the value of the DFE taps (i.e. – the DFE impulse response) will provide a direct indication of the time delay and magnitude of the post echo. For example, a post echo that is delayed by 1.2 µs will activate the sixth tap. See Figure 9-7. In the case of pre-echoes, the situation is slightly more complex since, in compensating for a single pre-echo, the feed forward equalizer FFE can create additional pre-echoes. For example, a single pre-echo at –0.6

μs at –10 dB will activate the –0.6 μs tap at –10 dB, in turn, creating other echoes at –1.2 μs and –20 dB, -1.8 μs and –30 dB and so on.

Several digital signal analyzers are capable of providing a display of equalizer tap values which may be used as an indication of the presence of reflections on the system. Figure 9-7 shows an example of a typical display for a 16-tap equalizer. The left side of Figure 9-7 illustrates a normal system condition in which the taps gradually decrease in amplitude as the distance from the Main tap increases. The right side of Figure 9-7 shows a condition in which a post echo of about 1.2 μs is present. Note that the sixth tap shows an increase in amplitude. If each division on the amplitude scale represents 10 dB, the echo amplitude would be about –15 dB. In normal operation, one would expect to see a display similar to the one on the left of Figure 9-7. A display such as that shown on the right would indicate a problem on the system or at the subscriber premises and should be cause for concern.

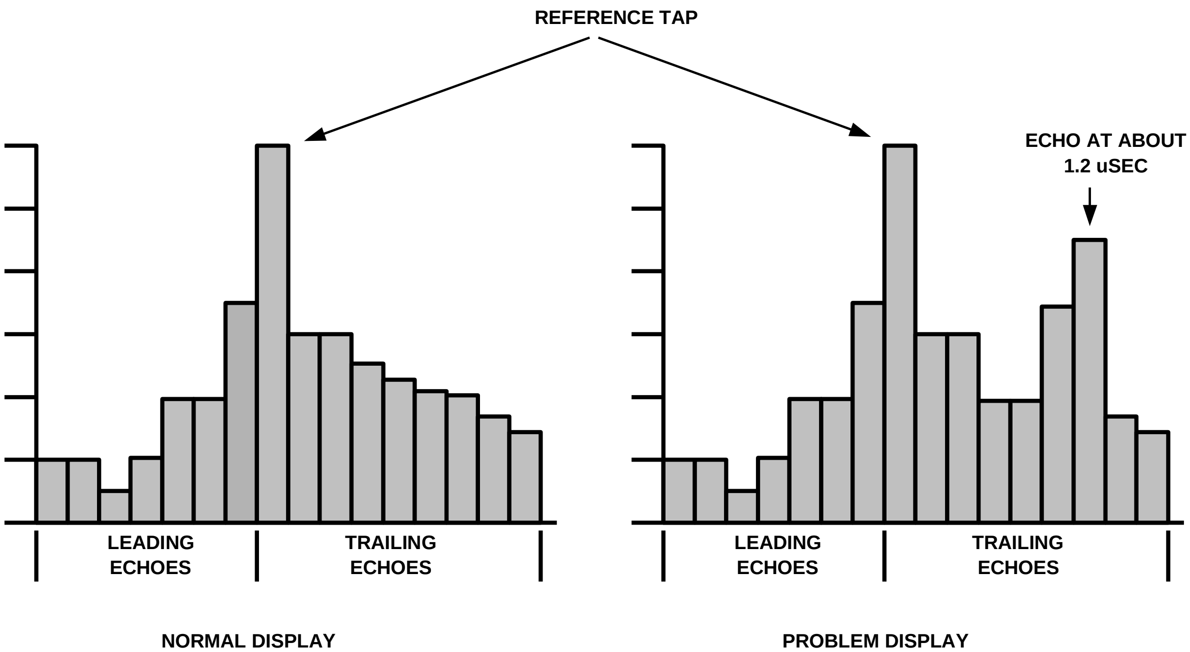


Figure 9-7: Adaptive Equalizer Tap Values Display

9.4 Reflections in Digital Signals

Discussion: Unlike analog transmission, the effect of reflections on a digital signal does not produce visible ghosting in the picture. Instead, the effect is a reduction in the MER due to errors resulting from Intersymbol Interference (ISI). Practical digital demodulators contain adaptive equalizers in order to cancel out the effects of reflections. In normal operation, the reflections on a cable system will have relatively short delays and will be easily compensated by the demodulator’s adaptive equalizer. If a reflection is within the equalizer’s range and is of significant amplitude, its presence can be detected by viewing the equalizer taps. The received signal spectrum can also be used to detect the presence of reflections.

PROCEDURE 1 – (In-service) Using a Digital Signal Analyzer to Detect Reflections

Discussion

Many digital signal analyzers include an adaptive equalizer graph display, sometimes called an equalizer stress display. The adaptive equalizer graph shows the instrument's internal adaptive equalizer taps and the taps' gains and time offsets (delays) relative to the main tap. The adaptive equalizer graph, when available, can be used to provide an approximate indication of the presence and extent of a micro-reflection. Some digital signal analyzers support display of a downstream-only adaptive equalizer graph, and some also support an upstream display. The latter usually requires that a separate continuous QAM test signal be transmitted in the upstream for the duration of the measurement, while the former uses existing downstream QAM signals.

Required Equipment

- Digital signal analyzer with capability to view adaptive equalizer taps
- Continuous QAM test signal source for upstream measurement.

Test Procedure

1. Follow the digital signal analyzer manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Follow the manufacturer's operating instructions to connect the Digital signal analyzer to the RF source of interest, and how to properly configure the instrument to display the adaptive equalizer graph for the specific signal being measured.
3. Identify the main tap on the adaptive equalizer graph display. The main tap is represented by the tallest vertical bar in the display. An adaptive equalizer graph showing a signal with minimal impairments is in the figure below. The main tap is the tall vertical bar left of center, just below "FFE DFE."

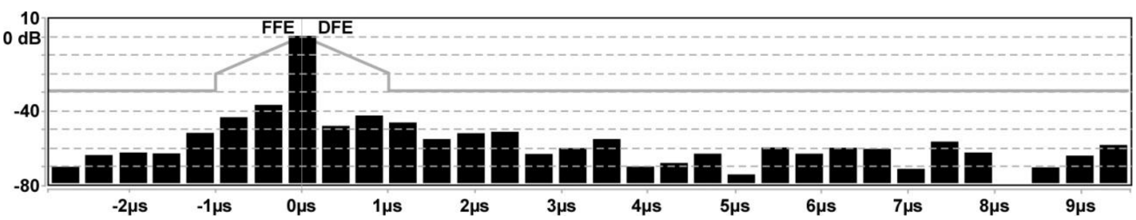


Figure 9-8: Digital Signal Analyzer Adaptive Equalizer Graph

4. Look for a tap or taps sticking up noticeably above other taps on the *right side* of the main tap (ignore the tap adjacent to the main tap). If one or more taller-than-expected taps is seen, that might indicate the presence of a micro-reflection.
5. Depending on the Digital signal analyzer being used, the adaptive equalizer graph may have the vertical axis labeled in decibels (dB), and the horizontal axis labeled in units of time such as microseconds (µs). The approximate amplitude of the micro-reflection relative to the incident signal can be read directly from the display, as the difference in decibels between the main tap and the tap being evaluated. A common way to express that relative amplitude is dBc. The micro-reflection's approximate time delay can be read from the horizontal axis. The example in Figure 9-9 shows an approximately -23 dBc micro-reflection with a time delay of about 2.5 µs.

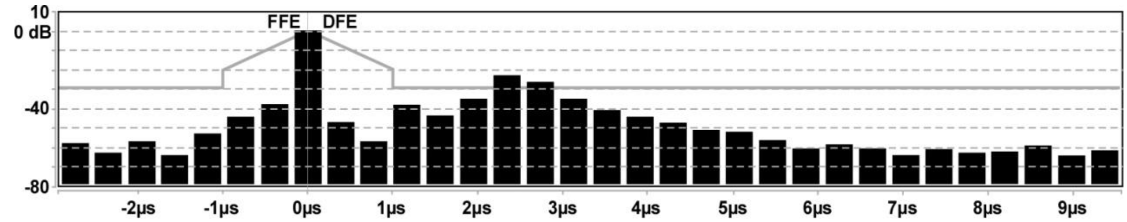


Figure 9-9: Micro-reflection Example

Figure 9-10 shows another example of a micro-reflection, this one approximately -34 dBc at just over 1 µs time delay.

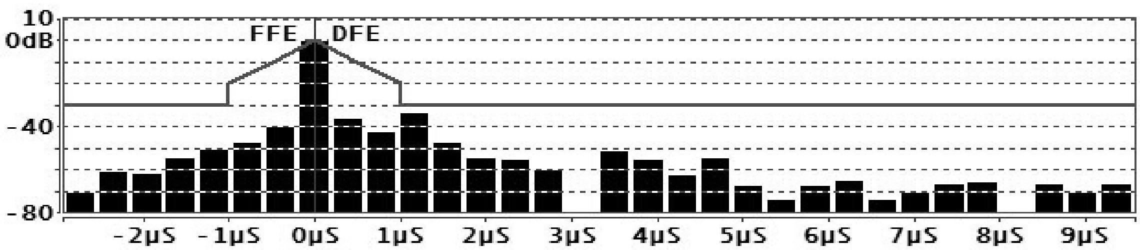


Figure 9-10: Another Micro-Reflection Example

Notes, Hints, and Precautions

A Digital signal analyzer's adaptive equalizer graph is derived from the instrument's adaptive equalizer coefficients. Resolution and accuracy are limited in part by the number of taps in the adaptive equalizer, the equalizer's tap spacing, and the type and severity of impairment(s) present in the signal path. Using an adaptive equalizer graph to characterize a micro-reflection is most effective when only one significant micro-reflection exists. If other channel response impairments such as in-channel tilt and/or rolloff are present, it may be impossible to determine if a micro-reflection is present using this method.

PROCEDURE 2 - Using a Spectrum Analyzer to Detect Reflections

Discussion: If a long reflection is present (i.e. – one that is delayed beyond the equalizer's time range), a spectrum analyzer can be used to measure the amplitude and delay of the reflection. A digital signal, after averaging, should show a flat spectrum. However, the presence of reflections causes constructive and destructive interference between the direct and reflected paths. This produces ripples in the digital spectrum. The peak-to-peak delay and peak to trough amplitude of the ripples can be used to determine the delay and amplitude of a reflection.

Required Equipment

- Spectrum analyzer with 75 Ω input and Marker Delta capability.

Test Procedure

1. Follow the digital signal analyzer manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Set the spectrum analyzer to display the digital signal. Decrease the frequency span as necessary to see if any ripples are present. In most cases, it will probably also be necessary to set the amplitude scale to 1 dB/division in order to get a good view of the ripple pattern. Signal averaging should be turned ON until display stability is reached.

3. Set the marker to the peak of one of the ripples. Activate the Marker Delta function and set the second marker to the next peak. Read and record the Marker Delta frequency.

4. Move the second marker to the trough of the ripple pattern. Read and record the Marker Delta amplitude in dB.

5. Calculate the delay of the reflection as follows:

$$\text{Delay } (\mu\text{s}) = \frac{1}{\text{Marker Delta (MHz)}}$$

6. Calculate the amplitude of the reflection in dB as follows:

$$\text{Amplitude (dB)} = 20 * \log\left(\frac{10^{\left(\frac{\text{Marker Delta (dB)}}{20}\right)} - 1}{10^{\left(\frac{\text{Marker Delta (dB)}}{20}\right)} + 1}\right)$$

Figure 9-11 shows an example of a spectrum analyzer display of a single reflection at 4 μs. and –10 dB.

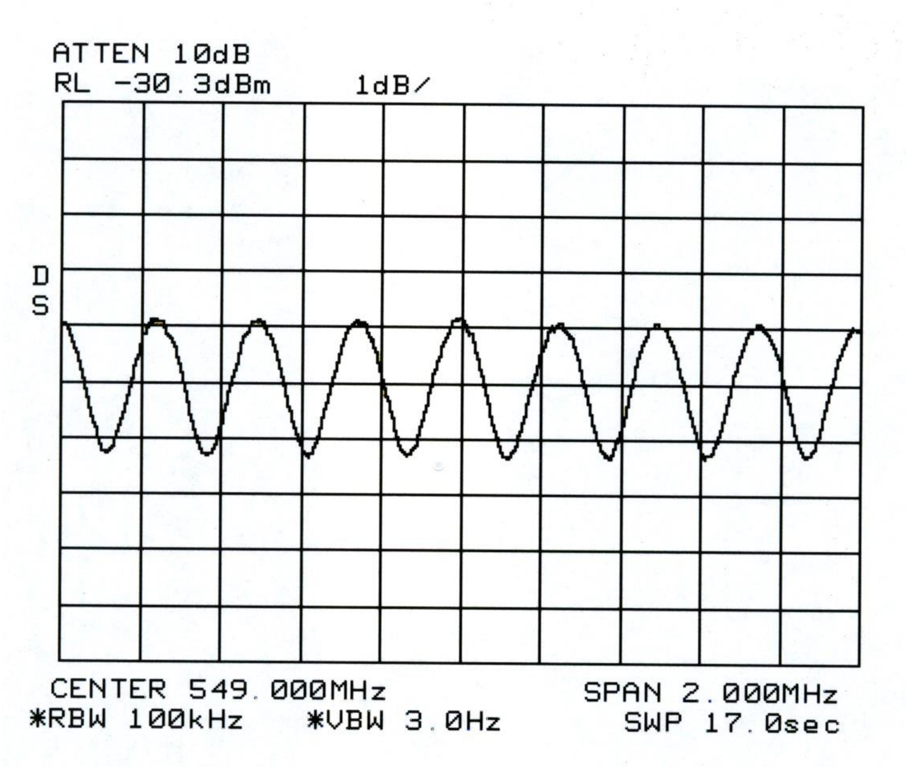


Figure 9-11: Spectrum of a Single Reflection at 4 µs. and -10 dB

9.5 Fault Location Using a Sweep System

Discussion: Broadband sweep systems are commonly used for testing of both the forward and return path frequency response of the cable television distribution system. Discontinuities in the impedance of the network, which may be caused by cracked cable, open terminator, etc., cause reflections in the signal. The voltage due to these reflected waves will add to the incident wave at some points along the transmission line and subtract at others. This effect creates a ripple frequency in the swept frequency response of the network, and it has been proven that the relationship of the distance to the impedance discontinuity to the ripple frequency is:

$$\text{Distance to Fault (feet)} = \frac{492 * V_P}{\text{Freq}} \qquad (1)$$

where:

V_P = Velocity of Propagation of the cable (expressed in decimal form)

Freq = Spacing of ripple peaks in MHz

By measuring the ripple frequency in the swept frequency response, the distance to the fault can be calculated.

Note: It is not the intent of this procedure to discuss the methods used for sweep response testing. This is covered in Section 6.2: “Frequency Response - Forward”. The emphasis in this section is to discuss how the swept frequency response may be used to identify the source of reflections.

An example of a normal sweep response for a 750 MHz system is shown in Figure 9-12.

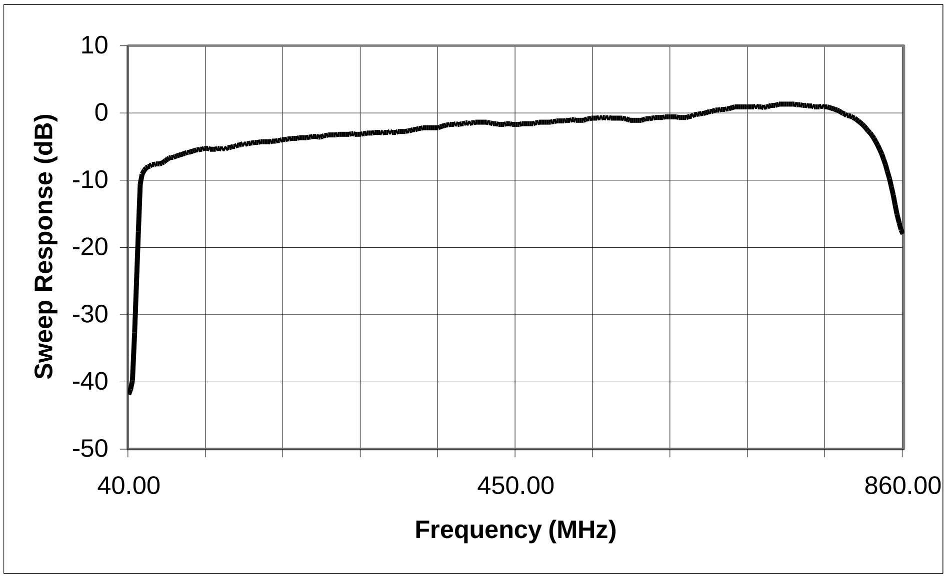


Figure 9-12: Sweep Response Without Reflections

Figure 9-13 is an example of a sweep response with ripples from a reflection present. By measuring the spacing of the ripple troughs, an approximation of the distance from the sweep analyzer to the impedance discontinuity can be calculated using equation (1).

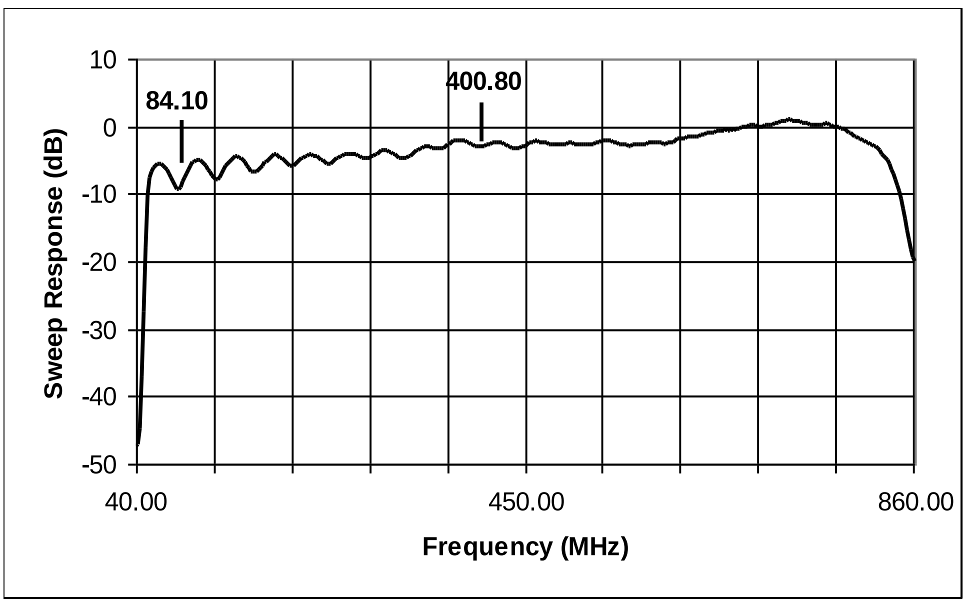


Figure 9-13: Sweep Response With Reflections

In order to get the best accuracy, you should measure the frequency span between a series of ripple troughs and divide by the number of ripples. Using the example in Figure 9-13, the frequency span between 8 ripple troughs is (400.80 - 84.10 MHz) or 316.70 MHz. The frequency span for a single ripple is (316.70 / 8) or 39.59 MHz. Assuming a normal hard-line coax V_P of 87%, the distance to the fault in our example is:

$$\text{Distance to Fault} = \frac{492 * 0.87}{39.59} = 10.8 \text{ feet}$$

Notes, Hints and Precautions

1) The peak-to-peak amplitude of the ripples will decrease with higher frequency due to the increasing loss vs. increasing frequency characteristic of coaxial cable.
2) If multiple discontinuities are present, the ripple frequency present may be the result of the dominant one or possibly the distance between two of the discontinuities.
3) Passive components, such as taps or directional couplers, may have poor enough return loss, relative to the return loss of the cable, to cause this type of reflection.
4) It is possible that amplifier problems may cause this type of response ripple, in which case the peak-to-peak amplitude of the ripples will typically remain constant with increasing frequency.

Aliasing from Digital Sampling

Frequency response sweep systems in use today use digital sampling techniques which limit the range of this measurement. In order to accurately determine the ripple frequency, the maximum frequency between samples of the swept frequency response must be < 50% of the ripple frequency being measured. In the example used above, the swept frequency response consists of 401 sample points. Therefore, the spacing between frequency samples is:

$$\text{Freq}_{\text{SAMPLE}} = \frac{(860 - 40)}{400} = 2.05 \text{ MHz}$$

The minimum ripple frequency which can be measured using this configuration is 4.1 MHz and therefore the maximum distance which can be measured is:

$$\text{Maximum Distance to Fault} = \frac{492 * 0.87}{4.1} = 104.4 \text{ feet}$$

Faults which are farther than 104 feet away using this sweep configuration will cause aliasing in the ripple frequency and actually appear to be closer. Quite often forward path sweep response test systems are configured to take one sample every 6 MHz. In this configuration, the minimum ripple frequency that can be measured is 12.0 MHz and the maximum distance that can be measured is:

$$\text{Maximum Distance to Fault} = \frac{492 * 0.87}{12.0} = 35.7 \text{ feet}$$

It is important that the user is familiar with the frequency resolution being used and the limitations of the measurement when using swept frequency response testing to identify faults.

Chapter 10 Analog Video

10.1 Depth of Modulation

Definition: Percent modulation of a video carrier using standard (type A5C negative) television amplitude modulation is the difference between the maximum and the minimum of the RF envelope amplitude expressed as a percentage of the maximum RF envelope level. The maximum RF envelope amplitude is constant and occurs during the synchronization pulses in the A5C modulation. The minimum RF envelope amplitude, therefore, corresponds to the whitest part of the picture.

FCC NOTE: There is no FCC requirement for Depth of Modulation. However, this measurement procedure is presented here because this setting can have a major effect on at least two of the fcc required colored tests.

Discussion: This setting can have a major effect on at least two of the FCC required color tests. The differential gain and differential phase tests {§76.605 (b)(11)(ii) and §76.605 (b)(11)(iii)} can only be properly performed if the color subcarrier is not clipped at all. Many modulators contain circuits to clip any portion of the video signal that would otherwise cause over modulation. If the depth of modulation is set even a small amount high, this clipper will cause the measurement of these color distortions to exceed the FCC requirements. Some modulators have lights to indicate over modulation. These lights generally are not sensitive enough to prevent the color distortions. Great care must be used in the setting of depth of modulation to eliminate detrimental effects in performing color tests.

Standard broadcast modulation depth is 87.5%. If deviations from this level occur due to measurement or other error, it is certainly preferable to err on the low side. Excessive modulation depth, even by small amounts, may cause video distortions, white clipping, and sound buzz.

The depth of modulation should also not be set too low. This can cause a decrease in signal-to-noise ratio and particularly an appearance of low brightness in the picture (a dark or dim picture).

The first procedure described presents a modulation measurement method which employs an RF Spectrum analyzer as the RF envelope demodulating device. This method should have wide application since virtually all system operators have access to these instruments. The second procedure uses a precision demodulator which is also internally calibrated. If the appropriate equipment is available, the second procedure is the preferred approach.

PROCEDURE 1 - Spectrum Analyzer

This adjustment requires that some peak white be a part of the video being measured. If it is not otherwise guaranteed that peak white is present, then an appropriate VITS test signal (such as the FCC combination or composite) can be used. Both of these contain a "white flag" that will show up as the lowest carrier amplitude. **Note**: Do not use the NTC-7 composite signal because the chrominance level exceeds the peak white level.

Required Equipment

- An RF spectrum analyzer with at least 300 kHz IF bandwidth, a display with a linear or calibrated decibel vertical axis, and zero span capability.

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Connect the RF or IF output signal of the modulator under test to the analyzer's RF input. Use the following analyzer settings:
   - Sweep time: 2 ms/div or adjust to display at least one complete picture field
   - IF Resolution Bandwidth: maximum, but at least 300 kHz
   - Frequency Span: zero (wider spans can be used initially for centering the carrier)
   - Video Field: If Video Triggering options are not provided, AC line triggering maybe used.
   - Video filtering: none

3. Tune the analyzer to the picture carrier and carefully fine tune for maximum display peak (positive) amplitude. Adjust the reference level for a nearly full-scale display.

4. At this point either the linear or log display modes of the analyzer can be used. If the log mode is used, the modulation percent is read directly as the decibel difference between the peak (sync tip or top of the video waveform) and the minimum of the video waveform. Use Table 10-1 to convert the dB ratio to percentage:

Table 10-1: Converting dB Ratio to Percent Modulation

Ratio	Percent Modulation
10 dB	68%
12 dB	75%
14 dB	80%
16 dB	84%
18 dB	87.5% (standard video percent modulation)
20 dB	90%

5. Adjust the modulator to give the dB ratio corresponding to the desired modulation depth.

6. If the linear analyzer display is used, adjust the sensitivity controls to place the waveform peaks (sync tips) on the highest line of the graticule. Since the bottom edge of the graticule is zero level, the display is now calibrated for 0% modulation (top) to 100% modulation (bottom). For example: If, with an 8 cm. vertical display, the minimum level is 1 cm. then

$$\text{Percent Modulation} = \left(\frac{8-1}{8}\right)\times 100\% = 87.5\%$$

Adjust the modulation depth to place the waveform minimums at the level which corresponds to the desired percent modulation.

PROCEDURE 2 - Demodulator and Waveform Monitor

This method uses a precision demodulator with a "zero carrier reference" feature and a waveform monitor. It also requires that peak white be present in the video being applied to the modulator under test. This is most easily done if a VITS signal is used which has a reference white bar. Both the commonly used FCC Composite and FCC Combination test signals have reference white bars.

Required Equipment

- A precision demodulator
- A waveform monitor
- Variable attenuator
- Spectrum analyzer

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. If the modulator has an RF or IF test port, connect the test equipment to this port as shown in Figure 10-1. Otherwise, it may be necessary to take the modulator off line or split the RF output of the modulator and connect one side of the split to the test setup. Note that some precision demodulators can be damaged by application of an excessively high signal input. Before connecting the modulator to the test equipment, set the variable attenuator to its maximum value. Also note that some video test equipment may have loop-through inputs. If this is the case, terminate the loop-through connection in 75 Ω.

3. Tune the demodulator to the channel to be measured. Adjust the attenuator until the spectrum analyzer shows that the input signal to the demodulator is at the correct level specified by the manufacturer. Set the zero-carrier reference pulse from the demodulator to the zero carrier (120 IRE) reference on the waveform monitor's vertical scale.

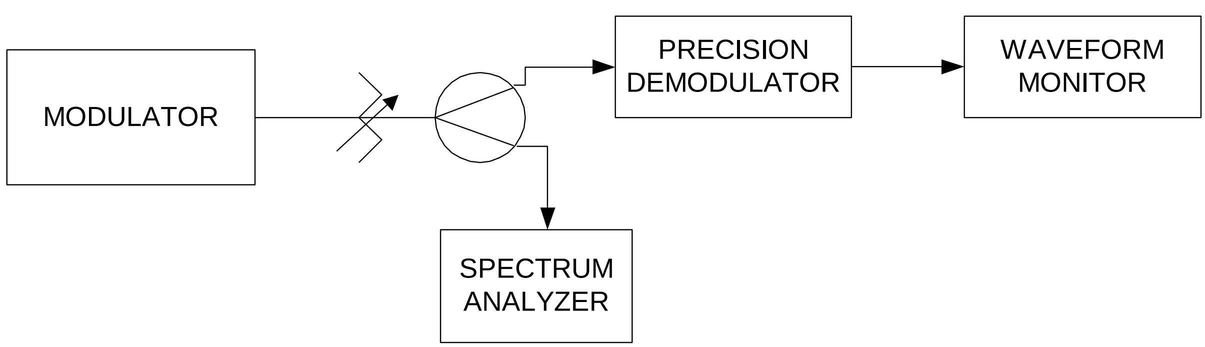


Figure 10-1: Procedure 2 Test Setup

4. Set the waveform monitor to display the VITS signal. Adjust the percent modulation until the white level is at 100 IRE. The sync tip should be at -40 IRE. If the sync tip amplitude does not extend to -40 IRE and the white level is at 100 IRE, do not readjust the depth of modulation to increase the sync tip amplitude to its correct level as this may result in clipping of the peak amplitudes.

10.2 Chrominance - Luminance Delay Inequality [FCC §76.605(b)(11)(i)]

Definition: The Chrominance-Luminance delay inequality caused by a headend system or component is defined as the change in delay time of the chrominance component of the signal relative to the luminance component after passing through this system. This parameter is also called Chroma Delay.

The following requirements apply to the performance of the cable television system as measured at the output of the modulating or processing equipment (generally the headend) of the system:

FCC §76.605(b)(11)(i): ***The chrominance-luminance delay inequality (or chroma delay), which is the change in delay time of the chrominance component of the signal relative to the luminance component, shall be within 170 nanoseconds.***

Discussion: Distortion occurs whenever a delay is introduced in one component of the picture signal without an equal delay being introduced in the other. The result is misregistration of the color and black and white parts of the picture.

A specialized video test waveform, the modulated 12.5T pulse, which is part of the FCC Composite test signal, is used to measure this delay. Where the input or output terminal of the headend system accepts or delivers a modulated RF (channel) signal a video modulator and/or demodulator must be included in the test equipment setup.

Required Equipment

- An NTSC video test waveform generator, which provides the FCC Composite test signal waveform. This generator is not needed if the channel to be tested already contains a suitable VITS.
- A video waveform test set
- A variable 75 Ω attenuator (two required for processor test)
- Video modulator and/or demodulator as required with appropriate channelization

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Use the appropriate setup of Figure 10-2. The headend channel to be tested may use a demodulator, a modulator, or an RF processor. The actual headend connections should be the RF input (or video input for a modulator) and the combined headend RF output (or video output for a demodulator).

   For a channel which uses demodulator--modulator RF processing, the units can be reconnected as in Figure 10-2 (a) and measured as a pair.

3. Set the video waveform generator to deliver an FCC Composite signal and adjust the modulator for the normal depth of modulation. This can be done as an in-service test using the VITS inserted into the regular programming instead of the full-field test signal.

4. With the demodulator set for normal video output level, adjust the waveform monitor to display the 12.5T pulse as in Figure 10-3. Select the Chroma/Luma delay measurement mode on the waveform measurement equipment. Note: most modern video waveform measurement sets will provide both chroma/luma delay and chroma/luma gain inequality in a single measurement. However, there is no FCC requirement on gain inequality.

5. If the test equipment is capable of signal averaging, select a minimum of 32 averages.
6. Follow the instructions in the operator's manual for completing the measurement. A typical display of chroma/luma delay is shown in Figure 10-4.

To determine the delay of the headend equipment alone, it is necessary to subtract the delay contributed by the test modulator and/or demodulator. In the absence of specific data on the test units, it will be necessary to assume that the test unit contributes delay equal to its maximum specification. Where a modulator-demodulator pair are used to test a processor channel, the mod-demod combination can be tested alone (with appropriate channelization) and its delay value subtracted from the total.

If this test is to be performed using VITS inserted into programming before arrival at the cable television facility, then the demodulator and/or waveform monitor can be used to measure the Chroma Delay present in the incoming signal. This can then be subtracted from the measurement made at the output of the modulating or processing equipment to arrive at the delay contributed by the cable system.

For example:

> If the incoming signal exhibits a ***delay*** of 50 nanoseconds, and the measurement after the headend shows a ***delay*** of 94 nanoseconds, then the incoming delay can be ***subtracted*** from the measured 94 to give a system delay of 44 nanoseconds.
>
> $$94 - 50 = 44$$
>
> In the same fashion, if the incoming signal exhibited an ***advance*** of 35 nanoseconds, and the measurement after the headend shows the same 94 nanoseconds, then the incoming ***advance*** must be ***added*** to the measurement for a system delay of 129 nanoseconds.
>
> $$94 + 35 = 129$$

NTSC Pre-Correction

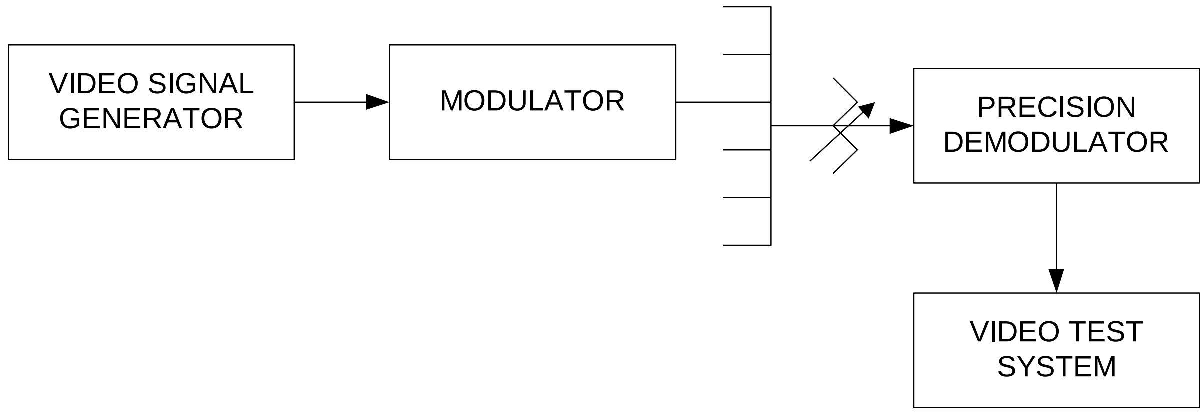
The System M broadcast transmission standard (which is used to transmit NTSC Video) was designed to accommodate the components and circuits available at the time. The filters (inside television receivers) that trap out the aural carrier from the picture could only be made from discrete components. These filters introduced a delay in the color subcarrier of 170 nanoseconds.

In order to compensate for this delay, the transmitters were required to add a phase "boost" or pre-correction of 170 nanoseconds.

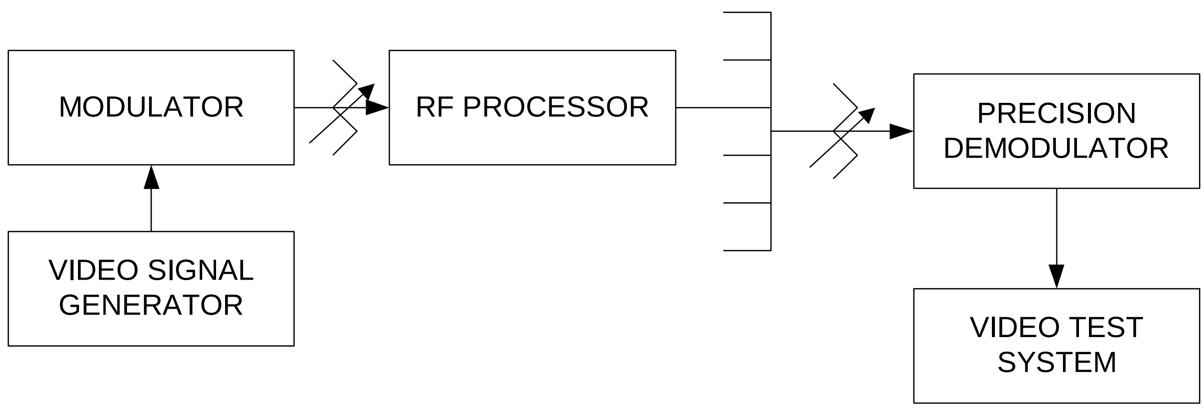
The measurements being made here all assume that this pre-correction is the reference for the delay being measured. Therefore, the modulator and demodulator must both use the standard NTSC 170 nanosecond pre-correction curve.

This means that the "sound trap" must be turned ON if the demodulator provides a choice. Also, the modulator must have its pre-correction turned ON if it is switchable.

It is only coincidental that the FCC specification is also 170 nanoseconds.



(a) Setup for a modulator channel



(b) Setup for an RF processor channel

Figure 10-2: Setups for Chrominance—(a) and (b) Luminance Delay Inequality

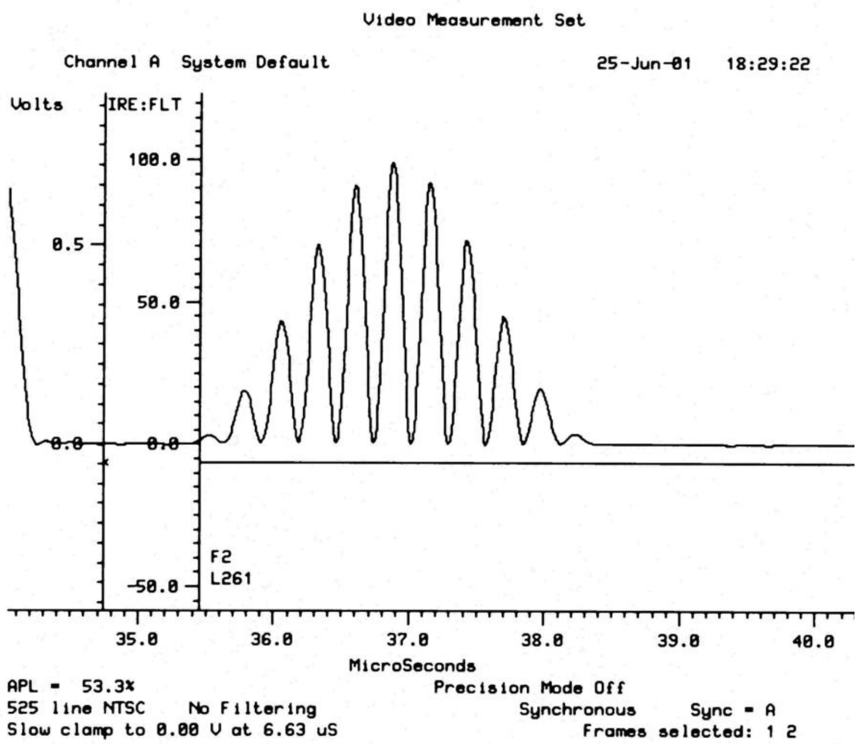


Figure 10-3: 12.5T Pulse

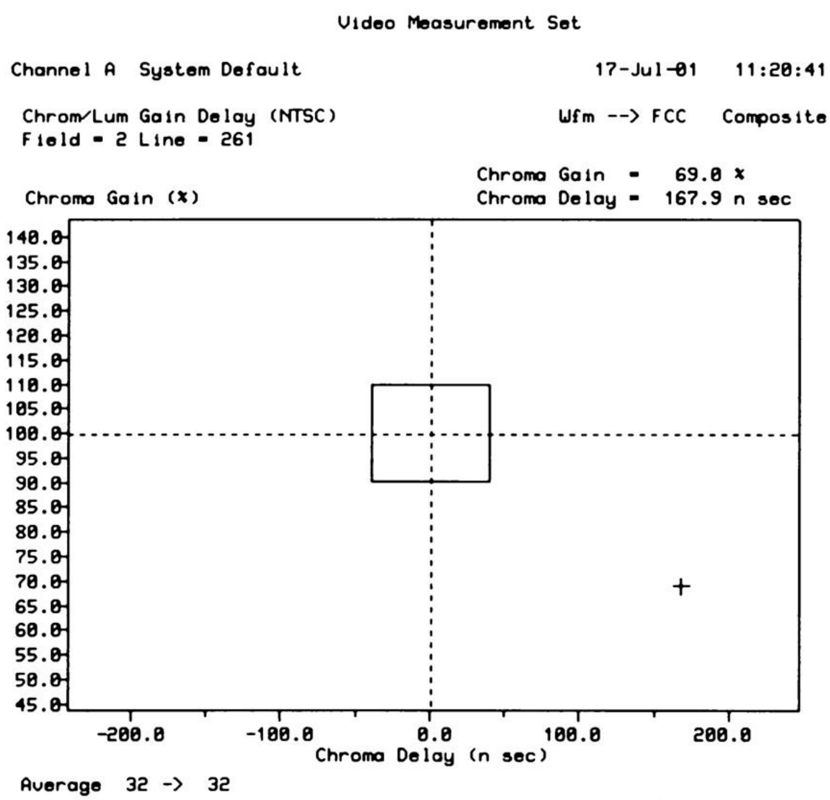


Figure 10-4: Chrominance-Luminance Delay Example

10.3 Differential Gain and Phase [FCC §76.605(b)(11)(ii), (iii)]

Modern video test systems are capable of measuring both differential gain and differential phase in a single measurement. Therefore, these measurements will be treated as a single test with specifics for each measurement described in the following subsections.

Note: There are two types of modulated staircase signals available for these tests. These are the FCC type, and the NTC-7 type. The FCC type should be used. The NTC-7 tests are intended for baseband video transmission, and contain subcarrier levels that exceed 100 IRE (whiter than white). When using a video test signal generator as shown in Figure 10-5, the FCC type should be used. For satellite signals, the NTC-7 VITS will probably exist. In either case, if the modulator under test contains a clipping circuit (90% modulation) ignore the fifth step in a 5-step signal or the ninth and tenth steps in a 10-step signal.

If an automated measuring system is used for these tests, it may require a very specific test signal to be used. The operating manual for the measuring system in use will give specific instructions.

The demodulator used for this test must have sufficient accuracy to ensure correct measurement. A precision demodulator should be used.

Differential Gain Definition: Differential gain is the change in amplitude of the chrominance signals as a function of the amplitude of the associated luminance signal. It is measured as the difference in amplitude between the largest and smallest segments of the chrominance signal, divided by the largest and expressed in percent. It may also be measured as a ratio, expressed in dB, of the largest segment of the chrominance signal to the smallest.

The following requirements apply to the performance of the cable television system as measured at the output of the modulating or processing equipment (generally the headend) of the system:

FCC §76.605(b)(11)(ii): ***The differential gain for the color subcarrier of the television signal, which is measured as the difference in amplitude between the largest and smallest segments of the chrominance signal (divided by the largest and expressed in percent), shall not exceed ±20%.***

Differential Phase Definition: Differential phase is the change in phase of the chrominance signal as a function of the amplitude of the associated luminance signal. It is measured as the greatest phase difference in degrees between each segment of the chrominance signal and the reference segment, which is the segment at the pedestal level (0 IRE).

The following requirements apply to the performance of the cable television system as measured at the output of the modulating or processing equipment (generally the headend) of the system:

FCC §76.605(b)(11)(iii): ***The differential phase for the color subcarrier of the television signal which is measured as the largest phase difference in degrees between each segment of the chrominance signal and reference segment (the segment at the blanking level of 0 IRE), shall not exceed ±10 degrees.***

Discussion: As indicated by these definitions, differential gain and phase are distortions that only have meaning when referring to color signals. Differential gain manifests itself as a change in the saturation or intensity of a color portion of the picture when only the brightness of that area changes. Similarly, differential phase manifests itself as a change in hue under the same conditions. These distortions occur most commonly in TV modulators and demodulators during the modulation and detection process. They may occur in heterodyne processors and strip amps, but generally, to a lesser degree.

To measure these distortions, the test signal must be in baseband video format. This means that to measure the differential gain and differential phase of a modulator, a precision demodulator is required. Likewise, to measure these distortions in a demodulator, a modulator is necessary. To make the measurement in a heterodyne processor or strip amp, both a modulator and a demodulator are needed.

The test signal most commonly used to make the measurements is a 5 or 10 riser stairstep with the steps varying between the blanking level and reference white (0-90 IRE units). Superimposed on these steps will be a chrominance signal of 20 IRE units peak-to-peak. If a clipper circuit is contained in the modulator under test to prevent modulation beyond 90%, then ignore information on the 5th step in a five-step signal or the 9th and 10th steps in a 10-step signal.

Required Equipment

- A video test signal generator capable of producing a modulated signal (FCC composite is recommended).

Note: This generator is not needed if the channel to be tested already contains a suitable VITS.

- A video test set with an internal chrominance bandpass filter
- A test modulator
- A precision demodulator

Differential phase measurements made with a demodulator that has both envelope and synchronous detection processes may yield widely varying results. Envelope detection is preferable for differential gain and phase measurements and adjustments.

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Equipment connections are similar to those shown in Figure 10-2. Connect the output of the video signal generator to the input of the modulator (normal video interface levels are 1 volt p-p) and select the FCC Composite signal as the test signal. This can be done as an in-service test using the VITS inserted into the regular programming instead of the full-field test signal.
3. Adjust the output level of the modulator to be within the unit's specified operating limits and with the use of a pad, ensure that the input levels to the demodulator are within that unit's specified limits (if a processor strip amp is being tested, check to ensure that similar conditions exist for its input and output levels).
4. Set the video test set to display the FCC Composite waveform. Adjust the Video Modulation control on the modulator to place the white level at 100 IRE. Verify that the subcarrier at the top of the stairstep is not clipped (see Figure 10-5). **<u>This setting is quite critical.</u>** See discussion in Section 10.1: “Depth of Modulation”.
5. Select the Differential Gain/Differential Phase test.
6. If the test equipment is capable of signal averaging, select a minimum of 32 averages.

7. Follow the instructions in the test equipment operator’s manual for completing the measurement. An example of a differential gain and phase measurement is shown in Figure 10-6.

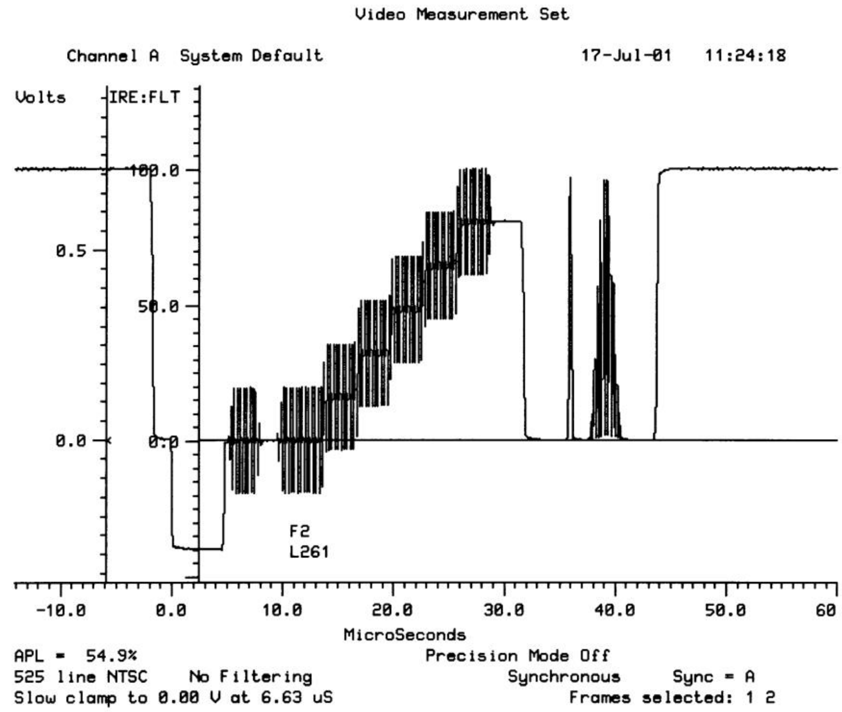


Figure 10-5: FCC Composite Test Signal

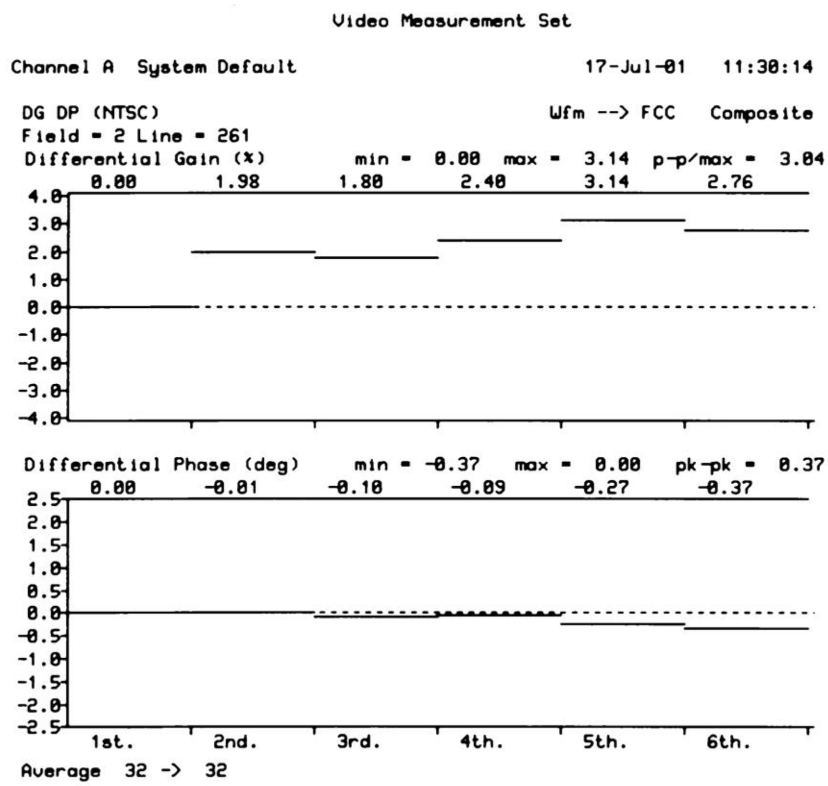


Figure 10-6: Example of Differential Gain and Phase Measurement

Notes, Hints and Precautions

It is desirable when making the differential gain measurements to vary the APL (average picture level) of the test signal from 10% to 90%. The worst-case measurement is then taken to be the differential gain of the unit under test.

As with differential gain, differential phase measurements should be made with the APL of the test signal varied from 10% to 90%. The worst-case measurement is then taken to be the differential phase of the unit under test.

It may also be useful to make the above tests under all conditions of specified input and output levels of the device under test.

It is obvious from the test setup that the differential gain and phase measured is not due only to the device under test. Whenever possible it would be desirable to have the differential gain and phase specifications for the test modulator or demodulator verified by another source.

In any event, if the resulting measurement indicates a low differential gain distortion level, say less than 2%, it is relatively safe to assume that the unit under test has a distortion level of less than 2%. Although possible, it is unlikely in the above situation that the differential gain characteristics of the devices involved would complement each other to the extent that such a low distortion is produced.

On the other extreme, when a medium to high differential gain distortion level is measured, say 20% to 30%, or greater, it is extremely difficult to determine the contribution of the individual units, unless some additional knowledge is obtained about one or more of the units.

When a processor or strip amplifier distortion is being measured, a careful observation of the test modulator--test demodulator back-to-back differential gain characteristics will allow this error to be subtracted from the readings made when the device under test is inserted in the test setup.

If this test is to be performed using VITS inserted into programming before arrival at the cable television facility, then the demodulator and video test set can be used to measure the differential gain present in the incoming signal. This can then be subtracted from the measurement made at the output of the modulating or processing equipment to arrive at the differential gain contributed by the cable system.

This subtraction must be done separately for each packet (a packet is the color subcarrier present on an individual step of the stairstep), since it is possible that the two measurements may have maximum error on different packets. This will result in 9 subtractions for a 10-step test signal (the first packet is the reference one).

Fortunately, most differential gain is usually the result of compression or clipping of the last packet (the one with the highest luminance). If this is the case for *both* the incoming signal *and* after the headend, then only this last packet need be subtracted, and it will provide the final answer.

For example:

> If the incoming signal exhibits a differential gain of +7%, and the measurement after the headend shows a differential gain of +11%, then the incoming gain can be subtracted from the measured +11% to give a system differential gain of +4%.

$$11 - 7 = 4$$

In the same fashion, if the incoming signal exhibited a differential gain of -2%, and the measurement after the headend shows the same +11%, then the incoming measurement must be *added* to the measurement for a system differential gain of +13%.

$$11 + 2 = 13$$

Generally, when differential gain distortion is present in a modulator, demodulator, or processor, differential phase distortion will be evident also. Differential phase distortion is visible in a color picture as a change in the hue of a certain area of the picture when only the brightness of that area was intended to change. Differential phase problems in cable headend systems are generally the result of a modulation or detection process with very little occurring in processors or strip amplifiers.

The measurement of differential phase suffers from the same problem associated with the measurement of differential gain. This is the inability of measuring this distortion in a modulator, demodulator, or processor without including in that process the distortion of a necessary test modulator, demodulator, or both.

The problem arises from the fact that the measurement must be made on a baseband video signal consisting of a 5 or 10 riser stairstep with the steps varying between the blanking level and reference white (0-90 IRE units). Superimposed on these steps will be a chrominance signal of 20 IRE peak-to-peak. If a clipper circuit is contained in the modulator under test to prevent modulation beyond 90%, then ignore the information on the fifth step of a 5-step signal or the ninth and tenth steps of a 10-step signal.

As is the case with differential gain, differential phase measurements also include the distortion present in the test unit. Under very low measurement conditions (less than 2 degrees) it is difficult to perceive that a cancellation process is occurring and it may be safe to assume that the unit under test has low differential phase distortion. Under conditions of high differential phase (greater than 10 degrees), it is extremely difficult to determine the contribution of each unit without some prior knowledge of one or more of the test units.

If this test is to be performed using VITS inserted into programming before arrival at the cable television facility, then the demodulator and video test set can be used to measure the differential phase present in the incoming signal. This can then be subtracted from the measurement made at the output of the modulating or processing equipment to arrive at the differential phase contributed by the cable system.

This subtraction must be done separately for each packet (a packet is the color subcarrier present on an individual step of the stairstep), since it is possible that the two measurements may have maximum error on different packets. This will result in 9 subtractions for a 10-step test signal (the first packet is the reference one).

Fortunately, most differential phase is usually the result of compression or clipping of the last packet (the one with the highest luminance). If this is the case for *both* the incoming signal *and* the signal measured after the headend, then only this last packet need be subtracted, and it will provide the final answer.

For example:

If the incoming signal exhibits a differential phase of +3 degrees, and the measurement after the headend shows a differential phase of +8 degrees, then the incoming phase can be subtracted from the measured +8 degrees to give a system phase of +5 degrees.

$$8 - 3 = 5$$

In the same fashion, if the incoming signal exhibited a differential phase of -2 degrees, and the measurement after the headend shows the same +8 degrees, then the incoming measurement must be *added* to the measurement for a system differential phase of +10 degrees.

$$8 + 2 = 10$$

10.4 TV Baseband Signal-to-Noise

Definition: Baseband Signal-To-Noise ratio in dB is the peak-to-peak value of reference blanking-to-white picture signal to the CCIR Rec. 567 weighted rms noise value in the 40 Hz to 4.2 MHz frequency bands (see discussion).

Required Equipment

- Video waveform generator
- Video test set with random noise measuring capability
- Variable 75 Ω attenuator

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. The choice of a video signal for this test is not critical, provided that a quiet line that does not contain video or data (i.e., a line in the vertical interval without test signals or closed captioning data) is selected for the measurement.
3. Equipment connections are similar to those shown in Figure 10-2. Adjust the demodulator input for the desired test level by using the variable attenuator placed between the modulator output and the demodulator input. A signal level meter or spectrum analyzer can be used to monitor the level.

   Note: Modulator must be set to 87.5% modulation.
4. Adjust the demodulator video output to one volt peak-to-peak using the waveform monitor function of the video test set.
5. Using the waveform monitor function of the video test set select a quiet line in the vertical interval that does not contain any video or data.
6. Select the appropriate highpass (40 Hz) and lowpass (4.2 MHz) filters. Select CCIR Rec. 567 (unified) weighting. Note: some video test equipment may not incorporate a 40 Hz highpass filter. In this case, select the frequency that is closest to 40 Hz. (e.g., 100 Hz).
7. Follow the instructions in the test equipment operator’s manual for completing the measurement. If the test equipment is capable of signal averaging, select a minimum of 32 averages.

An example of a video signal/noise measurement is shown in Figure 10-7.

Performance Objective: It is desirable to keep the noise generated by headend equipment low enough such that its effect on the overall cable system signal-to-noise performance is minimal. This means individual modulators and demodulators used for headend signal functions should have a 50 dB signal-to-noise performance as a minimum with a 55 dB performance preferred. The performance level would apply when the equipment is operating at its optimum signal level.

Modulators should easily meet a 55 dB signal-to-noise ratio performance level, however, some demodulators may only marginally meet that performance.

Discussion: This procedure is intended to measure the signal-to-noise performance of a modulator and demodulator operating together. The signal-to-noise performance of either unit can then be determined if the performance of the associated unit is known. If the performance is not known for either unit, then the measurement only indicates that the actual signal-to-noise ratio of both units is greater than the measured value. This is often adequate information for many applications.

The noise measurement is made by comparing the signal-to-noise measurement per the above definition. Standard test equipment is readily available for signal-to-noise measurements per the above definition.

Note: Unified weighting network to CCIR Rec. 567. This weighting filter which is used for all TV standards simulates the response of the human eye to noise. Since noise voltages of higher frequency are perceived by the eye as less disturbing, the attenuation of this filter increases with the frequency and the noise components of higher frequency are weighted less.

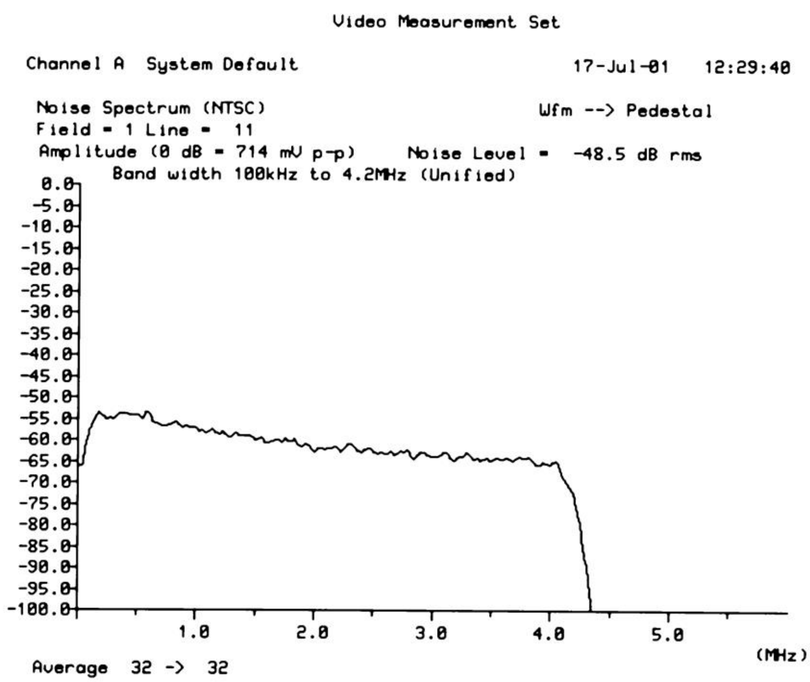


Figure 10-7: Example of Video SNR Measurement

10.5 TV Modulation Linearity

Definition: Modulation Linearity (also known as luminance nonlinearity) is the difference in amplitude between largest and smallest steps of a video signal divided by the largest step, (expressed in percent).

Discussion: Measuring the linearity of a television modulator requires a method of demodulating the signal with a device which has known linearity such as a precision demodulator. A synchronous demodulator will typically provide better linearity than an envelope detector demodulator and is therefore recommended for this type of test. A video signal containing a staircase, such as the FCC Composite signal, should be applied to the modulator. Modern video test systems will make a luminance nonlinearity measurement without the need to differentiate the steps. The following procedure is based on using a standard waveform generator and monitor and a precision demodulator used in synchronous mode.

Required Equipment

- Video waveform generator
- Video test set
- Synchronous demodulator
- Variable attenuator

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Equipment connections are similar to those shown in Figure 10-2. Adjust the demodulator input for the desired level using the variable attenuator placed between the modulator output and the demodulator input. A signal level meter or spectrum analyzer can be used to monitor the level.
3. Set the waveform generator to produce an FCC Composite test signal. Using the waveform monitor function, verify that the subcarrier at the top of the stairstep is not clipped (see Figure 10-6).
4. Select the Luminance Nonlinearity measurement function.
5. If the video test set is capable of signal averaging, select a minimum of 32 averages.
6. Follow the instructions in the test equipment operator’s manual for completing the measurement.

An example of a luminance nonlinearity measurement is shown in Figure 10-8.

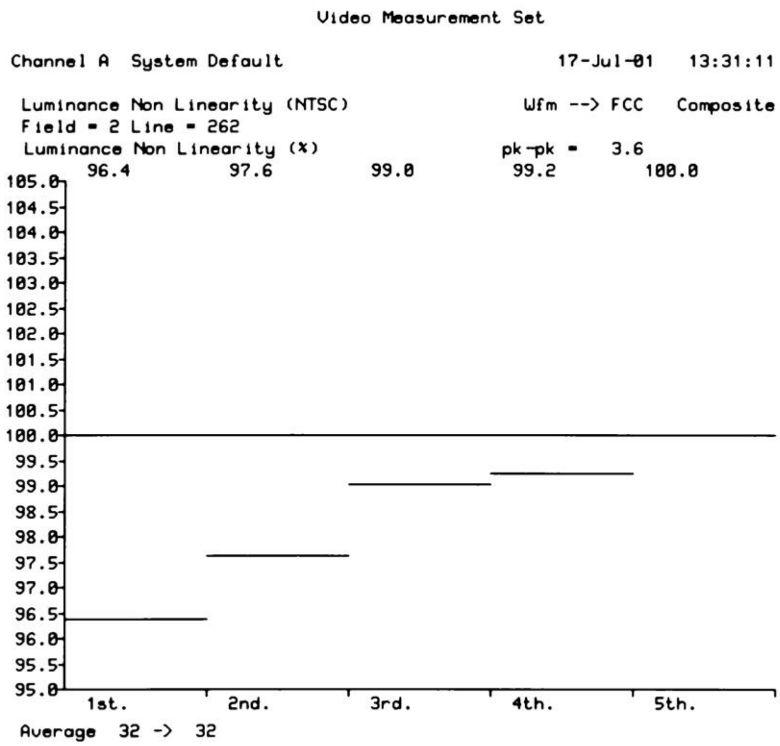


Figure 10-8: Example of Luminance Non Linearity

Performance Objective: The above test method can provide linearity test measurements which are accurate to within 1 percent. However, 10 percent linearity is a reasonable goal for modulators. Recently designed modulators will probably perform well within that limit while older modulators could probably only marginally meet a 10 percent maximum limit.

10.6 Short-Time Waveform Distortion (K-Factor)

Definition: Short-time waveform distortion is the change in waveform shape of a short duration pulse or short rise time transition caused by passage through a television facility. The K-factor is a standardized rating of the transient waveform distortion which relates to subjective picture quality.

The usual K Factor measurements are $K_{pulse/bar}$ (K-PB), K_{2T} and pulse/bar ratio. All of these are expressed as percentages. The overall K Factor is the largest value obtained from these measurements.

$K_{pulse/bar}$ is calculated from measurements of the pulse and bar amplitudes. The calculation is performed as follows:

$$K_{pulse/bar} = 0.25 * \left(\frac{\text{bar - pulse}}{\text{pulse}} \right) * 100$$

K_{2T} is a weighted function of the amplitude and time of distortion occurring before and after the 2T pulse.

Discussion: Short-time waveform distortion measurements utilize the sine-squared pulse and bar test waveform. For 4 MHz systems a sine-squared pulse with a 2T (0.250 µs) half amplitude duration is used. The spectrum of this pulse is almost entirely within the 4 MHz system bandwidth so that only the presence of non-uniform amplitude or phase characteristics within this bandwidth will cause the resultant pulse shape to be distorted. Differences in low and high frequency response will also change the amplitude of the pulse relative to the bar.

The numerical determination of the K-factor is made using an automatic measurement, which, in modern test equipment, replaces the specially engraved graticule formerly used on video waveform monitors.

Required Equipment

- A NTSC video test waveform generator which provides the 2T sine-squared pulse and bar waveform. These waveforms may be part of the FCC Composite test signal or a VITS signal, which also include other waveforms on the same horizontal line.
- A video test set capable of measuring K-factor.
- A variable 75 Ω attenuator (two required for processor test).
- A video modulator and/or demodulator, as required for the channel equipment to be tested, with suitable channelization.

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Equipment connections are similar to those shown in Figure 10-2. The headend channel to be tested may employ a demodulator, a modulator, or an RF processor. The actual headend connections should be the RF output (or video input for a modulator) and the combined headend RF output (or video output for a demodulator). For a channel which uses demodulator/modulator RF processing, the units can be measured as a pair.
3. Set the video waveform generator to deliver an FCC Composite test signal and adjust the modulator for normal depth of modulation. Verify that the subcarrier at the top of the stairstep is not clipped (see Figure 10-6).
4. Select the K-factor measurement on the test equipment menu.
5. If the test equipment is capable of signal averaging, select a minimum of 32 averages.
6. Follow the steps in the test equipment operator’s manual for completing the measurement.

An example of a K-factor measurement is shown in Figure 10-9.

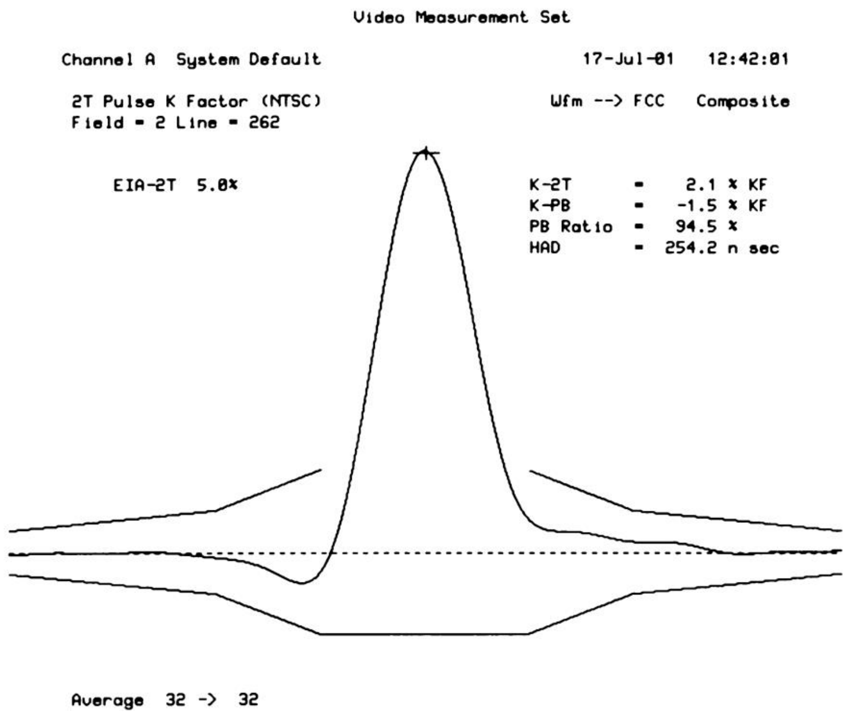


Figure 10-9: K-factor Measurement Example

Notes, Hints and Precautions

The test modulator/demodulator equipment used to support this measurement will contribute to the measured K-factor. Because of the time domain nature of this distortion, the K-ratings of the test modulator/demodulator cannot be subtracted from the total measured above even if accurately known. In general, it will be necessary to assume that all of the measured distortion is in the equipment under test, that is, that the distortion of the test channel is not greater than the total measurement. Where a processor channel is to be measured, first operate the test mod/demod alone and note their distortion. Then insert the processor between them and observe the change in distortion. If the change caused by the addition of the processor is small compared to the magnitude of the mod/demod distortion, then it is reasonable to conclude that the processor's K-factor is better than that of the mod/demod pair.

Use the variable attenuators to place the demodulator or processor input levels to match operating values. Whenever possible, set the input levels to match actual operating values.

Distortions to the sine-squared pulse & bar transition can be caused by variations in the amplitude or phase response within the 4 MHz video passband or the equivalent RF bandpass. Distortions caused by amplitude response errors are symmetrical around the pulse (or transition) while whose caused by delay variations (nonlinear phase) are anti-symmetrical. In general, both forms occur although non-uniform delay effects usually predominate.

Performance Objective

Good engineering practice calls for a K-factor of 4% or less at the system headend. Alternatively, the peak-to-peak amplitude variations preceding or following the T-step should not exceed 15 IRE units (15% of the bar amplitude) and the 2T pulse amplitude should be 86 to 119 IRE units. Current, delay - corrected headend processors, modulators and demodulators should be capable of exceeding this quality. Where supplementary selective devices, such as traps and bandpass filters are used with a headend unit, additional short-time waveform distortion may result. Generally, this is a preferable alternative to visible interference.

Where a channel is subject to additional short-time waveform distortion beyond the headend, as in reprocessing, better waveform quality at the point of origin is called for. These objectives do not include distortions which are present in over-the-air signals as received.

Chapter 11 Digital Modulation

11.1 Constellation Analysis

Although the Modulation Error Ratio (MER)[4] provides a good quantitative estimate of digital signal impairments, it does not provide any information regarding the cause of the impairment(s). Qualitative information regarding digital signal impairments is best obtained from the signal constellation. Transmission impairments affect all forms of Quadrature Amplitude Modulation in the same manner. The only difference is one of degree. (See Section 11.3: "Modulation Error Ratio (MER)" for more information on MER. See Section 16.8: "Signal Constellations" on for an explanation of the constellation.)

Required Equipment

- Digital signal analyzer capable of displaying the received signal constellation.

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Set the digital signal analyzer (DSA) to the desired channel. Select the constellation display on the DSA and observe the pattern of the dots on the display. If the DSA has zoom capability, it may be helpful to zoom in on a particular part of the constellation to obtain a closer view of the display. Constellation impairments can be determined by visual observation of the pattern. Refer to the following subsections for a description of possible impairment effects.

Impairment Effects

Constellation analysis is useful for determining the effect of the following impairments:

- Random noise
- Discrete frequency interference
- Reflections
- Phase noise
- Gain compression

Random noise produces a "smearing" of each point in the constellation. The amount of point spreading is proportional to the Carrier-to-composite noise ratio (CCN, or generically, CNR). Normally, some amount of point spreading will always be present since no system is totally noise free. However, a CNR that is too low for proper operation will produce enough spreading to make it difficult to distinguish between individual data points in the pattern. Figure 11-1 shows examples of constellations with moderate and severe random noise.

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[4] Depending on context, MER can be transmit modulation error ratio (TxMER) or receive modulation error ratio (RxMER). As well, MER can be computed before or after equalization, giving either unequalized MER or equalized MER.

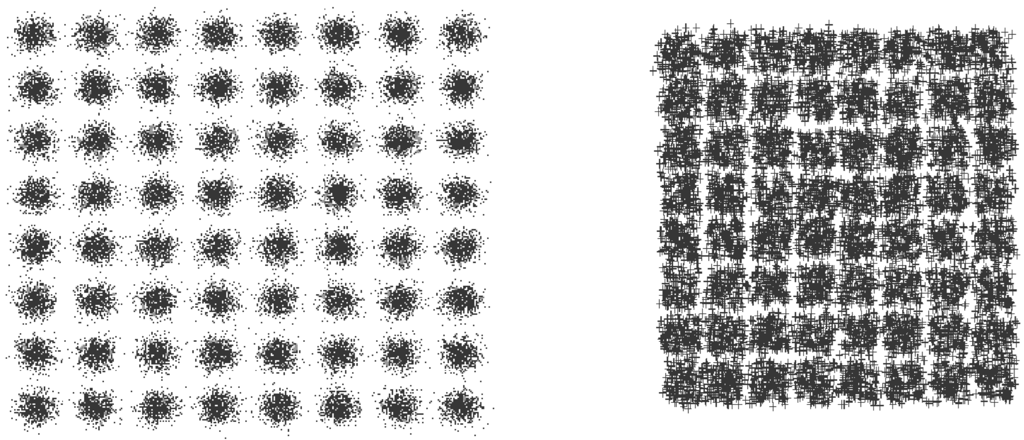


Figure 11-1: Constellation Impaired by Random Noise

A constellation diagram showing the effect of single CW carrier frequency interference is shown in Figure 11-2. The interfering carrier is a vector that is added to each point in the constellation. Since the phase of the interferer varies continuously with respect to the signal, the constellation points appear as circles whose radii are proportional to the voltage level of the interfering carrier.

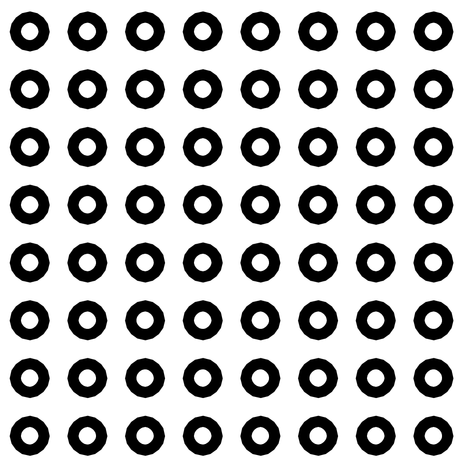


Figure 11-2: Constellation with single CW carrier interference

The effect of reflections on the signal constellation depends on the number of reflections, their respective amplitudes and their phase relative to the demodulated carrier. For a single reflection, the impairment is a miniature replica of the entire constellation around each point in the original constellation. The size of the replicated constellation is proportional to the amplitude of the reflection and the replicated constellation is rotated by an angle equal to the phase difference between the main signal and the reflection. Figure 11-3 illustrates an example of a constellation impaired by a single reflection with a phase of 45° relative to the main signal. Note that, in practice, the individual points in the replicated constellation will not usually be distinguishable since some amount of random noise will also be present. Reflections in cable systems are usually negligible except in cases of system problems such as cracked cables, bad connections, etc.

If the constellation shows that a reflection is present, more information on the nature of the reflection can usually be obtained by looking at the adaptive equalizer taps and/or the signal spectrum.

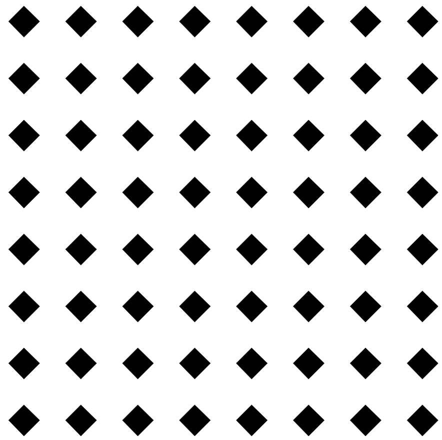


Figure 11-3: Single Reflection at 45 Degrees

Phase noise is largely a function of the stability of the local oscillator in either the headend equipment or the demodulator. When the receiver demodulates a signal with excessive phase jitter, the received constellation will oscillate about its center as shown in Figure 11-4.

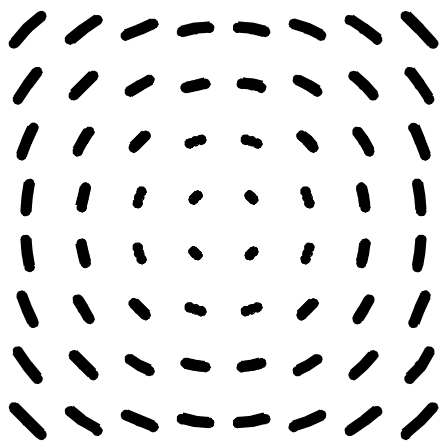


Figure 11-4: Phase Noise Impairment

Gain compression occurs when the digital signal is overdriven, causing it to operate in the nonlinear portion of the amplifier's transfer characteristic. The result is a barrel-shaped rather than square constellation. In severe cases, the points may form a circular pattern. Figure 11-5 shows an example of a constellation with gain compression.

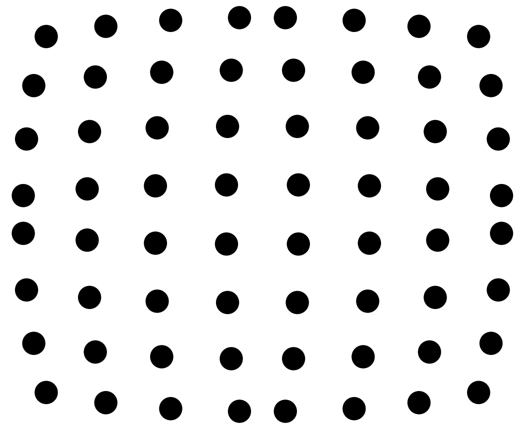


Figure 11-5: Amplitude Distortion due to Gain Compression

11.2 Error Vector Magnitude (EVM)

Definition: Error Vector Magnitude (EVM) is defined as follows:

$$\text{EVM (\%)} = \frac{\text{RMS Error Magnitude}}{\text{Peak Vector Magnitude}} \times 100$$

In order to obtain a valid number for EVM, it must be averaged over a statistically valid number of samples.

Discussion: Figure 11-6 illustrates EVM in terms of the vector difference between the ideal (reference) state position and the measured state position of a constellation point. Some signal analyzers may also provide a peak EVM value in addition to the rms value.

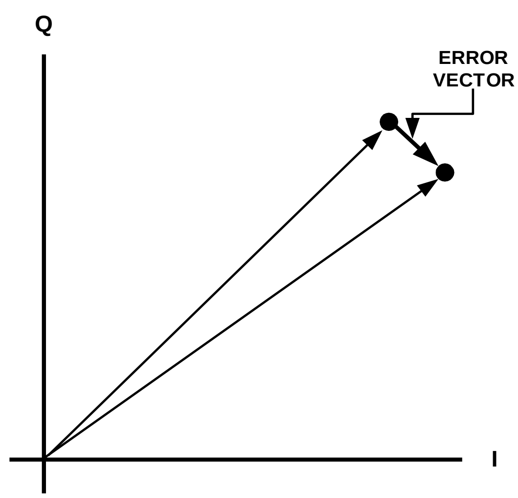


Figure 11-6: Error Vector Magnitude

Test Procedure

Error Vector Magnitude is measured in the laboratory rather than in the field. However, it is related to Modulation Error Ratio (MER). MER is a more practical quantity for field measurements of system performance and is discussed in the following section.

11.3 Modulation Error Ratio (MER)

Definition: Modulation error ratio is the ratio of average symbol power to average error power:

MER(dB) = 10log(average symbol power ÷ average error power)

In the case of MER, the higher the number, the better. Mathematically, a more precise definition of MER (in decibels) follows:

$$MER = 10\log_{10}\left[\frac{\sum_{j=1}^{N}\left(I_j^2 + Q_j^2\right)}{\sum_{j=1}^{N}\left(\delta I_j^2 + \delta Q_j^2\right)}\right]$$

where *I* and *Q* are the real (in-phase) and imaginary (quadrature) parts of each sampled ideal *target symbol* vector, and *δI* and *δQ* are the real (in-phase) and imaginary (quadrature) parts of each *modulation error* vector. This definition assumes that a long enough sample is taken so that all the constellation symbols are equally likely to occur.

In effect, MER is a measure of how “fuzzy” the symbol points of a constellation are, however to fully understand it, it will be related to other metrics that are, or have been used in cable.

MER and EVM

MER is converted to error vector magnitude (EVM) as follows:

$$EVM_{\%} = 100 \times 10^{-(MER_{dB} + MTA_{dB})/20}$$

where $EVM_{\%}$ is error vector magnitude (in percent), MER_{dB} is modulation error ratio (in dB), and MTA_{dB} is maximum-to-average constellation ratio (dB) from the following table:

Constellation (DS = Double-Square)	MTA Ratio for Constellation Symbols (dB)
QPSK and BPSK	0
16-QAM and 8-QAM-DS	2.55
64-QAM and 32-QAM-DS	3.68
256-QAM and 128-QAM-DS	4.23
1024-QAM and 512-QAM-DS	4.50
Limit for infinite QAM	4.77

Table 11-1: Maximum-to-Average Ratio for Various Constellations

Discussion: MER measurements are found on all field and lab digital instruments whereas EVM is not always provided in field test equipment.

MER is, as previously stated, the ratio of average symbol power to average error power. MER is analogous to signal-to-noise ratio (S/N or SNR) of a baseband digitally modulated signal. It measures the modulation impairments that affect the ability of the digital receiver to recover the data bits. In addition, it can be analyzed for phase and amplitude error components. High phase error component MER is usually caused by carrier frequency error or spurious interference. MER is most useful when considering digital video system margin or SNR issues.

As with EVM, MER includes all types of signal impairments such as the effects of noise, carrier leakage, and IQ level and quadrature imbalance (See 16.3 Adaptive Equalization Tutorial). Practical digital systems use filtering techniques that incorporate identical filters in the transmitter and receiver.

Therefore, MER measurements must be made after the receive filter or with test equipment that includes a matched filter. Since the receiver's adaptive equalizer compensates for channel response impairments, the shape of the adaptive equalizer reflects linear impairments in the channel through which the signal is transmitted. Thus, when an MER measurement is quoted, the use (or lack) of an adaptive equalizer should be stated.

There are practical limits when it comes to measuring MER. To compute MER, the test equipment must be able to demodulate the signal. If the signal quality is so poor that this process fails, the measurement result cannot be trusted.

Some Digital signal analyzers, set-tops, and cable modems may report MER as "SNR" or "SNR Estimate". The two abbreviations—MER and SNR—are often used interchangeably, so in most instances when "SNR" is reported it is in fact MER. In some instances, SNR is used to denote "digital" carrier-to-noise ratio (C/N or CNR). In practice, and as described in detail in Section 3.1 and Section 3.5, a cable network's noise floor will include a combination of thermal noise and other noise-like distortions, so carrier-to-composite noise ratio (CCN) is a more representative metric than just CNR. Under ideal conditions, equalized MER equals CNR, however the latter is true only if additive white Gaussian noise (AWGN) is the only significant impairment present on a system. In practice, this is not the case since other impairments also contribute to the MER. As such, using a spectrum analyzer to measure CNR is not necessarily an accurate method of measuring MER because only the effects of thermal noise are taken into account. The CNR result does not consider problems such as micro-reflections, CSO/CTB effects, and RF ingress. On a properly operating system, the MER value will usually be at least a few dB worse than the measured CNR. A good CNR in conjunction with a poor MER reading indicates that something other than noise is causing system problems and should be of concern.

Required Equipment

- Digital signal analyzer (DSA), or spectrum analyzer or signal level meter with QAM analysis capability, or digital set-top, or cable modem with on-screen/Web page diagnostic capability to display MER.

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. If a DSA or spectrum analyzer/signal level meter with QAM analysis capability is used, proceed as follows (refer to the instrument manufacturer's instructions):
   - Set the instrument to the desired channel
   - Select the MER display
   - Note whether the reported MER is an equalized or unequalized value.

If a digital set-top is used, proceed as follows (refer to the set-top manufacturer's instructions):

- Select the desired channel
- Select the set-top's diagnostic mode
- Select in-band diagnostics
- Read the MER (usually displayed as "SNR") in dB from the TV screen. Note that a set-top's reported MER value is almost always equalized MER.

If a cable modem or embedded multimedia terminal adapter (eMTA) is used, proceed as follows (refer to the cable modem or eMTA manufacturer's instructions):

- Ensure that a computer is connected via an Ethernet cable to the modem or eMTA.
- Enter the IP address for the modem's (or eMTA's) built-in diagnostic Web page (e.g., http://192.168.100.1) in the address bar of a suitable Web browser running on the computer
- Depending on the modem or eMTA, it may be necessary to select the appropriate page or menu item that displays the device's performance metrics. Read the MER (usually displayed as "SNR") in dB from the computer screen. Note that a modem's or eMTA's reported MER is almost always equalized MER.

Notes, Hints, and Precautions

Because MER is a digital computation performed on digital quantities in the receiver, it is by nature extremely accurate in itself. However, the measured value can be affected by many factors (statistical variation, unequal occurrence of symbols, nonlinear effects, linkage of carrier loop bandwidth to capture length, implementation loss MER ceiling, analog front-end noise, etc.). It is not unusual for the MER value measured with a given DSA to be different than what a set-top or cable modem reports. Furthermore, one is likely to see different reported values on different makes/models of DSAs on the same signal under identical conditions. These variations are normal and expected behavior, and are related in part to some of the previously listed factors, along with implementation loss differences among different QAM receivers, different adaptive equalizer implementations, the receivers' response to the presence of adjacent channels, whether the reported MER is equalized or unequalized, and so on. As well, the reported MER value by itself provides little insight about the type of impairments that exist.

Rather than emphasize a specific MER value, one should look for relative changes over time or from measurement point to measurement point (for example, from headend to node to end-of-line). MER should be used in conjunction with other measurements such as digital channel power, CNR or CCN, impairments that may be visible in the test equipment's constellation display, in-channel flatness, in-channel group delay, bit error ratio, and so forth when characterizing signal performance.

Ideally, MER should be as high as possible. Good engineering practice suggests that unequalized MER be maintained at least 3 to 6 dB or more above the lower E_S/N_0 threshold—in effect, the MER failure threshold—for the modulation format in use. The lower E_S/N_0 threshold is defined as the point at which the symbol error rate is in the range of 1×10^{-2} to 1×10^{-3}. The following table lists the approximate lower E_S/N_0 thresholds based on those symbol error rates, for several popular modulation formats used in cable networks.

Table 11-2: Lower E_S/N_0 thresholds for various modulation formats

Modulation Format	Lower E_S/N_0 Threshold
QPSK	7–10 dB
16-QAM	15–18 dB
64-QAM	22–24 dB
256-QAM	28–30 dB

11.4 Digital Group Delay

Definition

See Section 8.2 on measuring analog Group Delay.

Required Equipment

- Digital signal analyzer (DSA) with an adaptive equalizer display and frequency response / group delay capability

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Tune the DSA to the desired channel.
3. Set the DSA to view the adaptive equalizer tap display.
4. Follow the instructions in the DSA manufacturer’s user’s manual for measuring in-channel frequency response and group delay using the adaptive equalizer response.
5. The group delay result can be read directly from the display.

Note: Please see Section 16.3 for an in-depth discussion of the Adaptive Equalization and information available from the equalizer tap coefficients.

Chapter 12 Other Measurements

12.1 Subscriber Terminal Isolation [FCC §76.605(b)(9)]

Definition: Subscriber terminal isolation is the isolation in decibels between any two subscriber terminals in a cable television system.

FCC §76.605(b)(9): ***The terminal isolation provided to each subscriber terminal:***

> ***(i) Shall not be less than 18 decibels. In lieu of periodic testing, the cable operator may use specifications provided by the manufacturer for the terminal isolation equipment to meet this standard; and***
>
> ***(ii) Shall be sufficient to prevent reflections caused by open-circuited or short-circuited subscriber terminals from producing visible picture impairments at any other subscriber terminal.***

Discussion: FCC §76.605(b)(9)(i) allows the cable operator to use manufacturer's specifications of tap terminal isolation in lieu of actual testing. A copy of these specifications should be included in the semi-annual Proof-of-Performance test report. If these specifications are not available, the following test procedure should be completed at each test point or through a representative sample of system taps, on the bench.

PROCEDURE 1 - Without Sweep Generator

Required Equipment

- RF signal generator
- RF signal level meter or spectrum analyzer

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Refer to Figure 12-1.
3. Connect "A" to "B" with short jumpers.
4. Adjust the frequency of the signal generator to the video carrier of the channel to be measured.
5. Adjust the SLM or spectrum analyzer to read the level of the RF signal generator.
6. Adjust the level of the signal generator so that the signal level meter reads 18 dBmV.
7. Remove the jumper that was connected in step 3. Connect the equipment as shown in Figure 12-1, referring to Table 12-1.
8. Read the level on the signal level meter. The difference between the SLM reading and 18 dBmV represents the amount of isolation.
9. Repeat steps 3 through 8 for each channel.

If the isolation is less than 18 dB, an additional allowance equal to the attenuation of the drop cable is allowed. In any case, the isolation of the tap plus attenuation of the drop cable must be equal to or greater than 18 dB to meet the FCC rule.

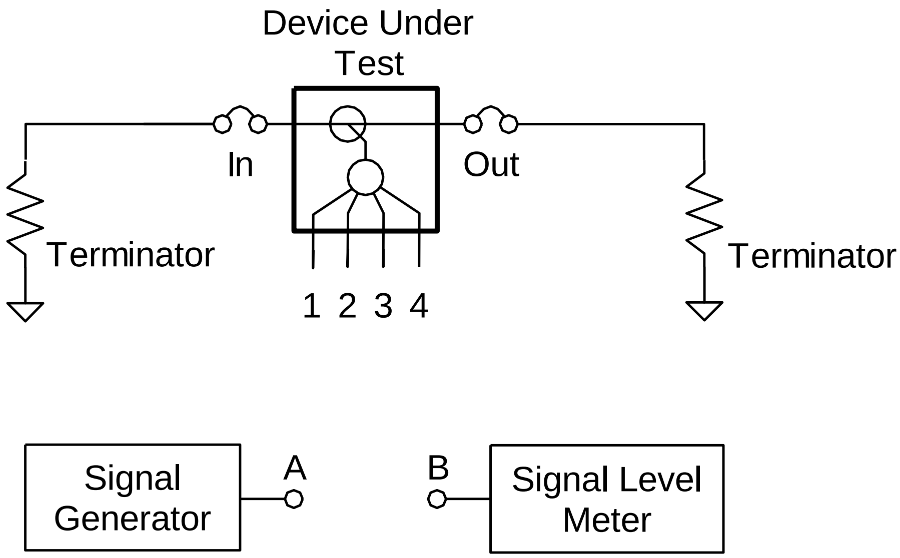


Figure 12-1: Subscriber Terminal Isolation Without Sweep Generator Test Equipment Setup

PROCEDURE 2 - With Sweep Generator

Required Equipment

- Sweep generator
- Oscilloscope, XY scope with 5 millivolt sensitivity
- A switchable attenuator.

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Refer to Figure 12-2.
3. Connect "A" to "B" with a short jumper cable.
4. Adjust the switchable attenuator to 30 dB.
5. Adjust the sweep generator controls to provide an RF output that covers the frequency bandwidth of the cable television system.
6. Adjust the sweep generator controls to provide 3 divisions of deflection on the oscilloscope.
7. Connect the device under test (DUT) as shown in Table 12-1.
8. Adjust the switchable attenuator to provide a maximum of three divisions at any one frequency.
9. Subtract the value of the attenuator obtained in step 8 from 30 dB. That number represents the isolation between DUT ports.

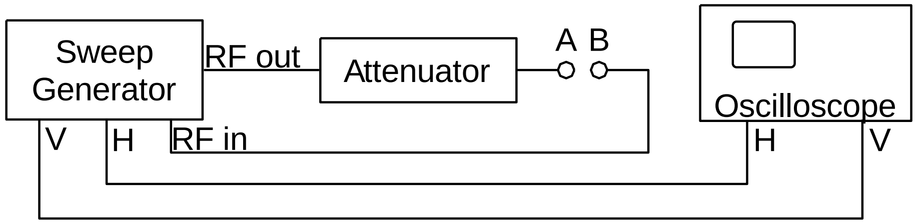


Figure 12-2: Subscriber Terminal Isolation With Sweep Generator Test Equipment Setup

	Input	Output	Tap # 1	Tap # 2	Tap # 3	Tap # 4
Test 1	T	T	A	B	T	T
2	T	T	A	T	B	T
3	T	T	A	T	T	B
4	T	T	T	T	A	B
5	T	T	T	A	B	T
6	T	T	T	A	T	B
T - indicates that the port or tap is terminated in 75 Ω						

Table 12-1: Subscriber Terminal Isolation

12.2 Signal Leakage [FCC §76.605(c)]

Definition: As used in Part 76 of the FCC rules, the term "leakage" refers to the undesired emanation of electromagnetic energy from the cable television system. FCC Rules and Regulations, Part 76 subpart K specifies the maximum allowable leakage and the conditions under which compliance may be verified.

FCC §76.605(c): ***As an exception to the general provision requiring measurements to be made at subscriber terminals, and without regard to the type of signals carried by the cable television system, signal leakage from a cable television system shall be measured in accordance with the procedures outlined in §76.609(h) and shall be limited as shown in table 1 to paragraph (c):***

Table 1 to Paragraph (c)

Frequencies	Signal leakage limit	Distance in meters (m)
Analog signals less than and including 54 MHz, and over 216 MHz	**15 µV/m**	**30**
Digital signals less than and including 54 MHz, and over 216 MHz	**13.1 µV/m**	**30**
Analog signals over 54 MHz up to and including 216 MHz	**20 µV/m**	**3**
Digital signals over 54 MHz up to and including 216 MHz	**17.4 µV/m**	**3**

Note: This standard practice is not intended for determining compliance with the CLI requirements specified in §76.611 but does demonstrate techniques that would be useful for collecting leakage data in order to calculate CLI.

Discussion: The signal leakage provisions of subpart §76.605(c) apply to all cable systems that serve 1,000 subscribers, or more, or to any system that distributes signals at frequencies not allocated for over-the-air television and FM broadcasting, regardless of size.

Systems with fewer than 1,000 subscribers that distribute signals only on standard over-the-air channel frequencies are not subject to the leakage requirements of §76.605(c) and are not required to perform compliance testing.

Any system regardless of its size, which distributes signals in the aeronautical bands (108 MHz to 137 MHz, and 225 MHz to 400 MHz) must meet the Cumulative Leakage Index (CLI) requirements explained in §76.611 and the quarterly monitoring provisions of paragraph §76.614.

The leakage measurement procedures described in this section conform to the general test methodology outlined in §76.609. Data collected by means of these procedures is adequate for determining compliance with the leakage limits stated in §76.605(c) as well as the 20 µV/m requirement of §76.614.

Procedures: The procedures that follow use a dipole antenna in conjunction with either a signal level meter or spectrum analyzer.

Dipole Antenna: Most commercial field strength measuring equipment operating above 25 MHz and below the microwave region use a resonant half-wave dipole as a probe or pick-up device. Below this range, a loop antenna is common. Above it, reliance is usually placed on a "standard horn" antenna. Test dipoles are often designed to "telescope" so that the length of its arms can be adjusted to the resonant length of the frequency measured.

The physical length of a practical antenna always is somewhat less than its electrical length. That is, a "half-wave" antenna is not one having the same length as a half wavelength in space; it is one having an electrical length equal to 180 degrees. Or, to put it another way, it is one whose length has been adjusted to "tune out" any reactance; therefore, it is a resonant antenna.

The antenna length required to resonate at a given frequency (independently of any dielectric effects) depends on the ratio of the length of the conductor to its diameter. The smaller this ratio is, the shorter the antenna will be for a given electrical length. For practical purposes, a close enough approximation for the physical length of a resonant half-wave antenna in inches may be taken to be:[5]

$$\text{Length (inches)} = \frac{(5905)(0.95)}{\text{f (MHz)}} = \frac{5610}{\text{f (MHz)}} \qquad (1)$$

Approximate lengths for the visual carrier frequencies in the VHF cable television bands are given in Table 12-2. If the manufacturer of a test dipole provides figures that differ from these, their figures should of course be used. For details on dipole construction, see “Appendix: Antenna and Balun Construction”.

[5] ARRL Antenna Book, published by the American Radio Relay League, Inc., Chapter 2, 15th Edition, Newington, CT.

Table 12-2: Dipole Length and Antenna Factors for Cable TV Channels

TV Channels	Visual Carrier (MHz)	Dipole Length (inches)	Antenna Factor (dB) for 75 Ω Termination	Antenna Factor (dB) for 50 Ω Termination
2	55.25	101.5	1.2	3.0
3	61.25	91.6	2.1	3.9
4	67.25	83.4	3.0	4.8
5	77.25	72.6	4.2	6.0
6	83.25	67.4	4.8	6.6
14(A)	121.25	46.3	8.1	9.9
15(B)	127.25	44.1	8.5	10.3
16(C)	133.25	42.1	8.9	10.7
17(D)	139.25	40.3	9.3	11.1
18(E)	145.25	38.6	9.6	11.4
19(F)	151.25	37.1	10.0	11.8
20(G)	157.25	35.7	10.3	12.1
21(H)	163.25	34.4	10.7	12.5
22(I)	169.25	33.1	11.0	12.8
7	175.25	32.0	11.3	13.1
8	181.25	31.0	11.6	13.4
9	187.25	30.0	11.8	13.6
10	193.25	29.0	12.1	13.9
11	199.25	28.2	12.4	14.2
12	205.25	27.3	12.6	14.4
13	211.25	26.6	12.9	14.7

	Antenna Factor	Antenna Factor (dB)
75 Ω termination	$0.021\ f_{MHz}$	$20 \log f_{MHz} - 33.6$ dB
50 Ω termination	$0.0257\ f_{MHz}$	$20 \log f_{MHz} - 31.8$ dB

Measurement of Field Strength with a Dipole: When a resonant dipole is used to measure an unknown field strength, E, that field can be related to the voltage, V, measured at the end of a properly terminated transmission line as shown in Figure 12-3. The dipole will respond, of course, only to the component of the electric field parallel to the length of the dipole. The input impedance at the center of a thin symmetrical dipole is 73 Ω. Antenna transmission lines used with most commercial field-strength measuring equipment are typically 75 Ω or 50 Ω unbalanced coaxial cables. A balanced-to-unbalanced transition, or "balun," is usually furnished as a part of the dipole assembly which gives the dipole a nominal output impedance of 73 Ω or 49 Ω. These values are sufficiently close to 75 Ω and 50 Ω that mismatch errors are insignificant. Relationships for the 75 Ω and 50 Ω cables are shown in Figure 12-3(a) and Figure 12-3(b) respectively.

When a 73 Ω half-wave dipole antenna is exposed to an RF field, the rms voltage, V, delivered to a 75 Ω load is:[6]

$$V = \frac{\left(5.58\sqrt{73}\,E\right)}{f} \qquad (2)$$

where:

V = the output voltage in microvolts rms (3)

E = the field strength in microvolts per meter (4)

f = the signal frequency in megahertz (5)

Conversely, if the voltage is known:

E	=	0.021 f V	for 73 Ω antenna	(6)
E	=	0.0257 f V	for 49 Ω antenna	(7)

Values of "antenna factors" (i.e., 0.021 f for 73 Ω and 0.0257 f for 49 Ω) are shown in Table 12-2. When cable loss and balun loss are not negligible, they should be added to the dipole factor. Cable loss in decibels per 100 feet is normally supplied by the manufacturer and can usually be found in various cable handbooks.

When using a 75 Ω signal level meter or spectrum analyzer calibrated in dBmV the following formulas can be used:

$$V(dBmV) = 20 * \log E(\mu V/m) + G - AF - 60 - L \qquad (8)$$

and

$$E(\mu V/m) = 10^{\frac{[V(dBmV) + AF + L + 60 - G]}{20}} \qquad (9)$$

where:

G	=	preamp gain in dB	(10)
AF	=	antenna factor in dB	(11)
L	=	cable loss in dB	

6 Technical Handbook for CATV Systems, Third Edition, Keneth Simons, pp. 103 - 104.

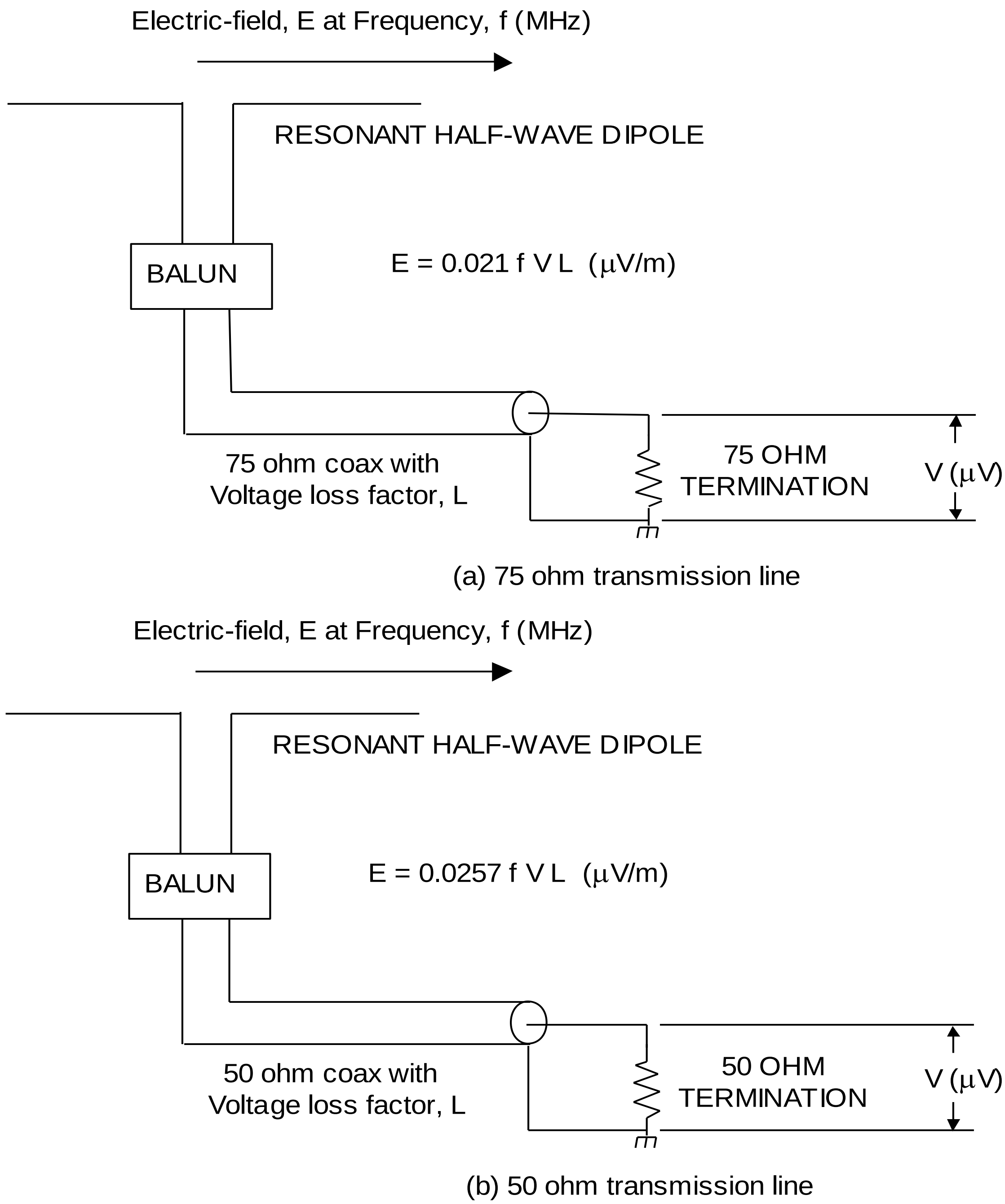


Figure 12-3: Dipole Field Strength vs. Output Voltage Relationships

PROCEDURE 1 - Signal Level Meter

The method for using an SLM is very straightforward; however, there are two difficulties which might be encountered of which the operator should be aware. The first of these occurs when the measurement is being made in the presence of high-level urban noise or non-cable related signals. Under such conditions, the interference might obscure the signal being measured, rendering the instrument ineffective. The second of these difficulties occurs when measurements are being attempted in the VHF high band. Many of the meters commercially available do not have the required sensitivity to measure 20 microvolts per meter. For this reason, measurements may have to be confined to either the low end of the VHF high band or made by employing a preamplifier ahead of the signal level meter.

Note: Extreme care should be taken when making measurements near AC power lines. Allowing the antenna or any part of the measuring system to contact power lines could result in severe injury or death.

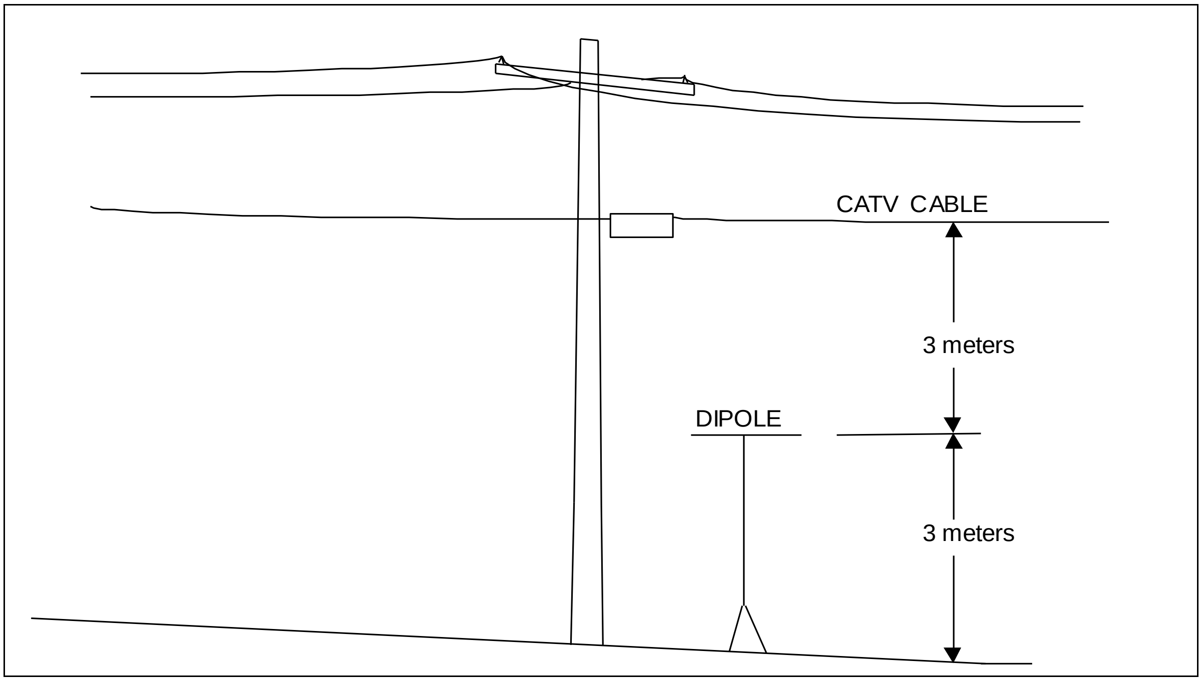


Figure 12-4: Measurement Dipole Placement

With the exception of the two possible difficulties mentioned above, the actual test procedure can proceed as follows. The field strength meter should be initially connected to one of the output ports of the system and peaked on the visual carrier of the channel to be checked. The meter is then disconnected from the cable and the test dipole connected to the input port of the meter as shown in Figure 12-5 (a). The test setup should be transported to the test site and positioned in the manner suggested in Figure 12-4. Starting with all of the attenuators switched in, they should be switched out until a maximum meter deflection is achieved and a direct reading of the field strength at ten feet is obtained. Because the signal for which the measurement is being attempted is in most cases so small, it is important to set the frequency for a peak meter reading directly from the cable each time a set of measurements is to be performed at a new frequency.

PROCEDURE 2 - Spectrum Analyzer

Perhaps one of the most useful instruments to a cable operator in making the required performance measurements on his system is the spectrum analyzer. A good quality spectrum analyzer with a frequency range of a least 5 MHz to 1002 MHz may be used to make several of the required measurements, including those related to cable leakage.

Essentially, the method of measurement for cable leakage with the spectrum analyzer is very similar to that outlined for a selective voltmeter, except that the analyzer is employed as the measuring device in lieu of a field strength meter. In making the electromagnetic field measurement, absolute rather than relative determination of signal strength is required. Most present-day spectrum analyzers employ integral amplitude and frequency calibration, thereby eliminating the need for external calibration devices. If necessary, an external, calibrated signal generator may be used. In either case (both of which are shown in Figure 12-5), the measured signal voltage is converted to field strength in microvolts per meter as discussed previously.

Employing the test setup shown in either Figure 12-5(b) or Figure 12-5(c), the analyzer should be adjusted to the visual carrier frequency of the channel being tested.[7] Further, the sweep width should be narrow enough to display only the power spectrum of the channel being investigated. (Although this is not critical, it will make the task of reading the measured values easier if the display dispersion is on the order of 1 MHz per division.) While the IF bandwidth can be set in the 100 kHz to 300 kHz range to approximate that of a signal level meter, improved rejection of nearby signals and increased sensitivity can be realized by using a narrower IF bandwidth. Care should be taken to set the bandwidth wide enough to capture the peak energy associated with the vertical sync pulse. The narrowest usable IF bandwidth for a particular analyzer is most easily determined by first starting with a setting of 100 kHz or greater and then reducing IF bandwidth noting the narrowest setting which does not produce a decrease in vertical sync amplitude.

In an urban environment, the level of man-made noise may affect or obstruct the measurement technique. With an IF bandwidth on the spectrum analyzer of 100 kHz or greater, the level of man-made noise may well exceed 20 microvolts/meter and would, therefore obscure the measurement.[8] This ignores the fact that the signal leakage probably would not cause harmful interference. A possible technique is suggested for making measurements in the presence of a high level of noise.

In order to perform this measurement, it is necessary to decrease the IF bandwidth of the spectrum analyzer so that less noise-energy is admitted. To do this without also omitting some of the energy in the signal to be measured, the power spectrum of the measured signal must be decreased. For this purpose, a CW signal generator may be employed with its output level set equal to that of the visual carrier of the signals normally carried on the cable system. Thus, the signal can be measured with a spectrum analyzer in the manner discussed previously, except that a much narrower bandwidth can be used.

Nature of Cable System Signal Leakage Fields: Signal leakage from a television cable system is likely to be both complex and different for each case, making generalization about its nature difficult. However, it may be assumed that leakage is frequently caused by a poor connection or a break, causing an RF discontinuity in the outer conductor. This acts as a source, driving a very long antenna, i.e., the cable's outer conductor. The source is likely to be "broadband" so that, if leakage occurs at one channel frequency, other channel frequencies are likely to be leaking as well. The field around the cable will consist of both a "near field" component, which diminishes as the inverse cube of the distance away from the cable, and a "far field component," which diminishes simply as the inverse of

[7] In some cases, a CW carrier is used instead of a modulated analog visual carrier for leakage testing. Make sure the CW carrier is set to the same peak envelope power as the modulated visual carrier.

8 ITT's Reference Data for Radio Engineers, Fifth Edition, cites the noise level for urban man-made noise at 100 MHz as being 40 dB above kTB, (Fig. 1.). With a bandwidth of 100 kHz, this is 14 microvolts across 50 ohms. In some urban areas, the noise level may be much higher than this.

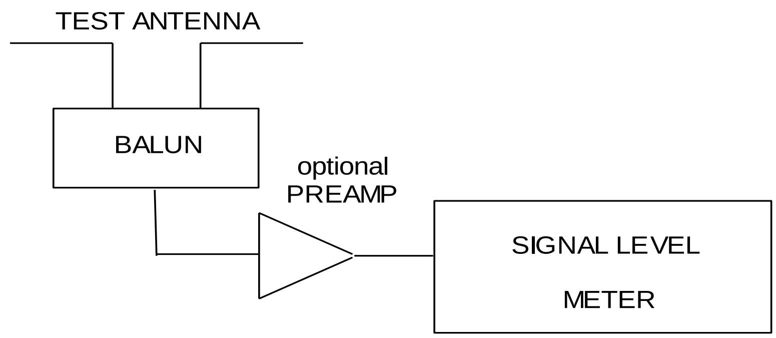
the distance. The relationships are likely to be altered, however, by the presence of reflections from other wires, the ground, and nearby vehicles and buildings.

Placement of the Dipole: §76.609(h)(3) of the FCC's rules suggests that the dipole used for measurements be placed 3 m above the ground and not closer than 3 m (10 ft) from the system component being measured. Because of the uncertain nature of the field and the difficulty of converting to equivalent values at 3 m when measurements are made at a distance greater than 3 m, it is recommended that the dipole, when practical, be placed exactly 3 m from the system component. This is illustrated in Figure 12-4. At the same time, it should be at least 3 m above the ground and 3 m away from other wires and cables. If the height above ground of the system component is less than 6 m (20 ft.) the dipole should be offset to a point where its center is both 3 m from the system component and 3 m above ground. The dipole should be moved along the cable maintaining the proper spacing as described above until a peak reading is observed. Then as the Commission's rules state, the dipole should be rotated in the horizontal plane to maximize the received signal.

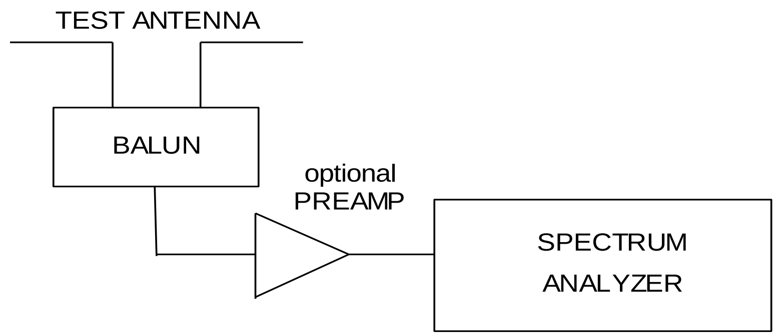
In many cases, there will be a telephone cable below the cable television cable, which will make compliance with §76,609(h) of the rules difficult. The dipole could be placed at the level of the cable television cable and offset. However, the measurement accuracy will probably be sufficient if the telephone cable is ignored and the procedure shown in Figure 12-4 is followed.

Measurements Above 216 MHz and Below 54 MHz: The advantage of the 3 m spacing when measuring leakage between 54 MHz and 216 MHz is that the measured results can be compared directly with the FCC limit of 20 microvolts per meter at a distance of 3 m. In most cases, measurements at a 3 m spacing will also suffice for frequencies above 216 MHz and below 54 MHz. However, in these cases, a conversion to equivalent values at 30 m (100 ft.) is necessary in order to determine whether the FCC limits, specified at 30 m, are exceeded. It is reasonable to expect that the "far field" component of the leakage at 30 m will be 1/10 the value of 3 m, and that the "near field" component at 30 m will be only 1/1000 of its value at 3 m. What is not likely to be known is the ratio of near and far fields (which is a function of the effective aperture of the leakage source). Nonetheless, it is quite unlikely that the field intensity at 30 m will ever be greater than 1/10 of the value measured at 3 m.

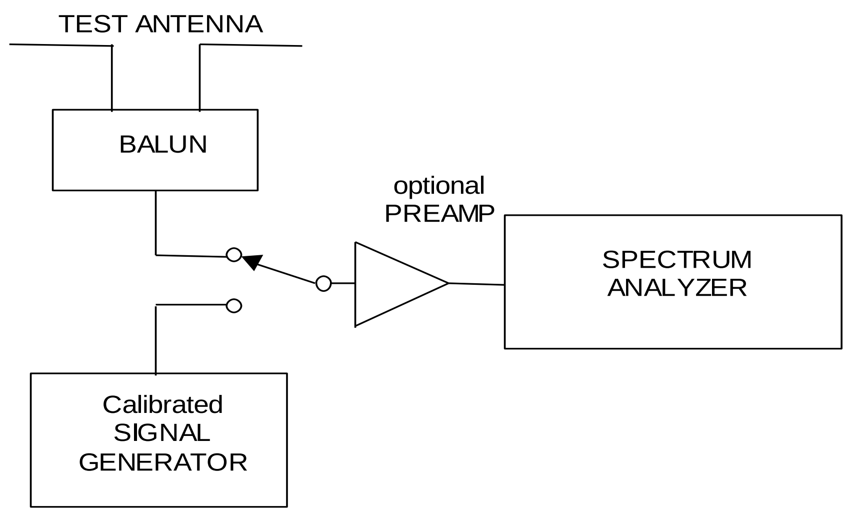
Based on this, it is suggested that, as a matter of convenience, all leakage measurements be made at a 3 m spacing, as illustrated in Figure 12-4. If measurements are made at frequencies above 216 MHz and below 54 MHz, the results should be considered within FCC limits if they do not exceed 150 microvolts/meter. If 150 microvolts/meter is exceeded, compliance with the FCC limits might still be demonstrated by moving the dipole out to a 30 m spacing and checking to determine whether the field is within the 15 microvolt/meter limit. Alternatively, the cable system owner or operator may prefer to take whatever action is required to bring the leakage level within a limit of 150 microvolts/meter at 3 m.



(a) Use of a Signal Level Meter



(b) Use of Internally Calibrated Spectrum Analyzer



(c) Use of Externally Calibrated Spectrum Analyzer

Figure 12-5: Test Equipment Connections

Notes, Hints and Precautions

It is recommended that, whenever possible, measurements be made on those channels that do not correspond with strong, over-the-air signals because of the difficulty of separating the two signals. However, if it is necessary to perform test measurements on channels that coexist with broadcast signals in order to investigate a specific complaint of interference, there are several alternatives from which the operator may choose in making the measurements.

Perhaps the easiest method for making these measurements is to wait until the broadcast signal is off the air, if in fact, that broadcast station does shut down at night. When the station is off the air, a test signal is inserted with a signal generator into the slot occupied by the channel to be tested. The signal generator is set at the frequency and level of the visual carrier of the signal normally carried in that slot. The field strength meter or spectrum analyzer is then taken into the field and the measurements made according to the procedures outlined earlier. The one obvious drawback to this method is that it may necessitate making measurements during the pre-dawn hours when the station is shut down.

An alternative method that is available to the operator with access to a spectrum analyzer allows the measurement to be performed during the normal working day. The procedure requires that the spectrum analyzer be able to synchronize its sweep generator to an external source (i.e., to have an external sync input) and be able to effectively reduce its spectrum width down to zero or, at least, a value equal to, or less than, the IF bandwidth. The test setup is similar to those shown in Figure 12-5 (a) and (b) except for the addition of an external sync source. This source can be a television receiver modified for this purpose by tapping off the horizontal sync pulse directly after the sync separator. A short lead is then used to transmit this sync pulse to the external sync input on the spectrum analyzer. Next, the spectrum width of the analyzer is narrowed as much as possible and the analyzer's sweep generator placed in the external sync mode. The analyzer is then tuned to the visual carrier frequency of the signal being measured. In this manner, the instrument becomes essentially a time reference measuring device. This means that the sync pulse of the cable signal will remain stationary on the screen, while the interfering over-the-air signal will tend to drift across the display. The peak of the stationary sync pulse is measured in the manner described under "PROCEDURE 2 - Spectrum Analyzer”, the absolute value of which indicates the radiated field strength.

Appendix: Antenna and Balun Construction

The test antenna is assembled from adjustable rods mounted on a push-on, rigid-plastic pipe coupling. The antenna elements are typical collapsible whips intended for CB use. The assembly is supported by 10 feet of standard one-half inch, rigid PVC-plastic water pipe.

The balanced dipole is connected to the 75 Ω coax cable through a balun to isolate stray ground currents and maintain symmetry (see Figure 12-6).

The No. 22 AWG insulated, twisted pair has approximately 75 Ω characteristic impedance. When wound on the ferrite toroidal core, sufficient impedance is produced to limit longitudinal currents to an acceptable value. The impedance match between coax and antennas is good from 50 MHz to 250 MHz and insertion loss is less than 0.3 dB. Because of higher insertion loss and questionable impedance match, the use of a conventional subscriber balun is not recommended.

The balun is constructed from the ferrite toroid core listed in the parts lists, and the twisted pair is made from ordinary No. 22 AWG insulated hookup wire. Five full twists per inch is proper for a characteristic impedance of 75 Ω. Using 8.5 inches of this twisted pair, wind 6 full turns on the ferrite core. Pull these tight and apply polystyrene coil dope. This assembly should be supported on insulated mounting strips supporting the “F” fitting at one end and the toroidal coil at the other. This is mounted inside the pipe coupling with the ends connected to the base of the telescoping whip antennas. These should be mounted also on the same pipe coupling.

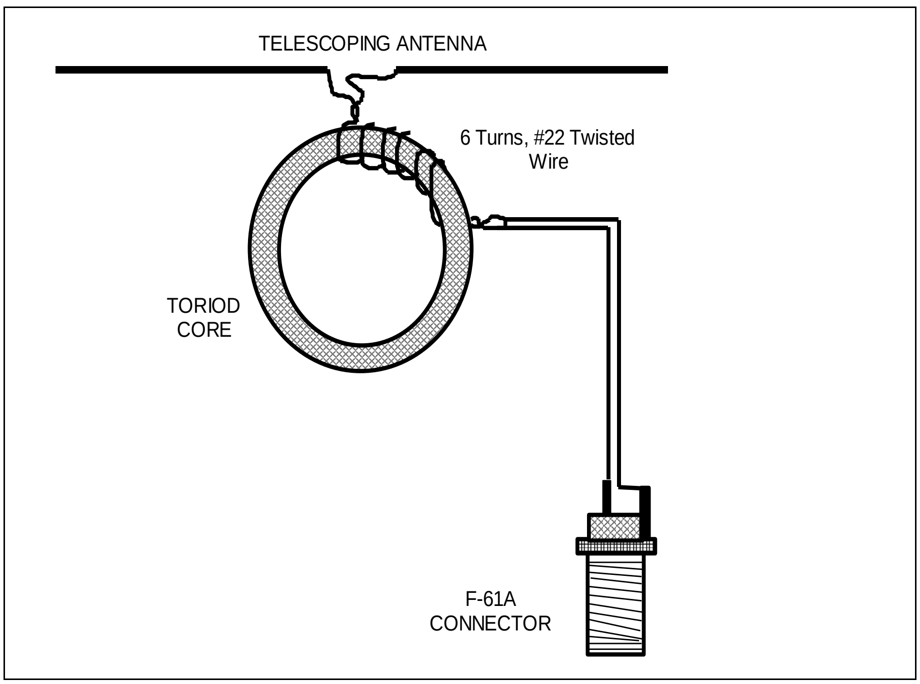


Figure 12-6: Dipole Construction

Parts and Materials Required

- (2) 5-foot sections of 1/2 inch rigid PVC plastic water pipe
- (4) Push-on couplings for above
- (2) Whip antennas – suitable whip elements available for CB radio replacement are composed of 9 to 11 sections and can be extended to 50 – 54 inches.

- (1) Ferrite toroid for balun
- Wire for balun No. 22 AWG solid plastic hookup wire
- (2) Bakelite mounting strips for balun
- (2) Ground lugs with 3/8 inch hole (as used for ground lugs on pots)
- (1) “F” chassis receptacle with mounting nut
- (1) 1-inch 4-40 nylon or brass machine screw with washers, lock-washers and nuts.

IMPORTANT SAFETY PRECAUTION
There is always the possibility, in these tests, that the antenna may come in contact with a power line. All exposed metal parts, such as cable connectors, should be taped or otherwise insulated to prevent accidental contact. The receiver should be grounded to the frame of the vehicle. When probing a cable, all parts of the antenna should be covered with insulation to prevent accidental contact.

12.3 Signal Leakage, All-Digital System

Leakage detector technology exists to enable both identifying and localizing sources of shielding integrity flaws allowing escape of QAM or other digitally modulated signaling currently endorsed by the cable industry. If that equipment is not available locally, the preferred method to ensure compliance with the various FCC rules, including §76.605(c) and §76.613, requires the inclusion in all cable systems of one or more test CW or television video carriers at some strategic spectral position within the downstream frequency allocation. Ideally, the selected location will be within the current FAA frequency band 108 MHz to 137 MHz. To avoid sensitive signal leakage receiver overload, it is further recommended that the lower boundary be increased to 115 MHz, narrowing the band of choice to the 115 MHz to 139 MHz range. Where a dedicated leakage receiver is used in combination with a CW carrier, take note of the following:

- Consult the leakage detector manufacturer to determine if the relative amplitude of the CW carrier versus equivalent analog video carrier requires adjustment to compensate for the leakage detector’s response to CW versus an analog video signal.
- The CW carrier should not be placed between two adjacent QAM channels because of the interference that could occur to those two QAM channels. Likewise, if a CW carrier is placed in a vacant channel slot that is adjacent to an active QAM channel, the CW carrier should be at least 1 MHz away from the QAM channel’s band edge to avoid interfering with the QAM channel.

Where either a CW or television video carrier is planned:

- Since 256-QAM haystacks signals are typically carried -6 dBc relative to what an analog TV channel would be if it were on the same frequency (and 64-QAM haystacks signals are typically -10 dBc), the CW carrier PEP or analog TV channel visual carrier PEP would be +6 dBc relative to a what a 256-QAM signal’s digital channel power would be if it were on the same frequency.

With this said, procedures and other information presently provided in Section 12.2 of this document remain valid and useful.

Note: Test equipment manufacturers now have available digital-compatible, multi-frequency leakage detectors. Most operate by injecting a special low-level test signal in between adjacent digital signals in the downstream spectrum. The leakage detector monitors and measures the test signal, and applies an amplitude offset correction to enable display of the equivalent field strength of the leaking digital signal. As the cable industry expands the RF bandwidth of the return path, that expanded upstream spectrum may overlap the 108 MHz to 137 MHz aeronautical band. Monitoring and measuring signal leakage field strengths of upstream signals is more complicated and difficult than downstream signals. One approach is OFDMA upstream data profile (OUDP) measurement capability for leakage monitoring/measurement of upstream orthogonal frequency division multiple access (OFDMA) signals in high-split architectures. That method is discussed in the 2020 Cable-Tec Expo Fall Technical Forum paper “Leakage In A High Split World: Detecting and Measuring Upstream Leakage Levels in a One Gpbs Symmetrical High Split Hybrid Fiber Coax Network” by John Chrostowski et al (https://www.nctatechnicalpapers.com/Paper/2020/2020-leakage-in-a-high-split-world). The method is also included in the latest version of the DOCSIS 3.1 CCAP OSSI Specification.

12.4 Longitudinal Sheath Current

Definition: Longitudinal Sheath Currents are power line neutral currents flowing longitudinally along the cable television cable sheath as a result of bonding the cable sheath to the power system neutral.

Discussion: In summary, longitudinal sheath currents are a result of the common grounding of cable television and power systems. Power line currents divide according to the parallel impedances of the power line grounding (neutral) and the cable sheath. The diverted neutral currents can be substantial and of magnitudes in the same order as the currents required for cable system powering. These currents flowing through the finite impedance of the cable sheath develop voltages that can add or subtract (depending upon relative phase) from the normal voltage powering the cable system power supplies. This can result in supply excitation voltages above or below the nominal range and hence seriously affect power supply operation. This effect was the subtle cause of many early cable system power supply failures.

The neutral current on the power system is theoretically zero; however, this is seldom the case. Unbalanced loads cause currents to flow in the neutral. For this reason, large unbalanced loads may be applied only at certain conditions or times; therefore observations should take place over extended periods in order to determine the maximum effect of the phenomenon. Since this effect is more or less steady state over limited periods (sometimes present for many hours) the result is often substantial operating increase or decrease of the applied voltage at the cable system power supply. These conditions can cause improper operation or damage to power supplies. It should be noted that this effect can occur even after a constant voltage transformer.

Required Equipment

- Clamp-on ammeter 300 A maximum capacity
- Recording facility for long time measurements

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. These measurements can be made by simply clamping the ammeter on the cable avoiding the strand and the power company neutral. This will provide a measurement of the effect at that particular moment and may well be able to correlate with a current malfunction. Longer term measurements with a recording apparatus will provide a more comprehensive look at the scope of the problem.

Chapter 13 Out-of-Service Measurements

13.1 Structural Return Loss and Cable Attenuation

Discussion: Cables have been manufactured to meet certain specified values of attenuation and cable systems are engineered using these values. If cable attenuation is higher than expected, system performance might be degraded with respect to output level, noise, distortion and dynamic range.

The accuracy of cable reel attenuation measurements is influenced by three factors: 1) the accuracy of length used in calculating loss per unit length, 2) the accuracy and precision of the test equipment, and 3) the accuracy of the temperature at which the field attenuation measurements have been made. Cable attenuation is given in manufacturers' specifications as a value at 68° F. RF coaxial cable attenuation varies (increases with increasing temperature above 68° F and decreases with decreasing temperature below 68° F) at the rate of 0.11% per degree Fahrenheit.

Measurement of Structural Return Loss and Cable Attenuation are component level tests that must always be done out-of-service. Since they are component tests, they are outside the scope of the Recommended Practices. For further information on these tests, consult the following documents:

- ANSI/SCTE 03 2016 American National Standard for Test Method for Coaxial Cable Structural Return Loss
- ANSI/SCTE 47 2007 Test Method for Coaxial Cable Attenuation

13.2 Coaxial Cable Fault Detection and Cable Length Determination

Definition: Coaxial cable fault detection (i.e. opens, shorts, impedance changes) to determine if a cable has been damaged during manufacturing, shipping or construction, and cable length determinations can be made using Time Domain Reflectometry (TDR) techniques.

Discussion: A TDR test is a qualitative measurement used only as a first test for locating damage and/or estimating cable length. It does not measure structural return loss.

Required Equipment

- Time Domain Reflectometer with
  - Scaled distance controls
  - Reflected signal amplitude control
  - Zero reference control

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. After verifying proper operational details, the following initial settings should be made:
   A. Adjust return signal amplitude (sensitivity) control to unity.
   B. Adjust the distance controls longer than the cable under test, e.g., 2500 (ft) or 25 with a multiplier setting of X100 for a 2000 foot length of cable. This accommodates adjustment of the sensitivity by positioning both the incident pulse and cable end reflection on the TDR display.

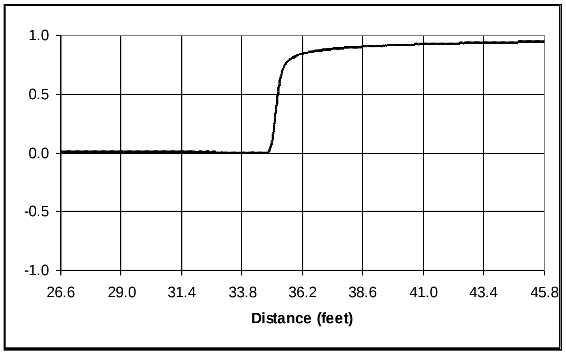
C. Adjust the leading edge of the incident pulse to two (2) divisions on the TDR display window with the zero-reference set control.
D. Adjust the cable dielectric control to match, as closely as possible, the manufacturers' specified velocity of propagation.
E. Adjust the cable impedance control equal to the expected cable characteristic impedance.
F. Adjust the signal pulse width control to deliver a 10 nanosecond (ns) pulse.

3. Connect cable under test to the TDR using an appropriate jumper cable of the same characteristic impedance as the cable being tested.
4. Adjust the reflected signal gain control so that the amplitude of the return pulse (reflection) of interest is of the same height as the incident pulse. Positive vertical deflections above the reference line indicate an increase in impedance and, likewise, negative reflections below the reference line portray a decrease in impedance.
5. To locate a point of impedance change more accurately, re-adjust the coarse distance control to a lower setting and adjust the zero reference setting as in step 2C. Turn the fine distance control until the leading edge of the reflected pulse is located on the incident pulse vertical graticule reference line. The distance dial reading times an existing multiplier setting gives the approximate length from the cable connector to the reflection. For cable length measurement, the open end of the cable provides the impedance discontinuity that reflects the pulse.

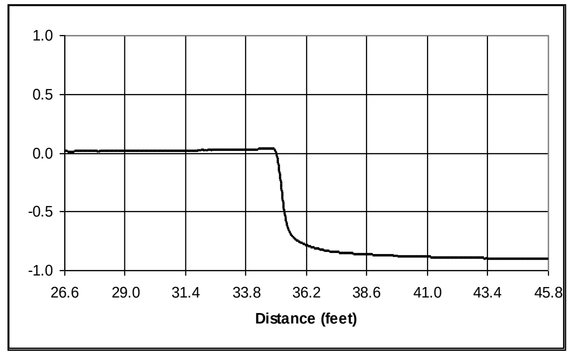
Note: Distance measurement accuracy is influenced by two factors:

1) Accuracy and stability of the TDR being used, and
2) Accuracy of the assumed cable impedance and velocity of propagation used for setting the TDR.

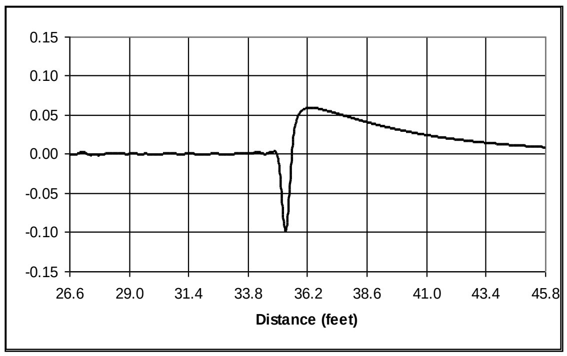
A ± 2 % distance error is normal and should be expected. Testing from both ends of the cable and averaging will improve the accuracy of the measurement. Figure 13-1 illustrates typical TDR measurement results for an open circuit, short circuit and impedance change.



Open Circuit



Short Circuit



Impedance Change

Figure 13-1: Example TDR Measurement Results

13.3 Automatic Gain Control Regulation of a Headend Heterodyne Processor or Demodulator

Definition: Automatic Gain Control Regulation is the accuracy, expressed in dB, to which the output level of each carrier under control is held for variations in its input level.

Required Equipment

- Television signal source at the input frequency of the processor or demodulator, with an output capability of at least 30 dBmV. A CW signal generator may be used if the unit under test uses peak AGC detection (see discussion below).
- Attenuator with 1 dB resolution and at least 60 dB of range
- Signal level meter or spectrum analyzer
- Signal source (either a video waveform generator and modulator, or over-the-air as described below)
- Oscilloscope or waveform monitor (required only for demodulator measurements)

Equipment Configuration

Figure 13-2a illustrates the test setup to be used with processors. The signal source, discussed below, is connected to the device under test (DUT) through the attenuator, which is used to vary the input level to the DUT. The preferred indicating instrument is usually a spectrum analyzer with settings as shown in the figure, but a signal level meter may be used. Connection A is used to calibrate the test setup, and connection B is used to make measurements.

Figure 13-2b and c show two alternative signal sources that can be used. Figure 13-2b is the preferred signal source. A video waveform generator is connected to a modulator. The preferred video waveform is a 10/90% bounce signal, but any waveform should be suitable for ordinary testing. Most modulators consist of an IF section, with individual outputs for the picture and sound carriers. The sound carrier has a level control facility shown as AT2, after which it is combined with the picture carrier, and the two are frequency converted to the output channel. If all of this facility is available in a suitable modulator, it may be used as-is. The sound carrier level control may be an adjustment on the front panel. It can be used but you will have to connect the spectrum analyzer (or signal level meter) as in connection A of Figure 13-2a each time you adjust the sound carrier level.

Figure 13-2c shows an alternate signal source that may be used if a modulator and video waveform generator are not available. An over-the-air signal is used, but it is necessary to trap the sound carrier and insert a carrier from a CW generator in order to test the processor for sound carrier level stability.

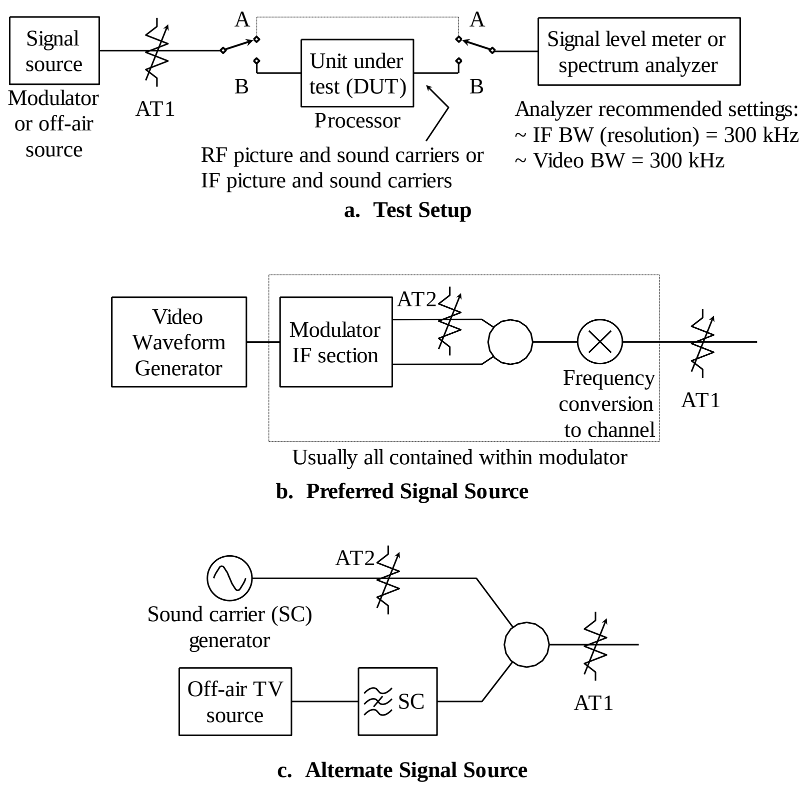


Figure 13-2: Test Configuration for Measuring AGC Performance of Headend Processors

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Determine from the manufacturer’s literature the minimum and maximum signal levels over which the unit under test is to operate. Start with the attenuator set so that the signal you supply is the average of the highest and lowest specified signal levels.

3. Connect the spectrum analyzer or signal level meter as at A of Figure 13-2a. Set AT1 so that you can adjust the picture carrier level between the nominal level and the highest and lowest levels

you wish to test. Set the picture carrier level to the center of the manufacturer's specified range. Set the sound carrier to the nominal desired level going into the processor. When the processor is used for over-the-air reception, this level will usually be 7-10 dB down from the picture carrier.

4. Connect the equipment as at B of Figure 13-2a. Use the RF or IF output of the DUT according to the test desired. Usually, you will be more interested in the stability of the RF output, but for some applications the stability of the IF output will be important. Tune the spectrum analyzer or signal level meter to the appropriate carrier frequency to measure the output level of the picture or sound carrier.
5. Adjust controls on the processor to set the output sound and picture carrier levels in the range specified by the manufacturer. Usually, you will be most interested in performance near the upper end of the range specified. Normally the sound carrier is set -15 dB from the picture carrier amplitude.
6. While monitoring the picture (or sound) carrier level, use AT1 to reduce the amplitude of the input signal to the minimum specified by the manufacturer. Then use AT1 to increase the amplitude of the input signal to the maximum specified by the manufacturer. Record any variation in picture carrier or sound carrier level. If unacceptable variation occurs, you may wish to adjust AT1 over finer steps to see what levels cause unacceptable operation. If appropriate, measure the output level as you vary the input level in smaller steps (e.g., 5 dB) over the specified range of the processor.
7. Tune to the sound carrier output and adjust the incoming sound carrier (relative to the picture carrier) level over the specified range and note any level change in the output sound carrier level.

Testing a Demodulator

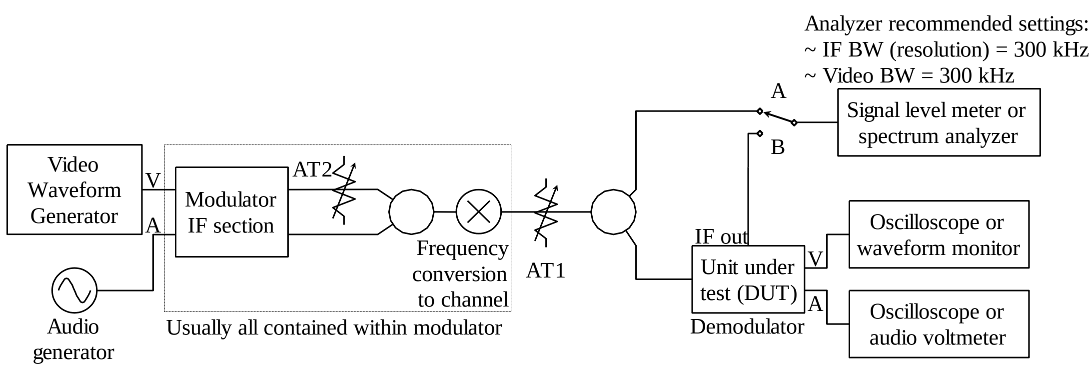


Figure 13-3: Test Configuration for Demodulator

Figure 13-3 illustrates the test configuration for a demodulator. The signal source is identical to that used to test a processor. Most demodulators have a composite IF output, which may be checked using the B hook-up to the spectrum analyzer or signal level meter. Otherwise, the important thing is to ensure that the video signal level does not change with changes in input signal level. This is measured using the oscilloscope or waveform monitor shown.

Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing. Set the waveform monitor to view an overlay of all video lines (usually this is called a line rate display), and observe the peak-to-peak amplitude of the signal as you adjust AT1 as described above.

If you are using live video as a test source rather than a signal generator, you can observe the peak-to-peak amplitude of the sync pulses rather than the peak-to-peak amplitude of the video signal. Convert the amplitude change to decibels using the formula

$$\text{Change} = 20 * \log \frac{\text{amplitude1}}{\text{amplitude2}} \text{ dB},$$

where amplitude 1 is that measured after changing the setting of AT1 and amplitude 2 is that measured with the attenuator set at nominal input level, as defined by the demodulator manufacturer.

An audio signal generator and oscilloscope or voltmeter is shown for measuring variation in sound with signal level. Normally this is not an issue, since the amplitude of the audio output is nominally independent of signal level in an FM system. The method for testing is shown, though under most circumstances you will not need to make this measurement.

13.4 Bit Error Ratio (BER)

Definition: . Bit Error Ratio (BER) is the ratio of bits in error to the total number of bits transmitted, received, or processed. Often referred to as bit error rate, though nontemporal. Bit error ratio is accepted as a standard metric of performance used to measure digital communications networks. Because of this it is used to measure performance of digital services in cable television networks.

Discussion:

Bit error ratio is commonly measured using a BER test set. A data source transmits a bit pattern – usually a pseudo-random binary sequence – through the device, system, or network being tested. The error detector has to either reproduce the original bit pattern or directly receive it from the bit pattern generator. The error detector compares bit-by-bit the original bit pattern with the one received from the device, system, or network being tested. Unless otherwise stated, BER requirements and tests (since at least the 1990s) apply to bit streams which are then encoded with FEC prior to transmission and decoded in or after the receiver; the BER requirements and tests on the received data are applicable after the FEC decoding. If there are measurements on received bits prior to complete FEC decoding this should be made explicitly clear in a discussion or report. The method just described is usually an out-of-service test, making it impractical to perform where service disruptions are not acceptable. How, then, are BER measurements performed in operating cable networks?

Field meters used by the cable industry don’t perform BER measurements the same way that a BER test set does. Instead, field meters perform an in-service measurement using an internal algorithm to derive a BER estimate based upon what the forward error correction is doing. The terms pre-FEC BER and post- FEC BER are widely used by cable test equipment companies and cable operators, with some variations in the terminology (e.g., “pre-FEC,” “pre BER,” or just “pre”). Many understand the terminology to generally mean the BER before and after FEC decoding fixes errors. However, some clarification is in order, since this assumption isn’t quite correct.

In a typical field meter being used to measure an ITU-T J.83 Annex B SC-QAM signal (e.g., DOCSIS 3.0 or earlier), what is called pre-FEC BER is estimated at the input to the Reed Solomon (RS) decoder but after the Trellis decoder, descrambler (de-randomizer), and de-interleaver. Post-FEC BER is estimated after Reed Solomon decoding. The estimates of BER made by the in-service decoders for ITU-T J.83 Annex B downstream SC-QAM signals are imperfect approximations (but are adequate for routine maintenance and troubleshooting in cable networks). In the BER test sets used for out-of-service testing, received bit errors are actually counted.

BER provides only a quantitative indication of system performance. It does not provide any information regarding the cause of the errors. Therefore, other parameters such as RxMER, Error Vector Magnitude (EVM), adaptive equalizer tap analysis and constellation analysis should be used in conjunction with BER measurements.

13.4.1 BER Measurement Using Random Data

Required Equipment:

- Pseudorandom data generator (may be contained in the digital modulator)
- Digital modulator with IF output (may contain an internal pseudorandom data generator)
- Digital demodulator and bit error ratio tester (BERT) or digital signal analyzer with BER measurement capability
- RF generator with capability to output a TTL/ECL clock at frequencies up to 50 MHz
- Digital capable RF upconverter

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Set up the equipment as shown in Figure 13-4.[9]
3. Select the desired modulation format.
4. Select the desired data pattern. A PRBS-15 or PRBS-23 pattern is recommended.
5. If an external BERT is used as input to the modulator, set the RF generator to the correct clock frequency. See Table 13-1 on the digital modulator, select the desired error correction. For 64-QAM and 256-QAM, this will be specified in ITU-T J.83, Annex B. The interleaving delay must also be selected.
6. Using a spectrum analyzer or digital signal analyzer, verify that the output of the digital modulator is set to the correct level for input to the RF upconverter. Refer to the manufacturer’s specifications for signal input level.
7. Connect the digital modulator to the RF upconverter. If applicable, adjust the upconverter to the digital mode. Using a spectrum analyzer or digital signal analyzer set the upconverter output to the optimum level. Connect the upconverter output to the system’s headend combiner through appropriate padding to produce the desired level in the system.
8. Connect the digital demodulator and BERT or the digital signal analyzer to the test point at the receive site. This will normally be at the end of a drop or at a point within the customer premises.
9. Tune the digital demodulator or digital signal analyzer to the desired channel and select bit error ratio mode on the BERT or the signal analyzer.
10. Allow the test equipment to average the BER over a minimum of 15 minutes or 100 errored seconds, whichever comes first. Read and record the average BER.

[9] The digital modulator and digital demodulator in the figure include FEC encoding and decoding respectively.

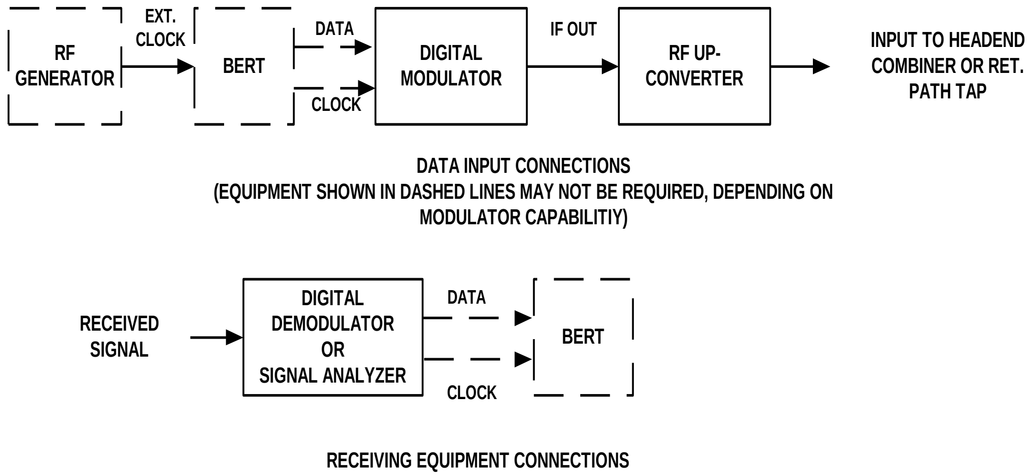


Figure 13-4: Equipment Configuration for BER Test

Table 13-1: BERT Clock Frequencies

Modulation Format	Clock Frequency (MHz)
QPSK (per DVS-167, Table 1)	1.488
QPSK (per DVS-167, Table 2.2)	2.976
QPSK (per DOCSIS 1.1)	See Table 13-2
16-QAM (per DOCSIS 1.1)	See Table 13-2
64-QAM	26.97035
256-QAM	38.8772

DOCSIS 1.1 specifies both QPSK and 16-QAM modulation for use on the return path. In addition, ten different levels of error correction or no error correction are permitted. Also, return path transmission operates in burst mode. Since BER measurement in burst mode is not practical, return path transmission at DOCSIS data rates should be done using continuous data. It will also be necessary to use a digital modulator that provides the same type of error correction that is specified by DOCSIS 1.1 (i.e. – Reed Solomon error correction with values of t = 1 to t = 10, where t = half the number of Reed Solomon parity bytes). Clock frequencies for DOCSIS 1.1 are shown in Table 13-2.

Table 13-2: BERT Clock Frequencies for DOCSIS 1.1 Return Path Data Rates

Symbol Rate (ksps)	QPSK Data Rate (kbps)	16-QAM Data Rate (kbps)	Clock Frequency (kHz)
160	320	640	See Note
320	640	1280	See Note
640	1280	2560	See Note
1280	2560	5120	See Note
2560	5120	10240	See Note

Note: For all data rates in Table 13-2, the clock frequency is calculated as follows:

Clock Frequency = Data Rate – 16t

The quantity 16t is used since there are 8 bits per byte and t is half the number of Reed Solomon parity bytes. For example, a 16-QAM transmission at a 640 kHz symbol rate using Reed Solomon t=4 error correction would require a clock frequency of 2560 – 16x4 = 2496 kHz. For no error correction, use the data rates as shown in Table 13-2.

13.4.2 Digital Signal BER Measurement Using Estimates from In-Service Decoders

Discussion: This method is presented as an alternative to measuring BER with pseudorandom data.

Required Equipment

- Digital signal, originating from a headend device such as an Integrated Receiver Transcoder, and up-converted for transmission over the cable multiplex
- Digital signal analyzer with BER measurement capability

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.
2. Connect the digital signal analyzer to the test point at the receive site. This will normally be at the end of a drop or at a point within the customer premises.
3. Tune the digital signal analyzer to the desired channel and select bit error ratio mode on the signal analyzer.
4. Allow the test equipment to average the BER over a minimum of 15 minutes or 100 errored seconds, whichever comes first. Read and record the average BER.

13.5 Dynamic Range

13.5.1 Bit Error Ratio (BER) in the Noise Notch (BNN)

Definition: The Bit Error Ratio in the Noise Notch technique is a performance-based procedure, which defines the digital operating window for a device over a specific range of input operating levels. This window is limited by two parameters, device carrier-to-noise on one extreme and by device compression or clipping, on the other.

By using an Additive White Gaussian noise (AWGN) source to simulate a fully loaded bandwidth condition, the device under test is evaluated via the bit error ratio performance of a digital signal, which is inserted into a notched portion of the noise source bandwidth.

Discussion: In the past, analog measurements comprised of single tone and multiple tones have been used successfully to determine the linearity of multiple channel analog cable television systems.

Whereas these methods are well documented, repeatable and time proven, these techniques are less than ideal for predicting digital performance of a device (or system) under test.

The intent of this technique is to measure true bit by bit performance of a return path digital signal under a range of worst case conditions. To accurately simulate the fully loaded digital band, full bandwidth loading is used. This is achieved by using a AWGN source, which is band shaped for the device to be tested. The use of a noise source as a full band signal accurately simulates the high peak to average power ratios encountered when using complex digital modulation techniques.

Within this bandwidth a bandstop filter is cascaded to this full band noise signal to provide a clean stopband. Within this stopband, an evaluation digital signal will be examined for bit error ratio. By choosing a digital signal which is representative of the most complex modulation type anticipated to be used in the signal path, a realistic worst case operational window can be determined. This information provides important design elements for future network planning and it is also useful in determining operational constraints of existing networks.

The procedure as described is for field use, which requires two BER analyzers. In a laboratory environment a single bit error ratio analyzer will suffice.

Required Equipment

- AWGN signal source
- 5 – 42 MHz bandpass Filter
- Two-way signal splitter/combiner
- Precision variable attenuator (1 dB step capable)
- Hybrid gain block 5-200 MHz
- T9 Bandstop Filter
- T9 Bandpass Filter (with bandpass larger than channel to be tested)
- Bit error ratio analyzer
- Precision modulator for QPSK and 16-QAM
- Precision demodulator for QPSK and 16-QAM
- Spectrum analyzer with channel power function

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. See Figure 13-5 for the test setup. An AWGN source is used to generate a bandwidth of flat noise for the given return bandwidth (i.e., 5 to 42 MHz).

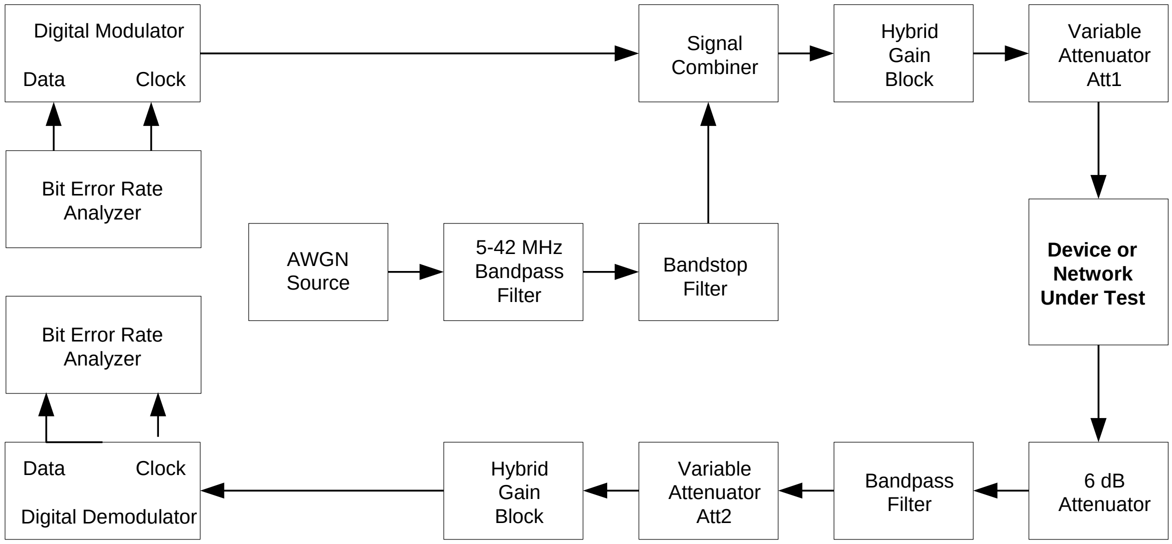


Figure 13-5: Block Diagram for BNN Measurement

3. The noise is bandstopped (trapped) in the center of this bandwidth using a suitable filter, which has sharp skirts and deep depth. The width of this bandstop should be larger than the channel width of the digital channel under evaluation, but not so large as to significantly remove the passband (i.e., <20%). The depth of the trap in decibels should be 10 dB greater than the minimum carrier-to-noise necessary for near error free transmission.

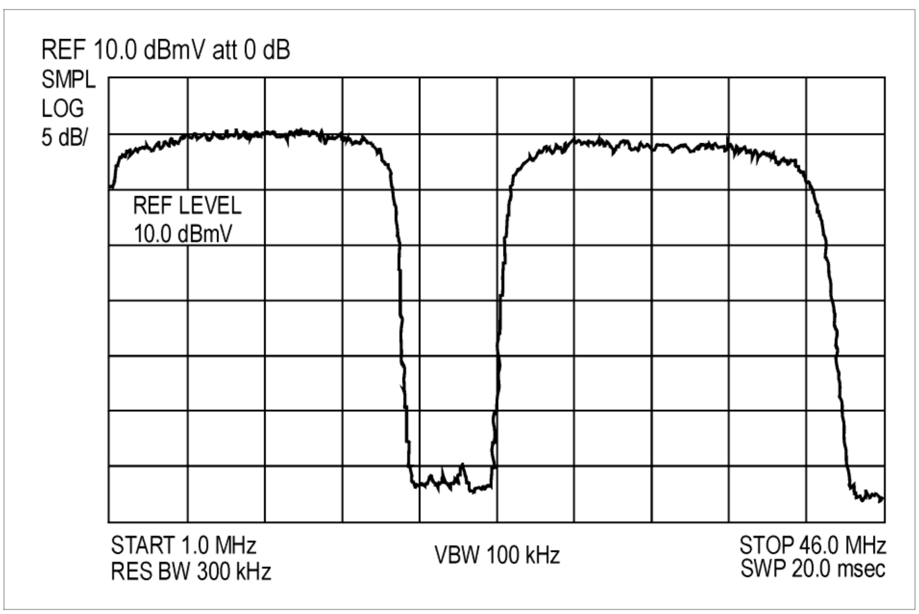


Figure 13-6: Bandstopped Flat Noise Source

4. Connect the reference digital test signal with its center frequency adjusted to fall in the center of the "trapped" bandwidth. The channel power of this test signal (as measured across its functional bandwidth, i.e., 1.6 MHz) should be set to be of the same channel power of the noise in that same bandwidth (i.e., 1.6 MHz).

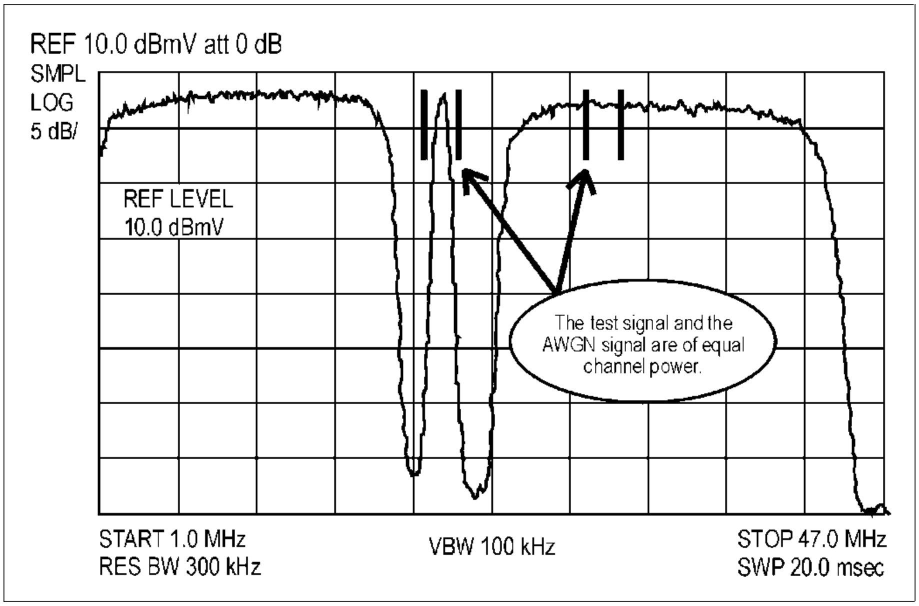


Figure 13-7: Combined Input Test Signal with Bandstopped Noise Source and Digital Test Signal

The digital signal under evaluation must be operated in a specific digital format (i.e. QPSK or 16-QAM) using a known Pseudorandom Binary Sequence (PRBS) which is provided by a suitable bit error ratio analyzer. The RF modulator and demodulator should operate with nearly theoretical performance especially with regard to carrier-to-noise. The modulator and demodulator should be operated with FEC disabled so as to determine true BER performance. The combined test signal is used as a signal source to the device under test with the digital signal being evaluated to determine the digital operational window for the device or network under test.

5. Record the BER as a function of total input power to the device or network under test and, referring to Figure 13-8 as follows:
   a) Adjust ATT1 to an input level to the device or system under test that is recommended by the manufacturer. See Section 16.4: "Return Plant Setup and Operational Practices" for more information on selecting the proper input level.
   b) Adjust ATT2 for the recommended input to the demodulator.
   c) Record the BER for either a period of two minutes or 100 error events, whichever occurs first. Record the associated total input power.
   d) Increase the combined input test signal by 1 dB (ATT1) and repeat steps b and c. Continue this step until a BER of 10^{-3} or loss of synchronization of the demodulator occurs. Remember to adjust ATT2 to maintain constant level to the demodulator. This completes the "clipping side" of the dynamic range curve.
   e) Return the combined input test signal as in step a. Decrease the combined input test signal by 1 dB and repeat steps b and c. Continue until a BER of 10^{-3} or loss of synchronization of the demodulator occurs. Remember to adjust ATT2 to maintain constant level to the demodulator. This completes the "noise side" of the dynamic range curve.

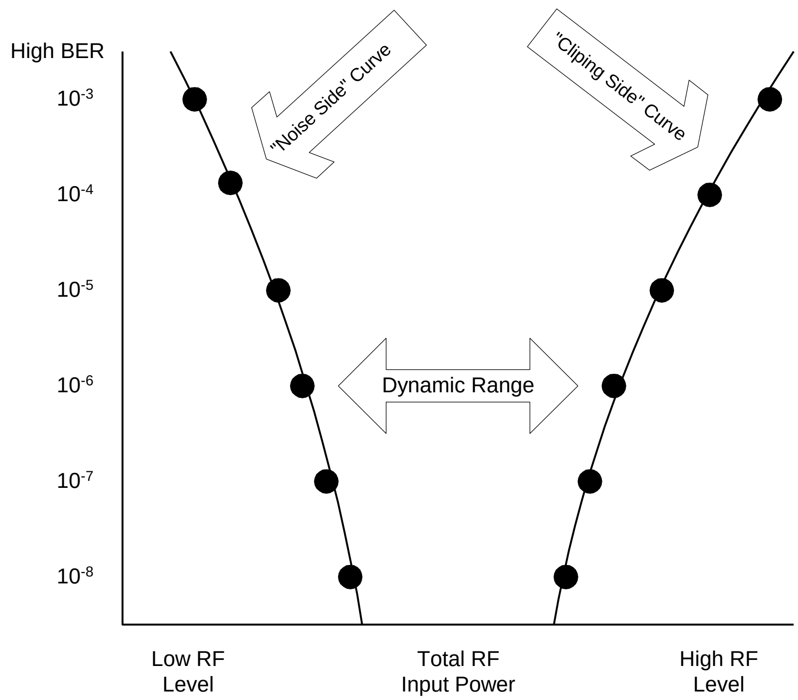


Figure 13-8: Dynamic Range Plot

Notes, Hints and Precautions

If no errors are encountered in the center of the operational range, this area is plotted as error free.

The digital operational window is measured as the dB range of total RF input power that is measured between some set BER, for example, 10^{-6} for QPSK and 16-QAM.

This test may be repeated at several different frequencies to fully characterize the bandwidth.

Results of this test will vary with the amount of uncontrolled interference that enters an operational system. This is most noticeable on the noise side of the dynamic range curve and will vary with time of the day and operating frequency.

13.5.2 Noise Power Ratio (NPR)

Definition: Noise Power Ratio (NPR) is a test method that examines the amount of noise and intermodulation distortion in a channel. The test signal is comprised of flat Gaussian noise waveform that is band-limited to the frequency range of the reverse path with a narrowband channel of the noise deleted by a notch filter. NPR is defined as the level of the signal relative to the level of the combined noise and intermodulation distortion in the channel. Essentially, NPR is the depth of notch.

Note: This procedure is only recommended for out-of-service measurements. These measurements will use the entire return spectrum and will cause disruption to any traffic that might exist on the plant during the tests. Before performing this test, the plant must be properly set up and aligned. For more information refer to Section 16.4: "Return Plant Setup and Operational Practices".

Discussion: The test is performed by injecting a block of noise into the plant (normally at the end of line), and measuring the resulting distortion at the headend. So that the level of the distortion can be measured, the injected noise has a notch in the spectrum, usually close to the middle. The injected noise is referred to as the "signal" (see Figure 13-9).

The notch (channel) will contain many types of noise. When the signal goes through nonlinearities (such as amplifiers, lasers, and photodetectors), distortion and clipping will result, which cause the signal to beat with itself. The result is a wide spectrum of noise that will build up under the signal (see Figure 13-10) and will be visible in the notch. This resulting noise (caused by distortion and clipping) is referred to as Intermodulation Noise (IMN). This intermodulation noise will increase as the signal level is increased. The ratio of the injected signal level to the level of the intermodulation noise is referred to as Carrier-to-Intermodulation Noise (CIN). Some of the noise under the signal, such as thermal noise, laser RIN, and shot noise, will not change with signal level.

When the signal level is high enough to cause significant distortion, intermodulation noise will dominate the noise level in the notch. When the signal level is low, thermal noise (along with other constant noises such as laser RIN and shot noise) will dominate. This procedure cannot differentiate between the two, since both appear as noise. Therefore, the resulting Noise Power Ratio will always contain the sum of both effects. The range of input levels that produces a desired NPR is called the "Dynamic Range" of the system. See Figure 13-11 for an example of a dynamic range plot.

Noise is used as the signal because, unlike analog AM video signals, which have a lot of energy in the carrier, complex digital modulations have a noise like spectrum. A full loading of these signals will closely resemble a block of noise, just as a forward spectrum fully loaded with AM video resembles a comb of CW carriers. For more information on the use of noise as a measurement signal, refer to Section 2.2.3: "Peak-to-Average Ratio" and Section 16.2: "Peak Voltage Addition".

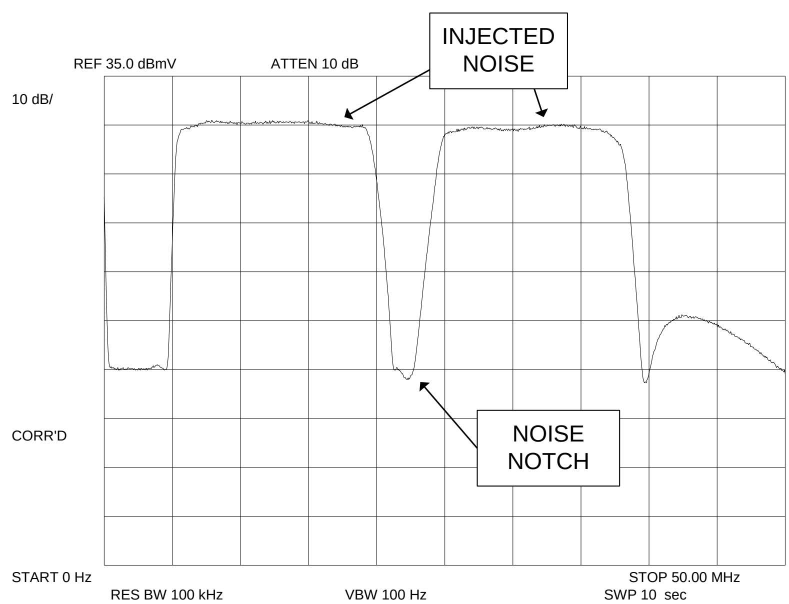


Figure 13-9: Spectrum of Injected Signal

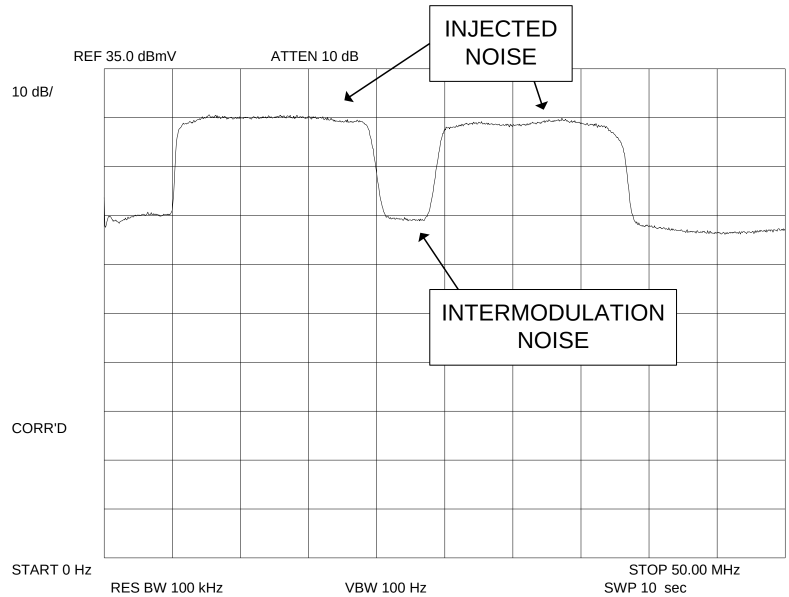


Figure 13-10: Signal with Intermodulation Distortion

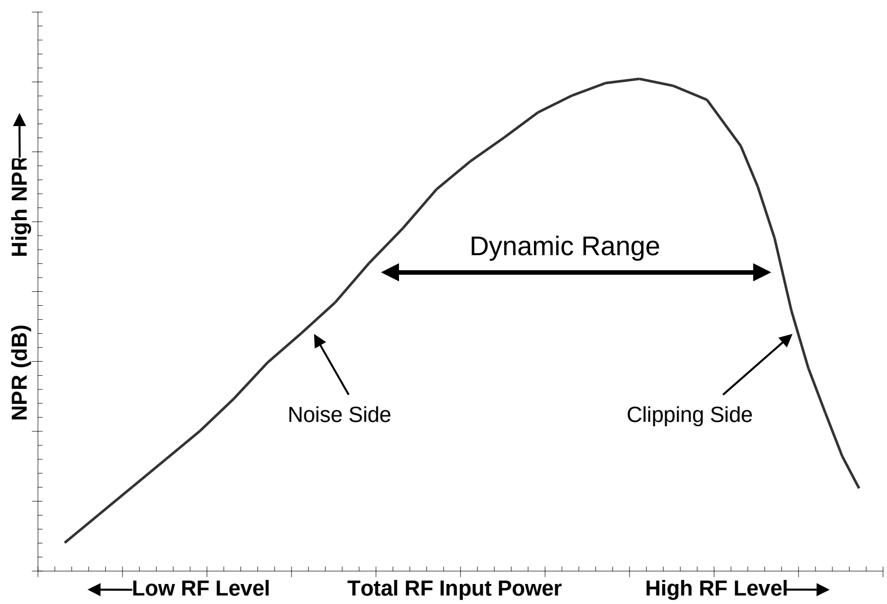


Figure 13-11: NPR Dynamic Range Example

Required Equipment

- Additive White Gaussian Noise (AWGN) source with a bandwidth at least as wide as the return spectrum
- 5 to 42 MHz bandpass filter
- Bandstop (notch) filter centered at mid-band frequency (approx. 22 MHz)
- Two-way signal splitter/combiner
- (2) Precision variable attenuators (0.5 or 1 dB step capable)

- Self terminating A/B switch
- RF power meter or spectrum analyzer with integrated channel power measurement capability
- Spectrum analyzer

Optional Equipment

- Amplifier, if required to obtain sufficient level at the injection point

Test Procedure

1. Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing.

2. Connect the injection equipment to the plant as shown in Figure 13-12. The setup illustrates a 5-42 MHz return plant. If the return plant has a different bandpass, then different filter frequencies should be used.

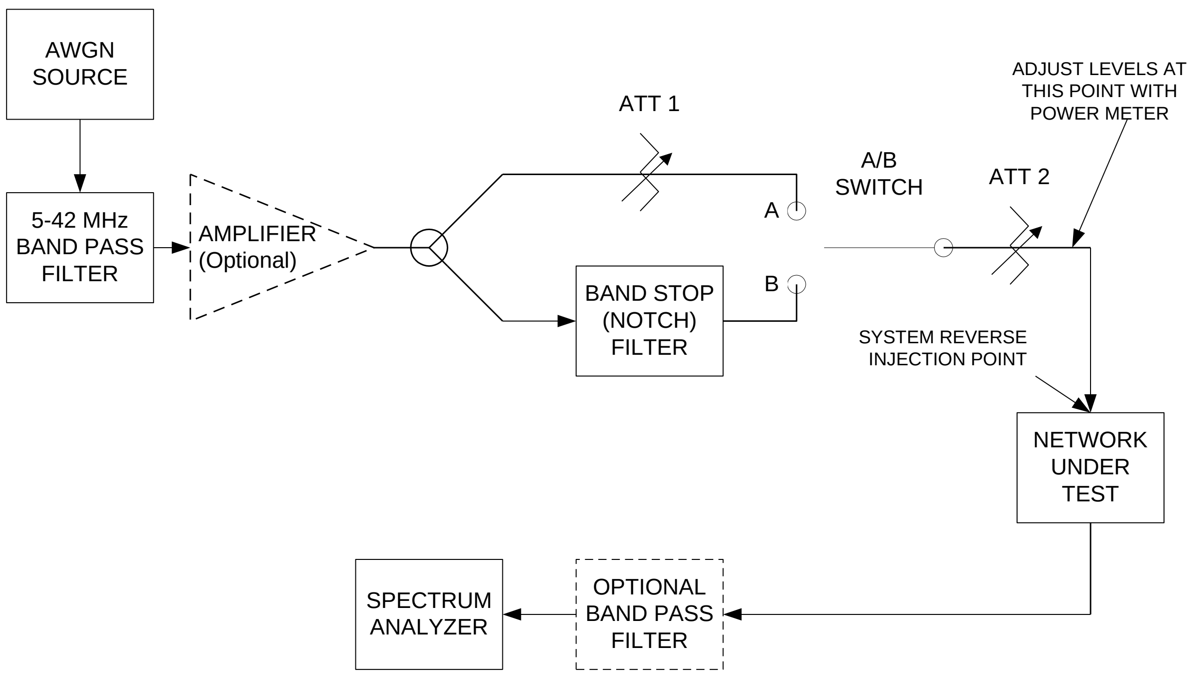


Figure 13-12: Equipment Connection

3. Turn on the noise source and connect the power meter to the output of ATT2 with ATT2 adjusted to minimal attenuation. Adjust ATT1 so that the power when the switch is in position A is equal to the power in position B.The channel power measurement function of a spectrum analyzer may be used for this step instead of the RF power meter provided that the spectrum analyzer is adjusted to accurately measure the total power across the entire return bandwidth.

4. Adjust ATT2 to the proper total power for the system reverse injection point (see “Notes, Hints and Precautions” for more information). In general, the proper attenuator setting will produce the level that hits the amplifiers (and the laser) at the nominal total power. See Section 16.4: “Return Plant Setup and Operational Practices” for more information on selecting proper signal levels for the return plant.

5. Connect the spectrum analyzer to the output of the system.

6. Set the spectrum analyzer as follows:
   - Center Frequency: Center of notch
   - Frequency Span: Approx. 10 MHz (see text)
   - Amplitude Scale: 10 dB/div
   - IF Resolution Bandwidth: Approx. 100 kHz (see Notes and Hints)
   - Sweep Time: Automatic for calibrated measurement
   - Input Attn: As required (see text)

The remaining steps in the procedure vary depending on the capability of the spectrum analyzer. Two methods are discussed below.

PROCEDURE 1 - Noise Marker (Preferred Method)

7. This procedure requires a spectrum analyzer with a noise measurement mode (sometimes referred to as a noise marker or a noise normalization).
8. Turn on the noise measurement marker.
9. Put the analyzer into the “sample” detection mode. (Note: On many analyzers, activating the noise measurement mode will automatically put the analyzer into the “sample” detection mode.)
10. Set the A/B Switch to position B.
11. Set the video filter to automatic (or off) and adjust the reference level so that the peaks of the noise never exceed the top of the display. The input attenuator should be in the automatic mode during this adjustment.
12. Measure the noise level by placing the marker in the center of the notch. Adjust the input attenuator of the spectrum analyzer high enough so that the analyzer does not cause additional intermodulation noise, and low enough so that the analyzer does not cause additional thermal noise. Do this by measuring the noise level in the bottom of the notch and adjusting the attenuator for a minimum reading. If a sufficient notch depth cannot be obtained, the optional bandpass filter (centered at the notch frequency) might be required to minimize spectrum analyzer distortion. If the marker reading is not stable, turn on video averaging or use a lower video bandwidth. (Note: On some spectrum analyzers, the noise marker reads all frequencies within about ±½ graticule of the marker location. Therefore, if the notch is not at least one graticule wide, readjust the frequency span setting.)
13. Verify that the analyzer noise floor is sufficiently below the noise notch level by either externally attenuating the signal (at the analyzer input) by at least 30 dB or by disconnecting the signal and terminating the analyzer input. If the drop is not at least 10 dB, correct the noise level recorded in step 12 by using the noise-near-noise correction table given in Figure 3-1.
14. Set the A/B Switch to position A and measure the signal level by placing the noise marker on the signal at the same frequency. If the marker reading is not stable, turn on video averaging or use a lower video bandwidth.
15. Calculate the NPR by subtracting the corrected noise level (step 13) from the signal level (step 14).

    NPR = Signal Level - Noise Level

    Note: When measuring noise with a spectrum analyzer, correction factors are usually required to correct for bandwidth; however, since this measurement is a relative measurement of one noise level vs. another noise level, corrections are not required.

16. Decrease ATT1 and repeat steps 10 through 15 to measure the clipping side of the dynamic range.
17. Return ATT1 to the initial level and then increase ATT1 settings and repeat steps 10 through 15 to measure the noise side of the dynamic range.

PROCEDURE 2 - Normal Marker (Alternate Method)

7. Put the analyzer into the “sample” detection mode.
8. Set the A/B Switch to position B.
9. Set the video filter to automatic (or off) and adjust the reference level so that the peaks of the noise never exceed the top of the display. The input attenuator should be in the automatic mode during this adjustment.
10. Measure the noise level by placing the marker in the center of the notch. It will probably be necessary to turn on video averaging to obtain a stable measurement. Adjust the input attenuator of the spectrum analyzer high enough so that the analyzer does not cause additional intermodulation noise and low enough so that the analyzer does not cause additional thermal noise. Do this by measuring the noise level in the bottom of the notch and adjusting the attenuator for a minimal reading. If a sufficient notch depth cannot be obtained, the optional bandpass filter (centered at the notch frequency) might be required to minimize spectrum analyzer distortion. If the marker reading is not stable, turn on video averaging or use a lower video bandwidth.
11. Verify that the analyzer noise floor is sufficiently below the noise notch level by either externally attenuating the signal (at the analyzer input) by at least 30 dB or by disconnecting the signal and terminating the analyzer input (remember to restart video averaging if it is on). If the drop is not at least 10 dB, correct the noise level recorded in step 10 by using the noise-near-noise correction table given in Table 3-1.
12. Set the A/B Switch to position A and measure the signal level by placing the noise marker on the signal at the same frequency. If the marker reading is not stable, turn on video averaging or use a lower video bandwidth.
13. Calculate the NPR by subtracting the corrected noise level (step 11) from the signal level (step 12).

    $$NPR = \text{Signal Level} - \text{Noise Level}$$

    Note: When measuring noise with a spectrum analyzer, correction factors are usually required to correct for noise bandwidth and log detection errors. However, since this measurement is a relative measurement of one noise level vs. another noise level, corrections are not required.
14. Decrease ATT1 and repeat steps 8 through 13 to measure the clipping side of the dynamic range.
15. Return ATT1 to the initial level and then increase ATT1 settings and repeat steps 8 through 13 to measure the noise side of the dynamic range.

Notes, Hints and Precautions

1. The notch in the signal from the setup in Figure 13-12 should be as deep as possible. The dynamic range of this measurement is limited by the depth of the notch. The maximum NPR measured will never be greater than the notch depth. In general, the notch should be at least 10 dB deeper than the desired measurement.

2. The width of the notch at the depth of interest must be at least as wide as the resolution BW setting on the analyzer. For instance, to use the recommended RBW of 100 kHz, the notch should be at least 100 kHz wide at the depth of interest.
3. Some NPR measurement procedures use the notched noise signal for all measurements, with the NPR being measured by comparing the level adjacent to the notch to the level inside the notch. That method has the potential for the following errors and should be avoided.
    a) The result is dependent on the flatness of the system being measured.
    b) The result depends on the repeatability of the marker frequency chosen from measurement to measurement.
    c) The power density of the noise signal with the notch is higher than the power density of the noise signal with no notch. Using the power density of the signal with a notch as the signal level reference leads to optimistic results.
4. If additional signal level is required, consider choosing a different injection point that has less loss to the nearest return path amplifier. The lowest loss location to inject into the plant is typically at a low value tap port.
5. If an amplifier is required to obtain sufficient signal level, the amplifier should be placed between the bandpass and notch filters as shown in Figure 13-12. No amplifier should be used after the notch filter because its distortion will reduce the depth of the notch. Be certain that any amplifier used has sufficient capability to produce the required level without compression. Any amplifier that follows the filters can reduce the peak-to-average ratio of the injected noise, which could produce unrealistically “good” results.
6. The injected noise (signal) should be flat. A level variation less than 2 dB across the band is recommended.
7. This procedure will not indicate whether noise or distortion is limiting the dynamic range of the system. For instance, a plant which is not correctly aligned for unity gain could simultaneously suffer from thermal noise in some spans and from CIN in other spans. To troubleshoot, first assure that the system is aligned correctly (see Section 16.4: “Return Plant Setup and Operational Practices” for more information). If performance is still less than desired, the system must be measured in smaller sections to determine where the problems are occurring.
8. Other contributors to distortion such as ingress and hum modulation might exist in the system and are discussed in other sections. If such impairments exist, they will add to the level in the notch and will lower the NPR result.
9. If the above procedure does not provide adequate dynamic range for measurements of high NPR, a bandpass filter may be added in front of the spectrum analyzer.

Chapter 14 Other Definitions

14.1 Bandwidth

Definitions

Channel Bandwidth: Channel bandwidth (or allocated bandwidth) is defined as the width of the total frequency spectrum allotted for signal transmission. For example, in U.S. cable systems, the forward path channel bandwidth for analog NTSC and ANSI/SCTE 07 SC-QAM signals is 6 MHz.

Occupied Bandwidth:

Generally speaking, “occupied bandwidth” is the width of the spectrum that actually contains the signal. Depending on context, any of several definitions may be applicable. For example, Rec. ITU-R SM.443-4 Annex 1 "Bandwidth measurement at monitoring stations" says “Unless otherwise specified in an ITU-R Recommendation for the appropriate class of emission, the value of β/2 should be taken as 0.5%.” The FCC rules in §15.403 state “For purposes of this subpart the emission bandwidth is determined by measuring the width of the signal between two points, one below the carrier center frequency and one above the carrier center frequency, that are 26 dB down relative to the maximum level of the modulated carrier.” The FCC definition is the "x dB” method, which is from Annex 2 of ITU-R SM.443-4.

With respect to DOCSIS signals, the following applies:

Occupied Bandwidth is terminology used in DOCSIS 3.1 PHY specifications, to a) define and describe the amount of spectrum “occupied” by downstream OFDM signals, and to b) describe and define the amount of spectrum “occupied” by the combination of OFDMA signals and legacy SC-QAM upstream DOCSIS signals.

For legacy DOCSIS SC-QAM channels, both upstream and downstream, the concept of “occupied” bandwidth is straightforward because the theoretical square-root raised cosine pulse shaping provides a well-defined non-zero power spectral density (PSD) for the signals. The spectrum with non-zero PSD is 1.25 times the symbol rate for the upstream DOCSIS SC-QAM signals, and is 6 MHz for the downstream 64-QAM and 256-QAM DOCSIS SC-QAM signals.

For OFDM, the theoretical non-zero PSD extends infinitely, therefore to provide a practical assignment for “occupied bandwidth,” the DOCSIS 3.1 PHY specification defined “modulated spectrum” for a defined amount of spectrum for each active (energized) subcarrier, and also defined a “taper” spectral region beyond the modulated spectrum to describe significant sidelobe energy present. In practice with OFDM, in addition to the theoretically ideal or perfect transmission, there will be spurious emissions. But just considering the theoretical PSD of the OFDM transmission, the DOCSIS 3.1 PHY spec defines the “taper region” which adds to the amount of modulated spectrum, which then serves to determine the “occupied spectrum” in the downstream. For purposes of defining occupied spectrum in the downstream, the PSD of the theoretical OFDM transmission beyond the taper region and modulated spectrum is ignored, so it is “neglected spectrum” for that OFDM signal.

For the DOCSIS upstream definition of occupied spectrum, with OFDMA, only the modulated spectrum is considered (spectrum defined for each energized subcarrier). There is no concept of “taper” spectrum for the upstream occupied bandwidth.

The DOCSIS definitions for occupied bandwidth are:

1) Downstream - The sum of the bandwidth in all standard channel frequency allocations (e.g., 6 MHz spaced CTA channels) that are occupied by the OFDM channel. The CTA channels which are occupied by the OFDM signal are those which contain any of the Modulated Spectrum and/or taper region shaped by the OFDM channels' transmit windowing, where the values for the taper regions are defined in Appendix V of the DOCSIS 3.1 PHY specification as a function of the Roll-Off Period. It is possible, but not problematic, for a CTA channel to be "occupied" by two OFDM channels.

2) Upstream - a) For a single OFDMA channel, the sum of the bandwidth in all the subcarriers of that OFDMA channel which are not excluded. The upstream occupied bandwidth is calculated as the number of subcarriers which are not excluded, multiplied by the subcarrier spacing. b) For the transmit channel set, the sum of the occupied bandwidth of all OFDMA channels plus the bandwidth of the legacy channels (counted as 1.25 times the modulation rate for each legacy channel) in a cable modem's transmit channel set. The combined bandwidth of all the minislots in the channel is normally smaller than the upstream occupied bandwidth due to the existence of unused subcarriers. The bandwidth occupied by an OFDMA probe with a skip value of zero is equal to the upstream occupied bandwidth.

Symbol Bandwidth: Symbol bandwidth (sometimes called symbol rate bandwidth) is defined for ANSI/SCTE 07 SC-QAM signals as the width of the frequency spectrum between the half power (-3 dB) points of the signal power spectrum. The symbol bandwidth in megahertz of a square root raised cosine shaped digital QAM signal amplitude response is equal to its symbol rate in Msymbols/second. For example, the symbol rate for a 6 MHz-wide ANSI/SCTE 07 256-QAM signal is 5.360537 Msym/second, and the symbol bandwidth is ~5.36 MHz.

Noise Bandwidth: Noise bandwidth is defined as the width of the frequency spectrum in which a noise power measurement is made (see for example Sections 2.2.1 and 13.5.2). Noise bandwidth must be specified when making such a measurement. Assuming the noise power is spectrally flat across the noise bandwidth, noise power measurements may be made in any practical bandwidth and then corrected to calculate the power in the desired bandwidth.

IF Resolution Bandwidth: IF resolution bandwidth (also known as intermediate frequency or IF bandwidth) is the pre-set or user-selected bandwidth of the IF filter that produces the narrowest bandwidth in the pre-detection IF chain (which may contain multiple frequency conversions) of an RF receiver or measurement device. This bandwidth is one of several factors that must be considered when measuring the power of noise or a noise-like digital signal. Since the resolution bandwidth is generally less than the channel bandwidth, a bandwidth correction is required to calculate the actual noise power or digital channel power.

Spectrum Analyzer/SLM Video Bandwidth: Video bandwidth is defined as the bandwidth of the post detection filter used prior to signal sampling in a digital analyzer, or prior to an analog analyzer’s display.

Baseband Analog Video Bandwidth: Bandwidth, in units of hertz (Hz), of an analog video signal. For example, for NTSC video this is defined as 4.2 MHz.

Discussion: The drawing in Figure 14-1 illustrates the difference between channel bandwidth, occupied bandwidth and symbol bandwidth, typical for a square root raised cosine shaped SC-QAM signal. As mentioned previously, for ANSI/SCTE 07 SC-QAM signals, the channel bandwidth is defined as 6 MHz (8 MHz in Europe) and the occupied bandwidth is based upon measured or specified spectral shaping (e.g., 26 dB down relative to the maximum level of the modulated carrier).

For ANSI/SCTE 07 SC-QAM signals the channel bandwidth and occupied bandwidth are approximately the same. Figure 14-1 shows graphically the symbol bandwidth being equal to the width of the power spectrum between the half power (-3 dB) points.[10]

The signal amplitude is shaped by the transmitter filter response. The receiver incorporates a filter whose amplitude response is the same (a "matched filter") which results in the lowest received error probability (i.e., an optimum receiver[11]).

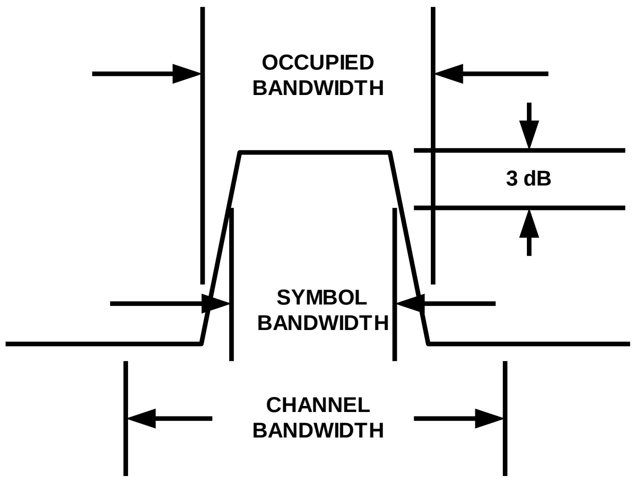


Figure 14-1: Illustration of Bandwidth Definitions applicable to a square root raised cosine shaped SC-QAM signal

The relationship between occupied and symbol bandwidths of SC-QAM signals may be calculated as follows:

$$B = R(1+\alpha)/q \qquad (1)$$

$$B_S = R/q \qquad (2)$$

where:

[10] Why is the symbol bandwidth equal to width of the frequency spectrum between the half power (-3 dB) points? The Fourier transform, H(f), of the square root raised cosine pulse shaping, h(t), when scaled for maximum value of unity, is $1/\sqrt{2}$ at a separation in megahertz equal to the symbol rate in megasymbols per second. Spectrum analyzers display the square of the absolute value of H(f), or in decibels, $20\log_{10}(|H(f)|)$. Therefore, the half-power (-3 dB) bandwidth of the square root raised cosine signal is equal to the symbol rate.

[11] Optimum Receiver Principle: An optimum receiver is designed to minimize the probability that a received symbol decision error occurs, that is, there exists no other receiver structure that can provide a lower probability of error. An optimum filter is used for acquiring a best estimate of the desired signal measurement made in the presence of AWGN, for example a matched filter. In an AWGN channel, the optimum filter is "matched" to the signal such that the filter frequency domain response is the complex conjugate of the signal frequency domain response, or equivalently the filter time domain response is the time reversed signal within the signal period. Thus, the matched filter being the complex conjugate of the filter used to shape the transmitted signal must mirror each other between the transmitter and receiver to optimally minimize the receiver probability of error.

B = occupied bandwidth in Hz

R = total data rate in bits/second

q = spectral efficiency in bits/second/Hz. (May also be expressed as bits/symbol)

α = filter excess bandwidth, expressed in decimal form. (Manufacturers usually specify α in percent)

B_S = Symbol bandwidth in Hz

Spectral efficiency is determined by dividing the bit rate (product of bits/symbol times symbols/second.) by the symbol bandwidth. The symbol bandwidth is primarily a function of the filter design in the modulator and demodulator. M-QAM transmits $q = \log_2 M$ bits on every cycle of the carrier. Therefore QPSK (4-QAM), 16-QAM, 64-QAM and 256-QAM have spectral efficiencies of 2, 4, 6 and 8 bits/second/Hz, respectively. Since it is not a double sideband modulated signal, 8-VSB has a spectral efficiency of 3 bits/second/Hz.

Sample calculations are shown in Table 14-1.

Table 14-1: Bandwidth Calculations

Modulation	R (Mbits/s)	q (bits/s/Hz)	α	B (MHz)	B_S (MHz)
64-QAM	30.34	6	0.18	5.97	5.057
256-QAM	42.88	8	0.12	6.0	5.36
8-VSB(1)	32.28	3	0.115	6.0	5.38
QPSK(2)	2.048	2	0.5	1.536	1.024
QPSK(3)	1.544	2	0.3	1.004	0.772
QPSK(4)	3.088	2	0.3	2.007	1.544

1. The actual symbol rate of 8-VSB is 10.76 Msymbols/second. However, since this is not a double sideband modulated signal, the symbol rate is divided by two for bandwidth calculations. Alternatively, one could use equations (1) and (2) for bandwidth calculation if the value of q were doubled.
2. Out-of-band transmission per ANSI/SCTE 55-1 2019, Table 1.
3. Out-of-band transmission per ANSI/SCTE 55-2 2019, Table 2-2.
4. Out-of-band transmission per ANSI/SCTE 55-2 2019, Table 2-2.

The total data rate, R, is the payload data plus whatever overhead data is required for error correction and synchronization. Manufacturers may not always specify a total data rate. However, payload data rate and symbol bandwidth should be specified. If the symbol bandwidth and the modulation method are known, the total data rate can be calculated.

In practice, two types of filtering may be used: root raised cosine and raised cosine filtering. In root raised cosine filtering, the transmitter and receiver each contain a filter whose frequency response is equal to the square root of the filter function. If a raised cosine filter is used, all filtering is done in the transmitter. Depending on which type of filtering is used, there will be a slight difference in the result obtained when measuring RF signal power because of the difference in signal slope at the edges. This difference is shown in Table 14-2.

Table 14-2: Root Raised Cosine/Raised Cosine Power Ratios

	RRC/RCOS Ratio (dB)	
Modulation	**Symbol (3 dB) Bandwidth**	**Occupied Bandwidth**
64-QAM	0.066	0.209
256-QAM	0.043	0.14
8-VSB	0.042	0.134
QPSK (ANSI/SCTE 55-1 2019,Table 1)	0.192	0.556
QPSK (ANSI/SCTE 55-2 2019, Table 2-2)	0.112	0.342

14.2 Noise Ratios

Definitions

E_b/N_0 is defined as the ratio of the signal energy per bit to the noise power spectral density (noise power in a 1 Hz bandwidth). E_b/N_0 is normally expressed in dB.

E_S/N_0 is defined as the ratio of the signal energy per symbol to the noise power spectral density (noise power in a 1 Hz bandwidth). E_S/N_0 is normally expressed in dB.

Carrier/Noise Ratio (CNR) is defined as the ratio of the digital channel power to the average noise power in the occupied bandwidth.

The noise ratios are related to each other via the following equations:

$$\text{CNR} = E_b/N_o * \frac{R}{B} \qquad (3)$$

$$E_S/N_0 = q * E_b/N_o \qquad (4)$$

$$\text{CNR} = Es/N_o * \frac{1}{1+\alpha} \qquad (5)$$

where the quantities q, R and B are as defined for equations (1) and (2).

14.3 Cluster Variance (CV)

Definition: Cluster variance is the ratio, in dB, of the average vector magnitude to the rms value of the spread in the constellation points. Cluster variance is equivalent to MER. Generally, digital signal analyzer manufacturers prefer using “SNR”, “SNR Estimate” or “MER”. All of these terms have equivalent meanings.

Chapter 15 BTSC Stereo and Cable Systems: Measurement Techniques and Operating Practices

"Unedited and retained for Historic Significance."

BTSC Stereo and Cable Systems:

Measurement Techniques and Operating Practices

NCTA Ad Hoc Subcommittee
on Multichannel Television Sound

Chairman: Alex Best

Preface

In January of 1984 the BTSC (Broadcast Television Systems Committee) committee of the Electronic Industries Association selected a television stereo system to serve as the standard for broadcast television in the United States. The format selected consists of a multiple subcarrier scheme developed by the Zenith Electronics Corporation and a noise reduction companding system developed by the DBX Corporation. In March of 1984 the FCC, stopping short of selecting this format as the only broadcast standard, 'protected' the system and granted permission for broadcasters to begin transmission. By protecting the standard, consumers would be assured that television receivers designed to receive the new standard could not be interfered with by broadcasters transmitting alternative schemes.

In August of 1982 the NCTA Engineering Committee, concerned about the capability of cable systems to provide quality transmission of the proposed stereo systems (Zenith, Telesonics, and EIA-Japan), formed an ad hoc subcommittee to investigate multichannel television sound. In September of 1982 the subcommittee sent a report to the Chairman of the EIA BTSC committee outlining the areas of technical concern. The cable industry also went on record as being opposed to any of the stereo formats being proposed. The opposition centered on the selection of a subcarrier scheme (very similar to the 30+ year old FM broadcast system) at a time when digital audio systems were becoming common in the consumer marketplace. More important there were cable carriage problems which were explained in detail in a report to BTSC committee.

In March, 1983 the NCTA subcommittee wrote a comprehensive test plan and hired a test engineer to measure the impact of cable equipment on the proposed stereo systems.

The testing was completed in September of 1983 and the results documented in a report titled “Multichannel Television Sound Report”. As a result of the efforts of the cable industry, plus others, the FCC granted a 'no must carry' status to the newly selected stereo system in February of 1985.

As of today approximately 50% of the 1,200 broadcast stations in the U.S. are transmitting in stereo. As a result 97% of U.S. households are reached by stereo TV broadcasts. With few exceptions, built-in stereo will be featured in more than 50% of the color sets offered by the major suppliers in 1989. Of the 200 million TV sets in the U.S., approximately 15% presently have built-in stereo capability. This should reach 50% by the year 1993.

With receiver penetrations reaching significant proportions, the cable industry is now accelerating the introduction of BTSC audio. There are presently some 12 or more satellite delivered programs which offer stereo audio, not to mention the over-the-air broadcast stations.

NCTA Ad Hoc Subcommittee on Multichannel Sound

After the conclusion of the test report in 1983, the BTSC subcommittee entered a phase of providing a clearinghouse for inputs on both successful and unsuccessful attempts by the cable industry to provide quality stereo reception to our customers. As a result of this effort, one fact became evident. The cable industry needed a comprehensive set of measurement procedures and operating practices to verify optimum performance for stereo encoding equipment plus ensure quality delivery through the remainder of the system. As a result the subcommittee reconvened its efforts in 1987. The following practices and procedures are the results of that effort.

Acknowledgments

The following members of the subcommittee were instrumental in providing guidance on the format and did in fact provide all of the inputs which make up the remainder of this document.

Jim Farmer
Scientific-Atlanta

Brian James
NCTA

Ned Mountain
Wegener Communications

Karl Poirier
Triple Crown Electronics

Luis Rovira
Scientific-Atlanta

David Sedacca
Scientific-Atlanta

Russ Skinner
United Artist Cable Corp

TABLE OF CONTENTS

1. Introduction

1a. Test Equipment

During the development of this report, much discussion took place on the quality of test equipment required to obtain meaningful results. The dilemma revolved around whether the subcommittee should recommend precision decoders and demodulators, knowing that many operators do not presently have access to this type of equipment. The alternative was to suggest the use of consumer type test equipment severely limiting the accuracy of the resulting measurements, and in many cases rendering the results useless. After all, in most instances what we are attempting to determine is the performance level of a high quality stereo encoder. Attempting to measure its performance through a device whose specifications are worse than the device(s) being tested will only lead to erroneous conclusions. Because of this, you will find that only precision test equipment is recommended in the “required test equipment” lists in each of the measurement descriptions.

1b. BTSC Mode Vs. Equivalent Mode

A second dilemma involved the issue of whether the measurements should be made with the equipment operating in the BTSC Mode or the Equivalent Mode. To help us understand equivalent mode, look at a block diagram of the BTSC system in Figure 15-1 and Figure 15-2. What sets Multichannel Television Sound (MTS) apart from conventional FM broadcast stereo is the BTSC noise reduction that makes possible buzz-free stereo operation with intercarrier sound detection. Unfortunately, the presence of this compressor and expander in the L-R channel that is not duplicated in the L + R channel creates a whole host of problems that do not exist in conventional FM stereo. FM stereo broadcasting is much simpler than MTS because the FM system is linear throughout.

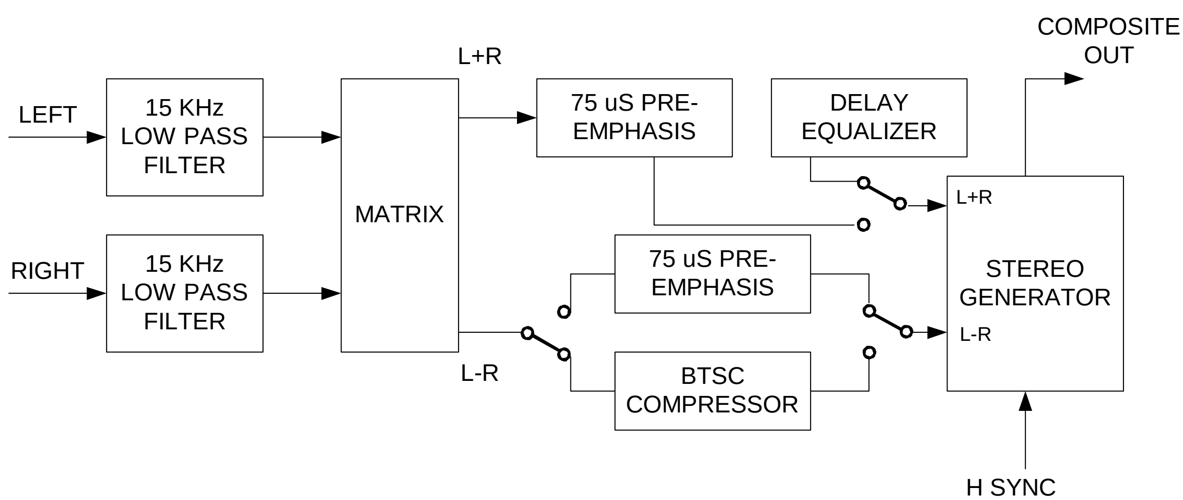


Figure 15-1: MTS Stereo Encoder

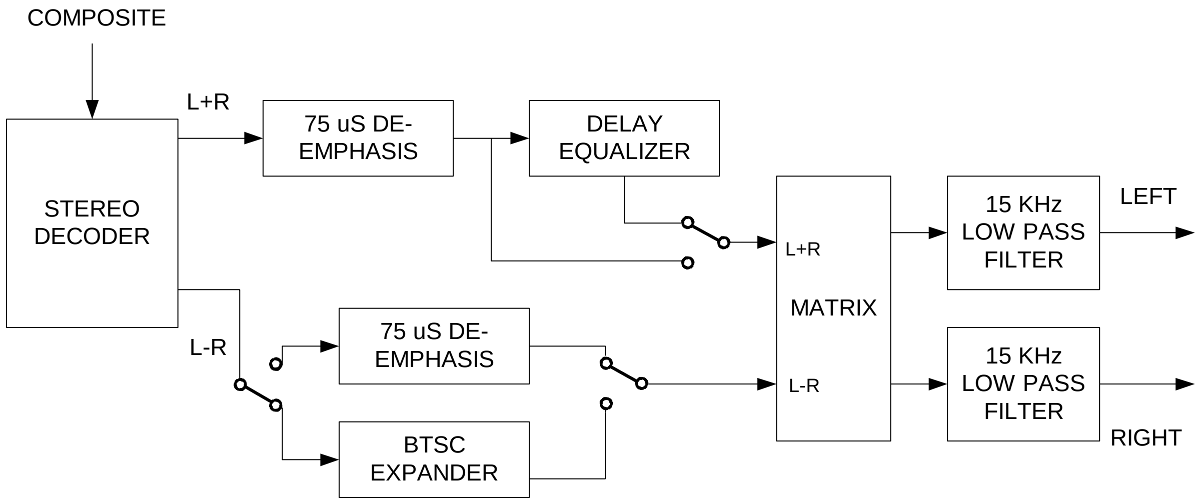


Figure 15-2: Stereo MTS Decoder

Measurements made in the BTSC mode (Expander/Compressor active in circuit), while more difficult to make accurately, gives us an indication of how the system actually operates in the real world. One of the major areas of concern is the level setting accuracy. For the expander and compressor to track one another correctly they must have a common reference. For this to occur the absolute FM deviation of a modulator must be accurate. The total allowable error budget, input to output, on the L-R channels is ±.1744 dB for 40 dB of separation and ±.550 dB for 30 dB of separation. Although many knowledgeable engineers strongly urge that stereo measurements (especially separation) be made with the encoder/decoder operating in the equivalent mode (expander/compressor replaced with 75 µs pre-emphasis/de-emphasis network), we have chosen not to do that for two reasons. First, not all BTSC encoders designed for cable applications have the capability of being operated in the equivalent mode. Second, interpreting the results of measurements made in this mode of operation is confusing when attempting to relate the numbers to real world operation.

1c. Audio Test Levels

Channel: Sum Channel After Pre-Emphasis (See Figure 15-3: TP 2)

Definition: 100% level in the Sum Channel after Pre-emphasis is the peak-to-peak voltage that causes the sound carrier to deviate ± 25 kHz around its center frequency.

Discussion: Eliminate the difference, pilot, and SAP signals. Now only the sum signal from TP 2 contributes to the BTSC output at TP 1, and therefore to deviation of the sound carrier. Change the peak-to-peak voltage level at TP 2 to get ± 25 kHz deviation of the sound carrier. The peak-to-peak voltage now at TP 2 is the 100% level for the Sum Channel After Pre-Emphasis.

Channel: Sum Channel Before Pre-Emphasis (See Figure 15-3: TP 3)

Definition: 100% level in the Sum Channel Before Pre-Emphasis is the peak-to-peak voltage that causes the sound carrier to deviate ± 25 kHz around its center frequency.

Discussion: Eliminate the difference, pilot, and SAP signals. Now the sum signal from TP 3 travels through TP 2 to the BTSC output at TP 1. No other signals contribute to the deviation of the sound carrier. Change the peak-to-peak voltage level at TP 3 to get ± 25 kHz deviation of the sound carrier. The peak-to-peak voltage now at TP 3 is the 100% level for the Sum Channel Before Pre-Emphasis.

100% level at TP 3 is frequency dependent. The pre-emphasis network that follows TP 3 applies more gain to high frequency signals than to low frequency signals. For example, a 400 Hz tone at TP 5 causes ± “X” kHz deviation of the sound carrier. Change the frequency of the source to 14 kHz. The deviation of the sound carrier increases greatly. Reduce the amplitude at TP 3 to once again get ± “X” kHz deviation. The peak-to-peak amplitude at TP 3 will now be about 16 dB smaller than before for the same carrier deviation.

Channel: Difference Channel After DSBSC Modulation (See Figure 15-3: TP 4)

Definition: 100% level in the Difference Channel After DSBSC Modulation is the peak-to-peak voltage that causes the sound carrier to deviate ± 50 kHz around its center frequency.

Discussion: Eliminate the sum, pilot, and SAP signals. Now only the DSBSC difference signal from TP 4 contributes to the BTSC output at TP 1, and therefore to deviation of the sound carrier. Change the peak-to-peak voltage level at TP 4 to get ± 50 kHz deviation of the sound carrier. The peak-to-peak voltage now at TP 4 is the 100% level for the Difference Channel After DSBSC Modulation.

Channel: Difference Channel After Noise Reduction Encoding (See Figure 15-3: TP 5)

Definition: 100% level in the Difference Channel After Noise Reduction Encoding is the peak-to-peak voltage that causes the sound carrier to deviate ± 50 kHz around its center frequency.

Discussion: The amplitudes of the signals at TP 5 and TP 4 are related by the “conversion gain” of the DSBSC modulation process. The energy at TP 5 extends from 50 Hz to 15 kHz. The same information is present at TP 4 in the frequency range between 16.4 kHz and 46.5 kHz.

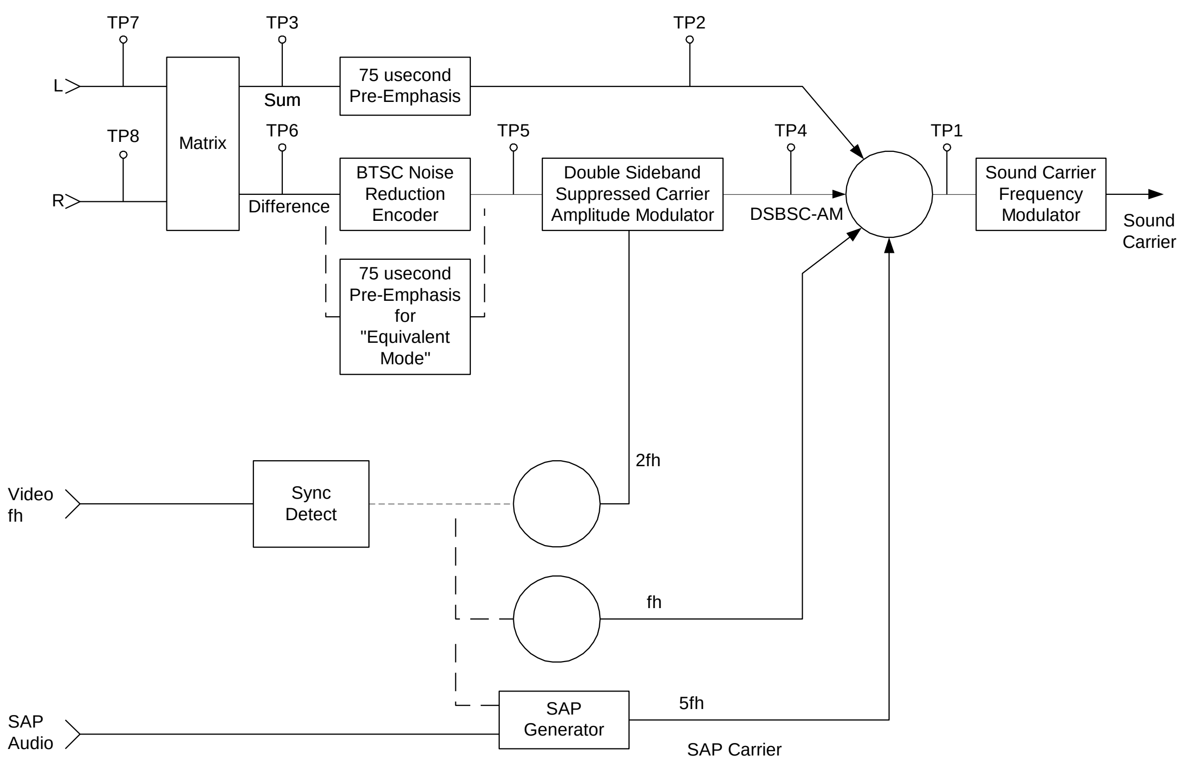


Figure 15-3: BTSC Encoder

Channel: Difference Channel Before Noise Reduction Encoding (See Figure 15-3: TP 6)

Definition: 100% “75 microsecond equivalent” level in the Difference Channel Before Noise Reduction Encoding is the peak-to-peak voltage that causes the sound carrier to deviate ± 50 kHz around its center frequency when the difference channel BTSC noise reduction system is replaced by a 75 microsecond pre-emphasis network.

Discussion: When the BTSC difference channel noise reduction encoder is engaged the signals at TP 6 and TP 5 are related by a complicated transfer function. The transfer function depends on the amplitude, frequency, and recent dynamic behavior of the signal at TP 6. Due to noise amplification and other effects, precise measurements at the output of the noise reduction encoder are very difficult to perform. A 75 microsecond pre-emphasis network is substituted for the BTSC noise reduction encoder to provide a system test mode. This is the “75 microsecond equivalent” mode.

Eliminate the sum, pilot, and SAP signals. Engage the 75 microsecond equivalent mode. Now the difference channel signal at TP 6 travels through TP 5 and TP 4 to the BTSC output at TP 1. No other signals contribute to the deviation of the sound carrier. Change the peak-to-peak voltage level at TP 6 to get ± 50 kHz deviation of the sound carrier. The peak-to-peak voltage now at TP 6 is the 100% “75 microsecond equivalent” level for the Difference Channel Before Noise Reduction Encoding

2. Measurement Techniques

2a. Signal-to-Noise Ratio

Definition: Audio signal-to-noise ratio is defined to be the ratio of the audio signal power output to the noise power in the entire audio passband.

Procedure: Required equipment:

- Audio Oscillator, 400 Hz, distortion less than 0.1%
- Audio Voltmeter, response flat within +/- 0.1 dB, 50 Hz to 15 kHz
- Precision BTSC stereo demodulator and/or decoder

1. Figure 15-4 is a block diagram showing proper test equipment setup for a measurement involving only an encoder and decoder.

   Figure 15-5 is a block diagram showing the addition of a 4.5 MHz or 41.25 MHz lF loop.

   Figure 15-6 is a block diagram showing the addition of a total headend RF loop.
2. Verify that the audio oscillator frequency is 400 Hz.
3. Apply the 400 Hz tone to both the left and right channel inputs of the stereo encoder.
4. Use the audio voltmeter to measure the audio levels at the output of the precision stereo decoder.
5. Adjust the input levels to the stereo encoder so that both the left and right channels are modulated at their "100% 75 microsecond equivalent input levels". (Were there is no pilot tone or multiplexed difference channel energy, these signals would cause the sound carrier to be deviated exactly ±25 kHz. With the pilot tone present, actual peak deviation is about ±30 kHz. The multiplexed difference channel contains some noise energy, and may contribute to instantaneous peak deviation in larger amounts.)
6. Note the rms amplitude of the 400 Hz tone at either the left or right channel output of the decoder (they should be equal). Call this value "$V_{100\%}$"
7. Remove the 400 Hz signal from both inputs of the stereo encoder. Terminate the encoder's audio inputs with 600 Ω resistors.
8. Measure the rms noise voltage now coming out of the precision decoder's left or right output (their levels should be very similar, if not equal). Call this value "V_{noise}."
9. Calculate sum channel signal-to-noise ratio:

   $$SNR = 20 \log (V_{100\%}) / (V_{noise})$$

Performance Objective: The BTSC signal delivered to the subscriber's receiver shall be capable of providing a sum channel signal-to-noise ratio, for sum channel peak modulation levels of ± 25 kHz, of no less than 55 dB, when demodulated and decoded by precision BTSC stereo equipment.

When a scrambling system that puts amplitude modulation energy on the sound carrier is employed, this measurement should show no less than 53 dB sum channel signal-to-noise ratio.

Discussion: In companded transmission systems, the noise power out varies as the receiver gain varies. The BTSC stereo system uses companding to transmit the stereo difference channel. However, difference channel energy is present in any transmission in which the left channel signal is not exactly equal to the right channel in frequency, amplitude, and phase.

The output audio noise power in the BTSC stereo system changes with the gain of the compander used in the difference channel. Output noise power is smallest when the expander in the receiver is applying its maximum attenuation. Output noise power is worst when the expander in the receiver is applying its maximum gain. This means that in the BTSC system, the output noise floor varies with program content - specifically the difference channel level. Passages with a lot of stereo information will have a worse signal-to-noise ratio than equivalently loud passages containing only mono information.

For these reasons, a straightforward measurement of signal-to-noise ratio for just the left channel or just the right channel is not available. Not only does the noise floor for a left-only or right-only transmission change with signal amplitude, the BTSC system also specifies a companding algorithm that adjusts gain based on audio signal frequency content. As test signals are swept in frequency or amplitude, the characteristics of the transmission system change.

Two measurements may be used to characterize the noise performance of the BTSC system. The sum channel signal-to-noise measurement procedure described here is used to evaluate output noise effects that are not a function of the companding process. A separate measurement of signal-to-buzz" ratio is described elsewhere. The signal-to-buzz measurement evaluates output noise including effects of the receiver's expander.

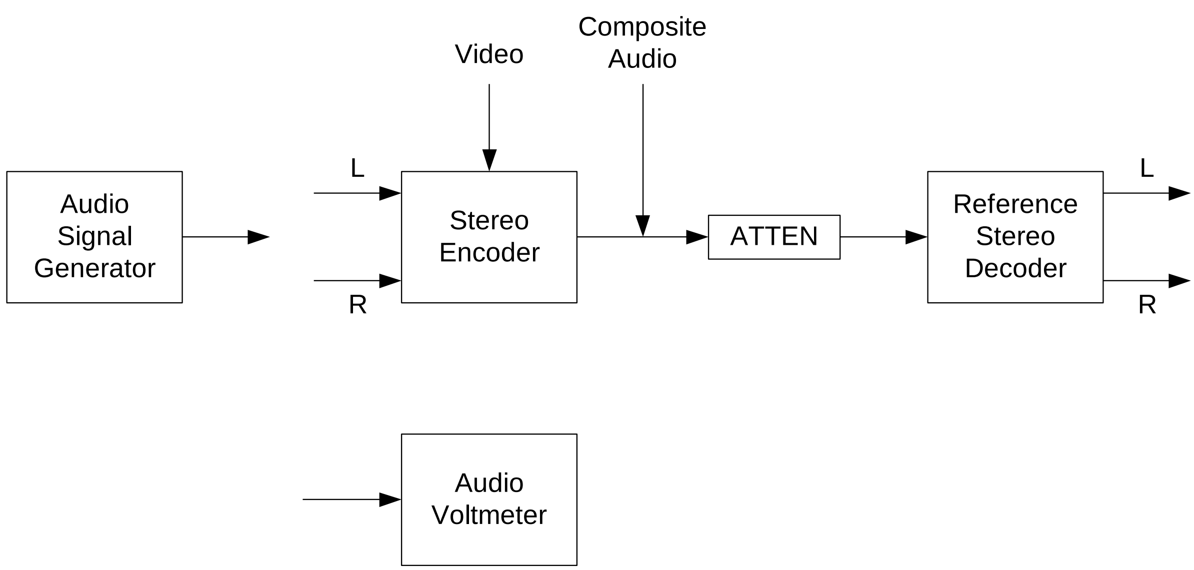


Figure 15-4: Basic Setup - No IF or RF Loop

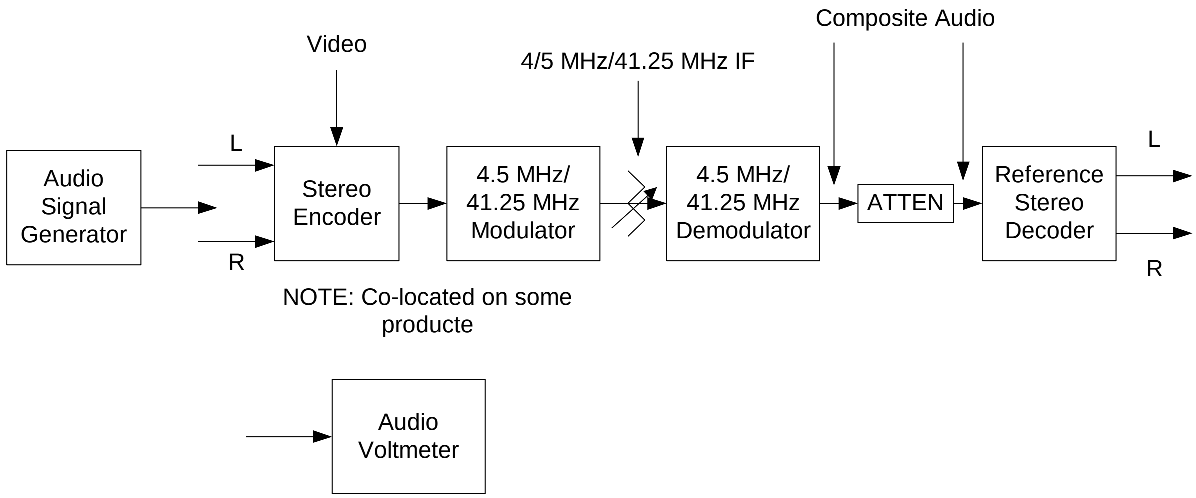


Figure 15-5: Test Setup with 4.5 MHz / 41.25 MHz IF Loop

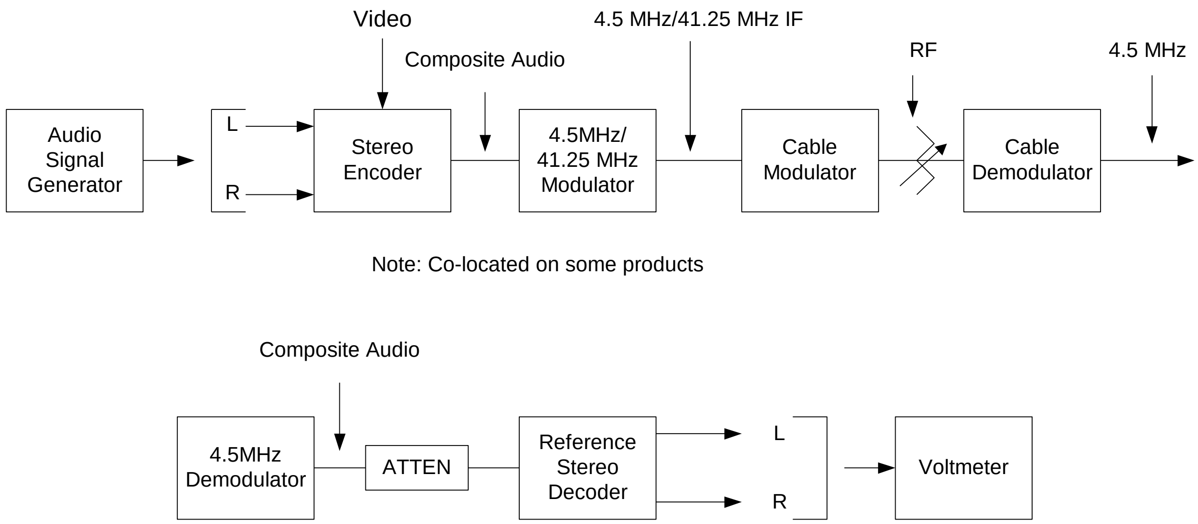


Figure 15-6: Test Setup with Full RF Loop

2b. Signal-to-Buzz Ratio

Definition: Signal-to-buzz ratio (S/B) is the ratio in dB of the peak-to-peak signal voltage divided by the peak-to-peak buzz voltage as seen using an oscilloscope at the output of a BTSC stereo decoder. The buzz measurement is made while the signal is present.

Procedure: Required test equipment:

- Audio oscillator
- Video signal generator capable of adequate stability for use with the stereo encoder
- Oscilloscope
- TV demodulator
- Stereo decoder

1. Figure 15-8 is a block diagram showing proper test equipment setup for a measurement involving only an encoder and decoder.

   Figure 15-9 is a block diagram showing the addition of a 4.5 MHz or 41.25 MHz IF loop.

   Figure 15-10 is a block diagram showing the addition of a total headend RF loop.
2. When a video waveform is needed in testing modulators or other devices requiring video, standard full-field color bars should be used.
3. Set the encoder and decoder to the BTSC mode.
4. Set the audio oscillator frequency to 1 kHz.
5. Connect the audio oscillator to the right input of the stereo encoder.
6. Adjust the oscillator level and encoder input sensitivity for 100% modulation of the right channel, as shown on the encoder front panel meter.
7. Connect the oscilloscope to the right output of the stereo decoder.
8. Set the sweep rate to .2 ms/div. Measure the peak-to-peak voltage of (S) of the sine wave. Measure the brighter sinusoid in the trace while ignoring the dimmer and more erratic buzz voltage. This measurement should not include buzz voltage.
9. Set the sweep rate to 20 ms/div. The signal sinusoid should appear as a wide sweeping band in the display. Increase the input sensitivity of the oscilloscope to expand the buzz that appears to be riding the top and bottom of the band as shown in Figure 15-7. Expand the top or bottom of the band, whichever is bigger. Measure the peak-to-peak buzz voltage (B) riding the signal sinusoid.

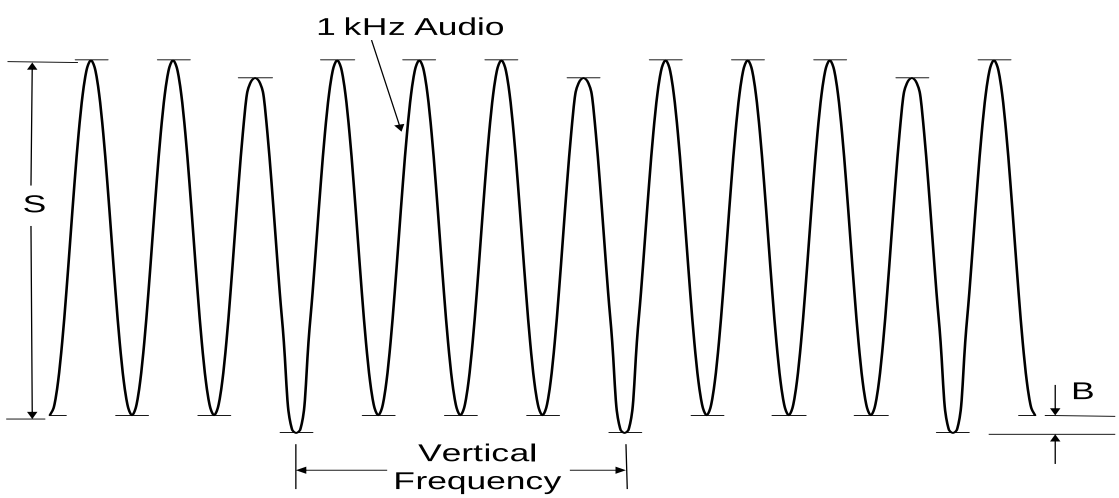


Figure 15-7: Example of Oscilloscope Trace

10. Calculate Signal-to-buzz ratio in dB using the following formula:

    S/B dB = 20*log (S/B)
11. Repeat procedure for the left channel.

Alternate Procedure: Required test equipment:

- Audio analyzer with SINAD capability (CCIR/ARM weighting optional) instead of audio signal generator listed above (HP 8903 or equivalent)
- All other equipment listed above

1. When a video waveform is needed in testing modulators or other devices requiring video, standard full-field color bars should be used.
2. Set the encoder and decoder to the BTSC mode.
3. Set the audio analyzer frequency to 1 kHz.
4. Connect the audio analyzer output to the right input of the stereo encoder.
5. Adjust the analyzer output level and encoder input sensitivity for 100% modulation of the right channel, as shown on the encoder front panel meter.
6. Connect the audio analyzer input to the right output of the stereo decoder.
7. Select the SINAD function on the audio analyzer and read signal-to-buzz directly from the display.
8. If desired, use CCIR/ARM weighting to get an indication of the perceptibility of the buzz.
9. Repeat procedure for the left channel.

Performance Objective: More than 35 dB unweighted should be measured using a precision demodulator and stereo decoder at the output of a complete system. In systems with scrambling a minimum of 27 dB should be measured. Measurements made with consumer equipment may produce lower numbers as discussed below.

Discussion: The results of this measurement depend almost completely on the ability of the test demodulator and decoder to reject video and AM on the sound carrier. A consumer demodulator/decoder may be more sensitive to this than precision equipment, and may be more advantageous in terms of finding system buzz problems. For this reason numbers obtained by this measurement should be regarded as comparative and not absolute. The performance objective listed above should be interpreted accordingly.

The measurements of signal-to-noise (SNR) and signal-to-buzz as described in this document represent two extremes for the BTSC system due to the dbx-TV noise reduction in the difference channel. Under conditions of no input signal (SNR measurement), the noise performance is apparently very good because the dbx-TV expander attenuates quiet signals. The signal-to-noise measurement then represents a “best case” for the system. When a large amount of difference signal is present (S/B measurement), the dbx-TV expander in the decoder applies gain to it, increasing the level of any buzz that may be in the difference channel. Thus signal-to-buzz represents a “worst case” measurement of the system.

It could be argued that signal-to-buzz is a highly pessimistic measure of true system performance. Then a full modulation signal is present, noise masking is most effective and the buzz is less noticeable.

Also, the nature of stereo television productions is such that almost all dialogue is located in center. Much audio is contained largely in the sum channel. Yet, in this measurement the test signal is applied to one channel only. This puts half the energy (before compression) in the difference channel causing the dbx expander to amplify. In real signals this much of the audio energy is rarely concentrated in the difference channel.

Note also that a peak-to-peak measurement of buzz gives no indication as to the frequency content or total buzz energy. Hence this measurement is not a good indication of how disturbing the buzz will be to a listener. The alternate procedure described, especially when used with weighting, addresses some of these problems. The disadvantage of the alternate procedure is that excessive distortion in the equipment can override the buzz measurement.

Despite these drawbacks, buzz is a real phenomenon that can be disruptive to stereo and should be considered. The measurement of S/B as described here provides some indication of its severity.

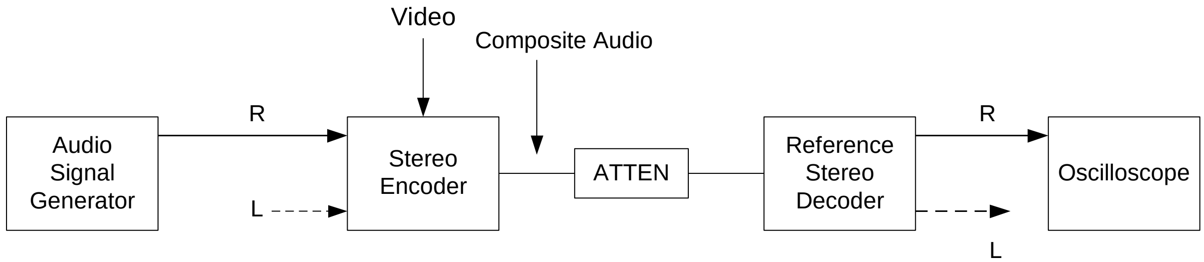


Figure 15-8: Basic Setup - No IF or RF Loop

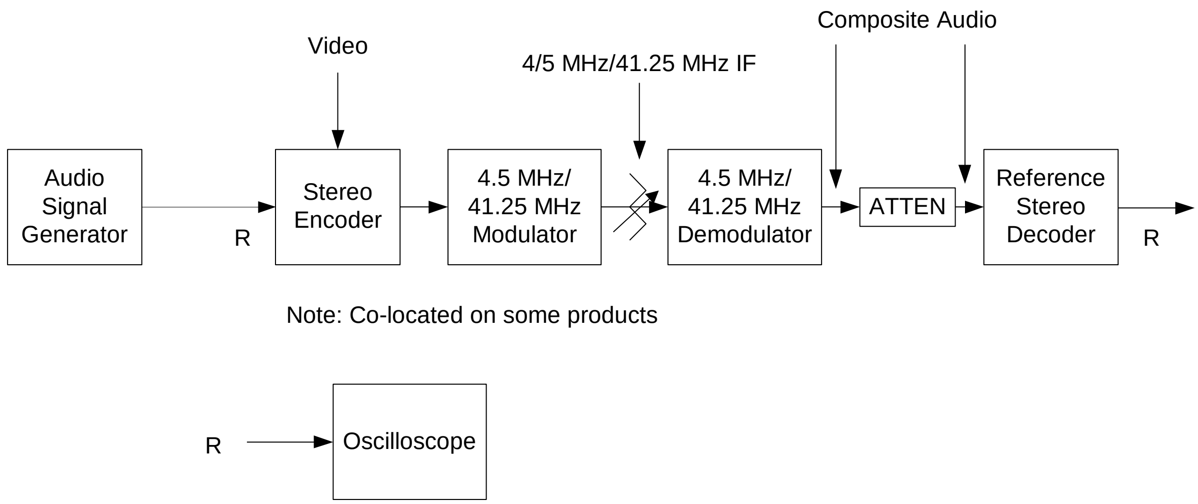


Figure 15-9: Test Setup with 4.5 MHz / 41.25 MHz IF Loop

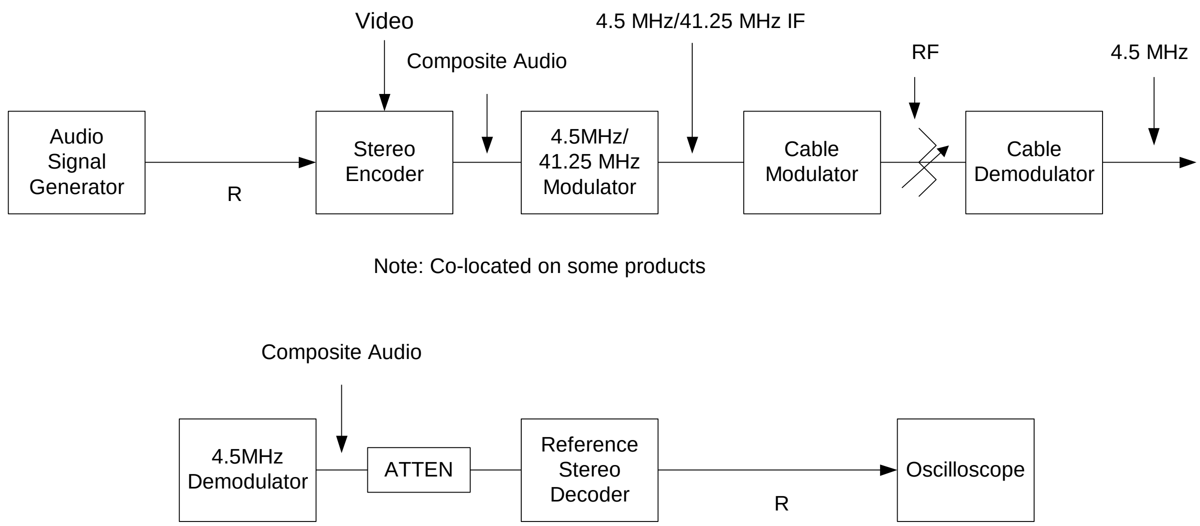


Figure 15-10: Test Setup with Full RF Loop

2c. Frequency Response of BTSC Stereo Transmission System

Definition: Frequency response of the MTS stereo system is the variation in system gain versus frequency from the input of the stereo encoder to the output of the test decoder. The frequency response measurement may include part or all of the cable distribution system.

Procedure: Required equipment:

- A Precision Audio Oscillator with frequency range of 50 Hz to 15 kHz and output distortion of less than 0.1%.
- A precision audio volt meter
- Audio Sweep Test Set

1. Figure 15-11 is a block diagram showing proper test equipment setup for a measurement involving only an encoder and decoder.

   Figure 15-12 is a block diagram showing the addition of a *4.5* MHz or 41.25 MHz IF loop.

   Figure 15-13 is a block diagram showing the addition of a total headend RF loop.

2. Connect the audio signal generator to the left channel input of the stereo encoder. Determine the input level required to obtain 100% modulation of the aural carrier using a 14 kHz test signal. Reduce the input level by a minimum of 17 dB for the tests. This will ensure that no active components are being overdriven and thereby affecting test results.
3. Connect the audio volt meter to the left channel output of the stereo decoder.
4. Measure the frequency response of the equipment under test. If an audio sweep system is used, then the response can be swept from 50 Hz to 15 kHz. The sweep speed of the test equipment must be set sufficiently slow to allow any transients to settle during the tests. If an audio signal generator is used, make test measurements at the following frequencies:

   50 Hz, 100 Hz, 400 Hz, 1 kHz, 3 kHz, 5 kHz, 8 kHz, offset of 10 kHz, 14 kHz, 15 kHz. The measurement at 15 kHz may be dropped if the equipment rolls off above 14 kHz.

5. Repeat the measurements with the input level increased in steps of 3 dB until 10 dB over 100% modulation level is reached. This will determine the dynamic response of the system under test.

6. If it is possible to remove the companding equipment from the circuitry of the modulator and demodulator then the test can be done using the 75 microseconds equivalent mode. Input level for this test should be set at the 17 dB below 100% modulation level determined in step 1.
7. Repeat measurements using the right channel as the input and output ports of the equipment.

Performance Objective: The frequency response, peak to valley, should be within three dB from 50 Hz to 14 kHz.

Discussion: Frequency response is a common audio measurement used to determine the transparency of the system to signal within the passband of the system. If the frequency response is greater than the three dB recommended then the subscriber may observe less than a satisfactory performance with the system.

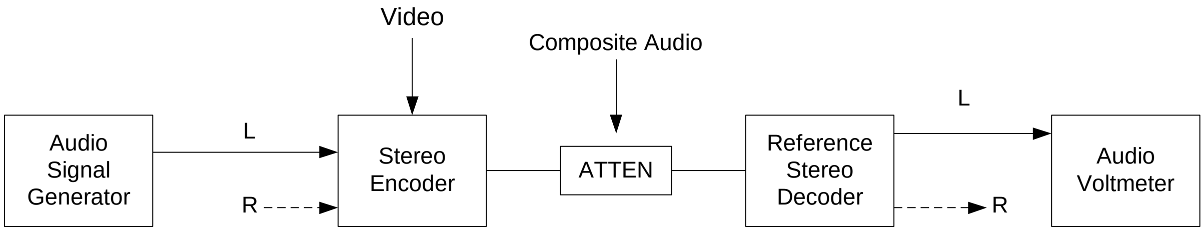


Figure 15-11: Basic Setup - No RF Loop

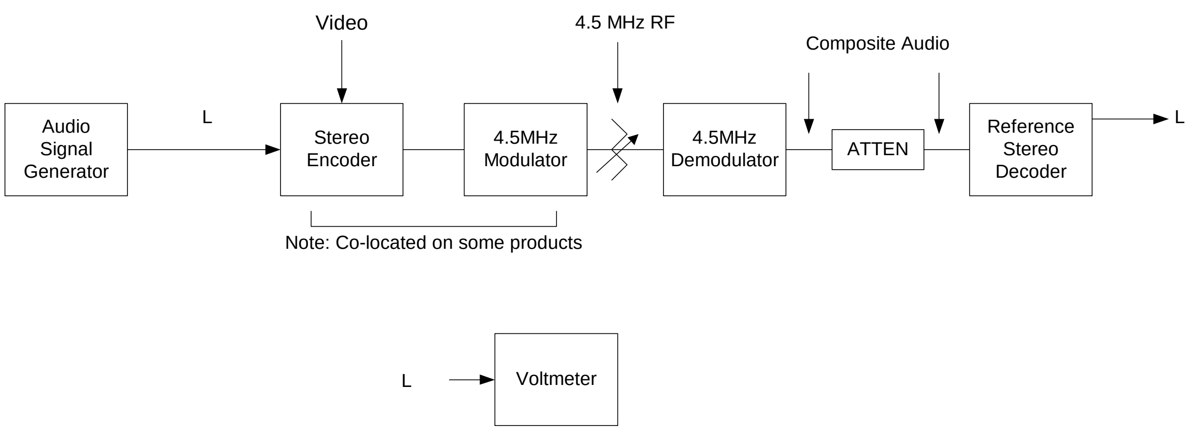


Figure 15-12: Test Setup with 4.5 MHz RF Loop

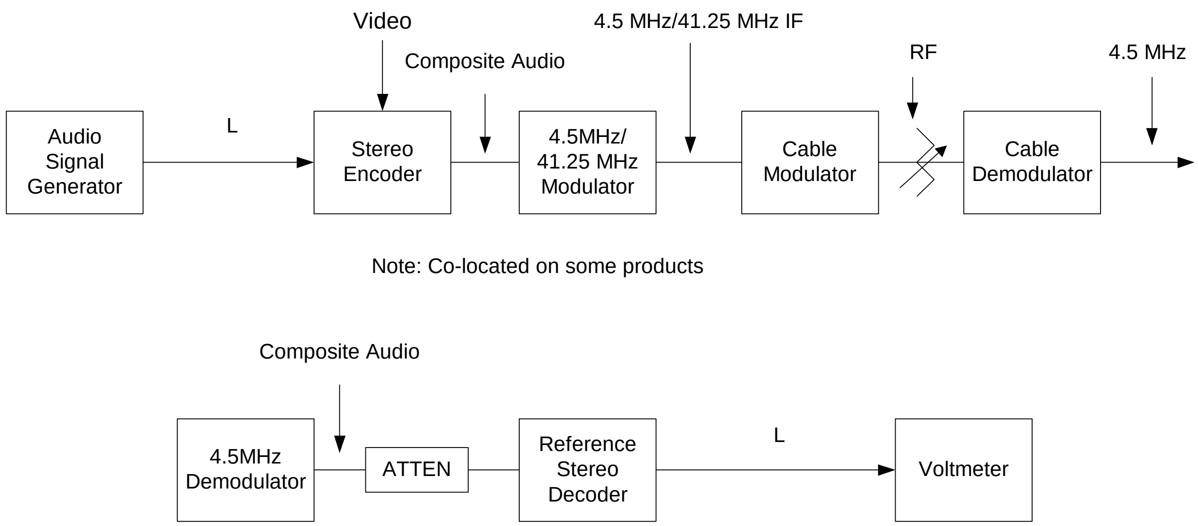


Figure 15-13: Test Setup with Full RF Loop

2d. Separation Measurements of the BTSC Stereo System

Definition: Stereo separation is the difference in output level between the demodulated left audio channel and the right audio channel exclusive of noise when only a left or right audio input channel is supplied to the stereo system including the encoder, distribution system and decoder. Separation is expressed as a voltage ratio in dB.

Procedure: Required test equipment:

- A precision audio oscillator with a frequency range of 50 Hz to 15 kHz with low distortion.
- A precision audio rms volt meter or distortion analyzer (Fluke 8840A, Tektronix 501A or equivalent).
- A precision demodulator (Tektronix 1450-1) and a reference decoder (Modulation Sciences SRD-1 or equivalent) for (30 to 40 dB) separation, or a high quality decoder or modulation monitor with an RF input, (TFT 850, Eiden or equivalent) for (30 to 35 dB) separation, or a high quality consumer stereo television or stand alone decoder (Recoton or equivalent) for (15 to 25 dB) separation. (Figure 15-16)
- Precision audio attenuator is needed (.01 dB steps) to set level at baseband between the encoder and precision decoder.

One of the most critical adjustments in the BTSC system is the accuracy of the composite baseband level feeding the reference decoder (Figure 15-14 and Figure 15-15). If the composite level is not set properly, then separation will suffer. If this level is in error by 0.28 dB, the best achievable separation is 30 dB (assuming an otherwise ideal system).

For measurements requiring a high degree of accuracy it also may be necessary to verify the proper performance of the reference decoder. A paper outlining this precedence has been published by John K. Chester of Modulation Sciences. The title is “Measuring Static Performance of BTSC Decoders.”

1. Install encoder, modulator, demodulator and decoder per manufacturers’ instructions i.e., interface may be composite baseband, 4.5 MHz or 41.25 MHz. (Figure 15-14 through Figure 15-16)

2. Use the Bessel null technique to set deviation of the modulator or a manufacturer's calibration tone that was referenced against Bessel null may also be used. Deviation error must be set to less than 1 percent.
3. It is best to perform frequency response, noise, and distortion test to verify encoder operation first.
4. Connect the audio signal generator to the left channel input on the encoder. Terminate the right channel input with a 600 Ω resistor. Set the generator frequency to 14 kHz sine wave. Set the level to approximately 100% modulation. Reset the generator to 300 Hz. The 300 Hz signal is now approximately set at 14% modulation due to the pre-emphasis or 17 dB lower than the level needed at 14 kHz to produce 100% modulation. Do not reset level to make the measurement only change frequency.
5. Connect the rms audio volt meter or distortion analyzer to the left channel output of the decoder. If the decoder has an output level control increase it to full output. Set balance controls to equal balance, turn off any built-in simulated stereo capabilities. Measure the output voltage of the 300 Hz sine wave signal from the decoder. This voltage is the left channel reference or 0 dB. Connect the voltmeter to the right channel of the decoder, measure the 300 Hz sine wave signal. This measurement is the separation from left to right channel. This level should be at least 20 dB lower than the level measured on the left channel.

Note: To convert voltage ratio to dB ratio use.

20*log [Ref. voltage (Lor R)]/[Meas. voltage (L or R)] = Separation dB

Example: 20*log [1500 my (L Ref.)]/[50mv (R Meas.)] = 29.54 dB

6. Repeat measurements by changing frequency only. Reverse audio input and measure right channel to left channel separation.

Performance Objective: Separation through the total system, to the subscriber's stereo television or decoder of 20 dB from 100 Hz to 8 kHz with a taper of 6 dB to 14 dB at 40 Hz and 14 kHz.

Discussion: The measurement of separation in the BTSC, multi-channel television sound delivery system is by far the most complex. The BTSC system uses the dbx companding system in the (L-R) difference channel to reduce noise. This channel is very sensitive to transmission error. For the system to deliver a high degree of separation, the dbx compressor in the encoder must track the expander in the decoder with a very high degree of accuracy. Any phase or amplitude error in the encoder, system or decoder, in the (L-R) channel relative to the (L+R) channel will cause a reduction in separation that will be very dramatic. In order to maintain 30 dB separation, amplitude error must be 0.3 dB or less and phase error of 3 degrees or less. To maintain 20 dB separation, amplitude error must be 1 dB or less and phase error of 10 degrees or less. This is for the total system including encoder, modulator, AML, cable plant, converter and the decoder in the consumer's television. The ability to measure separation is limited only to the accuracies of the test equipment and the care in which all parts of the system are matched for minimum amplitude and phase error. If you wish to measure an encoder with a *35* dB specification, a decoder whose performance has been verified must be used. If you wish to measure an encoder with a 20 dB specification, a high quality consumer decoder may be used. The test equipment used to measure any part of the system should be 10 to 15 dB better than the part of the system that is to be measured.

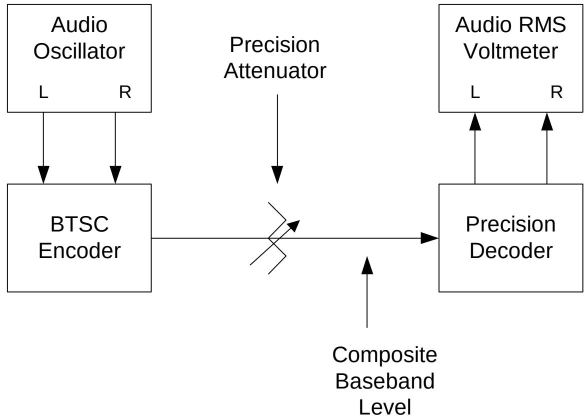


Figure 15-14: Separation Measurements of the BTSC Stereo System – Baseband

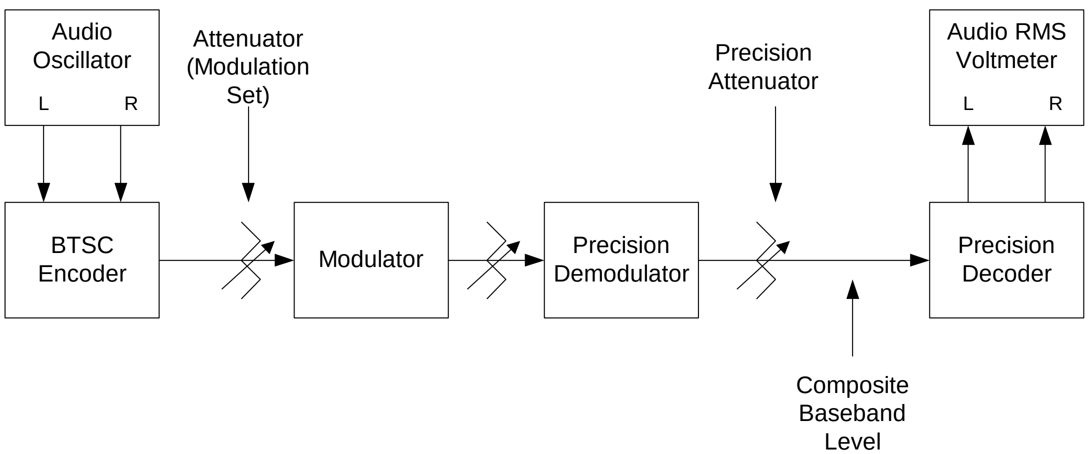


Figure 15-15: Separation Measurements of the BTSC Stereo System - 4.5 MHz

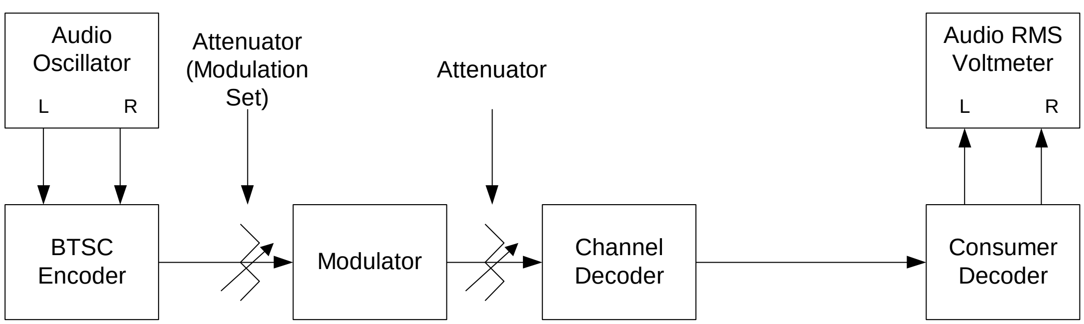


Figure 15-16: Separation Measurements of the BTSC Stereo System - 41.25 MHz

2e. Total Harmonic Distortion of BTSC Stereo Transmission System

Definition: Total Harmonic Distortion is defined as undesirable harmonic content of a modulating signal that is detected and presented at the output of a detector. With respect to stereo audio, it is the amount of unwanted input related signal. It can be detected at the output of a precision decoder when driven by the system under test from a precision signal generator.

Procedure: Required test equipment:

Basic Tests

- A precision audio oscillator with a frequency range of 50 Hz to 15 kHz and distortion output of less than 0.1%
- A reference BTSC stereo decoder capable of accepting the composite audio baseband output of a BTSC encoder
- A precision audio distortion analyzer

Tests involving RF loops at 4.5 MHz

- All equipment outlined under “Basic Tests”
- A precision 4.5 MHz demodulator

Figure 15-17 is a block diagram showing proper test setup of test equipment in this procedure for basic tests.

Figure 15-18 is a block diagram showing the addition of a 4.5 MHz RF loop.

Figure 15-19 is a block diagram showing the addition of a total headend RF loop.

1. Connect audio signal generator to audio distortion analyzer to verify that back-to-back distortion of test equipment is 0.1% or less at 50 Hz, 400 Hz, 1 kHz, *5* kHz, 8 kHz, offset of 10 kHz, 12 kHz, and 14 kHz. This should be done at level of +18 dBm if possible.
2. Connect all equipment per Figure 15-17.
3. Calibrate stereo encoder output to reference decoder input. Note that this is an extremely critical adjustment.
4. Measure total harmonic distortion on the distortion test set.
5. Connect audio generator to one channel only and terminate the other channel input with a 600 Ω resistor.
6. To measure total harmonic distortion at peak deviation at all frequencies, the following frequencies and relative levels should be used.

Frequency	**Level**
400	0
50	0
100	0
1000	-1
3000	-*5*
5000	-8
8000	-12
10000	-14
14000	-17

Table 15-1: Frequencies and Levels for Total Harmonic Distortion Measurement

This ensures that peak deviation will be at or slightly below the maximum allowed.

7. Measure distortion at low modulation levels by adjusting signal generator to reference in Step 6 (400 Hz - 0 Ret) and then reducing level by 20 dB. Distortion can then be checked at any frequency between 50 Hz and 14000 Hz.
8. Distortion measurement should also be made with RF loops shown in Figure 15-18 and Figure 15-19 if possible.

Performance Objective: At all levels and all frequencies, the measured total harmonic distortion should be 1.0% or less.

Discussion: Total harmonic distortion is a common audio measurement that is one indication of overall audio quality. Distortion levels on audio equipment should easily meet the 1% objective. The test method outlined exercises all aspects of the BTSC transmission system including the noise reduction system. If distortion measurements of less than 1% are recorded with this method, satisfactory performance will be experienced by the consumer BTSC decoder with regard to distortion.

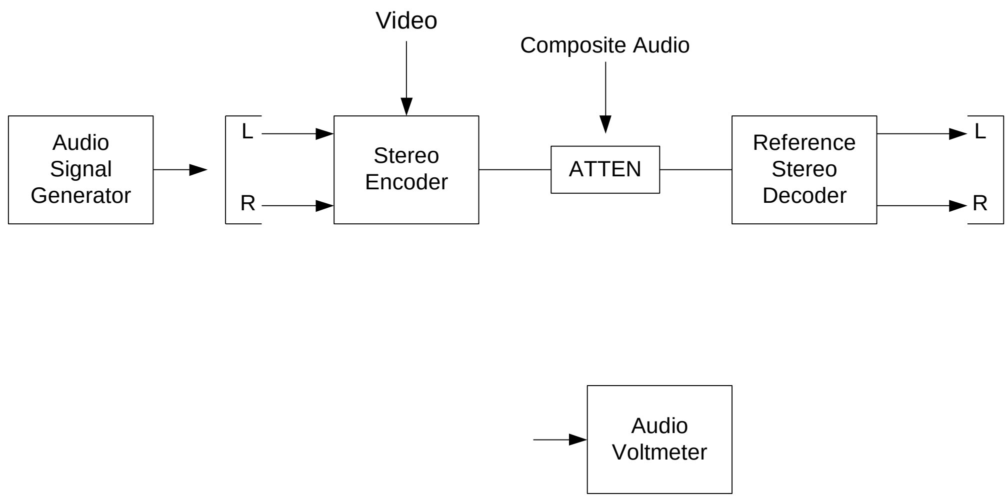


Figure 15-17: Basic Setup - No RF Loop

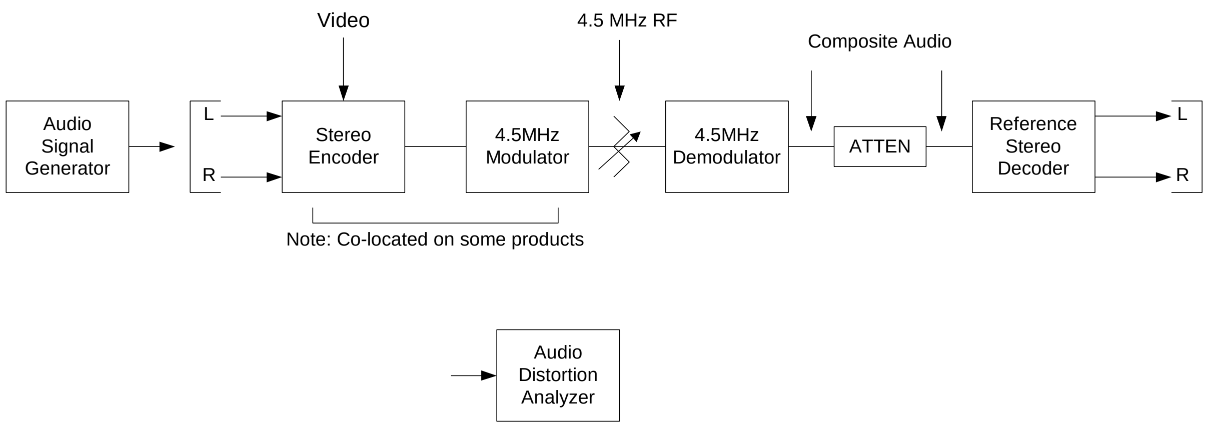


Figure 15-18: Test Setup with 4.5 MHz RF Loop

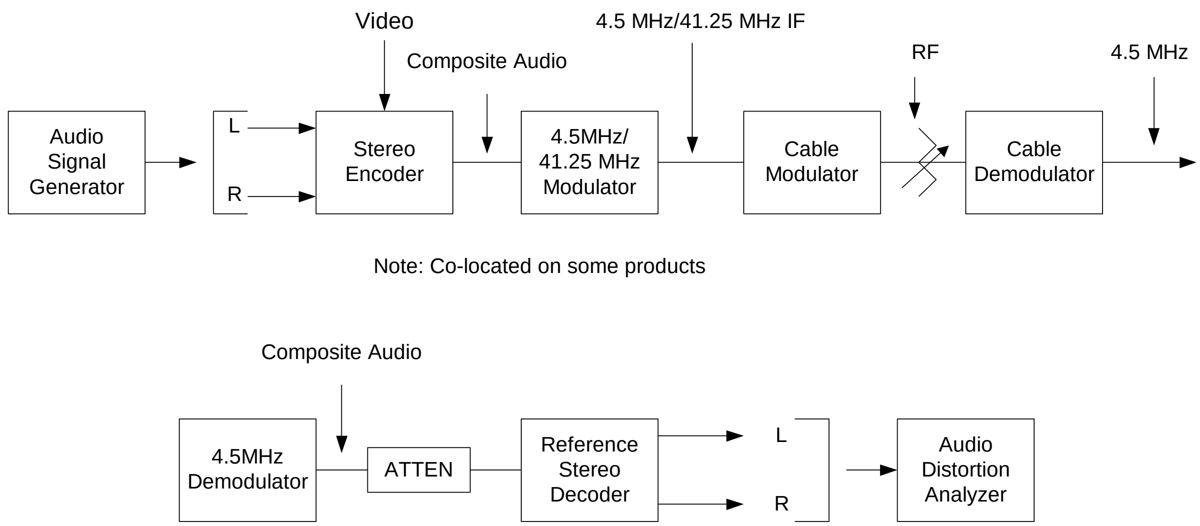


Figure 15-19: Test Setup with Full RF Loop

2f. Incidental Carrier Phase Modulation (ICPM)

Definition: Incidental Carrier Phase Modulation (expressed in degrees) is defined as Phase Modulation of the video carrier with changes in video input signal level, as the signal level varies from blanking to reference white (0 to 100 IRE).

Procedure: Required equipment:

- Video waveform generator capable of generating an unmodulated ramp or 10-step staircase. (Tektronix 1410 or equivalent)
- Oscilloscope capable of X~ Y operation. (Tektronix 475 or equivalent)
- TV demodulator with video and quadrature video outputs (Tektronix 1450-1 or equivalent)
- Two (2) 250 kHz low pass filters (Tektronix Part #015-0352-00 or equivalent)

1. Connect the equipment as shown in Figure 15-20.Connect the VIDEO OUT signal from the TV demodulator to the vertical (Y) input of the scope (as usual) through a 250 kHz low pass filter. The QUADRATURE OUT signal from the demodulator is connected to the oscilloscope's horizontal (X) input through a 250 kHz low pass filter (the 250 kHz filters are not required for this measurement, but they do make the display on the oscilloscope easier to read).
2. Drive the unit under test (modulator) with an unmodulated 10-step staircase or an unmodulated ramp video test signal.
3. Set the oscilloscope's vertical sensitivity (Y) to 0.2 volts per division and the horizontal sensitivity (X) to 20 mV per division (sensitivity might need to be adjusted for your preference).
4. If available, place the TV demodulator in the SYNC DETECTION mode.
5. The display on the oscilloscope should now resemble Figure 15-21. If no ICPM is present, the display will line up perfectly on the vertical axis. If ICPM is present, the displayed waveform will be tilted, as shown, away from vertical.
6. Calculate ICPM as follows:

$$\text{ICPM} = \arctan\left(\frac{\text{Horizontal Deflection}}{\text{Vertical Deflection}}\right)$$

where:

ICPM is in degrees

Horizontal deflection is in volts

Vertical deflection is in volts

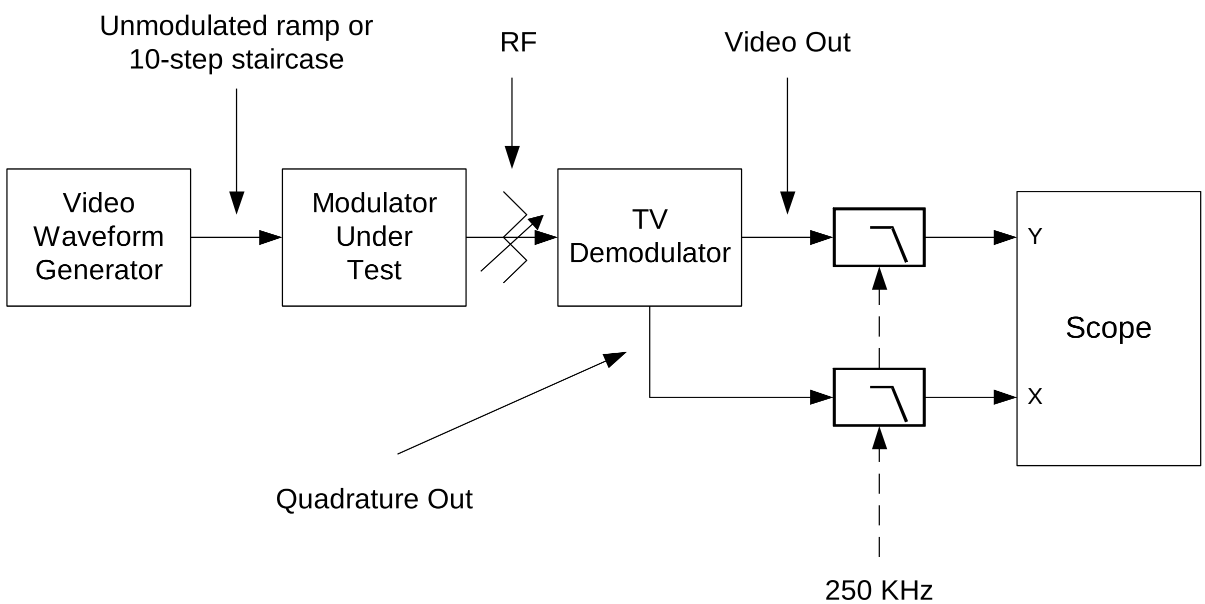


Figure 15-20: ICPM Test Setup

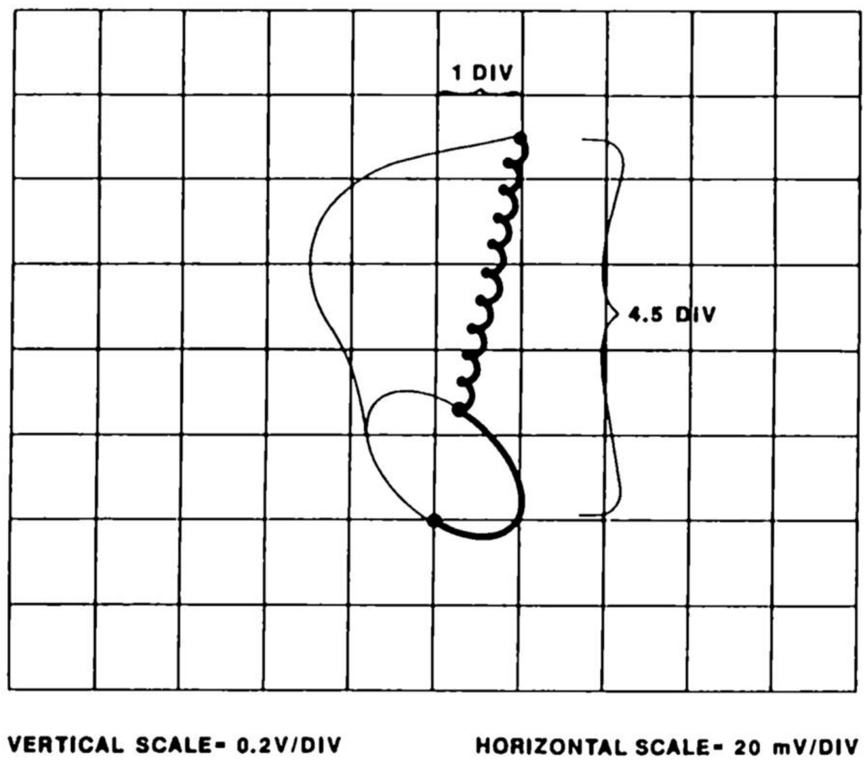


Figure 15-21: ICPM Sample Oscilloscope Display

3. Operating Practices

3a. Interconnecting at Baseband Audio

Discussion: The BTSC composite baseband signal is created by the stereo encoder. This signal contains energy in the band between 50 Hz and about 47 kHz. When SAP is used, energy components are present to about 90 kHz.

The aural carrier is frequency modulated by the BTSC composite signal. The deviation sensitivity of the aural carrier modulator must be precisely set to maintain the performance of the system. The frequency response of the modulator also affects performance.

In some installations, the aural carrier modulator may be included in the BTSC stereo encoder. The output from the encoder is then a frequency modulated aural carrier at 4.5 MHz or 41.25 MHz.

In other installations, the aural carrier modulator is in a separate chassis, with the video modulator. The BTSC composite baseband signal is sent to the modulator, where the aural carrier is created.

The procedure for setting audio modulator deviation below describes how to use the Bessel null technique to precisely set levels from the BTSC stereo encoder to the aural carrier modulator. This procedure applies to all system configurations—those with aural carrier modulators supplied within the BTSC encoder, and those meant to interface with an external modulator.

The Bessel null technique is the primary method of calibrating the deviation of a frequency modulated carrier. In addition, manufacturers may provide secondary measures of establishing BTSC composite levels. Such secondary techniques include built in calibration tones, built in voltmeters, and internal comparators with LED indicators. These techniques are provided to simplify system setup. They may require less equipment and time than the Bessel null technique. As long as the accuracy and precision of these secondary procedures can be verified by the Bessel technique, they are acceptable as a field operating practice.

Internally supplied modulators are expected to be calibrated in manufacture. External modulators must be calibrated upon system installation.

Procedure for Setting Audio Modulator Deviation: Use the Bessel null procedure to set the deviation sensitivity of the aural carrier modulator.

Procedure: Required equipment:

- A low distortion audio sinewave oscillator
- An audio voltmeter
- An audio frequency counter
- A spectrum analyzer

Connect the “BTSC composite baseband” output of the stereo encoder to the aural carrier modulator input (Figure 15-22). Determine what voltage amplitude at the “BTSC composite baseband” output represents 25 kHz deviation. Call this voltage “V_{25k}”.

V_{25k} may be determined in two procedures suitable for field use:

1. Some encoders are provided with a built-in calibration tone. Measure the amplitude of this tone as the encoder drives the modulator.

2. Consult the encoder manufacturer's data for V_{25k}. Check that this data is for the load impedance actually in use. Use the spectrum analyzer to observe the aural carrier created by the modulator. Adjust the analyzer to clearly resolve and display sidebands offset of 10 kHz away from the carrier.

Remove the "BTSC composite baseband" from the modulator input. Note the position of the unmodulated carrier on the spectrum analyzer display.

Set the frequency of the audio oscillator to 10396 Hz.

Use the audio frequency counter to verify that the tone is at a frequency of 10396 Hz.

Connect the audio oscillator to the input of the aural carrier modulator. Use the audio voltmeter to observe the amplitude of the 10396 Hz tone at this point.

Adjust the amplitude at the input of the aural carrier modulator so that it precisely equals V_{25k}.

Locate the "deviation sensitivity" control for the aural carrier modulator.

Adjust the deviation sensitivity of the aural carrier modulator. Observe the change in the spectrum analyzer display. The aural carrier should be surrounded by sidebands 10396 Hz away. When there is 25 kHz deviation, the aural carrier power will be zero. The sidebands will contain all the signal energy.

The deviation sensitivity of the aural carrier modulator is now calibrated. 25 kHz deviation will result when a signal of amplitude V_{25k} is applied to the modulator input.

Sources of Error in the Bessel Null Technique: The amplitude of the carrier of an FM signal is proportional to $J_0(\beta)$. $J_0(\beta)$ is the Bessel function of the first kind, zero order. β is the modulation index of the FM signal:

$$\beta = \frac{\text{deviation of carrier}}{\text{modulating frequency}} = \text{modulation index}$$

The amplitudes of the various sidebands around the carrier are given by the different Bessel functions $J_n(\beta)$, where n = 1,2,3... n indicates which sideband is being considered.

The carrier amplitude will equal zero when $J_0(\beta) = 0$. For this modulation index, the sidebands will contain the energy of the transmitted signal. For example, $J_1(\beta)$, $J_2(\beta)$, and others will be non-zero when $J_0(\beta) = 0$.

It is desired to set the peak frequency deviation of the carrier to 25 kHz. When a modulating frequency is also chosen, the modulation index *8* may be calculated. The smallest value of B for which $J_0(\beta) = 0$ is ≈2.4048.

Choose the modulating frequency so that $\beta = 2.4048$ when the deviation of the carrier is 25 kHz.

$$\text{Modulating frequency} = \frac{\text{deviation of carrier}}{2.4048} = \frac{25\text{ kHz}}{2.4048} = 10396\text{ Hz}$$

When a 10396 Hz tone modulates an FM carrier with kHz deviation, no energy is transmitted right at the carrier frequency. This effect can be observed on a spectrum analyzer. As the deviation is increased toward 25 kHz, the power at the carrier frequency decreases. The carrier disappears at 25 kHz deviation while the sidebands remain. At deviations greater than 25 kHz, power again appears at the carrier frequency.

The accuracy of the Bessel null technique is limited by the frequency accuracy of the modulating tone used, its distortion content, and the noise level of the spectrum analyzer display that is observed for the carrier null.

The effect of frequency accuracy of the modulating tone can be determined as follows:

At the first Bessel null of the carrier:

(Actual peak deviation) = (Actual modulating frequency)*(2.4048)

An error in deviation can be expressed in decibels as:

Error, dB = 20*log (Actual deviation/Desired devaluation)

Desired peak deviation = 25 kHz

Modulating Frequency (Hz)	Deviation (Hz)	Error (dB)
10360	24914	-0.030
10370	24938	-0.022
10380	24962	-0.013
10386	24976	-0.008
10390	24986	-0.005
10396	25000	0.000
10400	25010	+0.003
10406	25024	+0.008
10410	25034	+0.012
10420	25058	+0.020
10430	25082	+0.028

Table 15-2: Example Errors in Deviation in dB

Thus, if a modulating frequency of 10396, + /- 10 Hz is used, the peak deviation will be within 0.01 dB of 25 kHz when the carrier is completely nulled.

The effect of noise floor can be determined as follows:

Let the unmodulated carrier have a voltage amplitude of 1. Expressed logarithmically, its amplitude is 0 dB.

Absolute Amplitude	Amplitude (dB)
1.000000	0
0.316200	-10
0.100000	-20
0.031620	-30
0.010000	-40
0.003162	-50

0.001000	-60

Table 15-3: Amplitude in dB vs. Absolute Amplitude

As the carrier amplitude is reduced during the Bessel null procedure, the noise floor on the spectrum analyzer display limits the visibility of a true null.

For example, suppose the noise floor allows a range of 50 dB in observable carrier amplitudes. Then when the carrier appears to be nulled it may actually be only 50 dB down from its unmodulated amplitude.

Where it is desired that $J_0(\beta) = 0$, β may only be set such that $J_0(\beta) = -0.003162$, or $J_0(\beta) =-0.003162$, or $J_0(\beta) = 0.003162$.

The values of closest to the first zero of $J_0(\beta)$ that will satisfy these limits are $J_0(2.3987) = 0.003162$ and $J_0(2.4109) = -0.003162$.

The effect of nulling accuracy when using a modulating tone of exactly 10396 Hz is:

(Actual peak deviation) = (10396 Hz) * (Actual modulation index)

As before, the deviation error can be expressed as:

Error, dB = 20*log (Actual deviation/Desired deviation)

Carrier Null Depth (dB)		Limits on Peak Deviation	Error (dB)
60		24980	-0.007
	or	25020	+0.007
50		24937	-0.022
	or	25064	+0.022
40		24801	-0.070
	or	25201	+0.070
30		24375	-0.220
	or	25642	+0.220

Table 15-4: Deviation Error in dB vs. Carrier Null Depth

This chart shows that if a carrier amplitude range of 60 dB can be observed, deviation accuracy within 0.007 dB may be obtained. In most field environments such measurements are possible directly at the modulator's RF or IF output.

For a worst-case analysis, assume that error in the modulating frequency will affect results in the same direction as error in the depth of carrier null. Then the magnitude of the individual errors may be cascaded.

To achieve a deviation accuracy within 0.03 dB of 25 kHz deviation, it is sufficient to use a modulating tone of 10396 Hz, ±10 Hz, and set deviation to null the carrier frequency power at least 50 dB from its unmodulated amplitude.

Performance Objective: The deviation sensitivity of the aural carrier modulator must be matched to the operating output level of the BTSC encoder to within ±1 dB. An error of only 0.28 dB will limit the best achievable separation to 30 dB.

The amplitude response of the audio carrier modulator must be flat within ±1 dB from 50 Hz to 47 kHz. The phase response must be linear within ±3 degrees from 50 Hz to 47 kHz at carrier deviations up to 50 kHz.

For proper transmission of the SAP signal, the aural carrier modulator amplitude response must be flat within ±1 dB from 47 kHz to 100 kHz. Phase linearity should be within ±10 degrees from 47 kHz to 100 kHz.

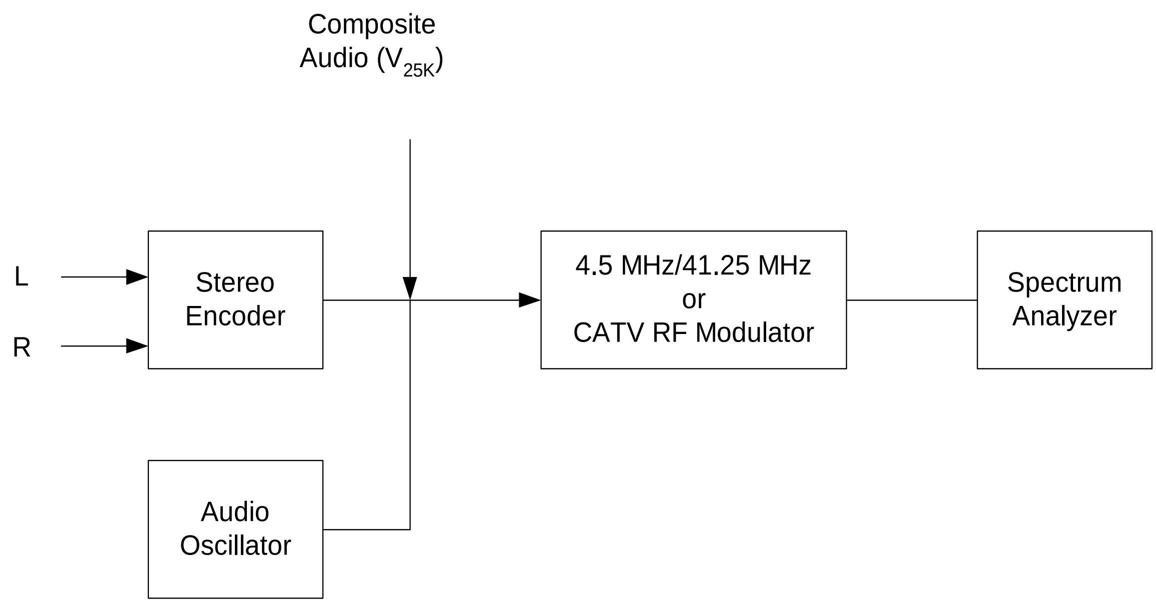


Figure 15-22: Test Setup for Setting - Deviation Sensitivity

3b. Interconnect with Modulators at 4.5 MHz

Discussion: The generation of BTSC stereo signals requires precise calibration of baseband audio levels. For this reason, it is very common to interface a BTSC encoder in such a manner that only non-critical adjustments must be made at the time of installation and on an on-going routine basis. Such is the case when the encoder contains a built-in 4.5 MHz subcarrier modulator. In this case, the encoder manufacturer is performing all critical baseband level adjustments as an internal part of the encoder. Installation of BTSC encoders with 4.5 MHz interfaces requires only two adjustments, both of which are non-critical

Procedure: <u>Individual left and right channel input level</u> - These two adjustments must be made to allow audio peaks of frequent recurrence to reach the maximum level specified by the manufacturer. There will usually be a peak level indicating device per channel as part of the encoder. By definition, if both are driven to peak level at the same time by the same source (L = R), then the encoder will be producing only L+R information plus the pilot tone. Stereo separation will not be affected by misadjustment of these controls. If set too low, low audio levels and low SNR can be expected. If set too high, audio distortion will result.

<u>4.5 MHz RF output level</u> **-** This adjustment determines the amount of audio carrier level that will be contained in the final signal. Most cable TV modulators are capable of accepting this 4.5 MHz signal

and using it as the audio source for a particular TV channel. There will usually be an aural carrier level adjustment control on the TV modulator as well as a 4.5 MHz output level control on the BTSC stereo encoder. These two adjustments will interact to some extent depending on manufacturer. As a general rule, the 4.5 MHz output level from the BTSC encoder should be set to the nominal input level specified for a particular TV modulator and final level adjustment made with the TV modulator aural carrier level control. As a starting point, the BTSC encoder should be adjusted to produce 100 millivolts peak-to-peak of 4.5 MHz energy at its output into a 75 Ω load. The encoder should then be connected to the TV modulator and the aural carrier level adjusted on the TV modulator to the level normally used in the plant. (Typically, it is 13-17 dB down from video carrier level.)

<u>Combined video +4.5 MHz and separate video and 4.5 MHz considerations</u> - Most BTSC encoders are capable of supplying the 4.5 MHz signal on a dedicated 75 Ω output or combining it with the video signal to present a video +4.5 MHz signal to the TV modulator. The method of interconnect will depend upon which type of signal the TV modulator is capable of accepting. As a general rule, it is advisable to keep the signals separate if possible. Several TV modulators are capable of this as shipped from the factory. Many can be field modified to interconnect this way with a factory approved modification. Ideally, there should be one, and only one, signal path that the 4.5 MHz signal can take through a TV modulator. Although sometimes satisfactory, the combined video +4.5 MHz signal can lead to interaction between video and audio that manifests itself in visible beats in the video signal. There are also some cases where combined signals have been reported to cause distortions to the BTSC stereo signal.

Summary

1. Interconnect BTSC encoders and TV modulators with separate video and 4.5 MHz signals if possible. Field modification of the TV modulator may be necessary to accomplish this.
2. Adjust 4.5 MHz level from the BTSC encoder to produce a 100 millivolt peak-to-peak signal into a 75 Ω load.
3. After interconnecting the BTSC encoder and TV modulator, make final aural carrier level adjustment with TV modulator aural carrier level control.

3c. Interconnect with Modulators at 41.25 MHz.

Discussion: The generation of BTSC stereo signals requires precise calibration of baseband audio levels. For this reason, it is very common to interface a BTSC encoder in such a manner that only non-critical adjustments must be made at the time of installation and on an on-going routine basis. Such is the case when the encoder contains a built-in 41.25 MHz subcarrier modulator. In this case, the encoder manufacturer is performing all critical baseband level adjustments as an internal part of the encoder. It remains the task of the operator however to ensure that the 41.25 MHz interface level is adequate to produce the desired visual/sound carrier ratio at the RF output of the modulator.

Procedure: <u>Individual left and right channel input level</u> - These two adjustments must be made to allow audio peaks of frequent recurrence to reach the maximum level specified by the manufacturer. There will usually be a peak level indicating device per channel as part of the encoder. By definition, if both are driven to peak level at the same time by the same source (L = R), then the encoder will be producing only L+R information plus the pilot tone. Stereo separation will not be affected by misadjustment of these controls. If set too low, low audio levels and low SNR can be expected. If set too high, audio distortion will result.

<u>41.25 MHz RF output level</u> - This interface scheme is used primarily when the modular being used has no provisions for accepting an external 4.5 MHz input. In many cases modifications are required

to the modulator for interfacing with an external 41.25 MHz source. Depending upon the particular brand of modulator, it is also possible that the actual interface level at 41.25 MHz will determine the visual/sound ratio at the RF channel output. In this case it is necessary that the 41.25 MHz interface level be properly adjusted and stable over time and temperature. In applications where the interface is applied to the input of the RF output converter, it is also important to ensure that the 41.25 MHz source contains no undesired spurious signal components (i.e., 45.75 MHz, 50.25 MHz, etc.). These components, when present at the output converter mixer input, could cause undesired spurious signals in the RF output spectrum.

It should also be realized that when interfacing at a frequency of 41.25 MHz, the visual/sound intercarrier frequency spacing will be determined by two different sources. One is the frequency accuracy of the 45.75 MHz video source and the other is the frequency accuracy of the 41.25 MHz audio source. It is important that frequency accuracy of the spacing between these two carriers be maintained at 4.5 MHz ±1 kHz.

Summary

1. Interconnect BTSC encoders and TV modulators with video and 4.5 MHz audio if possible. An acceptable alternative is to interface the encoder/modulator at 41.25 MHz. Field modification of the TV modulators may be necessary to accomplish this.
2. Adjust the 41.25 MHz level from the BTSC encoder to ensure that the desired visual/sound amplitude ratio at the RF output of the modulator is achieved (may also be determined by audio level control in modulator depending upon interface point with modulator).
3. Check to ensure that the visual/sound frequency separation is 4.5 MHz ±1 kHz.

3d. Online Checks

Discussion: Critical parameters of BTSC stereo performance include signal-to-noise ratio, signal-to-buzz ratio, frequency response, stereo separation, and relative phase between left and right signals. Quantitative performance standards and measurement techniques are given elsewhere in the NCTA standards for BTSC operating practices. Those techniques require interruption of service.

Presented here are methods for on-line checks of stereo performance without interruption of service. These methods provide qualitative indications of performance or allow subjective evaluation of various parameters. They are not a substitute for periodic proof-of-performance measurements. They provide a continuous method of monitoring stereo signals to obtain a reasonable amount of confidence in signal quality.

Procedure: Required equipment:

- BTSC stereo decoder (high-end consumer quality is sufficient)
- Headphones
- Oscilloscope with X-Y display mode

Perform the following listening tests through the headphones.

Signal-to-Noise: Place the stereo decoder in the “mono” mode. Listen to the dynamics of program audio. Observe program video. Buzz in the audio may be video dependent. Listen for changes in noise as scenes change. Adjust video modulation depth for possible improvements. Avoid video over modulation. Note that some noise may originate in program source material.

Buzz: Listen to the stereo decoder while switching the decoder between its "mono" and "stereo" modes. If buzz is audible only when listening in the "stereo" mode, the noise is in the "difference" channel. Difference mode buzz may be caused by some scrambling systems or by interference from video in the aural carrier demodulator.

Frequency Response: Listen to the stereo decoder in "mono" mode or to a conventional monophonic television receiver. If the audio sounds very tinny and the high frequencies are distorted, the aural carrier modulator may be adding pre-emphasis to the BTSC composite signal. The aural carrier modulator should be set to provide a flat frequency response, not the pre-emphasized response used in monophonic transmissions. Pre-Emphasis is provided by the BTSC stereo encoder. Alternatively, check the audio input connections to the stereo encoder. Improper grounding or incorrect impedance matching may cause frequency response degradations or hum.

Left and Right Channel Relative Phase: Listen to the stereo decoder in "mono" and "stereo" modes. The apparent loudness between the modes should not change very much. If the loudness drops severely when listening in the mono mode, the left and right channels may be out of phase. Check the audio path at or before the stereo encoder inputs.

Stereo Separation: Use the oscilloscope in the X-Y display mode. Connect the left channel output of the decoder to the "Yin input of the oscilloscope. Connect the right channel output of the decoder to the "X" input of the oscilloscope. Set the input sensitivities to be equal. When the signals on the left and right channels are the same, as during a mono transmission, the dot will move to create a diagonal line on the scope screen. The line will be angled 45 degrees from horizontal, traveling from the bottom left of the screen to the top right. (Do all scopes have the same polarity?)

If the left and right channels have equal but opposite signals, they are said to be "out of phase". The display will now show a diagonal line angled 135 degrees from horizontal, traveling from the top left of the screen to the bottom right. Mono receivers will produce little audio output under this condition.

During stereo broadcasts, the left and right channels will often have unequal signals. The display will show a "scribbly" pattern, usually tilted toward the axis of in-phase mono programs.

If the left and right channels are out of phase during a stereo broadcast, the "scribbly" pattern will be tilted toward the axis of the out-of-phase mono programs. Mono receivers will produce little audio output.

Another check for stereo separation in the BTSC system may be performed, though it disturbs online performance. Listen to the stereo decoder. Disconnect the right channel audio input from the BTSC encoder in the headend. The right channel audio should become greatly attenuated as heard in the headphones, while the left channel remains largely unaffected.

Performance Objective: Numerical standards for stereo performance are given elsewhere in the NCTA operating practices. The procedures given here are sufficient only to indicate the occurrence of catastrophic system failures or major changes in stereo performance. They are supplied as an aid to the recognition and troubleshooting of system failures.

3e. BTSC Operational Guidelines

Discussion: In most installations, the operation of BTSC stereo encoding may encounter difficulties caused not specifically by the encoder, but rather due to operating conditions. In most cases, these difficulties can be avoided through adherence to a few basic operating guidelines and practices.

The following section details some potential problem areas and recommended practices.

Video Sample: In order to operate a BTSC encoder, it is necessary to provide a video sample for pilot timing. This normally involves merely looping the program video through the encoder on its way from the source to the modulator. There are, however, several potential problems, particularly with satellite video and encryption systems.

Satellite Video: The video output of many satellite receivers may not be properly bandpass filtered. The lack of filtering may allow auxiliary subcarrier information to appear at the video sample detector of the stereo encoder. Depending on the number and level of subcarriers, the sync pulses may be masked and make phase locking of the pilot erratic. This can result in intermittent pilot lock loss and buzz or breakup of the sound. The normal cure for this effect is the installation of a video bandpass filter ahead of the encoder.

Encryption Systems: The video sample must have sync pulses intact, and must therefore be derived before any baseband sync suppression encryption. While this is normally no problem, it can be difficult if the final stereo encoder output is required as video/4.5 combined. In most encoders the same loop is employed for pilot sample and video/4.5 combining. These applications may require the use of external video 4.5 combiners.

Commercial Insertion: Video switching for stereo commercial insertion must occur before the Pilot Video Sample loop to the Stereo Encoder.

Input Handling: If stereo encoding were concerned with only a permanent discrete left and right source, few difficulties would be encountered. In many cases, however, source switching and stereo/mono switching may be required. The wide variety of operating circumstances to be encountered in headend layouts require particular attention to input source handling. Assuming that the proper input connections have been made, the following points should be noted: In certain cases, the inputs to the Stereo Encoder may be delivered as Matrixed (L+R, L-R) signals. Matrixed signals can tolerate very little phase or level imbalance without severely affecting the stereo reconstruction.

Should it be necessary to operate in this manner, particular attention should be paid to interconnect wiring. Interconnects should be made with short, identical wiring from source to encoder. Any buffer amplifiers required should have identical gain and frequency responses.

Once a BTSC encoder has been properly installed in a system, the adjustment of relative volume can no longer be performed by modulation deviation adjustments or by modulator baseband input level adjustments: It becomes necessary to adjust the BTSC encoder baseband inputs.

In the case of Encoders operating with discrete left and right source inputs, level imbalance between the sources will result in rotation of the stereo center to either side: an effect not unlike a stereo balance control.

In the rare case of Encoders operating with matrix (L+R, L-R) input sources, this adjustment becomes critical, and should be performed only with Programmer supplied test tones if possible. It may be required, particularly in the case of commercial insertion, to employ source switching. In these cases, several items should be considered. These are:

- level matching
- phasing
- DC biasing
- stereo/mono switching

In the case of transfer switching for commercial insertion, significant potential exists for phase reversal, left/right cross-over, and relative level imbalance. These problems can normally be avoided by careful attention to wiring details. It is, however, unavoidable that taped commercials will contain differing audio levels. In addition, commercials will vary between stereo and mono.

In cases where the commercial insertion program is mono, the commercial switcher should also drive the stereo encoder to mono mode rather than have subscribers with a mono signal and stereo indicator.

Level variations can be handled by automatic level control circuits either in the commercial insertion equipment or in the Stereo Encoder. It is imperative, however, that the ALC circuits be tracking, with constant equal variations, and be keyed from a combined L+R sample. Individual ALC circuits on left and right sources will cause stereo center rotation as the left or right level varies independently.

Commercial insertion switching may also cause a pop or bounce in the audio during transition. This can normally be minimized by reducing the residual DC present on the audio lines. It is also possible to reduce this effect through the use of balance transformers as switch interface connections if required.

Modulator Interface: There are two basic interface methods between Stereo Encoder and Modulator, although there are many minor variations on these methods. The interface will be either baseband, or modulated carrier.

Baseband Interface: In cable television applications, the baseband interface will rarely be employed, as very few existing modulators have the ability to hand 48 kHz baseband signals. The baseband interface also present calibration problems which most manufacturers have addressed through precision internal modulators.

The most common interface will involve modulated subcarrier, either at 4.5 or IF. This allows the manufacturer to provide a signal which has a proper present deviation, as well as the necessary bandwidth to deliver the BTSC signal. This interface avoids most of the BTSC inherent problems of modulator capability and precision setup. The modulated subcarrier will interface in one of the following three methods. Each of these methods has its own advantages and disadvantages, of which a few are:

The Video/4.5 combined input is the most commonly available 4.5 interface. The majority of these were, however, designed with the original 25 kHz deviation in mind and occasionally suffer bandwidth restrictions as well as other limitations which can cause distortions in the wide deviation BTSC signal. Several unique problems also arise when baseband video encryption is employed.

The separate video and 4.5 is probably the least difficult method and presents less potential for signal degradation. Unfortunately, very few modulators have a separate 4.5 input facility. It is, however, a very simple modification to implement and most manufacturers have modification information available.

The IF interface is very simple to implement provided that the modulator employs dual IF loops. One major difficulty is that when IF interface is employed, the intercarrier frequency stability becomes a function of both the video oscillator in the modulator and the audio oscillator in the encoder. With 4.5 interface, the intercarrier stability is solely a function of the encoder 4.5 MHz oscillator performance.

From an operational viewpoint, several practices and alignments can cause difficulty with BTSC signals in modulators.

While some items, such as ICPM, are in essence a manufacturing problem, other items are within the capability of the operator to control. Within the modulator, several processes may contribute to degradation of either buzz or separation performance.

Video spectral content: Ideally, the video response of a modulator should show a 30 dB cutoff between 4.2 and 4.5 MHz. This is rarely achieved or maintained. If the modulator is presented with a high frequency load, such as unfiltered satellite video or high level alphanumeric content, significant energy can be developed in the 4.5 MHz spectral area.

Alpha generators and titlers can present video over and undershoots which can significantly degrade buzz performance.

Narrow or non-symmetrical aural IF filters can remove sidebands from the aural carrier degrading separation. This can also occur in IF scrambling equipment.

Sync suppression scrambling systems may generate sufficient 30 Hz sidebands to create buzz problems. Care should be taken in alignment of suppression depth and timing.

Most commonly encountered problems with stereo encoder installation and some probable causes:

- Lack of Separation
  - Improper deviation setup
  - Poor modulator audio frequency response
  - Narrow or imbalanced 4.5 or audio IF. filters
  - Severe left vs. right input level or phase imbalance
- Buzz
  - Video over-modulation
  - Poor modulator video spectrum filtering
  - Video overshoot from alphanumeric generators
  - Pilot lock errors due to interference on video sample
  - Poor video sample clamping or stability
  - Gated sync suppression encryption
- Poor Quality Sound
  - Phase reversal on left or right input
  - Frequency response problems on left or right input
  - Overdriving left or right inputs
  - Modulator baseband frequency response problems

While it is not possible to cover all potential problems, the adherence to several basic recommendations, will alleviate the majority of difficulties.

  - Employ short, identical input connections, particularly with matrixed signals.
  - Observe + and - phase
  - Ensure a clean, stable video sample.
  - Employ video and separate 4.5 modulator interface, if possible.

- Maintain modulation depth of all components, including character generator over and undershoots, at 87± 1/2 % maximum. Reduce modulation depth slightly, if required, for buzz performance.
- Do not undertake arbitrary control adjustments unless proper measurement equipment is available.

3f. Transportation of Stereo Signals Over FM Links

Introduction: Field experience[1] as well as laboratory tests[2] have shown that the delivery of quality stereo television over a cable television system is feasible. The effect of the distribution system itself is slight.[2] The interface between stereo encoder and modulator, though critical, has received much attention and is well under control.[3] Scrambling is fairly well understood and at least some systems cause little degradation.[2] The transportation of stereo between hubs, satellite receiving stations, and headends, however, has until recently received very little attention. FM links are now available that are designed to carry BTSC stereo.

AML's and FM links are routinely used in cable television systems. When used properly this equipment is nearly transparent to video and audio signals. Care should be taken, however, when adapting existing systems to carry BTSC stereo. Older FM links, in particular, can cause significant degradation of the stereo signal if improperly used.

The FM Link: The newer BTSC-ready FM links typically take a 4.5 MHz audio carrier and multiply the frequency of this carrier by some factor to increase the frequency deviation as well as the carrier by some factor to increase the frequency deviation as well as the carrier frequency. They then combine this with baseband video as a subcarrier and frequency modulate a VHF carrier with this composite video-plus-subcarrier signal. The increased frequency of the audio subcarrier aids in the filtering required to separate video and audio, thus alleviating problems including buzz due to video in audio, and audio carrier multipath. Increasing the audio subcarrier deviation prevents the increased noise at the higher subcarrier frequency (due to FM triangular noise) from degrading the subcarrier-to-noise ratio.

The older FM links frequency-modulate one VHF carrier with video information and a separate VHF carrier with audio information. In upgrading such a system all that is necessary is to add one more carrier for a second audio channel. One carrier is then assigned to the left audio channel and the second to the right. Encoding of the audio into the BTSC format can be done after the link, with the encoder and modulator co-located. All the performance specified for the FM link is preserved.

To reduce the number of stereo encoders, or in systems that are near or at capacity for audio channels, it is tempting to avoid adding the extra audio channels. Most stereo encoders have built-in 4.5 MHz subcarrier modulators and provide video + audio subcarrier at their outputs. It would seem a simple matter to connect such an output directly to the video input of the FM link and avoid the separate audio channels altogether. Unfortunately the results could be quite marginal for the reasons to be discussed. The following precautions apply especially to the older FM links that were not designed specifically to carry the BTSC stereo signal.

<u>Video Clamps</u>: In sending video over FM links some sort of DC restoration may be used to recover the proper DC level of the video independently of APL variations. Video clamps normally act to restore the sync tips to a constant DC voltage. Although all clamp circuits use active components, some circuits are more "active" than others. These circuits detect the presence of the sync and then trigger active devices to saturate and clamp the signal path to some stiffly held voltage. Unfortunately, any subcarrier present in the signal also is effectively "shorted" to this voltage for the

duration of the clamping period. This produces an amplitude modulation of the subcarrier at the horizontal frequency (f_h) that can approach 100%. Any subsequent AM to FM conversion in the signal path or the consumer's receiver will then cause a discrete carrier at f_h which will add vectorially to the existing pilot. The resultant pilot will have a phase that is different from the phase of the original pilot. This phase error will cause the synchronous subcarrier detector to add a phase error to the recovered difference signal and thus seriously degrade stereo separation. Buzz can also be added to the audio due to inadequate AM rejection in the receiver's FM demodulator.

Bandwidth Considerations: Because the video channels are normally intended to carry only video, frequency response is usually specified only to 4.2 MHz. Though not necessarily present, the user should beware of sharp low-pass filters or transmission zeros in the signal path at frequencies above 4.2 MHz.

Before combining video with a 4.5 MHz subcarrier in any system, low-pass filtering of the video may be required to prevent spectral overflow of the video from interfering with the audio carrier.

A less obvious problem is that of over-modulation. FM links normally have bandpass filters at the output of the transmitter in order to protect adjacent channels. Selectivity is also provided in the receivers in order to protect adjacent channels. Selectivity is also provided in the receivers in order to reject adjacent channels. The bandwidths of these filters, along with the linear deviation range of voltage controlled oscillators (VCO), normally limit the deviation of the video carrier. With this upper limit in mind, manufacturers usually deviate as much as possible to preserve video signal-to-noise ratio. If a user then adds a subcarrier to the video signal without reducing the modulation sensitivity of the VCO, two things happen: deviation is increased by the added peak voltage of the video plus subcarrier (usually .1 (to .3) v p-p), and the highest modulating frequency increases from 4.2 MHz to about 4.6 MHz (stereo + SAP).

The resulting increase in bandwidth can exceed the bandwidth of the channel resulting in increased distortion of the audio and video.

Signal-to-Noise Ratio: The transmission of video plus audio subcarriers as a frequency modulated carrier is not without precedent in the cable television signal chain. Satellite delivery of signals is normally done this way, so much is already known about the technical tradeoffs involved. Early on it was discovered that better signal-to-noise ratio (SNR) could be obtained by using multiple audio subcarriers as opposed to one multiplexed subcarrier. The issue was more recently resurrected with the possibility of distributing stereo over the satellite already in the BTSC format. Most recently Mountain[5] showed that a substantial penalty in SNR would be incurred. The same consideration applies to an FM video plus multiplexed subcarrier signal, regardless of how it is transmitted.

Conclusion: Newer FM link equipment is designed to carry BTSC audio with little degradation. In older FM equipment, transmission of separate left and right audio information over individual audio channels is an excellent way to transport stereo.

In older FM equipment it is possible to transmit BTSC stereo as a 4.5 MHz subcarrier added to video over an FM link. By disabling active clamps, checking frequency response, filtering any spectral overflow and reducing deviation the BTSC signal should survive, though bruised with a reduced SNR. The problem is that the FM link is only a small part of a complex distribution system. No reasonable degradation budget could allow for any one part of the system to be marginal.

References:

1. Farmer, J., Sedacca, D., Williams, T., Jones, H., “Practical Experience with Stereo Cablecasting’ CED June 1986.
2. Rovira, L. “A Test of BTSC On The Cable”, CED December 1986.
3. Bowick, C., “The Importance of Setting and Maintaining Correct Signal and Modulation Levels in a CATV System Carrying BTSC Stereo Signals’; presented at the 1986 NCTA Convention.
4. Farmer, J., “RF and Stereo”, CED December 1985.
5. Mountain, N., “BTSC The Future of Stereo?”, CED June 1985.

4. Cable Error Budget (Separation)

Definition: A system separation budget is a calculation of the expected stereo separation through the entire cascade of headend, transportation link, distribution equipment and cable. The budget calculation is based on the required performance of the individual pieces of equipment. Measurements can be performed on the individual pieces of equipment to evaluate suitability or to initially decide on numbers for the budget.

Performance Objective: A minimum of 24 dB of separation (worst case) should be measured using a laboratory quality receiver and stereo decoder at the output of a complete system. This system performance, combined with a typical consumer decoder separation of 22 dB, will provide a worst case separation of 17 dB (typical separation of 21 dB) to the subscriber.

Performance objectives for the individual pieces of equipment depend on the budget for the system as described below.

Discussion:

Measurement Errors: In order to decide on the amount of degradation to allot to each piece of equipment or to verify if a particular piece of equipment meets the error budget, measurements of separation through that piece of equipment must be made. A measurement of separation is first made on the test encoder and decoder. Then a further measurement is made on the piece of equipment in question, embedded between the encoder and decoder (when testing the stereo encoder, the test encoder is not required).

The accuracy of this test depends on the quality of the test encoder and decoder as shown in Figure 15-23. The horizontal axis in this figure represents the separation of a device under test. The curved lines represent the separation of the test equipment used. Note that there is a top and bottom curved line for each value of test equipment separation. The vertical axis represents the error involved in measuring a given device with imperfect test equipment. The plot is used in the following manner

a. Locate the intersection of the vertical line corresponding to the true (known) separation of a piece of equipment with the top curved line for the test equipment separation. Read the error to the left of this intersection.

b. Locate the intersection of the same vertical line with the bottom curved line corresponding to the test equipment separation. Read the error to the left of this intersection.

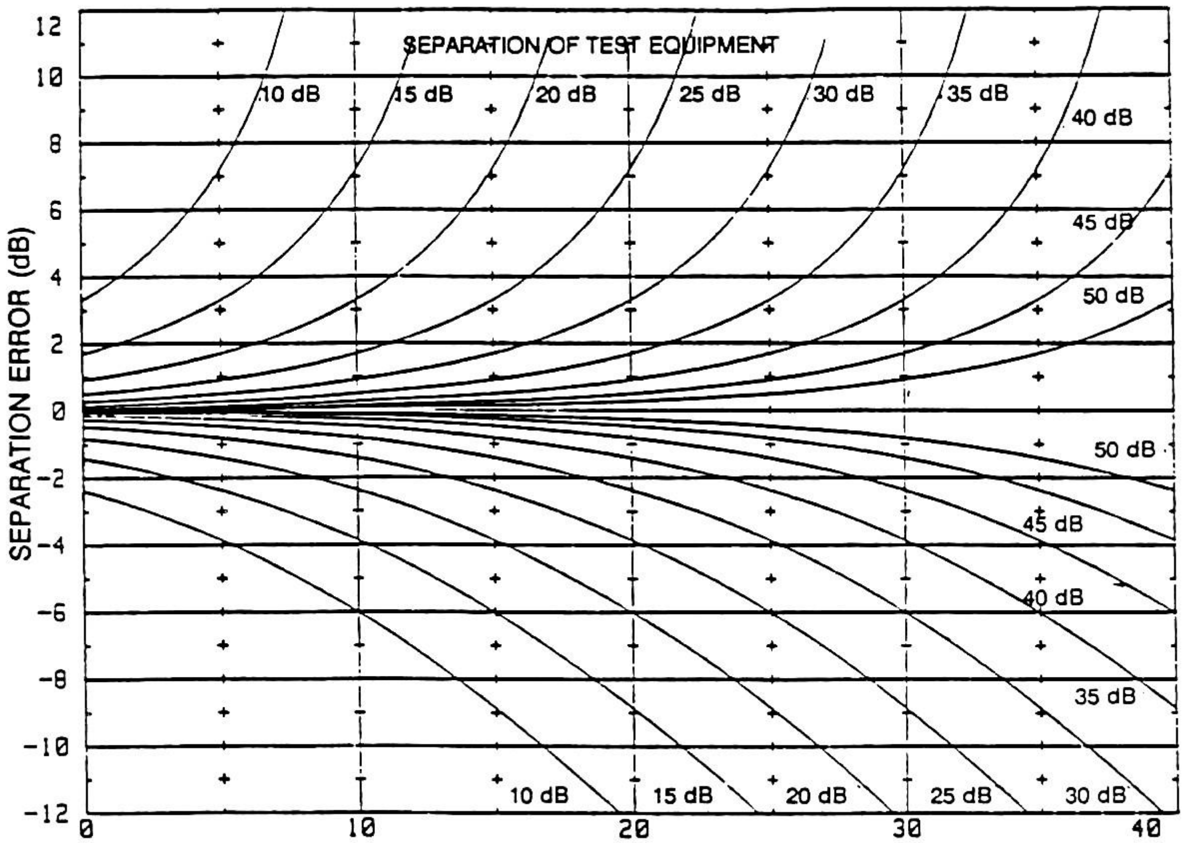


Figure 15-23: Separation of Device Under Test (dB)

The errors read from this plot represent the boundaries of the range of uncertainty. For example, a device is to be measured which has a known separation of 20 dB, using test equipment with a separation of 35 dB. Reading from Figure 15-23, the error involved is approximately

-1.4, + 1.7 dB

That is, the result of the measurement could be anywhere between 18.6 and 21.7 dB.

This graph clearly shows the futility of using anything but laboratory quality equipment in the measurement of separation. As a rule-of-thumb the combined test equipment should have a separation >15 dB better than that of the equipment under test.

In some cases, it is not possible to test one piece of equipment alone between an encoder and decoder. For example, a modulator cannot be embedded between an encoder and decoder without also including a demodulator. A signal processor or distribution system cannot be tested without a modulator and demodulator. Test demodulators are available (stereo versions of Tektronix 1450 or Scientific-Atlanta 62250) that are close enough to ideal that most separation degradation can be attributed to the modulator. Then, to measure the degradation of a signal processor or distribution system, use the modulator and demodulator along with the encoder and decoder as the test equipment.

In practical applications, it often occurs that the equipment to be tested is capable of higher separation than the test equipment within which it is embedded. A distribution system is a good example of this. Its separation degradation cannot be measured without using a modulator and demodulator. Yet, a good distribution system will be capable of higher separation than a modulator. Looking at Figure 15-23 we see that it is a waste of time to try to determine a number for the distribution system based on such a test. In such a situation, it is best to combine the effect of the modulator and distribution system and use this “subgroup” as one entry in the budget table. The separation of the combined equipment is known to the accuracy allowed by the encoder-decoder. An advantage of subgrouping is also that the uncertainty involved in calculating the effect of cascading the individual pieces (see following section) is eliminated by measuring the separation of the combined equipment. Unfortunately this approach loses the flexibility of having the individual pieces of equipment characterized.

In general, more degradation to stereo separation is seen in equipment that processes the audio signal or carrier than in broadband equipment that simply passes it. This is why a modulator might be expected to cause more degradation than a distribution system. An RF set-top converter should have better separation than a baseband converter, transportation equipment such as FM links or fiber optic links can degrade audio separation if BTSC stereo is passed through as a subcarrier on the video. It is also possible to degrade stereo separation with some scrambling systems.

Because of the many different ways that signals are processed in modern cable television equipment there is no substitute for testing each make and model of gear to be used in a system that must pass stereo.

Cascade Uncertainty: In predicting the effect of cascading two pieces of equipment, some uncertainty is incurred. This uncertainty is illustrated in Figure 15-24, where we consider cascading one “good” box with one having “worse” separation. Let the horizontal axis represent the separation capability of the equipment with the worse separation. Let the curved lines represent the separation capability of the better equipment. The separation of the cascade will be equal to the separation capability of the poorer equipment, within a tolerance. The tolerance is seen in the following manner.

a. Locate the intersection of the top curved line for the “good” equipment and the vertical line for the “worse” equipment. Read the error axis at the left of this intersection. This is the tolerance in the positive direction.

b. Locate the intersection of the bottom curved line for the “good” equipment and the vertical line for 7 the “worse” equipment. Read the error axis at the left of this intersection. This is the tolerance in the negative direction.

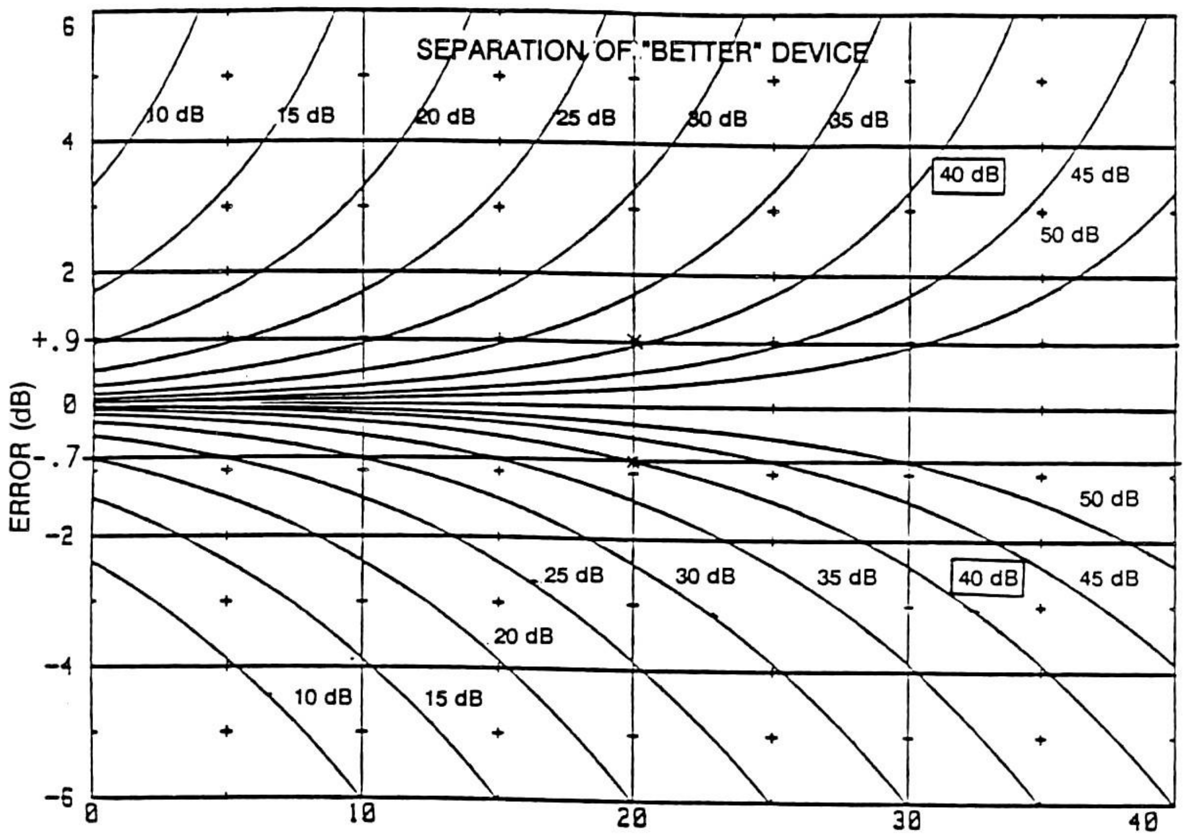


Figure 15-24: Separation of "Poorer" Device (dB)

For example, in Figure 15-24 are shown the results of cascading a 40 dB box with a 20 dB box. The cascade separation is approximately from 20- .7 to 20 + .9 or somewhere between 19.3 and 20.9 dB. We note that when cascading a "good" device with a significantly "worse" one, the separation of the cascade is very nearly that of the "worse" device.

The special case of cascading two pieces of equipment with equal separation is interesting. Note that the lower curved lines intersect the vertical lines of equal separation at -6 dB. Therefore, in the worst case the separation in such a cascade is degraded by 6 dB. The upper curved lines, though not obvious from this plot, actually approach infinity at the vertical lines of equal separation. This means that it is theoretically possible to get infinite separation from cascading two boxes of equal separation. It is unlikely for this to happen over a very wide range of frequencies, but can often be seen in a swept measurement as a "notch" in a separation plot.

Generalized Error Coefficient: Stereo separation in dB is a logarithm, and as such is not very useful in computing a cascade budget. For this reason the budget is done using a quantity derived from the measured stereo separation by taking the inverse logarithm and scaling to get a manageable number. This quantity is known as the generalized error coefficient (E_0). The term "generalized" refers to the fact that this coefficient includes the effect of any factor that can degrade stereo separation, since it is calculated from the measured separation without regard to how the degradation was caused.

Once the generalized error coefficient for each piece of equipment is known, it is possible to combine these to arrive at an expected error coefficient for the cascade. The cascade coefficient can then be reconverted to a separation number.

The individual error coefficients can be combined in various ways. As noted in the discussion of Figure 15-24, there is some uncertainty in calculating cascade separation. One approach is to calculate the worst case cascade combinations and budget for this number. This involves simply adding all of the error coefficients and reconverting the sum to dB separation. This is a conservative approach. Barring large measurement errors, this method will assure good results. The performance objective of 24 dB listed above is based on a worst case calculation.

A less conservative approach is to combine the error coefficients using a Root-Sum-of-Squares (RSS). This method entails squaring all of the coefficients, adding the squares, and taking the square root of the sum. This approach may be more realistic for a large cascade where none of the entries in the table have been combined as discussed previously.

Note that the stereo separation of a piece of equipment varies with frequency. Strictly speaking, this analysis applies to separation at one frequency, and should be repeated for all frequencies of interest. This could be done with a computer controlled measurement system that is also programmed to calculate the cascade results. Short of this, the separation of each piece of equipment over a wide range of frequencies should be measured. A frequency should be chosen at which the separation of all equipment is well-behaved and typical of the equipment. A mid-band frequency should be chosen as opposed to one at the low end, where separation is not as important, or the high end, where separation is more difficult to achieve.

For a further discussion on the theory of stereo separation budgeting, the reader is referred to the articles listed in the references at the end of this section.

Procedure: Measure the separation degradation of the individual pieces or subgroup of equipment to be used.

Calculate the generalized error coefficient for each using the formula:

$$E_{10} = 100 \times 10^{(-S/20)}$$

where S is separation in dB

Create a table showing each piece or subgroup of equipment, its stereo separation S(dB), its generalized error coefficient E_0^2(for use in a RSS calculation).

A sample budget table is shown below:

Equipment	% Contrib.	E_0	S(dB)	E_0^2	
Stereo encoder	22	3.11	30	9.66	
Modulator-demodulator or signal processor	8	1.13	39	1.28	In practice may be
Scrambler-SIT	7	0.99	40	0.98	measured as a subgroup
Distribution	4	0.57	45	0.32	
Consumer decoder	55	7.77	22	60.35	
System total	100	14.14		72.91	
			SQR	(72.91)	= 8.54

To get the worst case system separation, add up all of the entries in the E_0 column, and use the formula below:

$$S(dB) = -20 \times \log(E_0/100)$$

where E_0 is now the total of the column

For the system in the example, the worst case separation is:

$$-20 \times \log(14.14/100) = 17 \text{ dB}$$

To get the RSS system separation, add up all of the entries in the E_0^2 column, take the square root of the sum, and then use that number in the step 4 equation to get the "typical" system separation.

In the example system, the typical system separation would be

$$-20 \times \log(8.54/100) = 21 \text{ dB}$$

A column showing the percent contribution for each piece of equipment is also useful in visualizing the budget distribution. This number is simply the percent of the total E_0 that each piece of equipment contributes. Such a column is also shown in the example budget table.

References:

1. Gibson, JJ., "Accumulation of Stereo Separation Errors in Cascades of Subsystems Conveying L +R and L-R Over Different Paths'; publication of RCA Laboratories.
4. Rovira L.A., "Budgeting Stereo Separation In A CATV System'; IEEE Transactions on Consumer Electronics, Vol. CE-33, No.3, August 1987

PART III Additional Material

Chapter 16 Tutorials

16.1 Carrier-to-Noise Measurement through a Set-Top Converter

This tutorial is intended for those undertaking to determine the carrier-to-noise ratio delivered to subscribers through a set-top converter, based on computation of the carrier-to-noise ratio at the input to the converter, and the noise figure of the converter, as determined from the manufacturer or other source of authority. It is intended to be used with Section 3.3: "Measuring Noise of Systems using Converters".

The carrier-to-noise ratio (CNR) measured at the output of a set-top converter (or any other device for that matter) is a complex function of the incoming carrier-to-noise ratio and the noise figure of the device. Noise figure is a quantity which tells us how much noise is added over and above the thermal noise. Thermal noise is the theoretical minimum noise always present in any real system, regardless of how good we make it. The lower the noise figure, the less the carrier-to-noise ratio will be affected as the signal passes through the device.

The noise figure may be measured using rather specialized equipment and techniques not normally available in a cable system. It is possible to measure noise figure using a very clean signal source which has no noise over and above thermal noise, and measuring the carrier-to-noise out of the converter. Because the carrier-to-noise ratio will be so high, it will be especially difficult to measure accurately, so we don't normally recommend trying this with field equipment. Rather, it is better to rely on the manufacturer to characterize the converters for you.

When converters were added to to Part 15 of the FCC rules in the late 1980s, one of the requirements was to limit the output level from converters to a maximum of +15.56 dBmV for any input level up to +25 dBmV. This forced manufacturers to put delayed AGC circuits in the front end of RF set-top converters. At the same time, it was becoming the expectation in the industry that the converters should contribute less noise, so pre-amplifiers were added to reduce noise figure. The same architecture was used for baseband as well as RF converters (and was in some baseband units before the new rules went into effect).[1]

All of this complicates the way we would add noise generated in a set-top converter to the incoming noise. Let's consider first the simpler case, that of conventional (pre Part 15) converters without preamplifiers or AGC. Noise generated in a converter adds in a rather complex way to the noise coming in. We are interested only in the ratio of signal-to-noise, not in the absolute noise level. Figure 16-1 plots some examples of the way in which the noise adds. Along the X axis is plotted the signal level into the converter, and along the Y axis is plotted the carrier-to-noise ratio out of the converter. The three curves apply to carrier-to-noise ratios of 43, 45 and 47 dB on the cable, just before the converter. We have assumed that the converter has a noise figure of 13 dB.

[1] For a discussion of the architecture of old and new converters, see Farmer, J.O. and Cook, A.M., "The New Part 15 Set-top Requirements," CED Magazine, June 1990.

A common mistake is to assume that this means the output carrier-to-noise ratio is 13 dB higher than the input carrier-to-noise ratio. This is only true if the sole noise content in the input signal is thermal noise, and such is not the case by the time the signal gets to the home.

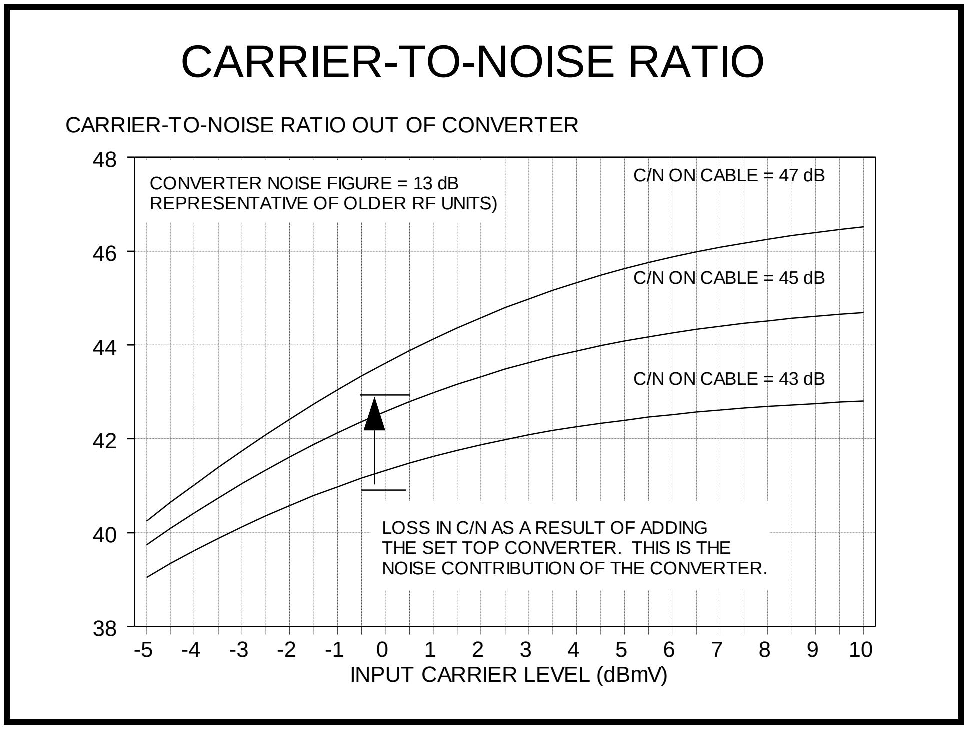


Figure 16-1: Set-Top CNR as a Function of Signal Level and Cable CNR, No Delayed AGC

Figure 16-1 shows the situation before the Part 15 rules were adopted. If we start with a carrier-to-noise ratio (CNR) coming in, of 43 dB (lowest curve), the ultimate (for now) FCC requirement, and have an input level of 0 dBmV, we see that the signal-to-noise ratio out of the converter has been degraded to about 41.3 dB. If we increase the carrier level into the box, the CNR asymptotically approaches the incoming 43 dB, but never quite gets there.

Note: Also the CNR is degraded more if the CNR coming in is higher. This is because the noise contributed by the converter makes more of a difference if there is less noise coming in. For an input of 0 dBmV and a carrier-to-noise ratio of 47 dB, the degradation is from 47 dB to about 43.6 dB, a loss of 3.4 dB. On the other hand, if the cable CNR is only 43 dB, the loss is from 43 to 41.3 dB, a loss of only 1.7 dB.

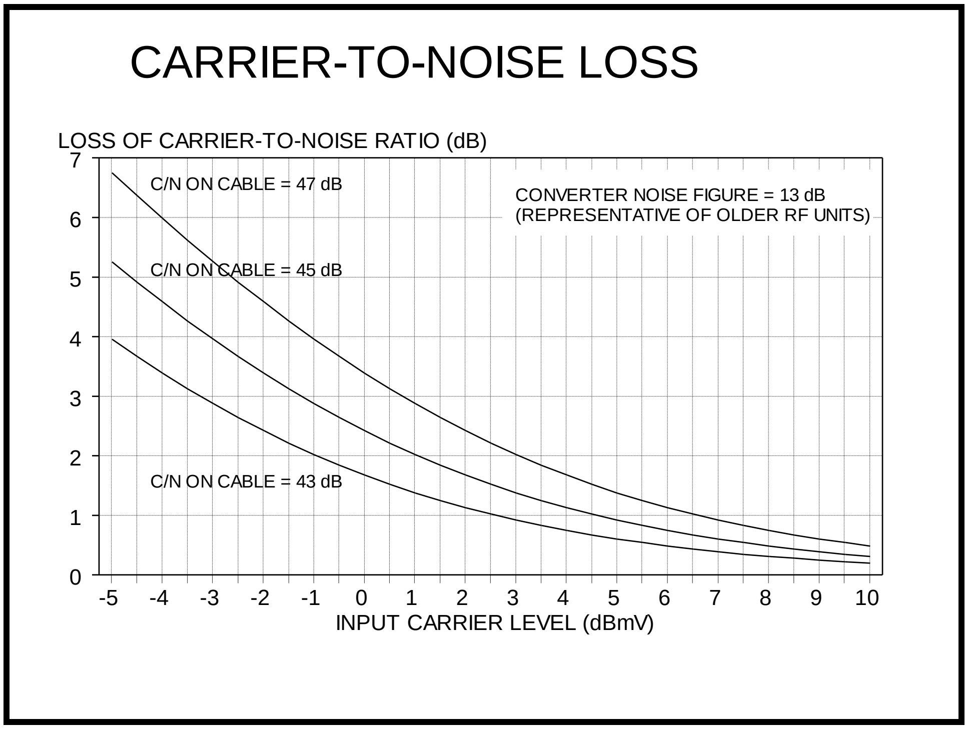


Figure 16-2: Loss of CNR Computed in Figure 16-1

We can plot this loss in CNR just as we plotted the CNR itself. Figure 16-2 is a plot for the same cases as we studied in Figure 16-1. Here we can see more easily that if the incoming CNR is 47 dB and the signal level is 0 dBmV, then the loss is 3.4 dB. If the incoming CNR is 43 dB at the same signal level, the loss is only 1.7 dB. From this chart you see that the reduction of CNR is a function of both the incoming signal level and the incoming CNR.

One could use Figure 16-2 with conventional set-top converters to compute the effect on CNR of adding the converter. However, it would be somewhat difficult to publish a set of curves for every case of noise figure, input level and CNR. In order to make the curves somewhat easier to use, the curve of Figure 3-3 is used. It is more versatile than is the Figure 16-2 of this part, because its X axis combines two things into one. In using Figure 3-3 for set-top converters, start with the measured input signal level and subtract the CNR measured on the cable. This is the number on the X axis. Go up from this to the curve corresponding to the noise figure of the converter, and go across to the Y axis, which is the same as the Y axis of Figure 16-2, the loss in CNR as a result of adding the converter.

The same procedure may be used for modern converters having delayed AGC, except that one must be aware of the possibility that the input signal level is above the delayed AGC threshold. In this case, the effective noise figure of the converter will be higher than assumed, resulting in poorer CNR out than would be expected.

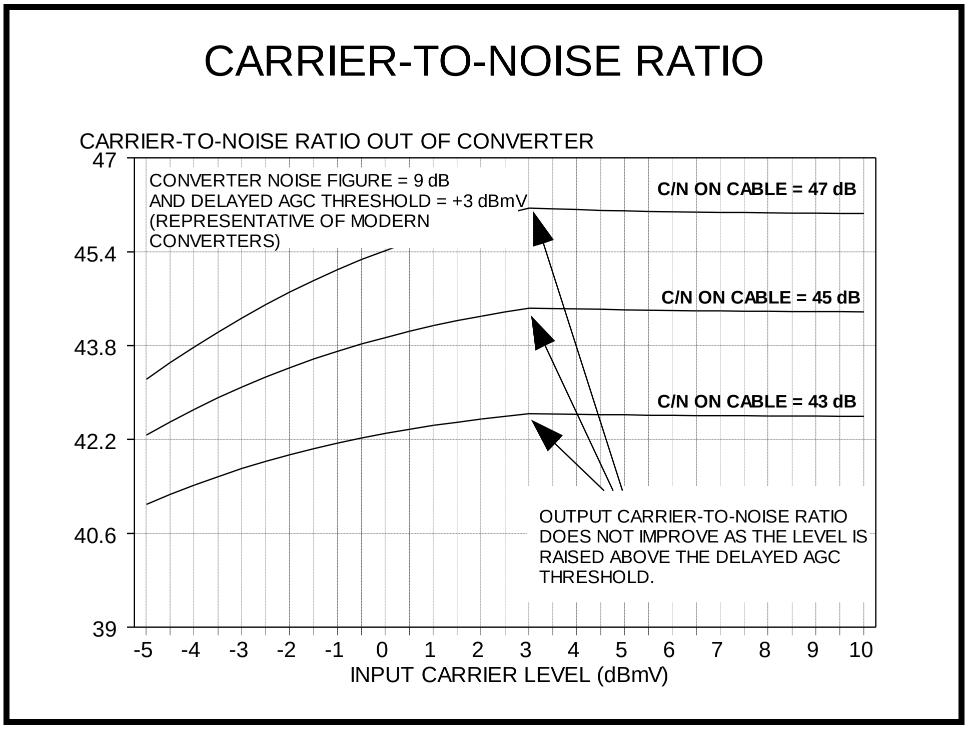


Figure 16-3: Set-Top CNR as a Function of Signal Level and Cable CNR, Delayed AGC Threshold +3 dBmV (Compare to Figure 16-1)

Figure 16-3 is the same as Figure 16-1 except that we have assumed typical characteristics for modern converters. We have assumed a 9 dB noise figure and a delayed AGC threshold of +3 dBmV. Though these are typical values, you should not use them in calculation: the only proper way to know the characteristics of the converters you are using is to either measure with equipment not normally available in the field, or to talk to the manufacturer. Note that above the AGC threshold, there is no increase in CNR as the signal level is increased.

Figure 16-4 is the same as Figure 16-2 except for the assumption of a modern converter. Figure 3-3 may be used in this case, but the noise figure used must be the noise figure of the device below threshold, minus the number of dB by which the input signal level exceeds the threshold.

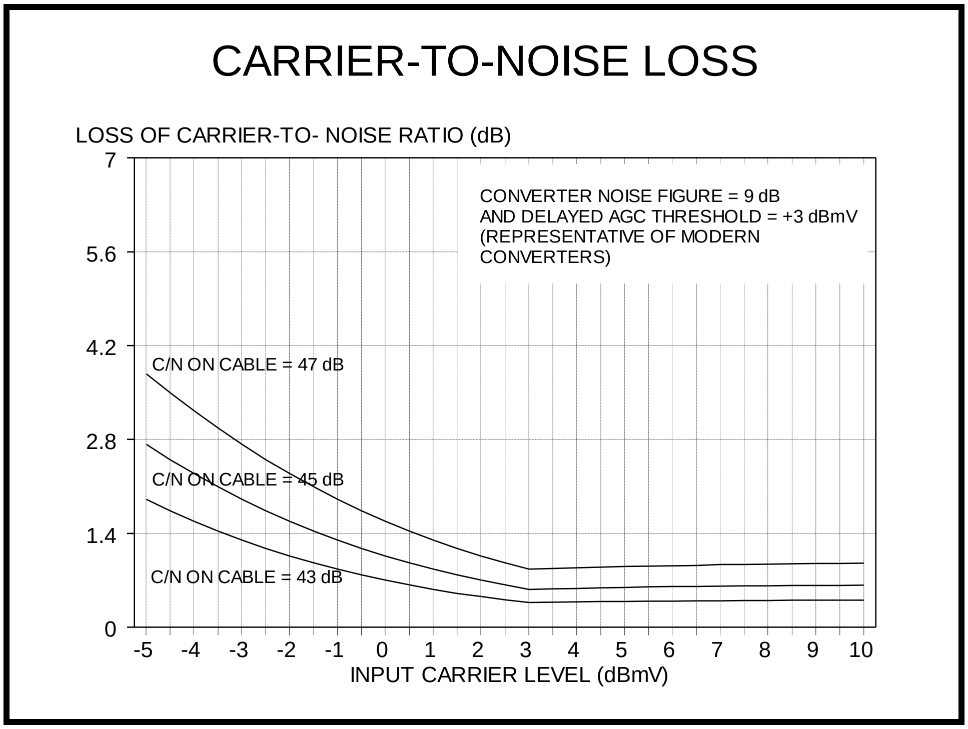


Figure 16-4: Loss of CNR Computed in Figure 16-3 (Compare to Figure 16-2)

16.2 Peak Voltage Addition

When multiple signals are combined in a single system the peak voltage of the combination is not simply the addition of the peaks of the separate signals. Furthermore the total average power obtained by adding the separate average powers is not an indicator of the composite peak.

A sinusoid (CW signal) has a well-known peak-to-rms ratio of $20*\log\left(\sqrt{2}\right)=3$ dB. A composite signal made up of *n* carriers of equal level and randomly phased, has a total average power *n* times the power of one carrier, or 10*log(*n*) dB greater than one carrier. The peaks add on a voltage basis (because at some time they will all add in phase), so the peak power is 10*log($2n^2$) dB greater than the average power of a single carrier.

In contrast to CW carriers, typical forward path signals, QPSK, VSB and QAM, have significantly higher peak-to-average ratios and most have broad spectra, filling most of their allotted channel. The forward path carries a multiplicity of such signals. Allocating individual signal levels proportional to channel width results in a composite signal with a nearly flat power spectral density and a high (about 6 dB) peak-to-average ratio similar to that of Gaussian noise.

A noise-like signal has a well-defined and measurable average power, but it doesn’t have a distinct peak value. There can be rare, but very high, signal excursions that stress the linearity capability of amplifiers and optical components. It is necessary to define a peak value for a noise-like signal in terms of a signal level exceeded with a certain probability (or fraction of the time). The probability or the fraction of the time should be based on the tolerance of the services to transmission impairments.

To associate some numbers with the above concepts, first consider a composite signal composed of 4 randomly phased CW carriers, each with average power, P_{av}. The average power of the composite signal is P_{av} + 6 dB. The peak power of the composite signal is P_{av} + 15 dB, and the peak-to-average ratio is 9 dB. Figure 16-5 illustrates how the peaks are related to the rms value of a sum of randomly phased CW carriers in terms of the probability that peaks exceed a given value. The infinite number of CW carriers case corresponds to the probability distribution for Gaussian noise.

One can see in Figure 16-5 that the sum of even a moderate number of sinusoids approaches the Gaussian noise probability distribution. The sum of digitally modulated carriers will approach the Gaussian noise probability distribution more rapidly as the number of such carriers increases. A Gaussian noise distribution exceeds its average by 13 dB with a probability of 10^{-5}. In this case we would say that the peak-to-average ratio for this Gaussian noise is 13 dB with the understanding that this ratio is exceeded very rarely, i.e. 10^{-5} (.001%) of the time.

From the above example it can be concluded that 4 CW carrier characterization does not adequately exercise the peak signal capacity of amplifiers or optical components which are intended to carry a composite power noise-like signal. This is due to the fact that, for the same total power, the noise-like signal has peak signal excursions about 6 dB higher than those of 4 CW carriers. In addition, one should not take a 1, 2, or 4 channel video specification and assume that the power sum of the specified video levels is a proper operating point for a composite noise-like signal.

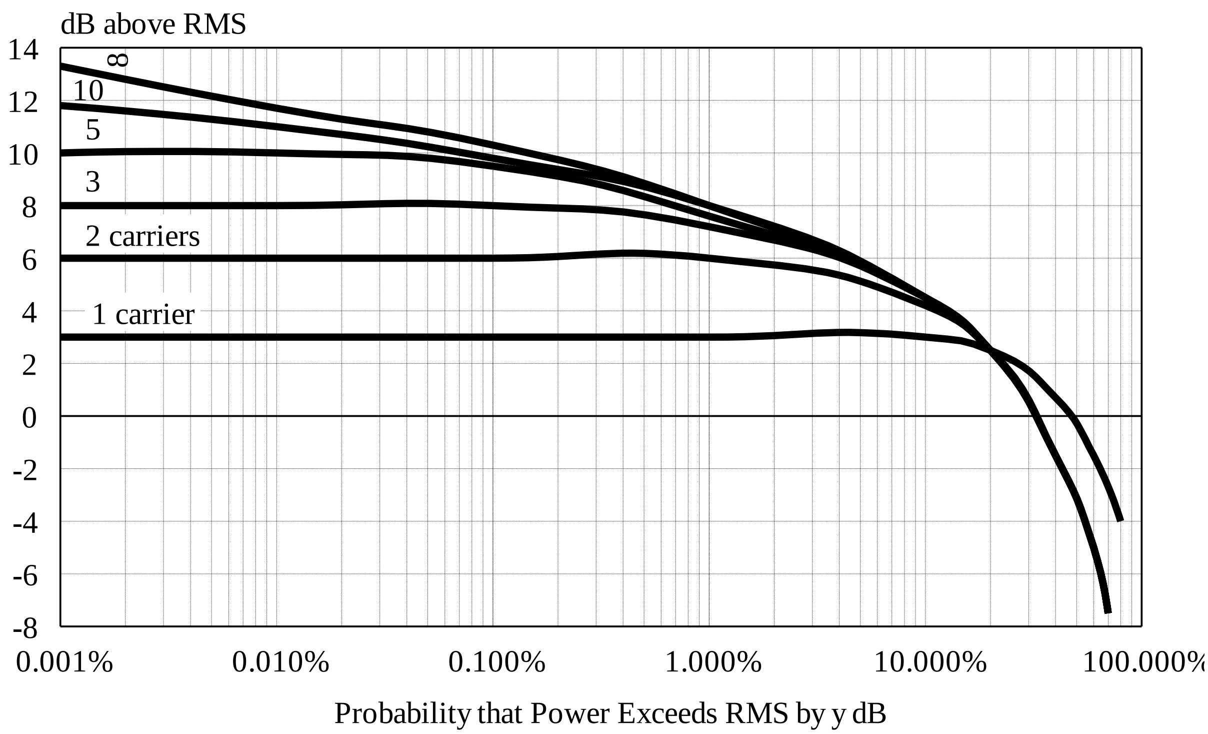


Figure 16-5: Peak Voltage Addition

Notes, Hints, and Precautions

A composite signal comprised of several QPSK carriers will have a peak-to-average ratio less than that of Gaussian noise, or less than 15 dB at a probability of 10^{-8}. For field testing purposes, a test signal comprised of 10 independent, equal level, CW carriers has a peak-to-average ratio of 13 dB. Such a test signal would stress the peak signal handling capability of reverse path active transmission components to within 2 dB of actual operating conditions.

16.3 Adaptive Equalization

Overview

The received QAM signal is degraded by impairments which can occur in the transmitter, the signal path and the receiver itself. These signal impairments include linear and nonlinear distortions. Linear distortions modify the signal regardless of the signal's amplitude and phase and all symbols are equally distorted. Linear distortion is caused by a deviation from a flat amplitude response and a linear phase response. Examples of linear distortions include variations in amplitude or phase/group delay response and echoes or reflections (micro-reflections) from impedance mismatches.

Nonlinear distortions vary from symbol to symbol. Examples of nonlinear distortions include intermodulation distortion (second and third order distortions), time dependent impairments such as burst events that cause compression or clipping, and distortions embedded in the signal such as gain or phase errors between in-phase and quadrature components. Both linear and nonlinear impairments contribute to the degradation of the signal.

Adaptive equalization compensates for linear distortions, but cannot compensate for nonlinear distortions. The adaptive equalizer is a digital filter, which has a response approximately equal to the inverse of the channel's actual frequency response. It is called "adaptive" because it optimizes its filter characteristics continuously, maximizing signal quality in a continuously variable environment.

All practical QAM demodulators contain an adaptive equalizer.

Description

An adaptive equalizer is made up of a tapped delay line that creates delayed versions of the main signal and previous outputs of the equalizer that are summed with the original received signal. The delayed signals are multiplied by coefficients that are adaptively adjusted and then summed to minimize symbol errors caused by linear distortions, thus eliminating intersymbol interference in the demodulated QAM symbols. This may be expressed mathematically as follows. The frequency response, $H_R(f)$, of the received signal is:

$$H_R(f) = H(f) * H_C(f)$$

where:

$H(f)$ = desired response

$H_C(f)$ = channel response

In order to get back to the desired response, the equalizer response $H_E(f)$ must be equal to the inverse of the channel response. That is:

$$H_E(f) = H(f) / H_R(f) = 1 / H_C(f)$$

The block diagram in Figure 16-6: Adaptive Decision Feedback Equalizer (DFE) Block Diagram shows an adaptive equalizer that is a combination of the following three sections:

- A feedforward equalizer (FFE) (also known as an FIR – Finite Impulse Response filter shown in Figure 16-7: Finite Impulse Response (FIR) Filter)
- A feedback section (resulting in what is known as an IIR – Infinite Impulse Response filter shown in Figure 16-8: Infinite Impulse Response (IIR) Filter)

- An ideal transmitted symbol decision (decision feedback) and a coefficient update algorithm (such as Minimum Mean Squared Error) using the error difference between the actual and ideal symbol decisions.

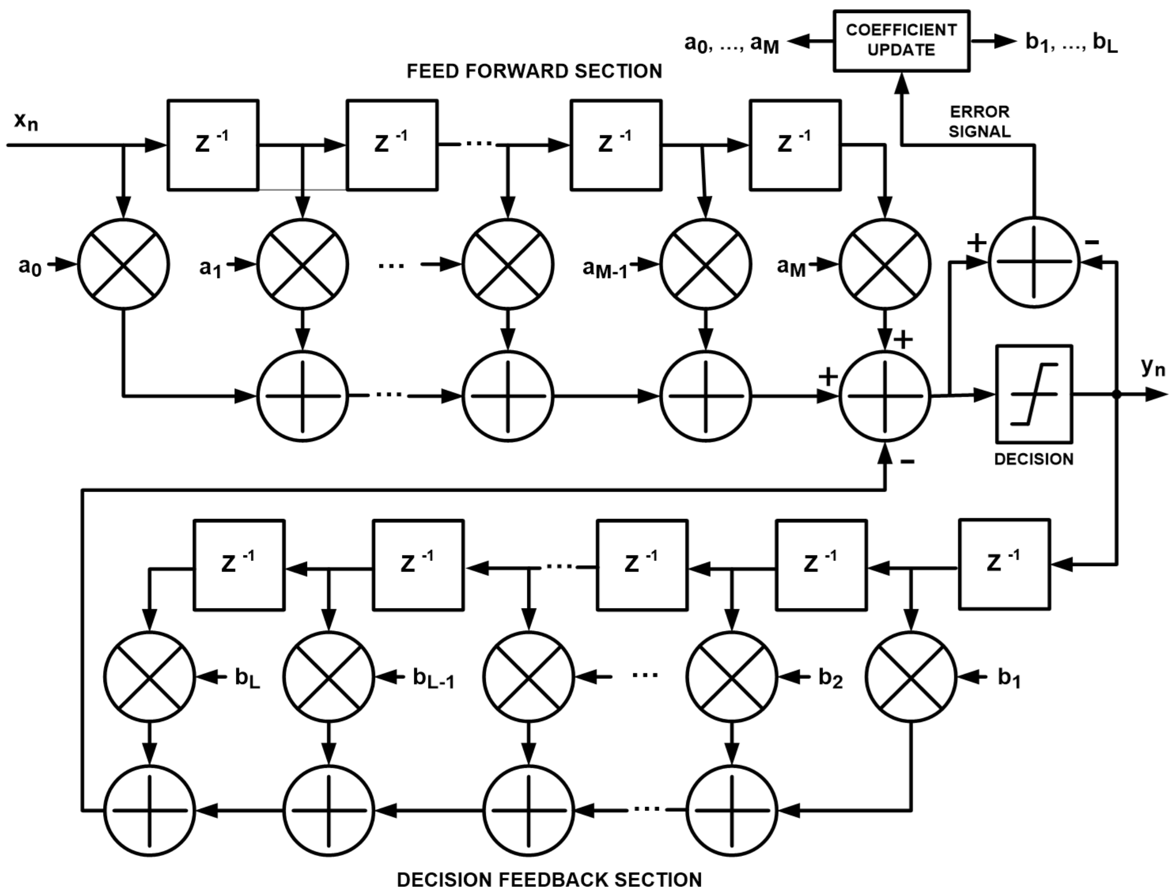


Figure 16-6: Adaptive Decision Feedback Equalizer (DFE) Block Diagram

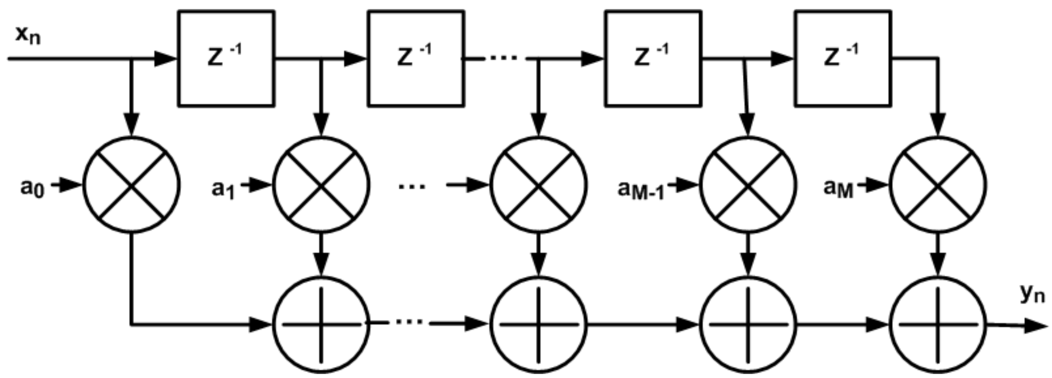


Figure 16-7: Finite Impulse Response (FIR) Filter

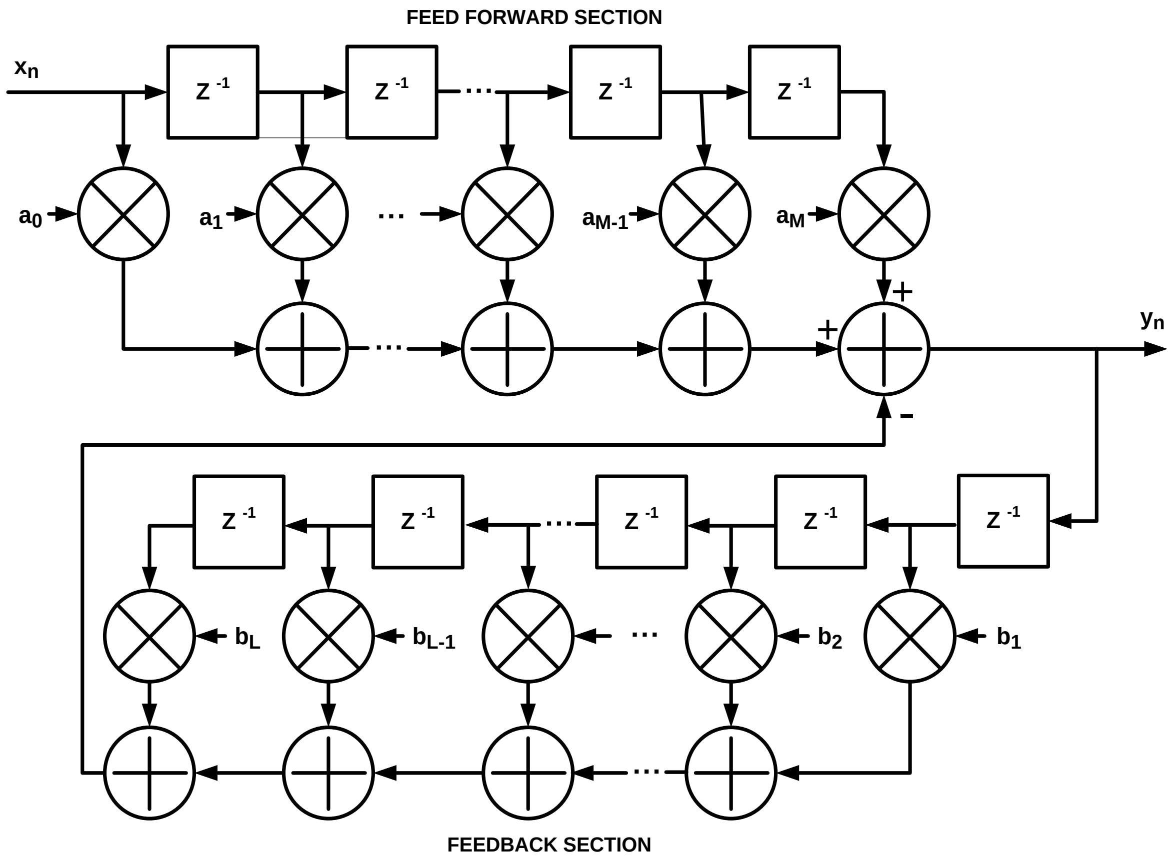


Figure 16-8: Infinite Impulse Response (IIR) Filter

Let's look in more detail at the operation of linear digital filters and the specific case of the adaptive equalizer with nonlinear decision feedback. The incoming QAM signal has been digitized and fed into a series of registers (or taps) at the symbol rate. Hence each tap contains a sample of the QAM signal for consecutive symbols. Taps following the main tap carry symbols received earlier, or past symbols relative to the main tap. Taps ahead of the main tap carry symbols received later, or future symbols relative to the main tap.

There are two types of generic digital filters made of a tapped delay line. To implement filtering, the signal value at each tap is multiplied by a coefficient (unique for each tap) and the sum of these weighted values generates the filtering effect. The two filters differ by the path taken by the weighted sum.

In the FIR filter shown in Figure 16-7: Finite Impulse Response (FIR) Filter, the weighted sum of delayed input signal samples is the output of the filter. It is called Finite Impulse Response because the output calculation, which is non-recursive, is directly dependent on the current and delayed input signal samples only. Therefore, removing the input signal will result in a finite duration non-zero output signal. The filter coefficients, a_0, a_1, …a_M correspond to the filter's impulse response and can also be referred to as the tap weights.

In the IIR filter shown in Figure 16-8: Infinite Impulse Response (IIR) Filter, an additional weighted sum is fed back to the input, and the output of the filter is the previously described FIR output minus the feedback (sum of weighted previous outputs). It is called Infinite Impulse Response because the output calculation, which is recursive, is dependent on a number of previous outputs. Thus, the current output depends on the current and delayed inputs and the recursive output feedback. Theoretically, due to the feedback loop, the output of the filter depends on inputs received recently and also some received much earlier than the time span of the filter. Therefore, removing the input signal will result in an infinite duration but diminishing output signal. For this IIR filter, there two sets of coefficients or tap weights, one set a_1, a_2, …a_M for the feed forward section at the top of the figure, and another set, b_1, b_2, …b_L for the feedback section of the filter at the bottom of the filter.

In a digital filter, if the coefficients are fixed, the response of the filter is fixed. But if the coefficients are modified, the response is also modified. This is what is done in an adaptive equalizer where the coefficients are adaptively modified to minimize linear distortions that result in intersymbol interference, the latter of which can cause errors in the received signal values and can also increase susceptibility to noise. Two types of adaptive equalizers are described below: Feed Forward and Decision Feedback.

FFE – Feed Forward Equalizer

The coefficients of an FFE are adaptive, the main tap coefficient is nominally the largest, and those of the others can vary considerably and are complex, that is, signed (±) and have in-phase (real) and quadrature (imaginary) components. The Main tap is the last one in an FFE. The FFE corrects for linear distortions such as pre-echoes and IF filter response. The FFE is essentially a time varying FIR filter. The limitation of this type of equalizer is the poor equalization performance in the presence of deep spectral nulls in the channel frequency response.

In ITU-T J.83 Annex B applications, the QAM signal is sampled at every symbol, then tap spacing equals symbol time spacing ~ 200 ns (1 / 5.056941 MS/s or 1 / 5.360537 MS/s) and an 8-tap equalizer compensates for echoes with a time delay up to about 1.5 µs (~3 µs for a 16-tap equalizer). This ability to compensate for echoes with specific maximum time delays is known as the span of the equalizer.

DFE - Decision Feedback Equalizer

This type of adaptive equalizer has two important differences when compared to an FFE. First, it uses an infinite impulse response (IIR) filter, where the sum of all weighted tap outputs is fed back to the main tap output such that the present filter output contains previously filtered outputs. In theory the filter response is (in part) dependant on the signal received an infinite time ago (hence the name infinite impulse response), but in practice the contribution of previous outputs diminishes quickly. The second difference is the "decision feedback" aspect because the feedback taps are not fed by the actual signal (as in an FFE) but by an idealized value decision. In other words, whatever value falls into the decision region is replaced by the ideal transmitted value. The nonlinearity of the DFE results from the nonlinear characteristic used in replacing the actual filtered values with the idealized transmitted values at the input to the feedback filter section. This results in the feedback filter section of the DFE operating on noiseless quantized levels whose output is free of channel noise. Thus the intersymbol interference contributed from these symbols can be cancelled out exactly at the output of the FFE by subtracting past ideal transmitted symbol values with appropriate weighting.

DFE output responses are much longer than FFE output responses. Long filter spans are required because echoes can have long time delays.

In both DFEs and FFEs, the tap coefficients are adaptively adjusted based on the results of an error reduction algorithm. Most adaptive equalizers used in QAM receivers adjust the tap values without using a known transmitted training sequence. This is known as blind equalization. The equalization process starts by setting all tap coefficients equal to zero with the exception of the main tap. As successive data symbols are input to the equalizer, the tap values are updated based on results of the error reduction algorithm. When the equalizer reaches convergence, the tap values can be used to characterize the micro-reflections on the system. (See Section 9.1 Micro-Reflection Overview, Section 9.3 Digital Adaptive Equalizer Impulse Response and also see the following section: Useful information from the Equalizer.)

Most adaptive equalizers are either T (i.e. – symbol period) spaced (also called synchronous designs) or T/2 spaced (also called fractionally spaced equalizers). Although, the fractionally spaced equalizer offers reduced sensitivity to sampling phase errors, it has a longer convergence time, is subject to coefficient drift and has higher complexity and cost.

Useful information from the Equalizer

The adaptive equalizer is a necessary component of the QAM demodulator. Useful information about the input signal can be extracted from the coefficients (or weighting factors) of the equalizer. As seen in section 9.3 Digital Adaptive Equalizer Impulse Response (and section 9.1 Micro-Reflection Overview), the timing position of a reflection can be visually approximated because “it sticks out” of the normal response. Distance and amplitude of the reflection can be estimated if we know the characteristics of the network (cable’s velocity of propagation, attenuation per unit length, termination locations, splitters, etc.) It is important to remember that the time resolution of an adaptive equalizer graph is the period of one symbol for a T-spaced equalizer, which for 256-QAM is 186.55 ns. Thus, in an 87% velocity of propagation cable, the resolution is limited to 48.68 m or 160 feet!

Other available information is the equalizer stress. We have seen that a fraction of future symbols (FFE) and past symbols (DFE) are added to the current symbol (reference or main tap) to cancel out linear distortions, hence the sum of all these fractions is equal (but opposite in sign) to the linear distortions within the QAM signal. We calculate the equalizer stress by summing up all of these complex fractions on a power basis. Let’s look at the following example in Table 16-1 with 2 FFE taps, the Main tap and 3 DFE taps with all coefficients normalized to the Main tap.

Table 16-1: Example Equalizer Stress Calculation

EQUALIZER STRESS				
Tap	In-phase coefficients	Quadrature coefficients	Amplitude of coefficient A = $\sqrt{I^2 + Q^2}$	Relative power = A^2
-2 FFE	+0.003 4	-0.006 9	0.008 32	0.000 0691 7
-1 FFE	-0.006 2	+0.007 8	0.009 96	0.000 0992 8
Ref.	1	0	1	1
+1DFE	+0.007 5	-0.008 8	0.011 56	0.000 133 69
+2 DFE	-0.002 8	+0.003 1	0.004 18	0.000 017 45
+3 DFE	+0.000 11	+0.000 086	0.000 138	0.000 000 019
Equalizer Stress = Sum of relative powers excluding Main tap:				0.000 319 6

Now let's see how to calculate the un-equalized MER from the values of equalized MER and the equalizer stress. The equalizer stress is equal to the ratio of the power of the ***linear*** distortions to the signal level. On the other hand, the equalized MER (i.e., the MER calculated after the Adaptive Equalizer) is essentially the ratio of the signal power to the power of nonlinear distortions, so we invert this (flipping the sign in dB) to get the ratio of the power of ***nonlinear*** distortions to the signal level. After converting to non-logarithmic values, the two distortion ratios with common denominator of signal power can be added, converted back into dB form, and then re-inverted to yield the un-equalized MER. An example of this process is shown in the table below.

Table 16-2: Example Un-Equalized Modulation Error Ratio Calculation

UN-EQUALIZED MER		
Equalizer Stress	-34.95 dB	0.000 319 6 relative power (linear distortions)
Equalized MER	37.8 dB	
Inverse EQ-MER	-37.8 dB	0.000 165 9 relative power (nonlinear distortions)
Total relative power		0.000 485 5 relative power (all distortions)
		-33.14 dB
Un-Equalized MER		**33.14 dB**

As shown in the Tutorial in Section 16.14 the equalizer is a filter and the tap coefficients determine the impulse response of this filter in the time domain. Also, one can convert from the time domain impulse response to the frequency domain amplitude and phase responses versus frequency of this filter. The phase response can be translated into group delay variation response by calculating the derivative of the phase variation versus frequency.

This conversion from time to frequency domain uses the Discrete Fourier Transform (DFT), but one needs to know exactly the transfer function of the filter. This information may not be readily available and each QAM demodulator has its own equivalent equalizer implementation of Figure 16-6: Adaptive Decision Feedback Equalizer (DFE) Block Diagram. Most test instruments with QAM demodulation incorporate the appropriate conversion and display amplitude response and either phase or group delay responses.

If the test instrument provides the phase response $\varphi(\omega)$, the corresponding group delay variation GD can be calculated as:

$$GD = -\frac{d\varphi(\omega)}{d\omega} = -\frac{1}{2\pi}\frac{d\varphi(f)}{df}$$

where ω is the frequency in radians/second, which is equal to $2\pi f$ where f is the frequency in Hz.

When transforming from phase to group delay, the frequency separation df (also known as "aperture") is very important. If the aperture is too narrow, then the group delay varies widely and we see too much detail and lose the major trend. If the aperture is too wide we lose details of the group delay response. A reasonable (experimental) trade off is 8% to 12 % of the bandwidth.

To illustrate aperture, let's take an example. One drives from New York to Los Angeles, recording GPS coordinates along the trip. He wishes to plot a graph of his heading. A huge aperture would be the heading calculated between NY and LA, but it gives no indication on the actual trip. Going to the other extreme, if the aperture is reduced to ½ a mile, there will be headings in all directions as the road twists around corn fields and mountains and goes through interchanges. There are too many details (noisy) and there is no way to reconstruct the itinerary.

A heading from NY to Chicago, then Chicago to Denver and finally Denver to LA gives more information. The aperture can be reduced to show more intermediate cities and points of interest. With a reasonable size for the aperture, we put the headings on a map and figure out the itinerary.

It is important to remember there are several limitations to these responses.

1) The bandwidth is equal to the reciprocal of the symbol rate because the time steps are exactly that of the symbol rate. It does not cover the full channel bandwidth.

Table 16-3: Symbol Rate versus Channel Bandwidth

Symbol Rate versus Channel Bandwidth in ITU-T J.83 Standard				
Standard	**Annex B, 64-QAM**	**Annex B, 256-QAM**	**Annex A**	**Annex C**
Symbol rate	5.056941MS/s	5.360537MS/s	6.952 MS/s	5.217MS/s
Time Step	197.748 ns	186.5485 ns	143.8435 ns	191.681 ns
Resp. BW	5.056941MHz	5.360537MHz	6.952 MHz	5.217MHz
Channel BW	6 MHz	6 MHz	8 MHz	6 MHz
% coverage	84.28 %	89.34 %	86.9 %	86.95 %

2) The resolution of the frequency response is the bandwidth divided by the number of samples used in the DFT which may be much larger than the number of taps of the filter (see PART IIChapter 9). So increasing the number of time domain samples (sampling period equal to the tap spacing) in the DFT provides as much resolution as desired at the expense of an increased number of computational operations.

Consider the following example of the calculation of the channel amplitude and group delay response from the equalizer filter coefficients. The channel combines the main (desired) signal with four micro-reflections given in the following table:

Table 16-4: Channel Amplitude and Group Delay for Main Signal and Reflections

Signal:	Main		Micro-Reflections		
Amplitude (dB):	0	-10	-15	-20	-30
Delay (μs)	0	0.5	1.0	1.5	4.5

A spectrum analyzer trace of the 256-QAM signal transmitted through the channel distorted by these micro-reflections is shown in the following figure:

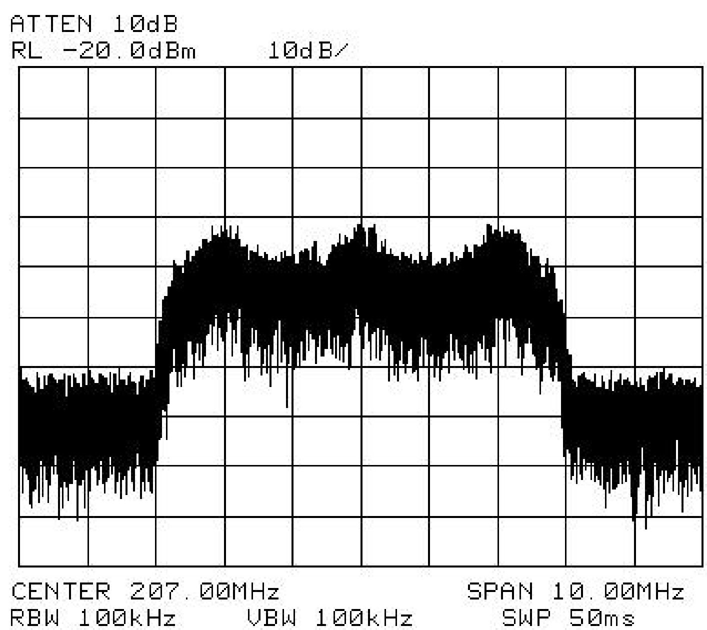


Figure 16-9: Spectrum Analyzer Trace of Distorted 256-QAM Signal

The receiver equalizer coefficients (normalized for unity DC gain) for this distorted channel are shown in the following table:

Table 16-5: Receiver Equalizer Coefficients for Distorted Channel

FFE Coefficients Position:	0	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20
Real:	-25	197	20	-25	374	-202	330	-335	728	-1088	1437	-1752	2455	-3125	4360	-5517	8169	-12027	22474	-60967	544248
Imaginary:	134	45	134	-43	134	1	178	-88	311	-176	488	488	665	-752	1241	-1284	1905	-2479	3721	-5978	1
DFE Coefficients Position:	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21
Real:	-25651	27231	45212	-20203	33431	6547	-2886	18948	-1823	613	-362	258	-317	170	-185	125	-185	170	-185	170	-229
Imaginary:	-1020	1106	-887	442	-134	-90	43	43	-134	87	-90	-90	-46	-1	-46	-1	-46	-1	-46	-1	-90
DFE Coefficients Position:	22	23	24	25	26	27	28	29	30	31	32	33	34	35	36						
Real:	303	-583	5573	170	-185	170	-140	125	-52	170	-52	37	37	37	-7						
Imaginary:	-1	43	43	-1	43	-46	-1	-1	43	43	-46	-46	-46	-1	-1						

The resultant calculation (see the Tutorial in Section 16.14 below) of the amplitude responses of the equalizer, the unequalized channel (inverse of the equalizer) and the equalized channel (equalizer * unequalized channel) are shown below:

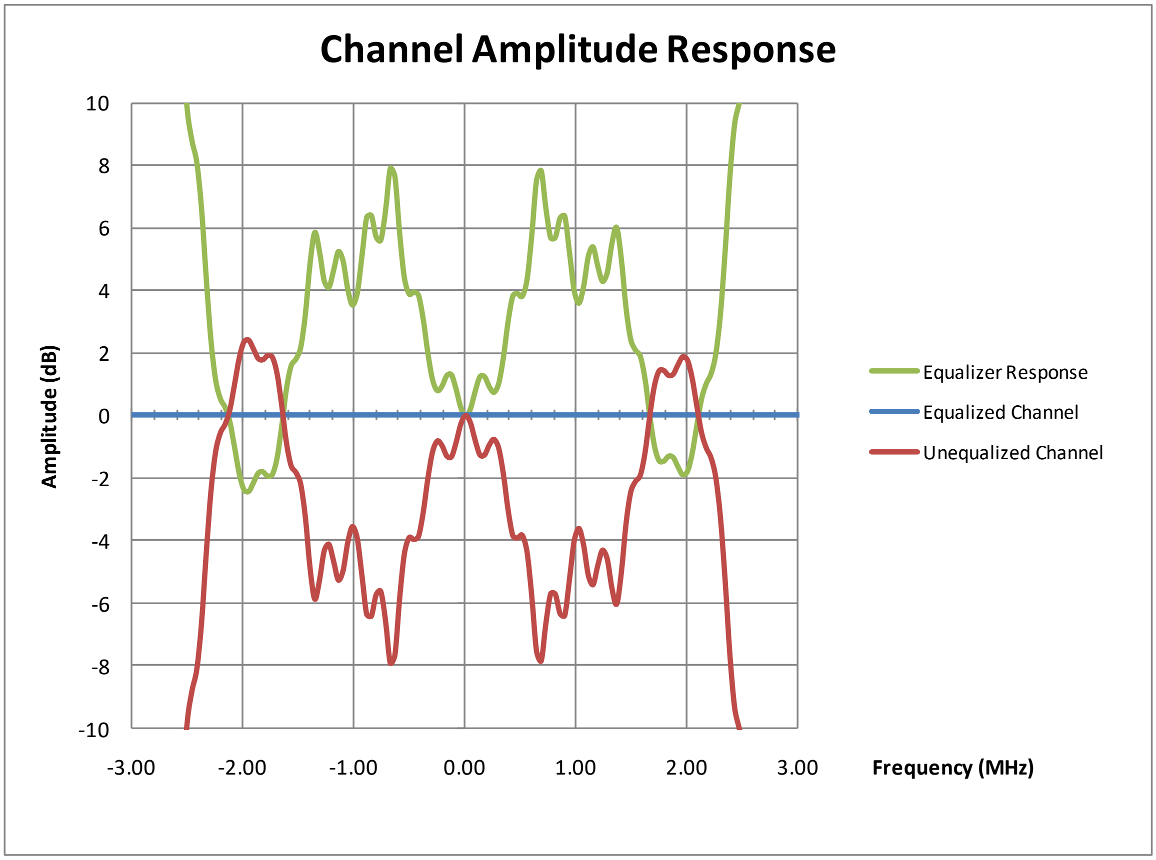


Figure 16-10: Spectrum of Unequalized Channel, Equalizer Response and Resulting Equalized Channel

The resultant calculation of the group delay of the unequalized channel is shown below:

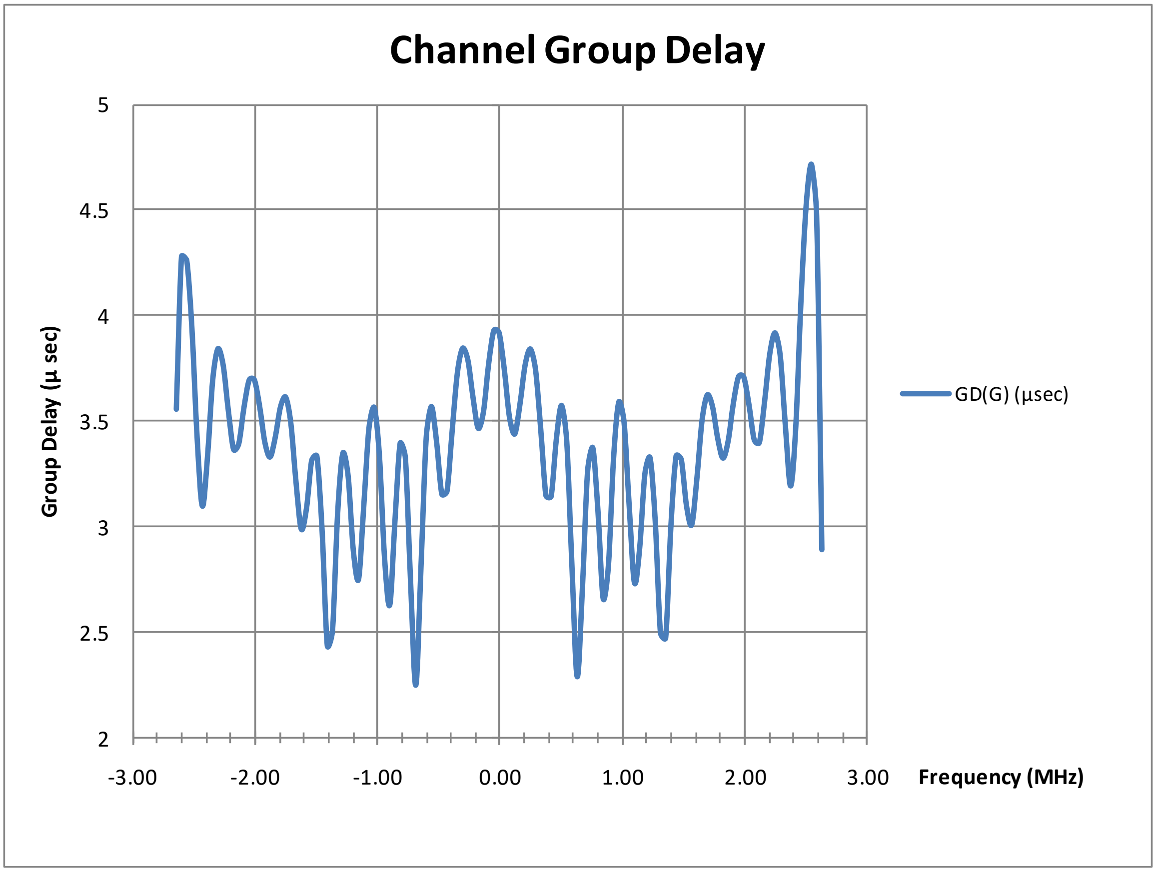


Figure 16-11: Group Delay Calculation of Unequalized Channel

The group delay response that has been low pass filtered (with a one-quarter Nyquist band filter) to reduce the aperture as previously discussed is shown below:

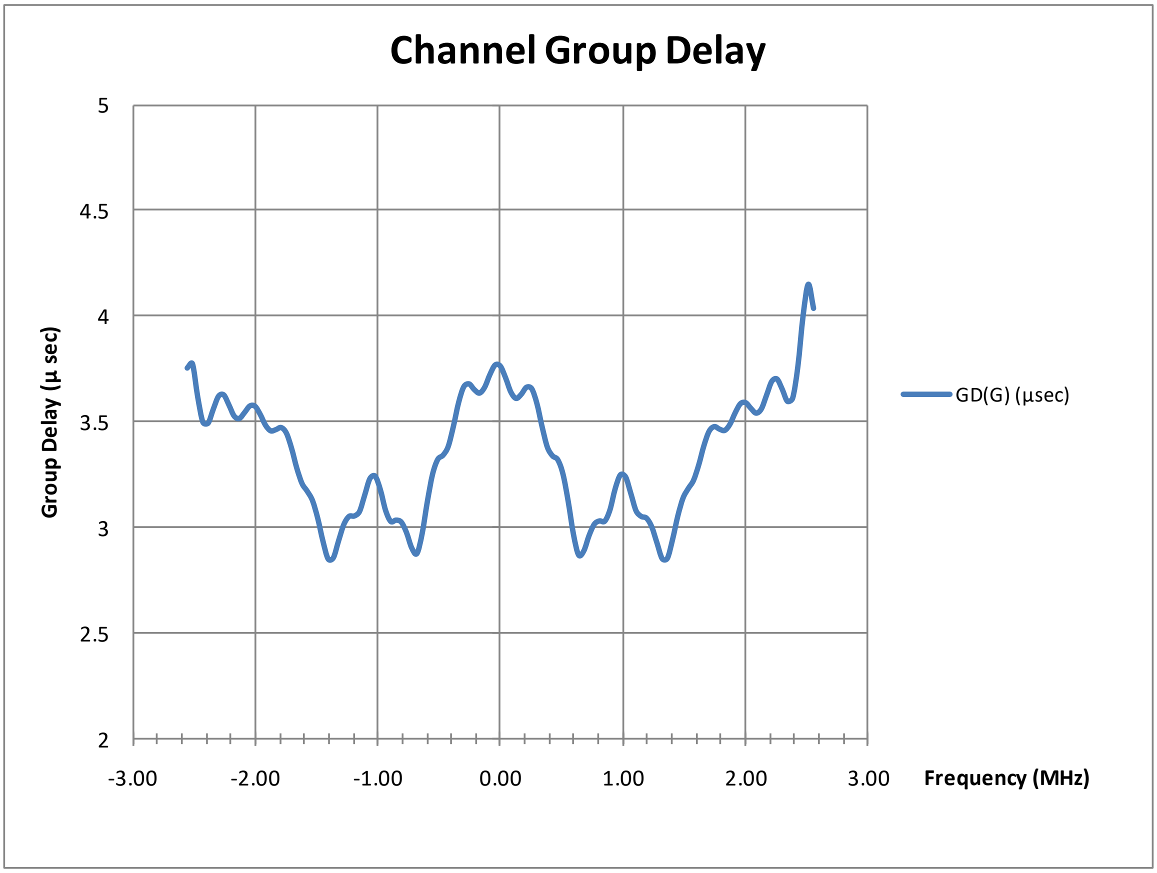


Figure 16-12: Group Delay Response with Reduced Aperture

16.4 Return Plant Setup and Operational Practices

Introduction

Although the recommended practices included in this document only refer to measurements, a prerequisite to meaningful measurements is a properly aligned plant. Therefore, the following setup and operational procedures are included as guidelines for operators. No specific operating parameters are suggested, but rather an organized process for setting those parameters.

The upstream plant differs in several critical ways from the downstream situation, even though they share the same physical distribution network. In the downstream direction, for instance, signals are usually present continuously and at well-defined power levels. The channelization scheme, bandwidth and modulation are also well-defined. Finally, each amplifier has but a single input.

In the reverse direction, by contrast, signals may be of a wide variety of bandwidths and modulation types and may be carried only intermittently and at varying power levels. Operating frequencies may vary from time to time, also. Finally, due to network splitting, each amplifier may receive signals from several inputs. All of these variable factors require that the procedures used to operate the upstream plant differ from those used in the downstream direction.

Another difference is that, while amplifiers in the downstream direction are aligned by adjusting their output signals to predetermined levels, in the upstream direction the plant is adjusted so that input signals are equal. The operator must decide whether to use the actual upstream amplifier module inputs or the external amplifier ports as the **Reference Points** for upstream alignment and setting operating levels. Figure 16-13 is a diagram of a typical trunk/bridger station showing the logical locations of each of the possible Reference Point choices. Each has its advantages and limitations:

Amplifier Upstream Module Input Reference Point

Many amplifier manufacturers use the same internal amplifier modules for all amplifier stations, regardless of the number of downstream output ports. Thus, all modules have similar noise figures, gains and signal handling capability. It can be shown that, when such stations are cascaded as they are in bridging node configurations, setting amplifier module input levels equal results in the minimum noise addition and distortion buildup.

Using this procedure, however, requires that different test signal insertion levels be used for different amplifier configurations, depending on how the upstream signals are combined internally. Also, in cases where, for instance, trunk bridger ports have a high internal combining loss in the upstream direction, higher upstream transmitter power levels will be required to adequately drive return amplifier modules. This may make drop configurations in some homes (especially those fed from high value taps near the amplifier output) more difficult, as the upstream losses must be minimized.

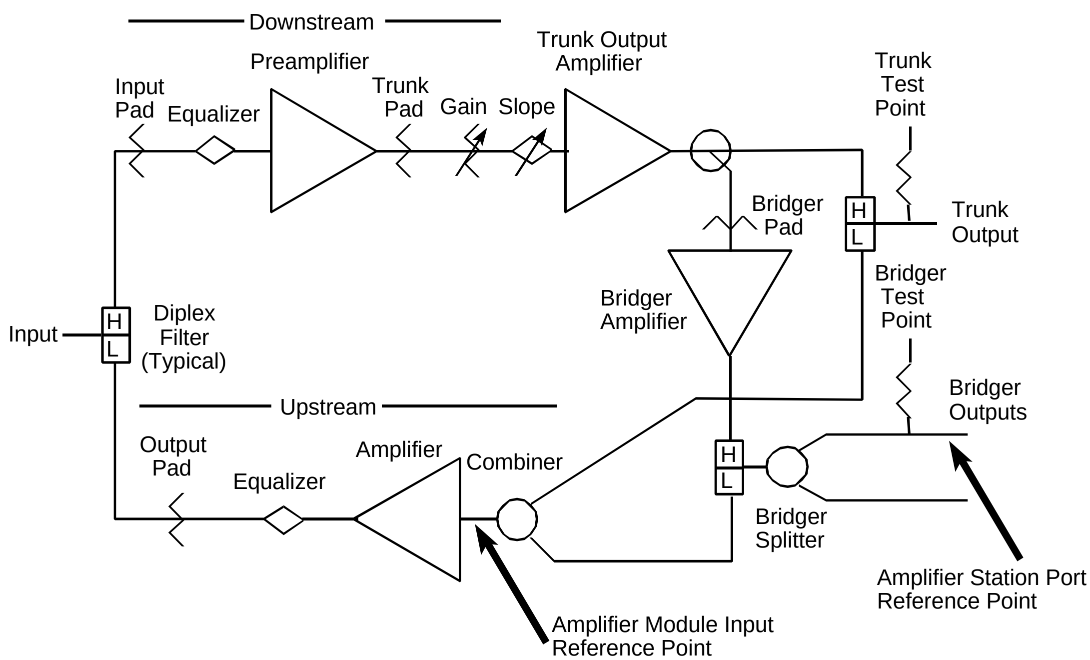


Figure 16-13: Typical Amplifier Station Showing Reference Point Choices

Amplifier Station Port Reference Point

Using one of the external amplifier output (upstream input) ports as the Reference Point has the advantage of simplicity in that test insertion levels are the same throughout the plant (assuming identical test port attenuation values), and also the advantage that upstream transmitter levels will not have to be higher in tap strings which are fed by multi-output amplifiers.

A major disadvantage to this scheme is that some upstream modules will have levels which are lower than others. For instance, a four-output system amplifier will generally have about 7 dB of additional combining loss between external ports and the upstream amplifier module. Thus, the noise contribution of this station will be five times as high as it would be if it were being driven at the "full" input level. The impact of this additional noise is, of course, dependent on the percentage of amplifiers which have multiple downstream outputs and the amount of excess loss in each amplifier.

Operators choosing to use an amplifier station port as the Reference Point should be aware of the fact that levels from all upstream input ports must be matched after internal combining in an amplifier station. This will result in external port levels which will differ by the difference in upstream combining losses.

Whichever method is chosen, operators should understand the compromises involved.

Steps to properly setup and adjust the upstream plant

Follow the manufacturer's recommendations for appropriate factory calibration requirements, warm-up time and field calibration before performing these tests. These steps should be taken immediately prior to the commencement of testing. None of the steps in properly aligning an upstream plant is unnecessarily complex, but all are essential:

1) First, each type of line equipment, including coaxial amplifiers and the fiber node equipment, must be characterized. Most critically, it is necessary to know the signal loss from the test point that will be used to inject upstream test signals to the Reference Point chosen in the last section - either input of the reverse amplifying module or the external amplifier port chosen as the reference. In the case of a simple line extender, this may be approximately equal to the test point loss (typically 20 or 30 dB) in either case, but a trunk amplifier or distribution amplifier will have additional losses to the module input due to combining of signals from several coaxial legs. Also amplifiers differ in test point losses, whether they are directional or non-directional, and the placement of test points in the circuit. Some provide separate upstream test points while others provide common test points.

   For operators choosing external port Reference Points, the input port with the highest upstream combining loss (say the bridger port or ports in a trunk/bridger station) is usually chosen as the Reference Point as this minimizes upstream transmitter level variation. In that case, the levels at ports with less combining loss (say the trunk port) will have to be reduced by the difference in combining loss. The result of this difference is that the levels of upstream signals in the trunk line will be lower than those in the distribution line. When the upstream alignment is performed at this station, either the normal test signal must be inserted at the bridger test port, or a level which is reduced by the difference in combining losses must be inserted into the trunk line test port to get the proper alignment levels in the plant.

2) Second, the operator must determine how much total input power will be allowed for all intentional signals in the return band. **This may be different for the upstream optical transmitter than for the upstream coaxial amplifiers**. Consult the manufacturers' data for this information.

   Some amplifier and optical transmitter manufacturers may specify a total power handling capability. In some cases, however, peak composite signal voltage, rather than total average power may limit levels. Refer to Section 16.2: "Peak Voltage Addition" for a detailed discussion on choosing maximum total power levels based on maximum signal voltages.

However it is determined, for future reference; call this total amplifier input power handling capability P_1. Note that P_1 may need to be adjusted as a function of the depth of the cascade.

See Table 16-6 **and** Figure 16-14 **for a summary of the power relationships and references used in this section.**

Table 16-6: Level References and their Relationship

System Location	Total Power Handling Capability	System Test Signal Level	Specific Upstream Service Signal level
Amplifier Input Reference Point	P_1	P_3	P_5
Upstream Optical Transmitter Input	P_2	$P_4 = P_3 - (P_1 - P_2)$	
Headend Optical Receiver Output	P_8	$P_9 = P_8 - (P_2 - P_4)$	$P_6 + A$
Headend Optical Receiver Test Point		P_0	$P_6 = P_0 - (P_3 - P_5)$
Specific Service Data Receiver Input			P_7
Headend Optical Receiver Test Point Attenuation: A = P9 - P0			
Required Gain from Headend Optical Receiver to Specific Service Data Receiver Input: B = P7 - (P6 + A)			

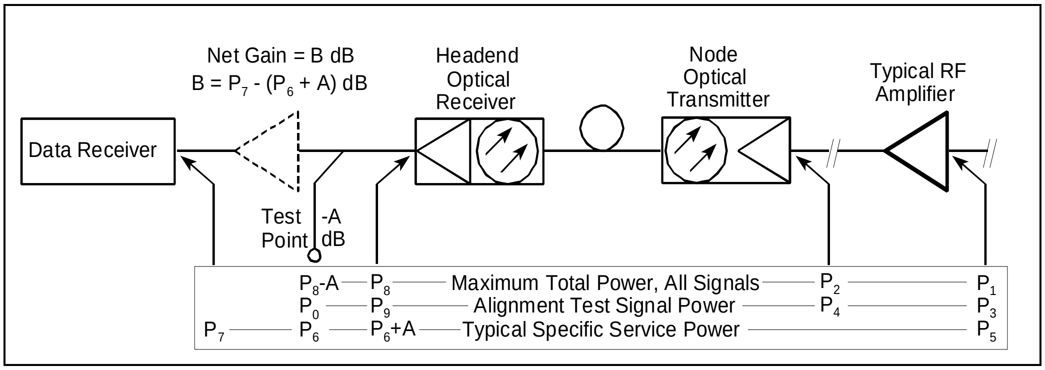


Figure 16-14: Symbolic Diagram of Upstream Showing Level Reference Points

Upstream laser transmitters vary more than coaxial amplifiers in their recommended input power, with some rated for a single carrier and others for four or more. This is due partially to the different transmitter technologies used: Fabry-Perot (FP) vs. Distributed Feedback (DFB) lasers, cooled vs. uncooled and optically isolated vs. non-isolated. Call the total input power handling capability of the laser transmitter P_2 in dBmV. **P_2 may or may not be equal to P_1.**

Finally, headend optical receivers may have post amplifiers with limited power handling capability. Since optical losses of links will vary, setting optical transmitter input levels may not be sufficient to guarantee that the headend optical receiver output level will be non-distorted. Therefore, the total output power handling capability of the headend receiver must also be determined based on the manufacturer’s ratings. This tutorial will refer to receiver total output power capability as P_8.

The consequences of designing for too much total input (and therefore output) power is that distortions will be excessive, including both intermodulation distortion and laser clipping. Additionally, non-intentional signals such as ingressing carriers and electrical transients will be more likely to drive the system into saturation. If the total power is set too low, then individual signals will be received at the headend with unnecessarily degraded carrier-to-noise ratios.

Note also that one reason that the total power level must be set well below the limiting level for a single carrier is that the peak voltage levels with multiple signals will be higher than that from a single signal, and the peak-to-average ratio will increase with the number of signals and will be higher for some types of modulation than for others. Thus, operators may wish to set the total power limit based on a probability density analysis related to clipping-level voltage rather than simply total average power. Note that P_1 may need to be adjusted as a function of the depth of the cascade. Alternatively, the total power level can be selected based on the results of a Noise Power Ratio test (see Section 13.5.2: “Noise Power Ratio (NPR)”). For more information on peak-to-average ratios, see Section 2.2.3: “Peak-to-Average Ratio”.

3) Pick standard reference levels at the upstream Reference Points and upstream laser transmitter input which will be used for system alignment.

A common industry practice is to use +20 dBmV at amplifier Reference Points (external ports or module inputs), but operators may choose to use some other value. Whatever level is chosen, the total test signal power should be at or below the total input power loading value, P_1, chosen above and sufficiently above the system noise so that accurate results can be obtained. This section will refer to this test level as P_3 in dBmV

The reference input level for the laser transmitter should differ from the amplifier level by the difference in the total input power levels chosen in step 2. For example, if the total input power for the amplifier modules is to be +26 dBmV and the total input power for the upstream transmitter is to be limited to +16 dBmV, the difference in reference test power levels must be 10 dB. Thus, if +20 dBmV is chosen as the reference level for alignment of the coaxial amplifiers, then the reference level at the return laser input must be +10 dBmV. This tutorial will refer to the laser test reference input level as P_4 in dBmV. Mathematically,

$$P_4 = P_3 - (P_1 - P_2).$$

4) The first step in system alignment is to insert a signal within the node so that the return laser is driven at the P_4 level. This signal may consist of a single carrier, several carriers, or a sweep signal. As with downstream testing, sweep signals will allow more accurate alignment and reveal problems that may not be apparent if testing is done at only a few frequencies, while use of multiple, unmodulated carriers will allow direct observation of distortion products.

   Depending on the design of the node, it may be possible to inject the test signal directly into the upstream optical transmitter, possibly by removing a jumper cable. In other cases, it may be necessary to use the node forward path output test point. In the latter case, if this is the initial alignment of a new upstream system, all other inputs to the laser transmitter should be disconnected. The best method of doing this is to insert high values of output pads in all the return amplifiers directly driving the node. This preserves the impedance matches on all the lines. Isolating the node allows the performance of the upstream optical link to be evaluated independently from the coaxial distribution system.

5) Evaluate the signal at the headend upstream monitoring point for the node being tested. If the node has been isolated, as described above, the fiber optic link can be tested for its noise contribution, amplitude flatness and other parameters in this step. Note that if the CNR is to evaluated relative to manufacturer's specified performance, it will be necessary to correct for any level difference between the test signal and the specified operating conditions and also to measure noise in the specified bandwidth.

6) Measure the upstream receiver output level. If the optical receiver/post amplifier provides the means for adjusting internal gain, either through adjustments or plug-in pads, set the receiver output level, P_9, so that it will be operating optimally under conditions of full signal loading. This level can be determined as follows:

$$P_9 = P_8 - (P_2 - P_4)$$

   This adjustment compensates for the variation in upstream link losses.

   If the optical receiver output is not adjustable, then measure the upstream output level P_9 and calculate what the output level P_8 would be if an upstream amplifier were driven with a single carrier whose level was equal to the full input power handling capability of the amplifier, $P_{1:}$

$$P_8 = P_9 + (P_2 - P_4)$$

7) Carefully measure and record the level of the signal(s) or sweep trace at the headend monitoring point. This reference level, which will be lower than P_9 by the attenuation of the test coupler, A dB, will be used for all subsequent node alignment. Call this level P_0.

8) Restore the node to normal operating conditions.

9) Align the coaxial distribution plant, working from the node outwards (beginning with the node itself, unless the node output test point was used to insert the test signals in step 4), using the following procedure at each amplifier station:

- First, insert a test signal into the reverse insertion point determined in step 1) above. The inserted signal should have a level such that the power at that amplifier's Reference Point will be P_3. For example, if a given line extender has an actual loss of 31 dB from the common output test point to the Reference Point and the P_3 level is chosen to be +20 dBmV, then the inserted test signal(s) should be at a level of +51 dBmV.

- Now adjust the upstream output pad and equalizer for the station being tested so that the headend monitoring point signal is just equal to P_0. This is most conveniently done if the field technician can directly observe the signal at the headend. Special purpose upstream sweep equipment provides this visibility by transmitting that information back to a combination test signal generator/display that the technician uses. Lacking that, some spectrum analyzers provide an NTSC video output representation of their screen display which can be connected to a modulator on a spare downstream channel. Then the field technician can use a portable television set to observe the headend signal.

- Continue to the next downstream amplifier in cascade. Where the network splits, amplifiers can be done in any order, so long as all amplifiers in the direct path between the amplifier being aligned and the node are done first.

If this procedure is followed correctly, the result will be that the signal level at the chosen input Reference Point of every amplifier between any insertion point and the node will be identical, and that the upstream optical transmitter levels (assuming an HFC plant) will differ from the Reference Point levels by exactly the difference in total allowable input power. If the chosen Reference Point is the upstream amplifier module inputs, it also results in each amplifier contributing equally to the total thermal noise in the system.

Choosing operating levels for specific services

The next step is choosing the operating level for the various services which will share the upstream transmission path. This is simple if just one service is contemplated, as signals can be run anywhere between where CNR is adequate and where the signals are unacceptably compressed. In the more general case, however, the operator may want to retain future flexibility by assigning power levels to each service in such a way that the total power is below the power handling capability of the amplifiers and upstream optical transmitter unless the entire usable bandwidth is occupied.

While there is no general industry agreement on how much power to allocate to each service, one popular method is to assign power on the basis of occupied bandwidth, sometimes called "Constant Power per Hz." Since the thermal noise power in a given channel is also proportional to bandwidth, this results in all services experiencing the same CNR, provided that:

- The noise power is flat across the return spectrum, and
- The noise susceptibility bandwidth of each service is the same as its occupied bandwidth.

Performance with respect to discrete interfering carriers is more difficult to predict. To the extent that ingressing carriers are randomly distributed with respect to both frequency and level, the probability of an ingressing carrier falling within the susceptibility bandwidth of a service occupying a narrow bandwidth will be lower, however any ingressing carrier will be larger relative to the desired signal, due to the lower assigned signal level. Thus the overall probability of interference will be dependent on the statistical distribution of interfering signals and may be better or worse for narrow-band, lower-level signals.

As an example, if the total amplifier input power, P_1, is equal to +26 dBmV and the usable return system bandwidth is 35 MHz, then the allowable power density is

$$+26 - 10 \log (35 \times 10^6) = -49.4 \text{ dBmV/Hz.}$$

A upstream signal occupying a 2 MHz band would then be run at a level of

$$-49.4 + 10 \log (2 \times 10^6) = +13.6 \text{ dBmV.}$$

While this seems an inherently “fair” method of power allocation among services, it has its limitations:

- As discussed previously, and in more detail in Section 1.4, upstream signal levels are primarily limited by peak voltage excursions, rather than total input power. Thus, a single unmodulated carrier may have the same average power as ten modulated carriers carried at 10 dB lower level, but will generally have a much lower peak voltage. In recognition of this, the maximum total power used in the above calculation must include a defined “headroom” for peak-to-average voltage ratio, or the maximum total power should be determined with a signal that has a high peak-to-average voltage, such as band-limited noise.

- While the thermal noise floor may be nominally independent of frequency, the sum of noise, distortion products and ingress is certainly not, with the result that some frequencies are more degraded than others.

- Not all services may use the same modulation scheme, and thus may require different CNRs. A simple frequency shift keyed (FSK) modulated carrier, for instance, is more robust than a 16 level quadrature amplitude modulated (16-QAM) signal. Logically, therefore, we should operate the latter signal at a higher level to get comparable performance.

- Not all services will tolerate the same error rate. Voice telephony signals, for instance, typically offer no opportunities to recover from lost data, while Internet data packets can be re-transmitted if damaged. Similarly, some protocols may include forward error correction (FEC) which is able to correct some errors.

- Services utilizing frequency agile upstream transmitters (as a means of moving away from potentially interfering ingress signals), may be assigned exclusive use of a portion of the upstream spectrum which is much greater than the channel bandwidth. The difference between channel bandwidth and service bandwidth will not be used for services and, thus, carrier levels could be higher.

Whatever method is used, the signal of each service should be assigned a power level at the amplifier input Reference Point. Call this level for a given service P_5.

Determining upstream transmitter dynamic range

Once an operating level is defined for a given service, the upstream transmitters, installation configurations and plant design must be analyzed to make sure that the desired levels are achievable.

In the downstream direction, it is common to set signal levels at all subscriber taps nominally the same (with a tolerance for available increments of tap values and for the unavoidable variations across the spectrum). §76.605(b)(3) through §76.605(b)(5) of the FCC's rules set the allowable limits on levels of downstream analog video signal levels and their variations. Section 2.1: "Analog Signal Power" beginning on covers measurement techniques for determining compliance with those rules.

Since the loss of both passive components (usually) and coaxial cable (always) increase with frequency, the upstream loss from subscriber's outlets back to the last amplifier station will be less than the loss for the downstream signals. Not only that, but the loss from subscribers who are more distant from the last amplifier be much less (as much as 20 dB) than from subscribers fed from the first tap. If upstream amplifier module inputs, rather than external ports, are chosen as the Reference Points for system alignment, then an additional variation will result from the variation in internal amplifier losses in the upstream path.

The total loss from upstream transmitter to amplifier input reference point should be analyzed over the full range of expected configurations to determine the required transmit levels. If these are not within the range of the selected terminal equipment, then either plant, drops, terminal equipment or operating levels must change. Some options for reducing the required upstream transmit range are:

- Selectively replace high value taps with lower values to lower maximum required upstream transmit levels
- Install in-line equalizers to increase the minimum required upstream transmit levels. In new designs, avoid low value taps for the same reason and also to reduce the sensitivity to ingress.
- Where allowed by adequate downstream signal levels, use unequal splitting arrangements in houses to favor the data transmitter and thus lower the maximum required upstream transmit levels.
- Selectively deploy drop amplifiers with two-way gain to lower maximum required upstream transmit levels.

Where it is not practical to reduce upstream losses sufficiently that transmitters will be able to reach amplifier input reference points at the desired level (P_5) under extremes of temperature and operating variations, then either different terminal equipment should be selected or the operator must accept a lower-than-optimum system operating level for the service supplied through the selected equipment. If the level must be lowered, the CNR and carrier-to-ingress ratios will be degraded relative to optimum transmission conditions in the upstream system, though they may be adequate for the service in question. The amount by which the level needs to be lowered will depend, at least in part, on the tolerance of the headend data receivers to varying input levels.

Setting upstream RF transmitter levels

Upstream transmitter levels must vary from transmitter to transmitter if the system operating levels are to be uniform. If transmitted levels are all the same, as is the case with some two-way set-top boxes, then the signal from some homes may cause severe overloading, while others are so low as to be noisy when received at the headend. This was sometimes tolerable when two-way boxes were the only users of the upstream system, used FSK modulation which is unaffected by signal limiting, and were polled to prevent more than one box from transmitting at a time. It is clearly a problem when multiple services will share the spectrum and are required to not interfere with each other.

Many modern upstream transmitters are agile in power level and controlled from the headend so that their levels as received in the headend data receiver are approximately the same. Setting up such a system is discussed in the next section. Where transmitters are fixed in power, they must be adjusted or padded externally so that the levels received at the headend are appropriate for the service. The proper headend test point level, P_6, for a service whose amplifier input operating level is P_5 (see previous section) is

$$P_6 = P_0 - (P_3 - P_5).$$

For example, if the headend test point level was +12 dBmV when the system was aligned with a Reference Point signal of +20 dBmV and the desired operating level for a given service is +5 dBmV at the same point, then the corresponding headend test point level used to set the upstream transmitters should be $P_6 = +12 - (20-5) = -3$ dBmV.

Setting up systems utilizing power-agile transmitters

If upstream transmitters have a transmit level that is controlled by the headend data receiver (resulting in a condition known as a "long loop AGC"), as is common with cable modems and telephony systems, then the headend must be carefully configured if the proper operating levels are to result. In order to do this, the operator must understand all the losses and gain stages between the headend optical receiver output and the headend test point and the data receiver input. In particular, if the loss from the headend optical receiver output to its test point is A dB, then the actual optical receiver output level corresponding to test point level P_6 is $P_6 + A$ dBmV. If the data receiver will control all upstream transmitters so that they hit the receiver at level P_7, then the operator must assure that the net signal gain, B, between each optical receiver output and data receiver input is

$$B = P_7 - (P_6 + A) \text{ dB}$$

(if the result of the calculation is negative, then attenuation must be inserted, rather than gain).

Note that, unless all optical receivers were adjusted to equal test signal outputs in step 6, the net gain between each optical receiver and its corresponding data receiver must be independently adjusted.

If too much gain is inserted between the optical receiver and the data receiver, the result will be lowered signal levels for that service in the plant and degraded CNR. If too little gain is inserted, the result will be that the upstream transmitters will be turned up too high and may overload the plant.

Summary

Although this process may seem complex, the steps are individually quite simple:

- First, determine the power handling capability of the upstream amplifiers, optical transmitter and headend receiver (assuming the system is an HFC network, otherwise only the amplifier capabilities need be considered).

- Second, pick a test level for alignment and set all amplifiers such that the input levels in the aligned plant are equal and the input to the upstream optical transmitter differs from the amplifier levels by the proper amount;

- Third, assign a power level to each carrier that optimally allocates the total system power handling capability among services. Where the selected upstream transmitters are unable to reach the optimal power from all locations, a lower level may be assigned to the service, at the expense of reduced CNR and carrier-to-interference ratios.

- Fourth, for manually adjusted upstream transmitters, adjust transmit levels so that the headend received levels are appropriate for that service;

- Fifth, for headend-controlled upstream transmitter power levels, adjust the net gain or loss between the optical receiver output and data receiver input so that the plant levels will be correct when the upstream transmitters are controlled by the data receiver.

If these procedures are followed, the plant will operate optimally, with the services operating in a mutually non-interfering manner, while obtaining the maximum possible CNR consistent with adequate headroom to accommodate operating variances and ingressing signals.

16.5 Upstream Noise Prediction

Description: The noise power in a 4 MHz bandwidth from a 75 Ω resistor into a 75 Ω load (which has no noise of its own) at 68 °F is -59 dBmV.

Noise Power (dBmV/Hz)

$$= -59 - \left[10 * \log(4 \times 10^{6})\right] = -125 \text{ dBmV/Hz}$$

Total Noise Power

$$N_{TH35} = -125 + \left[10 * \log\left(BW_{T}\right)\right] = -125 + \left[10 * \log\left(35 \times 10^{6}\right)\right] = -49.6 \text{ dBmV}$$

where:

$BW_{T} =$ the total upstream bandwidth (35 MHz in this example)

$N_{TH35} =$ the thermal noise in a 35 MHz bandwidth

RF Coaxial Amplifier Noise

If, in the coaxial portion of the network, the return amplifiers have equal noise figures of 10 dB, the equivalent input noise power will be -39.6 dBmV. For a 32 amplifier configuration, the amplified thermal noise power funneled by the coaxial portion can be calculated using the following equation:

$$N_{COAX} = N_{TH35} + NF + \left[10 * \log(M)\right] = -49.6 + 10 + \left[10 * \log(32)\right] = -24.6 \text{ dBmV}$$

where:

$N_{COAX} =$ the total amplifier noise power at the node station input

$NF =$ the amplifier noise figure

$M =$ the total number of amplifiers in the node

This is an increase of 15 dB at the input of the upstream node.

The above example assumes all amplifiers have the same noise figure. In the more general case, several different station configurations, each with a different noise figure, may exist in a node. In addition, the effective noise figure of a station will depend on whether the Amplifier Upstream

Module or the Amplifier Station Port was chosen as the alignment reference point (see Section 16.4: Return Plant Setup and Operational Practices for more details). Assuming that several station types exist and that all noise figures are calculated relative to the reference point, the total noise contribution can be calculated as follows:

1. Calculate the input noise power of each station type using the following equation:

$$N_X = -125(\mathrm{dBmV/Hz}) + 10 * \log(BW_T) + NF_X$$

where:

$N_X =$ the input noise power of an upstream amplifier station type

$BW_T =$ the total upstream bandwidth

$NF_X =$ the noise figure of an upstream amplifier station type (specified or calculated relative to the reference point used for alignment)

2. Calculate the total noise contribution of the coaxial plant using the following equation.

$$N_{COAX} = 10 * \mathrm{lLog}\left[\left(M_1 \times 10^{\frac{N_1}{10}}\right) + \left(M_2 \times 10^{\frac{N_2}{10}}\right) + \left(M_3 \times 10^{\frac{N_3}{10}}\right) + \ldots\right]$$

where:

$N_{COAX} =$ the total amplifier noise power at the node station input

$N_1, N_2, N_3 \ldots =$ the input noise power of each upstream amplifier station type

$M_1, M_2, M_3 \ldots =$ the quantity of each corresponding upstream amplifier station type

Optical Link Noise

The actual noise power at the output of the optical link is the total coaxial noise combined with the upstream optical link noise. The optical link noise is not subject to funneling additions and must be determined independently from measurements or manufacturer's specifications. The combining must also take into consideration the difference in total power between the coaxial portion, and the laser transmitter input.

In order to add the optical link noise to the RF coaxial amplifier noise, the optical link noise will be calculated as if it occurred at the node station input. An estimate of the noise contributed by the optical link can be made using the following equation.

$$N_{OPT} = C_2 - (C/N)_2 + \left[10 * \log\left(\frac{BW_T}{BW_2}\right)\right] + (P_1 - P_2)$$

where:

N_{OPT} = the total equivalent input optical noise power (in dBmV) normalized to the node station input

C_2 = the laser transmitter input signal power (in dBmV) for a specified or measured CNR

$(C/N)_2$ = the specified or measured CNR for some bandwidth and the C_2 signal power

BW_T = the total upstream bandwidth which is the same as used for the coaxial analysis

BW_2 = the bandwidth under which the reference CNR was measured or specified

$(P_1 - P_2)$ = the difference between the coaxial reference point and laser transmitter inputs, respectively. Note that the terms P_1 and P_2 are fully defined in Section 16.4: "Return Plant Setup and Operational Practices".

The $(P_1 - P_2)$ factor is required to normalize the optical noise power for the difference in operating levels in the coaxial amplifiers versus the laser transmitter.

Combined Noise (Coaxial and Optical)

With the normalized optical noise determined, the total equivalent input noise can be calculated using:

$$N_T = 10 * \log\left[10^{\frac{N_{COAX}}{10}} + 10^{\frac{N_{OPT}}{10}}\right]$$

where:

N_T = the total equivalent noise power normalized to the node station input.

N_{COAX} = the total amplifier input noise power at the node station input.

N_{OPT} = the total equivalent optical input noise power as determined above.

Figure 16-15 is a plot of the total noise power for different quantities of amplifiers and compares the coaxial noise (N_{COAX}) calculated earlier with the combined coaxial and optical noise. This example assumes the optical link performance is approximately equivalent to a 10 amplifier cascade (N_{OPT} is equal to -29 dBmV). An additional trace has been provided with 10 dB of noise in excess of the thermal noise in the coaxial portion of the network.

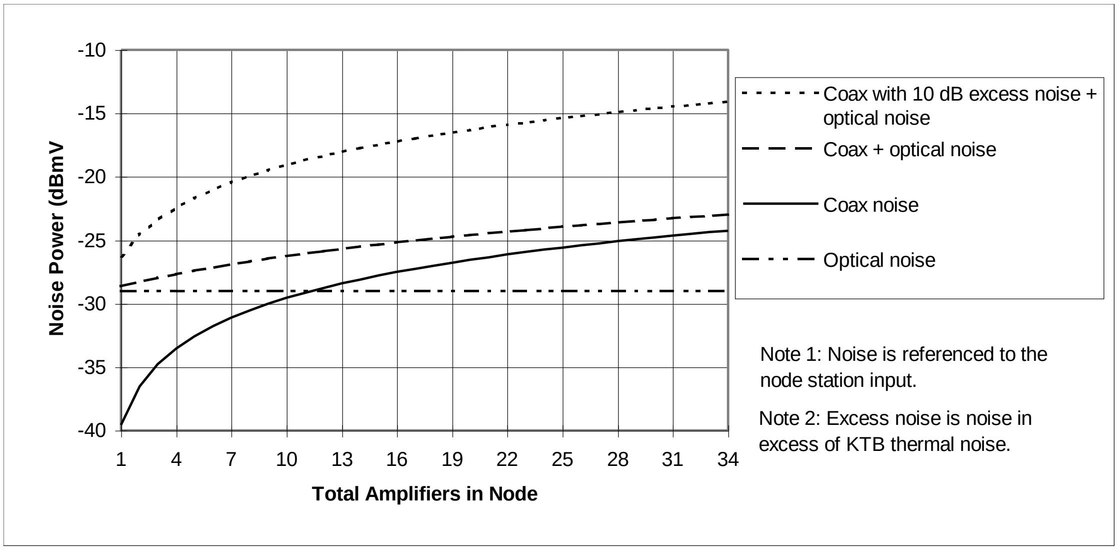


Figure 16-15: Return Noise Power Considerations (35 MHz BW)

If channel power is allocated per Hz, then for any given bandwidth channel, the carrier-to-noise ratio (which can be directly related to the modulation error ratio or BER) is nearly the same as the total power to total noise ratio as long as total noise is at least 10 dB lower than total power. The low power operating limit will be a function of amplifier noise funneling and the upper operating limit will come from laser clipping or amplifier compression. If the total link noise is calculated as above, then CNR of any service can be determined once the network size, characteristics of the amplifiers and optical link, and operating levels are known.

16.6 Technical Considerations in the Delivery of Satellite Signals to Cable Headends

Purpose

This section will present the following information on delivery of signals via satellite to a cable television headend:

1. General information on satellites and their configuration
2. Information on both analog and digital signals delivered via satellite
3. A discussion on satellite link design that includes the impairments a satellite signal will encounter
4. Receive earth station configuration and performance
5. An example satellite link budget

General Satellite Information

The name the FCC has given to the service that is used for the distribution of television programming (as well as other signals) is the Fixed Satellite Service (FSS). The FSS is covered in Part 25 of the FCC Rules and Regulations. The satellites in this service are arranged in an orbital belt 35,786 km above the equator. The orbital belt is known as the domestic satellite arc, and it extends from 70 degrees West longitude to 135 degrees West longitude. Each satellite in the belt is separated from its neighbor by 2 degrees. This orbit has been selected since it will enable the satellites to appear at a fixed position in the sky, relative to the receiving station on the earth, for the whole day. This simplifies the configuration of the satellite antenna at the receiving location because it will not have to move continuously to track the satellite's position This orbit is known as a Geosynchronous orbit since it is synchronous with the earth's rotation.

The approved frequency bands for use in the FSS are shown in Table 16-7.

	Uplink	**Downlink**
C Band	5.925 to 6.425 GHz	3.7 to 4.2 GHz
Ku Band	14.0 to 14.5 GHz	11.7 to 12.2 GHz

Table 16-7: Fixed Satellite Service Frequency Bands

The uplink frequencies are those used by the service provider to transmit programming to the spacecraft. The downlink frequencies are those used by the spacecraft to transmit the programming to the cable television headend. The signal received by the satellite is not demodulated on the spacecraft. It is simply frequency translated from the uplink frequency to the downlink frequency and amplified for re-transmission to the cable television headend.

The 500 MHz of available bandwidth on a spacecraft is divided, with the use of filters on the satellite, into channels called transponders. Each transponder has both an uplink and a downlink frequency associated with it. Spacecraft using C Band have 24 transponders with filter bandwidths of 40 MHz each. The center frequency of each transponder is separated from the adjacent transponder by 20 MHz. However, in order to avoid interference between adjacent transponders, the polarization of the signal is alternated between adjacent transponders as shown in Figure 16-16.

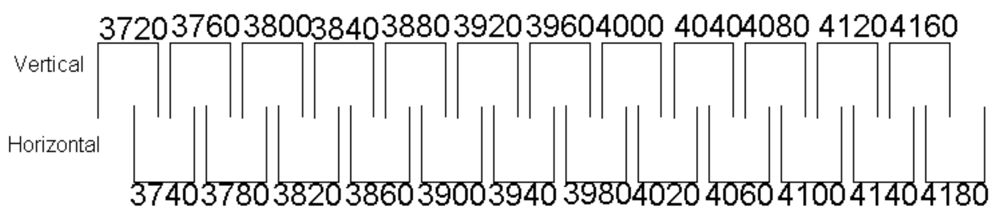


Figure 16-16: Typical C Band Frequency Plan

Spacecraft using Ku Band do not have a standardized bandwidth for transponders. The transponder filter bandwidths could be 30 MHz, 60 MHz, or some combination of both on a given spacecraft. Therefore, the number of Ku transponders per spacecraft will vary and be dependent on the particular spacecraft design. Additionally, the polarization of the transponder does not necessarily alternate between adjacent transponders. For example, all transponders in the downlink may have a like polarization, or half may be vertically polarized while the other half are horizontally polarized. A Ku Band frequency plan is shown in Figure 16-17.

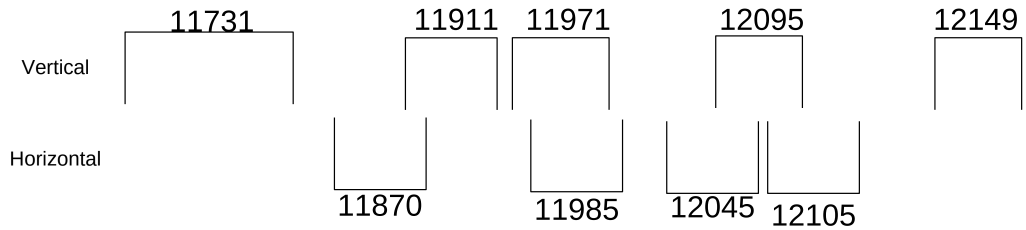


Figure 16-17: Typical Ku Band Frequency Plan

While the filter bandwidth of the transponders used on the spacecraft is as listed above, the useable bandwidth for a particular signal accessing the transponder will be less. This is done to give a certain amount of guard bandwidth to each transponder. For C Band the usable bandwidth for a 40 MHz transponder is 36 MHz. For Ku Band, the usable bandwidth for a 30 MHz and 60 MHz transponder is 27 MHz and 54 MHz, respectively.

Each of the transponders on the spacecraft has its own amplifier. On C Band spacecraft these amplifiers are Solid State Power Amplifiers (SSPA) which use GaAs FET devices for the amplification process. The most recent versions of these amplifiers have output powers of 25 to 35 Watts. On Ku Band spacecraft the amplifiers use a Traveling Wave Tube (TWT). These devices have power outputs which are typically in the 60 Watt range, but some can have as much as 120 Watts per transponder. Neither the SSPA nor the TWTA is a linear amplifier. Since the power subsystem on the satellite is of limited capacity, the power amplifiers that are used must have reasonably good efficiencies. Both SSPA's and TWTA's can provide good efficiencies. Additionally, to achieve the largest amount of output power possible these amplifiers are operated at or near their saturated output powers. This amplifier operating point is in a nonlinear range so the modulation scheme of the signals that access the transponder must be tolerant of the nonlinearity.

Signal Classification

Both analog and digital signals are available via satellite to the cable operator. The modulation scheme for each is chosen appropriately to accommodate the special needs of satellite transmission. Each signal will be described below.

Analog Satellite Signals

Analog television signals delivered via satellite use Frequency Modulation (FM). Each signal carries one NTSC program and can carry multiple audio signals. The video and audio signals modulate an RF carrier set to the center frequency of the transponder. The baseband video signal extends from DC to 4.2 MHz, and the audio subcarriers are typically placed at 5.5 MHz, 6.2 MHz, and 6.8 MHz. Not all subcarriers are necessarily in use on every transponder. A subcarrier at 7.1 MHz is a special subcarrier that is used for uplink identification. This subcarrier is called the Automatic Transmitter Identification System (ATIS) subcarrier. It carries information, encoded with Morse code, about the uplink such as its call sign and telephone number. This information can be used to aid in the identification of an interfering station.

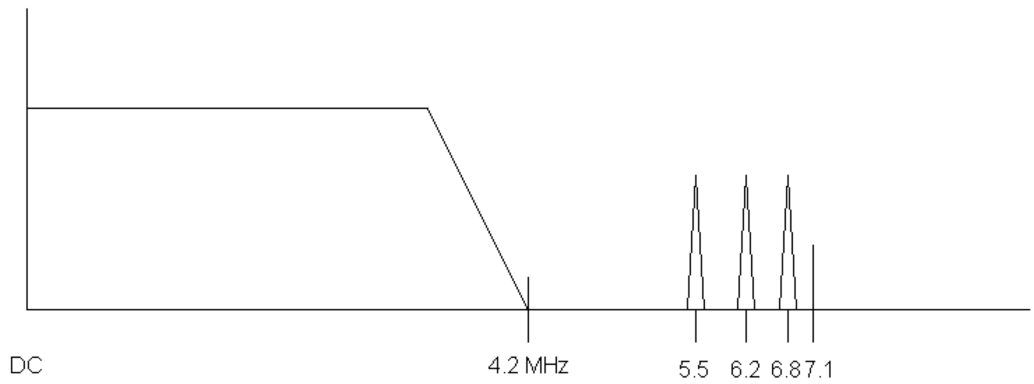


Figure 16-18: Typical Composite Baseband Satellite Signal

The deviation of the RF carrier, by both the video and audio subcarriers, is set to a level that ensures the total signal bandwidth is not more than the useable bandwidth of the transponder. The video signal deviates the carrier to a larger degree than the audio subcarriers. Each audio subcarrier is itself frequency modulated, with the audio information, before it modulates the RF carrier. Typically, the audio subcarriers deviate the RF carrier 2 MHz. The composite amplitude of the subcarriers must be controlled so as not to deviate the RF carrier more than this amount. Therefore, as the number of subcarriers used in a given transponder increases, the amplitude of each subcarrier must be reduced in level so the composite subcarrier amplitude does not over deviate the RF carrier. The video signal deviates the RF carrier 10.515 MHz. Additionally, with a C Band carrier, a signal known as Energy Dispersal also modulates the RF carrier. This signal is a 30 Hz triangular wave that deviates the RF carrier 1 MHz. It is used to ensure that the downlink signal at C Band does not interfere with terrestrial microwave signals that occupy the same frequency range as the downlink signal. Energy Dispersal is not used at Ku Band.

The composite deviation of the RF carrier (Δf_c) may be characterized as the Root Sum Square of the deviations of the RF carrier by the individual baseband signals.

$$\Delta f_c = \sqrt{(\Delta f_1)^2 + (\Delta f_2)^2 + (\Delta f_3)^2} \text{ [MHz]} \qquad (1)$$

where:

Δf_c	=	composite deviation of RF carrier
Δf_1	=	deviation of RF carrier by video signal
Δf_2	=	deviation of RF carrier by the composite audio subcarrier signal
Δf_3	=	deviation of RF carrier by Energy Dispersal

Inserting the numbers for all the deviations listed in the paragraph above, the peak deviation for the RF carrier can be calculated.

$$\Delta f_c = \sqrt{(10.515)^2 + (1)^2 + (2)^2}$$

$$\Delta f_c = \sqrt{110.565 + 1 + 4} = \sqrt{115.25}$$

$$= 10.750 \text{ MHz}$$

Occupied Bandwidth of an Analog Satellite Signal

The occupied bandwidth of an FM signal can be calculated using Carson's Rule which states the following:

$$BW=2(\Delta f_c + f_m) \ [MHz] \quad (2)$$

where:

BW = occupied bandwidth in MHz

Δf_c = composite deviation of RF carrier in MHz

f_m= highest instantaneous modulation frequency in MHz

The value for Δf_c in the case of the FM satellite signal is 10.750 MHz. As was previously calculated this is the peak composite deviation of the RF carrier. The value of f_m is based on the highest frequency subcarrier, in this case the ATIS subcarrier. However, since all the subcarriers are FM they cannot be treated as discrete signals with zero bandwidth. They have, albeit small compared to the total television signal, an occupied bandwidth of their own. Therefore, the value of f_m is equal to the sum of the highest subcarrier frequency plus the maximum deviation of that subcarrier. For the ATIS subcarrier the FCC Rules & Regulations specify a maximum deviation of 25 kHz. That will make f_m equal to 7.125 MHz.

Knowing the values of Δf_c and f_m the value of the BW for the FM television satellite signal can be calculated.

$$BW = 2(10.750 + 7.125) \text{ MHz}$$

$$BW = 35.75 \text{ MHz}$$

Notice that the occupied bandwidth of the signal is slightly less than the useable bandwidth of a C Band transponder. While a Ku Band satellite can have a 54 MHz useable bandwidth transponder, an FM video signal is not deviated to a greater amount. It would still occupy 35.75 MHz when used at Ku Band. However, in general, Ku Band transponders are not used to deliver FM television to headends.

Digital Satellite Signals

Unlike analog modulation, where the baseband information continuously varies between two levels with respect to time, with digital modulation the baseband information is binary (i.e., the "0's" and "1's") which has two discrete levels with respect to time. In order to modulate an RF carrier, the baseband information must vary one or more properties (e.g., amplitude, phase, or frequency) of the RF carrier. With analog modulation the RF carrier properties are varied in a continuous manner; however, with digital modulation the carrier properties are varied to discrete carrier states. Each of the discrete states is known as a symbol, and the number of symbols transmitted per second is called the symbol rate. A particular digital modulation scheme can be classified according to the property of the RF carrier that is varied and to the number of baseband bits that are assigned or "mapped" to each symbol. This form of digital modulation is known as M-ary modulation. The value of M would be equal to the number of discrete states or symbols in the modulation scheme and is determined by the number of bits that are mapped to each symbol. It can be represented mathematically as in (3):

$$M = 2^n \quad (3)$$

where

n is the number of bits per symbol

If, for example, the value of n is 1 then M would equal 2, and the modulation scheme would be referred to as Binary modulation since a carrier property is varied between two states. If the value of n is 2, then M would equal 4 and the modulation scheme would be referred to as Quadrature modulation since a carrier property is varied between four states

To arrive at the symbol rate for the signal, simply divide the baseband bit rate by the number of bits per symbol used with the particular modulation scheme.

$$R_s = R_B / n \quad \text{[Msymbols/second.]} \qquad (4)$$

where:

R_s is the symbol rate

R_B is the baseband bit rate in the same units as R_s

n is the number of bits per symbol

As will be seen later the symbol rate will aid in determining the occupied bandwidth of the signal.

The modulation scheme used with satellite delivered digital television signals is M-ary Phase Shift Keying (M-ary PSK) modulation. With PSK the baseband information varies the phase of the RF carrier in discrete states. The two most encountered forms of M-ary PSK on a satellite link are Binary Phase Shift Keying (BPSK) and Quadrature Phase Shift Keying (QPSK). Since with BPSK M=2, there are two possible RF carrier states (or symbols) that can be selected. One baseband bit is mapped to each of the two symbols causing the RF carrier phase to be shifted between the two states. The RF carrier phase difference will be 180 degrees. With QPSK modulation since M=4, there are four possible RF carrier states that are 90 degrees apart. Two baseband bits are mapped to each of the four symbols. Table 16-8 and

Table 16-9 show a possible symbol mapping of both BPSK and QPSK.

Table 16-8: BPSK Symbol Mapping

Baseband Bit	RF Carrier Phase
0	0 degrees
1	180 degrees

Table 16-9: QPSK Symbol Mapping

Baseband Bit	RF Carrier Phase
00	45 degrees
01	135 degrees
10	225 degrees
11	315 degrees

If a digital television system had a baseband bit rate of 29.30 Mbps, a BPSK modulated signal would have a symbol rate of 29.30 Msymbols per second; however, a QPSK modulated signal would have a symbol rate of 14.65 Msymbols per second. If the system was adjusted so both a BPSK and a QPSK signal had the same symbol rates, the QPSK signal could carry twice the baseband bit rate as the BPSK signal, or 58.60 Mbps.

Occupied Bandwidth of a Digital Satellite Signal

In the previous section it was stated that combinations of baseband bits select which symbol is to be transmitted. The baseband bits themselves are not directly applied to the modulator. Each bit or group of bits is first converted to an analog pulse. This pulse is then filtered to give it a very unique shape. The shape is chosen to aid in detection of the pulse at the receiver and to minimize overlap between pulses. The filter also aids in containing the overall bandwidth of the modulated signal. The shape of the filter used has a Raised Cosine shape. A key parameter of the filter that affects signal bandwidth is the filter roll-off factor. This factor is represented by the Greek letter alpha, α, and it has a value between 0 and 1. The lower the value of α the steeper the filter roll-off will be.

To determine the occupied bandwidth of a digital satellite signal, the symbol rate, R_S, must first be calculated. Once R_S is known the bandwidth of the signal can be calculated as follows:

$$BW = R_S (1+\alpha) \quad [\text{MHz}] \tag{5}$$

where:

α is the roll-off factor of the filter

R_S is the symbol rate in MHz

Therefore, if a digital signal had a symbol rate of 29.30 Msymbols per second and a roll-off factor of 0.23, its occupied bandwidth would be equal to 36 MHz.

$$BW = 29.30(1+0.23)$$

$$BW = 36 \text{ MHz}$$

As it turns out this is the useable bandwidth of a C Band transponder. Therefore, if the signal used BPSK, the C Band transponder would have a capacity of 29.30 Mbps. But, if QPSK was used, the capacity of the transponder would be doubled to 58.60 Mbps.

A very good approximation for the bandwidth of a digital signal is the 3 dB bandwidth, and as it turns out the 3 dB bandwidth is equal to the symbol rate of the signal.

Satellite Link Design

In this section information will be presented that can be used in the determination of the performance of both analog and digital signals in a satellite system. The information should aid in the design of the receiving earth station at the headend.

There are two links involved in a satellite system. The uplink path from the program provider's earth station to the satellite, and the downlink path from the satellite to the receiving earth station (e.g., the cable television headend). Only calculations for the downlink path will be presented in this section since the operator has no control over the uplink. However, the uplink's performance will affect the downlink performance and comments will be made where appropriate.

In the General Satellite Information section above, information was presented on spacecraft configuration. That information will be useful in determining the link performance of the satellite signal at the headend. As with any link calculation, there are five parameters that must be known or calculated to determine the link performance. These five are:

1. The transmitter's radiated power
2. Any losses on the path between transmitter and receiver
3. The amount of Noise and Interference in the link
4. The receiving system performance
5. The signal-to-noise ratio required by the receiver in order to demodulate the signal correctly

Once all five quantities are known, the link performance can be calculated easily.

Satellite Radiated Power

The power radiated by the spacecraft is a function of both the output power from the transponder amplifier, and the antenna gain of the downlink antenna on the satellite. While the calculation of this number is not difficult, it does not have to be performed by the link designer. The information is easily found in reference material for spacecraft. One source is the World Satellite Almanac. This book will present the radiated power for each transponder on all spacecraft in the domestic arc. The data will be presented as Effective Isotropic Radiated Power (EIRP) in dBW. The unit dBW refers to the power in decibels relative to 1 Watt. The number listed in the reference will be the saturated power of the transponder.

Path Attenuation

Various forms of attenuation in the radio path between the satellite and the receiving earth station will degrade the signal at the receiver. Some of the more prominent forms of path loss will be presented here. These loses will need to be taken into account when the link budget is calculated.

Free Space Loss

When an RF signal propagates between a transmitting antenna and a receiving antenna the signal will be attenuated. This attenuation is known as the Free Space Loss or, more correctly, the Spreading Loss. It is called Spreading Loss because as a signal travels from a transmitting antenna toward a receiving antenna the wave spreads out. This is analogous to the spreading of the waves in a pool when a rock is tossed into the pool. At the receiving antenna, due to the spreading, not all of the original signal is available for capture by the antenna so this results in an apparent loss in signal level. The amount of attenuation is both frequency and distance dependent and can be calculated using the formula in (6).

$$\text{Loss} = 92.5 + 20*\log(F_{DL}) + 20*\log(D)\ \ [\text{dB}] \qquad (6)$$

where:

F_{DL} is the downlink frequency in GHz

D is the distance from the earth station to the satellite in km

While the satellites in the domestic arc are located 35,786 km above the equator, when the earth station is located at a point on the earth other than at the equator the distance to the satellite will be greater than 35,786 km. However, for the latitudes encountered for cable television headend installation in North America, the correction is small and only produces a difference <1 dB in Free

Space Loss. Therefore, setting D equal to 35,786 is a valid assumption in the calculation. Table 16-10 contains FSS downlink frequencies and the corresponding Free Space Loss (FSL) values.

Table 16-10: Free Space Loss vs. Downlink Frequency

C Band Frequency (GHz)	FSL (dB)	Ku Band Frequency(GHz)	FSL (dB)
3.70	195.0	11.70	205.0
3.95	195.5	11.95	205.1
4.20	196.0	12.20	205.3

Polarization Mismatch

Since satellites use two polarizations (i.e., Horizontal and Vertical), the polarization of the antenna at the receiving earth station must be adjusted properly. If the antenna is not adjusted correctly for the desired transponder polarization, the signal from the spacecraft will be attenuated due to the mismatch in the polarization.

Rain Attenuation

In addition to the link attenuation caused by the Spreading Loss, a rain cell in the path between the earth station and the satellite will attenuate the signal power at the receive earth station. This is due to raindrops both scattering and absorbing the radio wave energy. The amount of loss is a function of radio frequency, sky temperature, rain rate, and path length. Therefore, the harder it rains the greater the attenuation will be for a given frequency and path length. Signals in the Ku Band will experience a much larger attenuation than signals in the C Band. The equation is (7) calculates the approximate amount of attenuation per kilometer of path length:

$$\alpha = aR^b \quad [\text{dB/km}] \tag{7}$$

where:

R is the rain rate in mm/hr

a and b are frequency and temperature dependent constants

To get an approximation of the total path loss the value found in (7) must be multiplied by the total path length in kilometers. Researchers have spent numerous years studying rain cells in order to calculate a value of rain attenuation for a specific size rain cell. The attenuation models arrived at by the researchers are based heavily on statistics. The two most widely used rain attenuation models are the Crane Rain Attenuation Model and the CCIR Model. The models use very similar equations for calculating the amount of attenuation, but they differ in the value of path length used in the equation.

The Crane Model is more often used for North America and will, therefore, be reviewed in this sub-section. The Crane Model divides the world into four climate regions: Polar, Temperate, Sub-tropical, and Tropical. Each of the four regions is further sub-divided into two regions based on the amount of rain in each region. Table 16-11 shows the eight sub-regions that the Crane Model uses.

Table 16-11: Crane Rain Rate Climate Regions

Polar	Temperate	Sub-tropical	Tropical
Tundra (Dry,) A	Maritime, C	Wet, E	Moderate, G
Taiga (Moderate), B	Temperate, D	Arid, F	Wet, F

Crane labels each sub-region with a letter from A to F. Some sub-regions are further sub-divided based on the particular rain rate in the sub-region. That is, a particular sub-region may be considered Temperate, but due to differences in rain rate within the region it is further divided. This is done to increase the accuracy of the attenuation calculation due to the differing rain rates in a particular sub-region.

The equation used to calculate the amount of rain attenuation is as follows:

$$L_R = (aR_p^b / \cos\theta c * [(e^{Ubd} - 1)/Ub - (X^b e^{Ybd}/Yb) + (X^b e^{YbD}/Yb)] \text{ [dB]} \quad (8)$$

where:

a,b = Specific rain attenuation coefficients

R_p = point rain rate for the sub-region

θ = elevation angle of the earth station antenna

d,U,X,Y = empirical constants that depend on R_p

D = projected surface path length of the satellite-to-ES path

The magnitude of R_p in each sub-region is based on statistical studies of point rain rate. The value of R_p chosen to use in the equation is based on the link availability required for the link. The more link availability required for a given link the greater the attenuation experienced on the link. Typical values of downlink rain attenuation, in dB, for different frequency bands and link availabilities are presented in Table 16-12. These calculations were done for an earth station located in the D2 rain zone.

Table 16-12: Rain Attenuation [dB] for a D2 Rain Zone

Link Availability	99.99%	99.95%	99.00%	97.00%
C Band	0.2	0	0	0
Ku Band	10.5	0.8	0.4	0.1

As can be seen in Table 16-12, for a link availability of 99.99% Ku Band has significantly more attenuation than C Band. The rain rate associated with 99.99% availability for the D2 zone is 47.1 mm/Hr as compared with 0.9 mm/Hr for 97.00% availability. As a point of reference, a 99.99% link availability implies that the link will be unavailable for 52.56 minutes out of the entire year; however, an availability of 97.00% implies the link will be unavailable for 15,768 minutes out of the entire year. A complete step-by-step procedure for using the Crane Model is presented in" Appendix: Calculation of Rain Attenuation Using the Crane Rain Attenuation Model".

In addition to increasing the attenuation on a satellite link, a rain event will also depolarize the satellite signal. Since the earth station antenna is adjusted to have a specific polarization to receive the signal from the satellite, the depolarization caused by the rain event will add additional loss to the satellite link since the rain is effectively rotating the electrical polarization of the satellite signal. This loss is typically much less than 1 dB for earth station latitudes in North America.

Sources of Noise and Interference in a Satellite System

As with any communication system, a satellite system will experience noise and interference from both internal and external sources. Noise can come from both the internal electronics used in the earth station and from external sources. Interference can come from signals on neighboring spacecraft, neighboring transponders on the same spacecraft, and from intermodulation products from the same transponders.

Noise in a communication system is in the form of Additive White Gaussian Noise (AWGN). This noise will have a power spectral density that is flat across a large frequency range. A discussion on the specifics of AWGN is beyond the scope of this paper; however, it will be sufficient to state that AWGN includes thermal noise produced by the random motion of electrons in a conducting medium, solar noise and cosmic noise. The amount of power delivered to a matched load by an AWGN noise source is dependent on the temperature of the noise source and on the bandwidth over which the noise is measured. Noise power can be calculated using the formula in (9).

$$N = kTB \quad \text{[Watts]} \tag{9}$$

where:

k is Boltzmann’s Constant ($1.38e^{-23}$ J/K)

T_S is the temperature of the noise source in kelvin

B is the bandwidth in Hz

The formula in (10) can be used to determine the noise performance of a system of cascaded components that is driven by an AWGN noise source. The noise source is not the only contributor of noise to the system, though. Each of the components in the cascaded system contributes noise to the system. The total noise in the system will be referenced to a particular component’s input. Because of this fact the total noise from all of the components can be considered to be produced by a fictitious noise source connected to the reference terminals. All of the components in the cascade are then considered to be noiseless. The fictitious noise source can be represented by a noise source with an equivalent noise temperature, $T_{e.}$ The total noise temperature of the system is then equal to the sum of the noise from the input noise source, T_S and the noise from the equivalent noise source, T_e.

$$T = T_S + T_e \quad \text{[kelvin]} \tag{10}$$

where:

T_S is the noise temperature of the input noise source

T_e is the noise temperature of the equivalent noise source

In order to calculate the value of T_e from a cascade of components as in Figure 16-19 use the formula in (11).

$$T_e = T_{e1} + T_{e2}/G_1 + T_{e3}/(G_1G_2) + \ldots + T_{eN}/(G_1G_2\ldots G_N) \tag{11}$$

where:

T_{en} is the equivalent noise temperature of the n^{th} component in the cascade in kelvin

G_n is the gain of the n^{th} component in the cascade as a ratio

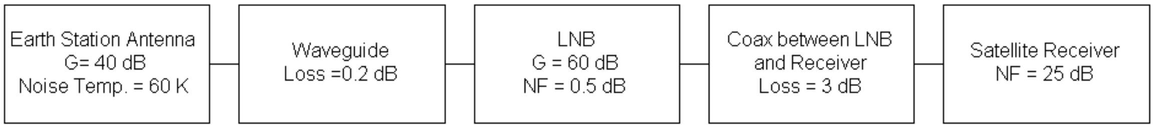


Figure 16-19: Typical Receiving Earth Station Cascade

The concept of noise temperature may be foreign to some readers, but the concept of a component's Noise Figure (NF) may not be. Noise Figure is the measure of the signal-to-noise ratio at the system input to the signal-to-noise ratio at the system output. NF can be related to noise temperature easily through the formula in (12).

$$T_e = (F-1)\,T_0 \quad \text{[kelvin]} \qquad (12)$$

where:

T_e is the equivalent Noise Temperature of the component in kelvin

F is the Noise Figure of the component expressed as a ratio

T_0 is the ambient temperature, typically 290 K

Also note that the NF of a component which has loss, such as a piece of waveguide, is simply its loss in decibels.

Earth Station Configuration and Performance

The receiving earth station must have high performance capability since the signal from the satellite will be very small in amplitude. Additionally, the receiving system at the earth station must contribute very little noise to the signal so the signal-to-noise ratio will not be reduced. The receive system, in its most basic form, consists of an antenna, a Low Noise Block Converter (LNB), and a receiver. These three components are in cascade and are interconnected using coaxial cable and waveguide. Figure 16-19 is a block diagram which is representative of the cascade of components in a basic earth station.

Since there can be a long cable run between the LNB and the satellite receiver, any signal loss will need to be minimized. This is accomplished by converting the C Band or Ku Band frequencies from the satellite to lower frequencies. The lower frequencies will experience less attenuation for a given cable length. The frequencies that the LNB has at its output are in the 950 to 1,450 MHz range.

A figure of merit that is used to determine the performance of the receiving earth station is the antenna gain-to-noise temperature ratio, G/T. The larger this value the better the earth station will be at receiving the weak signal from the spacecraft. The parameter G is the gain of the earth station antenna and the parameter T is the equivalent noise temperature of the receiving system both referenced to the input of the LNB.

Antenna Performance

The gain of the antenna used for reception of the downlink signal at the cable television headend should be high enough to overcome the significant attenuation in the link between the satellite and the earth station. It should have reasonable mechanical tolerances, and its antenna pattern should be precise.

The typical antenna used for downlink reception is a parabolic dish. This type of antenna is from a family of antennas know as Reflector Type antennas. The basic operation of the antenna involves a feed system that is placed some distance away from a parabolic shaped reflecting surface. The reflecting surface can be solid or made of a wire mesh. The gain of the antenna can be calculated using the formula below. This formula assumes that the antenna has a circular shape.

$$G = 10*\log(\eta(\pi fD/c)^2) \quad [dB] \qquad (13)$$

where:

η is the illumination efficiency of the antenna, (55% to 65%)

π is the value pi

f is the frequency in Hz

D is the antenna diameter in meters

c is the speed of light in m/s

The illumination efficiency of the antenna is a function of the type of feed system used for the antenna, the surface roughness of the antenna reflector, and any blockage in the aperture of the feed by mechanical supports on the antenna. As can be seen from an examination of the equation if the diameter of the antenna is increased, for a fixed frequency, the gain will increase. Similarly, if the frequency of a signal is increased, for a fixed size antenna, the gain will also increase. Table 16-13 lists some typical gains for antennas that would be encountered at a headend.

Table 16-13: Typical Antenna Gains of Earth Station Receiving Antennas

Ant Size [m]	C Band (3.95 GHz) [dB]	KU Band (11.95 GHz) [dB]
1.8	35.50	45.20
3.0	40.00	49.62
3.8	42.00	51.70
4.6	43.70	53.30

The radiation pattern from a parabolic antenna should be in the form of a pencil beam where the maximum gain is concentrated in a narrow beam that is pointed at the spacecraft. Gain of the antenna in any direction other than that of the main beam is known as a sidelobe of the antenna pattern. If the antenna has significant gain outside of the main beam (i.e., high gain sidelobes), the earth station will experience interference from neighboring satellites to the desired satellite. This is especially important since spacecraft only have 2 degrees of separation between them. Also, if the shape of the reflector is deformed, the precision of the antenna beam will be compromised. This can cause difficulty in correctly aiming the antenna at the spacecraft.

System Noise Performance

In a satellite receiving system the antenna can be treated as the AWGN noise source connected to the input of the cascade of components in the system. The noise temperature of the antenna (i.e., T_A) is dependent on factors such as the elevation angle at which the antenna is pointed and the quantity and magnitude of the sidelobes in the antenna pattern. For simplicity, with the typical antenna found at a headend in North America assume an antenna noise temperature of $T_A = 30$ K. The equivalent noise temperature of the cascade of other components is referenced to the input of the LNB.

G/T Evaluation

Now that a method for antenna gain calculation and noise temperature calculation has been presented, it is an easy exercise to calculate the earth station figure of merit, G/T. Recall that the G/T value is referenced to the LNB input in Figure 16-19. Using the formula in (14) the total noise temperature may be calculated.

$$T = T_A / L_{WG} + (L_{WG} - 1/ L_{WG}) T_0 + T_e \qquad (14)$$

where:

T_A is the antenna noise temperature, 30 K

L_{WG} is the loss of the waveguide connecting the antenna feed system to the LNB

T_0 = 290 K

T_e is the equivalent noise temperature of the cascade from the LNB to the receiver

The gain of the antenna referenced to the LNB input is simply the gain of the antenna minus the loss of the waveguide in dB. Therefore, the G/T of the receive system is the difference between the G and T in dB. This calculation for G/T is based on clear sky conditions.

When a rain event is occurring the sky temperature that the earth station sees is increased due to the warm rain cell. This will raise the value of T_A and consequently lower the G/T of the system during the time of the rain event. The increase in sky temperature caused by the rain cell can be calculated as in (15).

$$\Delta T = T_R (1-1/ L_R) \qquad (15)$$

where:

L_R = attenuation associated with the rain event

T_R = temperature of the rain cell

This new value of T can be substituted in the G/T calculation to arrive at a new figure of merit for the earth station. Therefore, a rain event affects the received signal in three ways. It will lower the signal power received at the earth station due to increased link attenuation and depolarization of the signal, and it will decrease the G/T figure of merit of the earth station due to the increase noise temperature of the sky during the rain event.

Sun Outages

Sun outages (also called solar transit outages) occur twice a year, in Spring and Autumn, when the Sun moves across the equator. Relative to an earth station on the ground this effectively places the Sun behind a Geosynchronous satellite. At that time the earth station antenna pointing at the satellite will “see” a very large amount of RF energy from the Sun. This energy will manifest itself as thermal noise entering the earth station antenna. The noise will degrade the performance of the earth station causing the CNR of the signal to decrease. The amount of degradation and the duration of the outage are dependent on the earth station antenna’s size. Smaller diameter antennas will suffer from longer, more severe outages due to their larger beamwidths and lower gains. The outages will occur during March and October for about one week with duration of a few minutes per day. The onset of a Sun outage can be seen in the satellite downlink as an increase in block errors for a digital signal or the addition of black and white “sparklies” in an analog picture.

Required Carrier-to-Noise Ratio

Analog and digital satellite signals have different requirements with respect to carrier-to-noise. This is due to the differences in the modulation used for both signals. However, once all of the parameters such as spacecraft EIRP, path loss, and antenna G/T are known it is a simple exercise to calculate the carrier-to-noise at the earth station receiver.

Analog Satellite Signals

For analog satellite signals the carrier-to-noise ratio is a measure of the amount of carrier power received at the earth station divided by the amount of noise power measured in the IF bandwidth of the receiver. The typical value used for the receiver IF bandwidth is 36 MHz. Use formula (16) to calculate the carrier-to-noise ratio.

$$CNR = EIRP - PL + G/T - 10*\log(BW) + 228.6 \ \ [dB] \qquad (16)$$

where:

CNR is the carrier-to-noise of the received signal in dB

EIRP is the Effective Isotropic Radiated Power of the transponder in dBW

PL is the path loss in dB

G/T is the earth station antenna gain-to-noise temperature ratio in dB/K

BW is the IF bandwidth of the receiver in Hz

The link carrier-to-noise ratio is actually the power summation of the uplink carrier-to-noise and the downlink carrier-to-noise. However, since a cable system operator has no control over the uplink carrier-to-noise and this value is typically much larger than the downlink carrier-to-noise, a good approximation of link performance can be obtained by simply calculating the downlink carrier-to-noise. The path loss value, PL, in the formula can be the sum of all losses in the path (i.e., spreading loss, rain attenuation, pointing loss, etc.).

Digital Satellite Signals

For digital satellite signals there is no carrier present. The signal power is evenly spread across the occupied bandwidth of the signal. Instead of using carrier-to-noise as a quantifier of link performance the quantifier Energy per symbol-to-Noise density, E_S/N_0, is used. The noise density is simply the noise power measured in a 1 Hz bandwidth. The receiver IF filter bandwidth can be set to the same value as the symbol rate bandwidth of the digital signal. This is the same as the 3 dB bandwidth of the signal. When this is done the E_S/N_0 of the digital signal can be calculated using the formula in (17); however, E_S/N_0 must be substituted for CNR. If a bandwidth other that the symbol rate bandwidth is used for the calculation use formula (17).

$$E_S/N_0 = CNR + 10*\log(BW) - 10*\log(R_S) \ \ [dB] \qquad (17)$$

where:

E_S/N_0 is the Energy per symbol-to-Noise density of the digital signal

CNR is the carrier-to-noise of the received signal in dB

BW is the bandwidth selected for the calculation in Hz

R_S is the symbol rate of the digital signal in Hz

Conclusions and Example

This section of the tutorial has presented enough basic information to enable an engineer to determine the performance of an earth station at a headend. An example will now be presented in order to demonstrate to the reader the process.

Example of a Link Budget for an Analog Satellite Signal

A C Band analog FM video signal is to be received at a headend. The following parameters are assumed:

1. Spacecraft EIRP is 45 dBW
2. The earth station antenna is 3 M in diameter and uses an LNB with a Noise Figure of 0.5 dB and a gain of 60 dB
3. Waveguide Loss is 0.2 dB
4. Coax Loss is 3 dB
5. Receiver Noise Figure is 25 dB and the threshold CNR of the receiver is 7 dB
6. Clear Sky conditions (i.e., no rain event is occurring)

The first step will be to calculate the earth station G/T. This can be done using the formulas in (11), (12), and (14). Use formula (12) to find the noise temperature of each of the components listed.

1. The noise temperature of the LNB is 35.4 K {T = (1.122-1)*290}
2. The noise temperature of the Waveguide is 13.7 K {T = (1.05-1)*290}
3. The noise temperature of the coax is 290 K {T = (2-1)*290}
4. The noise temperature of the receiver is 91,416 K {T = (316-1)*290}

Once the noise temperatures are known the equivalent noise temperature of the cascade consisting of the LNB, the coax, and the receiver can be calculated using (10).

$$T_e = 35.4 + 290/1{,}000{,}000 + 91{,}416/(1{,}000{,}000*0.5)$$
$$= 35.6 \text{ K}$$

This result shows that it is the equivalent noise temperature of the complete cascade is only slightly higher than the LNB by itself. This shows that it is the LNB that sets the overall noise temperature of the cascade; therefore, it should be chosen to have a low noise temperature and a high gain.

Now that the equivalent noise temperature of the cascade has been calculated, the system noise temperature is found using (14). Assume an antenna noise temperature, T_A, of 60 K.

$$T = 60/1.05 + (0.95/1.05) *290 + 35.6$$
$$= 355.1 \text{ K}$$

Converting to decibels using 10*log(T) the value is 25.5 dB-K. The value of G/T is arrived at by simply subtracting 25.5 dB-K from the antenna gain.

$$G/T = 40 - 25.5$$
$$= 14.5 \text{ dB/K}$$

Now proceed to calculate the downlink CNR using (16). Assume a clear sky (i.e., no rain attenuation) path loss.

$$CNR = 45 - 195.5 + 14.5 - 75.5 + 228.6$$
$$= 17 \text{ dB}$$

Since the receiver threshold is 7 dB, a received CNR of 17 dB implies that there is 10 dB of link margin in this link.

Link margin can be defined as the difference, in dB, between the received CNR of the satellite signal at the earth station and the threshold CNR of the earth station. The link margin is used to provide protection against a rain fade.

When designing a receiving earth station, one of the values that the designer must know is the required link availability. From that value a link attenuation, during a rain event, can be calculated using the Crane or CCIR Rain Models. The earth station must be built to have, as a minimum, a link margin equal to the value calculated by the Rain Model calculation. This margin can be achieved by increasing the size of the antenna, and therefore, its gain, or lowering the noise figure of the LNB, or both.

Appendix: Calculation of Rain Attenuation Using the Crane Rain Attenuation Model

This appendix will take the reader through the step-by-step process of calculating rain attenuation using the Crane Global Rain Attenuation Model.

To begin the earth station designer must know the desired link availability for the satellite link, and the location within the United States that the earth station will be constructed. Once those pieces of data are known the designer will use **Figure** 16-20 to determine the rain zone sub-region in which the earth station is located. Then find the point rain rate for the required availability and sub-region in Table 16-14.

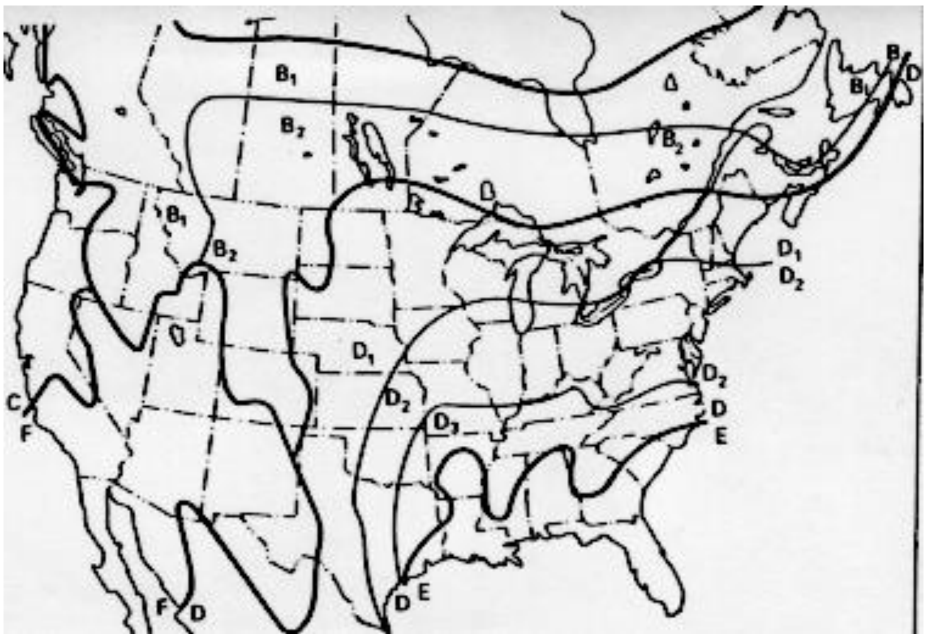


Figure 16-20: Crane Rain Attenuation Model Rain Zones
(from Radiowave Propagation in Satellite Communication by L. Ippolito)

Table 16-14: Point Rain Rate Distribution

Availability,%	A	B	C	D_1	D_2	D_3	E	F
99.999	24	54	80	90	102	127	164	66
99.998	24	40	62	72	86	107	144	51
99.995	19	26	41	50	64	81	117	34
99.990	15	19	28	37	49	63	98	23
99.980	12	14	18	27	35	48	77	14
99.950	8.0	9.5	11	16	22	31	52	8.0
99.900	6.5	6.8	7.2	11	15	22	35	5.5
99.800	4.0	4.8	4.8	7.5	9.5	14	21	3.8
99.500	2.5	2.7	2.8	4.0	5.2	7.0	8.5	2.4
99.000	1.7	1.8	1.9	2.2	3.0	4.0	4.0	1.7
98.000	1.1	1.2	1.2	1.3	1.8	2.5	2.0	1.1

The first calculation that needs to be performed is the projected surface path length using formula (A-1).

$$D = H(p) - G / \tan \theta \qquad (A-1)$$

where:

D = the projected surface path length

H(p) = 0 degree Isotherm height of the rain cell

G = the earth station ground elevation above mean sea level

θ = the earth station elevation angle

The 0 degree Isotherm Height can be found in Figure 16-21 for the particular link availability required and earth station latitude. The formula in (A-1) is applicable for earth station elevations great than 10 degrees. When the earth station elevation angle is less than 10 degrees, the curvature of the earth will effect the calculation, and the formula in (A-2) must be used.

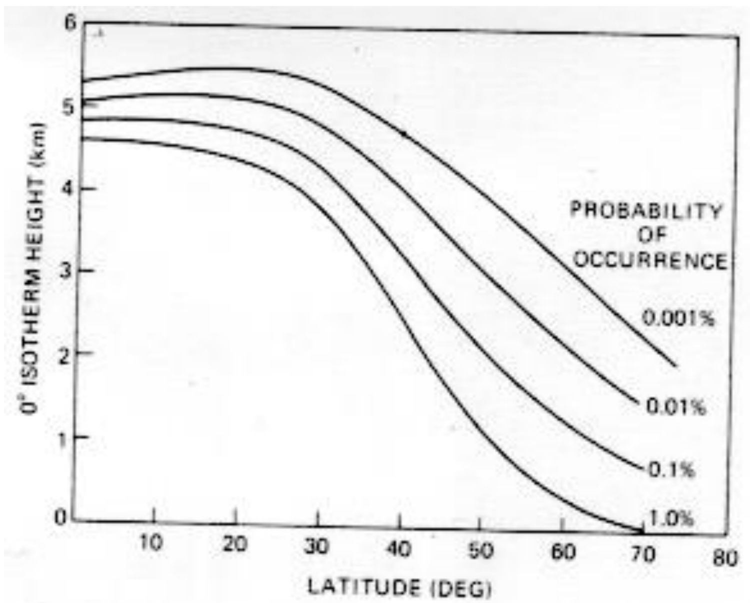


Figure 16-21: 0 Degree Isotherm Height from Radiowave Propagation in Satellite Communication by L. Ippolito

$$D = R \sin^{-1}[\cos \theta / H(p) + R(\text{square root}((G + R)^2 \sin^2 \theta + 2R(H(p) - G)$$
$$+ H(p)^2 - G^2) - (G + R)\sin \theta)] \qquad (A-2)$$

where

R = effective radius of the earth, 8,500 km

The next step involves using Table 16-15 to arrive at values for a and b given the frequency of operation and polarization of the earth station.

Frequency, GHz	a_h	a_v	b_v	b_h
1	0.0000387	0.0000352	0.912	0.880
2	0.000154	0.000138	0.963	0.923
4	0.000650	0.000591	1.121	1.075
6	0.00175	0.00155	1.308	1.265
7	0.00301	0.00265	1.332	1.312
8	0.00454	0.00395	1.327	1.310
10	0.0101	0.00887	1.276	1.264
12	0.0188	0.0168	1.217	1.200
15	0.0367	0.0335	1.154	1.128

Table 16-15: Specific Attenuation Coefficients

At this point the values for d,X,Y, and U are calculated using formulas (A-3 through A-6)

$$d = 3.8 - 0.6 \ln R_p \qquad \text{(A-3)}$$

$$X = 2.3\, R_p^{-0.17} \qquad \text{(A-4)}$$

$$Y = 0.026 - 0.03 \ln R_p \qquad \text{(A-5)}$$

$$U = \ln (X\, e^{Yd}) / d \qquad \text{(A-6)}$$

where

R_p = the rain rate found from Figure 16-20 in step one of this procedure

Once all of the preparatory steps are complete the value for the rain attenuation can be calculate choose the correct formula from (A-7a, A-7b, or A-7c).

If 0 < D <= d:

$$L_R = (aR_p^b / \cos\theta c * [(e^{Ubd} - 1)/Ub] \qquad \text{(A-7a)}$$

If d < D <= 22.5:

$$L_R = (aR_p^b / \cos\theta c * [(e^{Ubd} - 1)/Ub - (X^b e^{Ybd} / Yb) + (X^b e^{YbD} / Yb)] \qquad \text{(A-7b)}$$

If D > 22.5:

Calculate L_R with D = 22.5 but use a rain rate at a value of p’

P’ = p[22.5/D]

16.7 The Relationship of Cable System Carrier-to-Noise to the Baseband Video System Signal-to-Noise

The relationship of carrier-to-noise in a cable system to the signal-to-noise in the video section of a standard TV receiver will be discussed. This derivation helps to relate the use of carrier-to-noise measurements in a cable system to picture quality. Although a good carrier-to-noise measurement does not guarantee a good quality picture, a poor carrier-to-noise measurement guarantees a poor quality picture.

In the cable system:

$$\frac{\text{Signal}}{\text{Noise}}(\text{in dB}) = 10\log\left(\frac{S_{rf}}{N_{rf}}\right)$$

where $\mathbf{S_{rf}}$ is the signal power in the cable system:

$$S_{rf} = \frac{(0.707\ \text{V peak})^2}{75}$$

and $\mathbf{N_{rf}}$ is the noise power in a spectrum of "b" MHz bandwidth centered near the reference signal.

$$N_{rf} = b\left(\frac{V_n^{\ 2}}{75}\right)$$

where **b** is in megahertz (MHz) and $\left(\frac{V_n^{\ 2}}{75}\right)$ is the noise power in a 1 MHz band.

In the Video Section of a TV Receiver:

(Assuming the TV receiver does not itself contribute to the noise)

$$\frac{\text{Signal}}{\text{Noise}}(\text{in dB}) = 10\log\left(\frac{S_v}{N_v}\right)$$

where $\mathbf{S_v}$ is the power related to the peak-to-peak video voltage $(\mathbf{V_v})$ and **R** is the impedance of the video circuit being measured, then:

$$S_v = \frac{V_v^{\ 2}}{R}$$

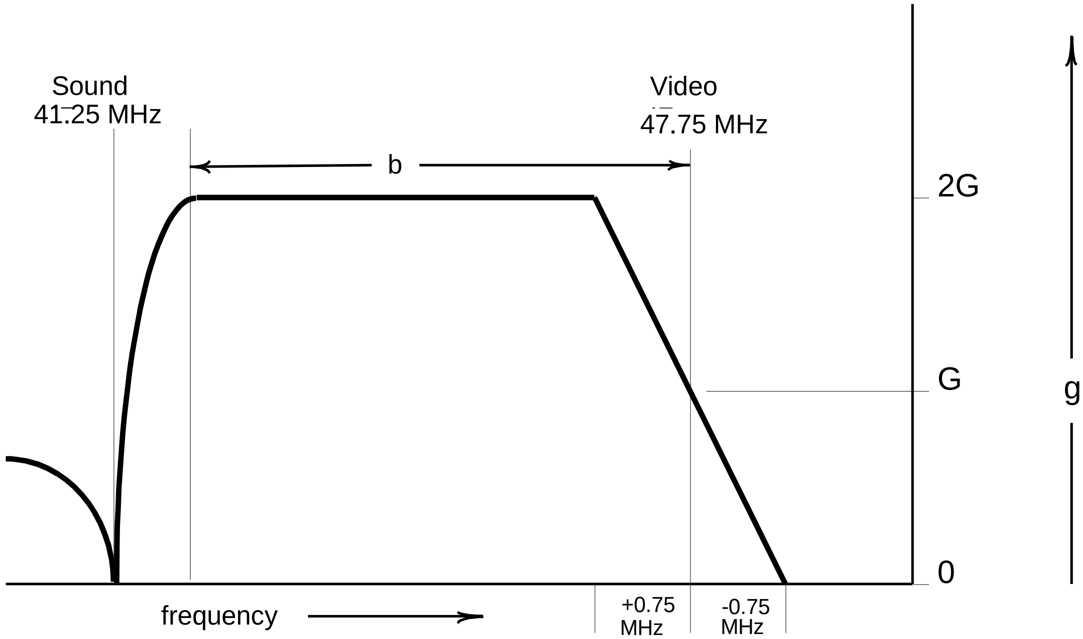


Figure 16-22: Typical TV IF Response

Figure 16-22 shows the response shape of a typical television receiver IF where **g** is the voltage gain of the receiver at any frequency within the band of the channel and **G** is the specific gain at the visual carrier frequency. Note that the spectrum appears reversed due to the IF frequency conversion. From Figure 16-22 it can be seen that:

$$g = 2G \quad (\text{from} + 0.75 \text{ MHz to b})$$

and

$$g = 2G(0.5 + 0.667\ f)$$

when **f** is any frequency expressed in MHz between -0.75 MHz and +0.75 MHz.

Since

$$S_v = \frac{V_v^{\ 2}}{R}$$

and we know that the peak-to-peak swing of the RF carrier is from $\mathbf{V_{peak}}$ to $\mathbf{0.125\ V_{peak}}$ and at video, $\mathbf{V_v}$ is from black level (0 IRE) to white level (100 IRE) which is 0.714 of the voltage from SYNC tip to white level, then

$$S_v = \frac{\left(0.714\ G\left(0.875\ V_{peak}\right)\right)^2}{R}$$

To find N_V (video noise power):

We know that at any frequency in the passband, the video noise voltage per MHz is equal to gV_n. From Figure 16-22 it can be seen that the video noise is equal to $2GV_n$ from +0.75 MHz to "**b**" and that video noise voltage is equal to **2G(0.5 + 0.667 f)** from -0.75 MHz to +0.75 MHz.

$$N_v = \frac{(2GV_n)^2}{R}(b - 0.75) + \int_{-0.75}^{+0.75} \frac{[2G(0.5 + 0.667f)V_n]^2 df}{R}$$

$$N_v = \frac{4G^2V_n^2}{R}(b - 0.75) + \int_{-0.75}^{+0.75} \frac{4G^2V_n^2(0.5 + 0.667f)^2}{R} df$$

$$N_v = \frac{4G^2V_n^2}{R}\left\{(b - 0.75) + \int_{-0.75}^{+0.75} (0.5 + 0.667f)^2 \, df\right\}$$

If we substitute

$$u = 0.5 + 0.667f$$
$$du = .667df$$
$$df = 1.5du$$

Then, our equation becomes

$$N_v = \frac{4G^2V_n^2}{R}\left\{(b - 0.75) + 1.5\int_{0}^{+1} u^2 \, du\right\}$$

This is an integral of the form

$$\int X^m dX = \frac{X^{m+1}}{m+1}$$

$$N_v = \frac{4G^2V_n^2}{R}\left\{(b - 0.75) + 1.5\left[\frac{u^3}{3}\right]_0^1\right\}$$

$$N_v = \frac{4G^2V_n^2}{R}\left\{(b - 0.75) + 1.5\left[\frac{1}{3} - 0\right]\right\}$$

$$N_v = \frac{4G^2V_n^2}{R}\left\{(b - 0.75) + 0.5\right\}$$

$$N_{v} = \frac{4G^2 V_n^{\,2}}{R}\left\{(b - 0.25)_{\text{MHz}}\right\}$$

The above says that the noise power contribution on the IF slope of a receiver from 0.75 MHz below the carrier to 0.75 MHz above the carrier is equivalent to the noise power in a one-half (0.5) MHz bandwidth occurring on the flat top portion of the receiver response.

To determine the ratio between the RF Carrier-to-Noise and the baseband video Signal-to-Noise, the following is given:

$$\text{Ratio} = \frac{S_{rf}/N_{rf}}{S_v/N_v} = \frac{S_{rf}}{N_{rf}} \bullet \frac{N_v}{S_v}$$

$$\text{Ratio} = \left(\frac{\frac{(0.707\,V_{peak})^2}{75}}{\frac{V_n^{\,2} b}{75}}\right)\left(\frac{\frac{4G^2 V_n^{\,2}(b-0.25)}{R}}{\frac{((0.714)(0.875\,GV_{peak}))^2}{R}}\right)$$

$$\text{Ratio} = \left(\frac{0.5\,V_{peak}^{\;\;2}}{V_n^{\,2} b}\right)\left(\frac{4G^2 V_n^{\,2}(b-0.25)}{0.39G^2 V_{peak}^{\;\;2}}\right)$$

$$\therefore = \frac{S/N_{rf}}{S/N_{video}} = \frac{5.12(b - 0.25)}{b}$$

Since the previous expressions are power ratios, we can take the log of both sides of the equation and express them as:

$$dB = 10 * \log\left(\frac{S/N_{rf}}{S/N_{video}}\right)$$

For this example let the video noise bandwidth **b** = 4 MHz, then:

$$S/N_{rf(dB)} - S/N_{video(dB)} = 10 * \log\left(5.12\left(\frac{3.75}{4}\right)\right) = 6.81\,dB$$

$$S/N_{video(dB\ -\ unweighted\ )} = S/N_{rf\ (dB)} - 6.81dB$$

These calculations have not allowed for the fact that baseband CCIR noise measurements use a weighting filter that increases the SNR at video by reducing the noise power by a factor of 6.81 dB.

Weighted $S/N_{Video\ (CCIR)} = S/N_{rf\ (dB)} + 6.81dB$

Weighted $S/N_{Video\ (CCIR)} = S/N_{rf\ (dB)} - 6.81dB + 6.81dB$

$$\therefore \text{Weighted } S/N_{Video\ (CCIR)} = S/N_{rf\ (dB)}$$

The fact that in the above equations, using a 4 MHz reference for both RF and video, the SNRs are equal is a mathematical coincidence.

Using different noise bandwidth references for RF and video allows expression of a more general equation.

$$S/N_{rf\ (dB)} \ - \ S/N_{video\ (dB)} \ = \ 10 * \log_{10} 5.12 \ + \ 10 * \log\left(\frac{W \ - \ 0.25}{b}\right)$$

$$S/N_{rf\ (dB)} \ - \ S/N_{video\ (dB)} \ = \ 7.09 \ + \ 10 * \log\left(\frac{W \ - \ 0.25}{b}\right)$$

where "**W**" is the video noise bandwidth reference and "**b**" is the RF noise bandwidth reference.

This equation is identical to the general equation used by J. J. Gibson of RCA in a report to "Members of Working Group on Noise - CTAC Panel I," dated January 20, 1974. It gives the relationship of $\mathbf{CNR}_{\mathrm{RF}}$ to un-weighted $\mathbf{SNR}_{\mathrm{Video}}$ for the various noise bandwidth considerations. One also must determine the weighting factor for the CCIR filter at the frequency of the video noise bandwidth reference.

As an example, let **W** = 3 MHz for video and "**b**" = 4 MHz for RF;

Then:

$$S/N_{RF} \ - S/N_{Video} \ = \ 7.09 \text{ dB} \ + \ 10 * \log\left(\frac{3 - 0.25}{4}\right)$$

$$S/N_{RF} \ - S/N_{Video\ (unweighte\ d)} \ = \ 7.09 \text{ dB} \ - \ 1.63 \ = \ 5.46 \text{ dB}$$

The CCIR weighting filter at 3 MHz reduces the noise power by 5.61 dB.

$$S/N_{CCIR\ (3\ MHz)} = S/N_{Video\ (unweighte\ d)} \ + \ 5.61 \text{dB}$$

$$S/N_{CCIR\ (3\ MHz)} \ = \ S/N_{RF\ 4\ MHz)} \ - \ 5.46 \text{ dB} \ + \ 5.61 \text{dB}$$

$$S/N_{CCIR\ (3\ MHz)} \ = \ S/N_{RF(4\ MHz)} \ + \ 0.15 \text{ dB}$$

In addition to this Section and the Gibson report, a 1974 NCTA Technical Paper entitled "The Relationship Between the NCTA, EIA, and CCIR Definitions of Signal-to-Noise Ratio" by T.M. Straus covers signal-to-noise.

16.8 Signal Constellations

Before defining parameters such as Error Vector Magnitude and Cluster Variance, it is useful to review some aspects of the received signal constellation. In order to understand how a QAM signal constellation is formed, it is helpful to examine the Quadrature Amplitude Modulation process. Figure 16-23 is a simplified block diagram of a QAM modulator.

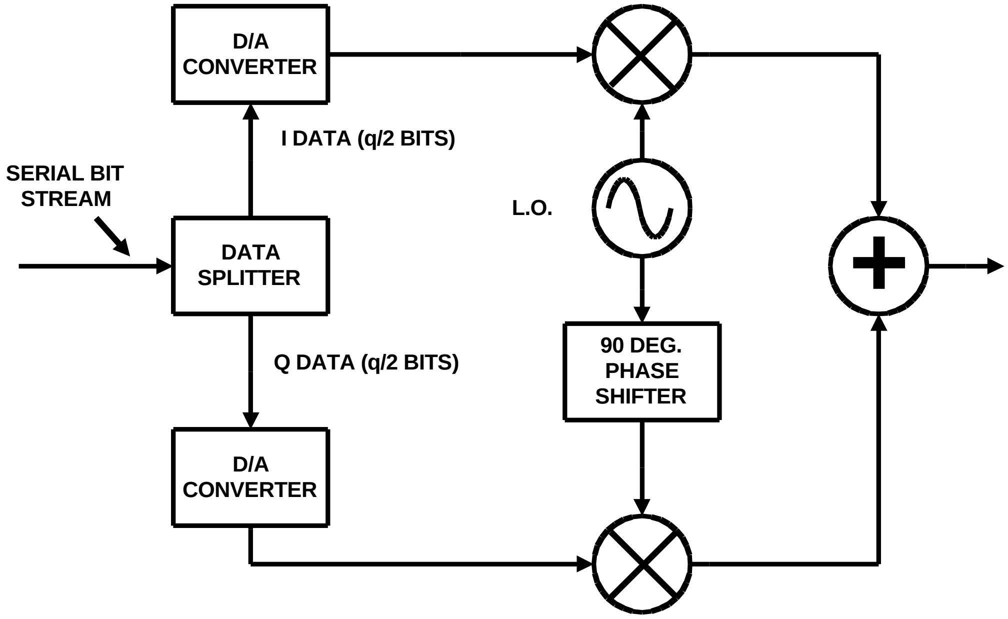


Figure 16-23: QAM Modulator Block Diagram

The QAM signal is generated by splitting the input serial data stream into in-phase and quadrature (I and Q) data streams. The I and Q streams are then partitioned into groups of q/2 bits and these groups are converted to $2^{q/2}$ analog levels. After filtering, the analog levels are modulated on in-phase and quadrature carriers, respectively. The modulated carriers are then summed to produce the QAM output.

Since each phase of the carrier can transmit $2^{q/2}$ data levels, there are $(2^{q/2})^2 = 2^q$ = M possible data combinations in an M-QAM signal. The phase and magnitude of any data point can be represented in polar or rectangular coordinates as a discrete point in the I-Q plane. If all possible combinations of I and Q data are plotted in phase space (i.e. – the data on the Q carrier is plotted against the data on the I carrier), the result is a constellation. The value of I is a plot based on the amplitude of the signal on the I channel at the time the signal modulates the I carrier. The corresponding value of Q is plotted based on the same criteria for the Q channel. In terms of a CRT display, the I signal drives one axis of the scope trace, the Q signal drives the other and the CRT beam is only turned on at the sampling instant.

An example of a 64-QAM constellation is shown in Figure 16-24.

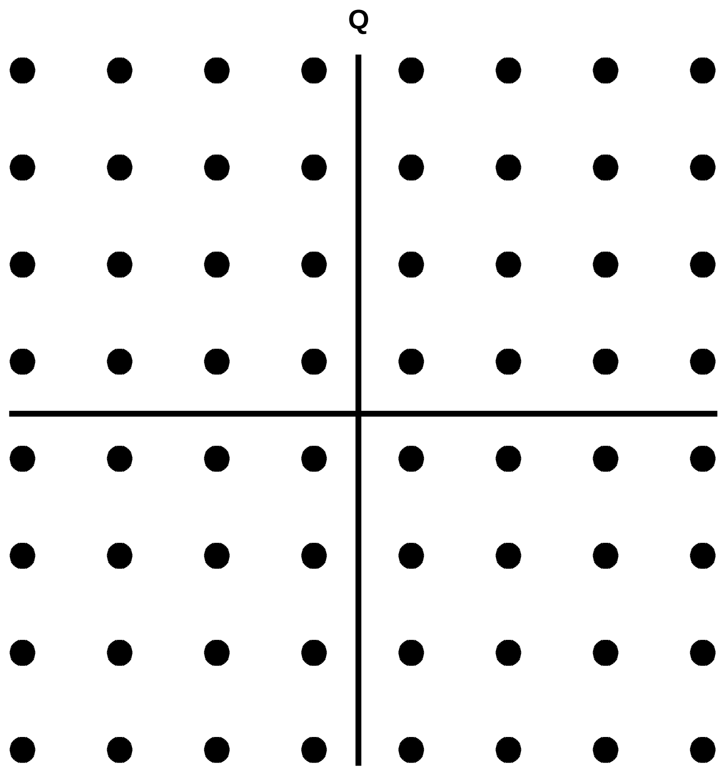


Figure 16-24: 64-QAM Constellation

If the modulation and carriers were perfect, the result would be a rectangular pattern of M dots on the screen, where M is the value of M-QAM. Since the real world is less than perfect, the dots cluster around an area. The way the dots cluster provides information about the modulation or demodulation quality.

16.9 Composite Measurements of Digital Performance

All three of these signal quality measurements Error Vector Magnitude (EVM), Modulation Error Ratio (MER) and Cluster Variance (CV) require the digital test receiver to recover the data bits transmitted. After the correct data bits are known, the perfect phase and amplitude variations of the carrier for the data transmitted can be reconstructed for any modulation format. The perfect modulation can then be subtracted from the actual modulation recovered to yield the error in phase and magnitude of the received transmission.

The signal flow for modulation error calculations is shown in Figure 16-25.

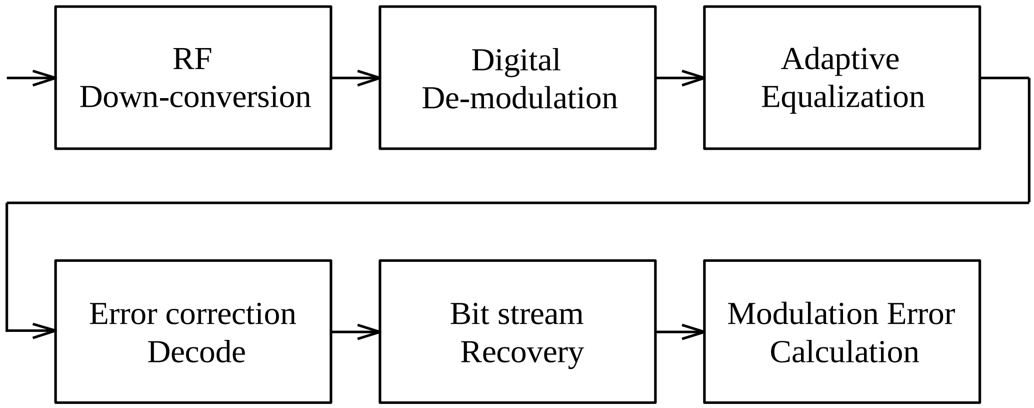


Figure 16-25: Modulation Error Calculation Signal Flow

The RF down-conversion can be done using set-top converters or higher performance spectrum analyzer conversions. The trade-offs, as usual, are cost and performance. Dynamic range for these measurements is limited by several factors. Higher quality down-conversion and demodulation increases the likelihood of accurate symbol recovery. Modulation error measurements reach a sensitivity limit when bit errors in the recovered data cause errors in the reference baseband modulation.

Gain flatness, group delay and phase noise in the data receiver's conversion chain will limit the dynamic range of the measurement at both ends. The adaptive equalization and FEC will be limited in their ability to correct for channel errors on impaired signals, thus causing increased difficulty in recovery of errored symbols. As an example, a tuner with poor performance may not be able to accurately recover the symbols from an impaired signal with 18 dB MER and may only be able to measure MER to 20 dB before losing lock. These conversion chain impairments will also mask high quality signals and limit the dynamic range on the high end of the measurements.

When evaluating the performance of a new measurement instrument, the dynamic range should be verified with known impaired signals and reference clean signals of different amplitudes.

16.10 Notes on Noise and Distortion Contribution from Non-Traditional Mechanisms

This discussion is intended for the practitioner concerned with greater precision in noise and distortion subjects and measurements for diagnostic, link budget analysis and verification, or other concerns not associated with end user satisfaction in daily practice.

The introduction of digital signal formats and fiber optic transportation systems into daily practice has brought forward some threshold-level contributions, significant at the level normally associated with traditional error budgets for both noise and intermodulation. These are discussed here in a cursory form; rigorous detail is found in the referenced literature.[1,2,3]

[1] Ciciora, Farmer, Large; Modern Cable Television Technology; Morgan Kaufman 1999.

[2] Agraval: Nonlinear Fiber Optics; Academic Press, 1989.

[3] Hamilton, Stoneback; The Effect of Digital Carriers on Analog CATV Distribution Systems; NCTA Technical Papers, 1993

Noise contribution can be found from Composite Intermodulation Noise (CIN) and Interferometric Intensity Noise (IIN).

CIN is the intermodulation component resulting from the presence of digitally modulated carriers, and is similar (in appearance and statistically) to random thermal noise for the digital modulation commonly found on cable systems. It is influenced by amplifier level and linearity, and number and location of digital carriers.

IIN is the resultant of post-detection scattered light in the optical path and is sensitive to splice integrity, fiber integrity, linewidth, and length and loss of the optical link.

Distortion Contribution can be found in CSO and CTB (to a lesser extent) measurements, resulting from nonlinearities in the fiber and interaction with other link components, and is sensitive to link length, linewidth, laser chirp, and fiber dispersion, in addition to traditional variables such as levels and loading.

16.11 Notes Concerning the Use of Scrambled Channels for Signal Level Measurement

We cannot recommend using scrambled channels for signal level measurement because the peak amplitude of a scrambled channel is not constant, being a function of the details of how a signal is scrambled and the video content. The best way to measure the amplitude of a scrambled signal is to turn off the scrambling during the level measurement. When that is not possible, such as during a 24 hour test, it may be possible to measure the peak level when observing over several scenes. Most analog scrambling systems will maintain a peak video level that is constant, provided that there is at least some black (usually) or some white (rarely) content to the picture. You will have to observe the reading long enough to capture the peak reading, which may depend on the picture content. Also, the relative level compared to adjacent unscrambled channels may or may not be read accurately when the picture is scrambled. This can be determined at the headend by measuring signal level with and without scrambling turned on. The issues are explained in this section.

To illustrate the problems, we shall use a simple 6 dB sync suppressed video signal, which is presented several ways. The video is assumed to be a ramp, which progresses from 0 IRE on the left side of the screen, to 100 IRE on the right. The last example will change the video to progress from 50 to 100 IRE, to illustrate the sensitivity of scrambled signals to picture content. We shall show that the peak amplitude of a scrambled video signal is not constant, and therefore, should not be used for distribution system AGC.

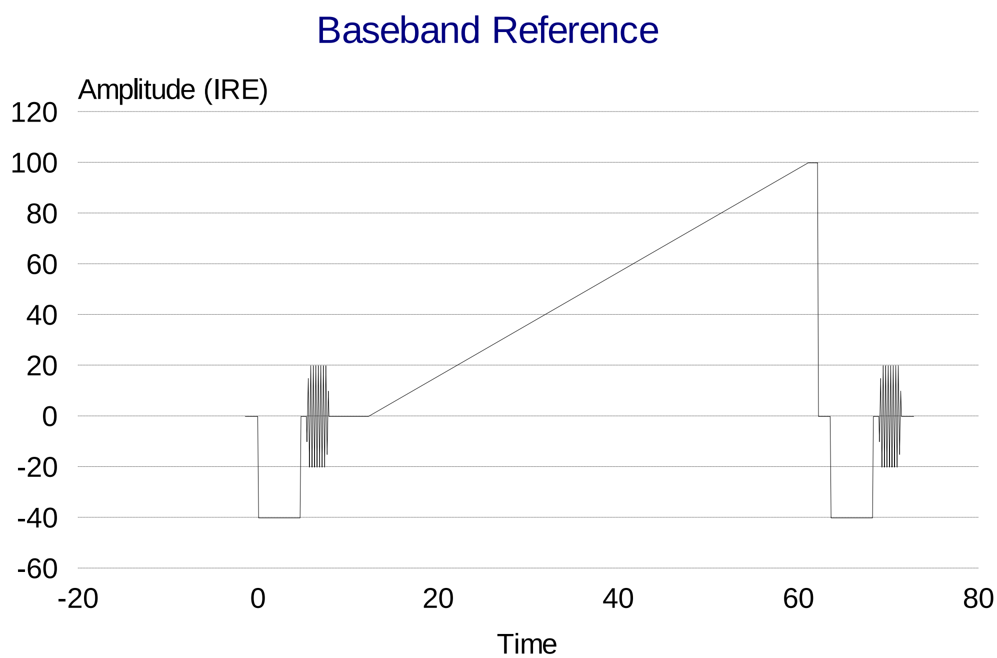


Figure 16-26: Reference Signal Used in All Except Last Example

Figure 16-26 illustrates the signal which is to be modulated onto a carrier. Note that it is a ramp progressing from 0 to 100 IRE. The remaining figures are based on this ramp, except for the last figure, which is similar, but based on a ramp that progress from 50 to 100 IRE to illustrate the effect of picture content on the signal. The figures below are all derived from this waveform by the exact mathematics of an NTSC signal, and the definition of how it is modulated onto a carrier. The scrambling assumed is precisely 6 dB sync suppression, in which the horizontal blanking interval (HBI) is attenuated by 6 dB. For the purpose of illustration, it is assumed that the suppression interval slightly overlaps the ramp. This is a rather common occurrence.

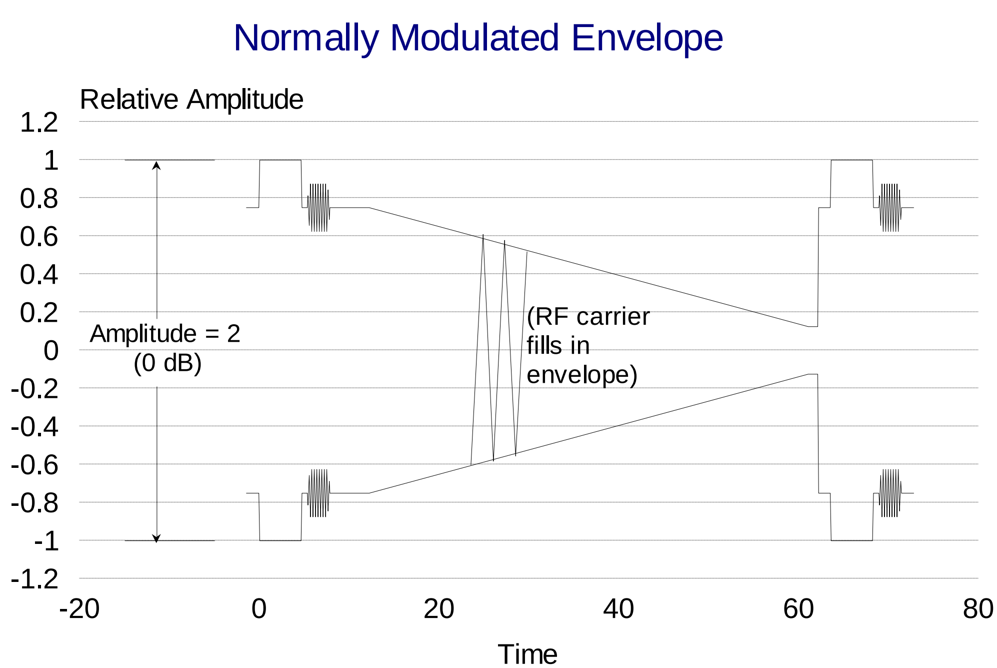


Figure 16-27: Waveform of Figure 16-26 Modulated Onto an RF Carrier

Figure 16-27 shows the waveform of Figure 16-26 modulated onto an RF carrier. The top and bottom of the modulated envelope are shown. The amplitude is normalized to a peak amplitude of 1, or a peak-to-peak amplitude of 2. This amplitude is taken as a 0 dB reference for subsequent illustrations.

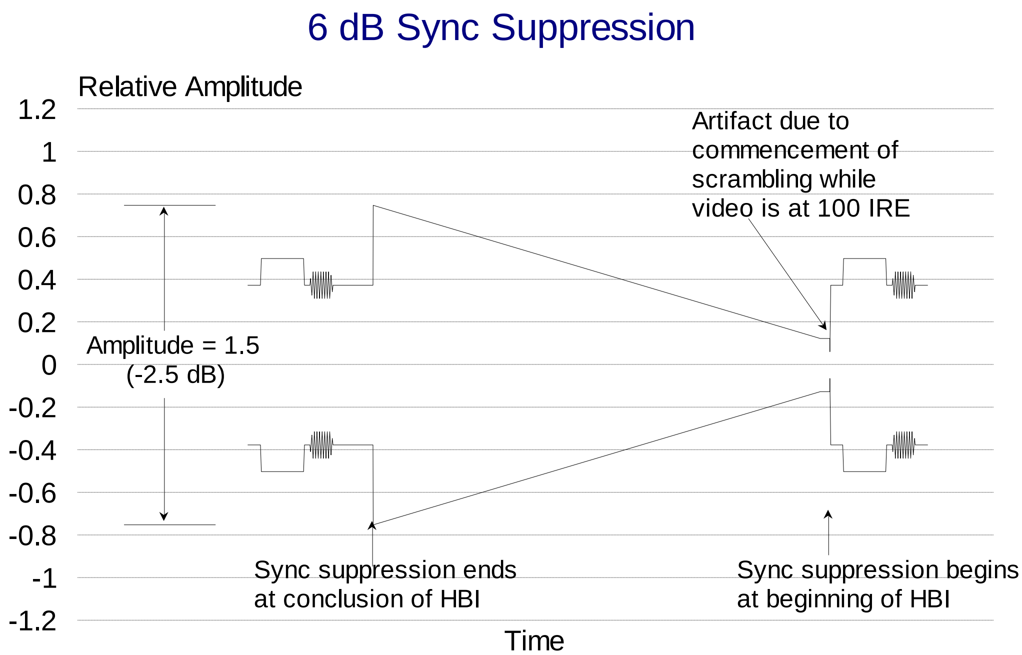


Figure 16-28: Signal of Figure 16-27 Sync Suppression Scrambled

Figure 16-28 illustrates the waveform of Figure 16-27 sync suppression scrambled. The system gain during the HBI is reduced by 6 dB; i.e., to one-half of the gain of Figure 16-27. The gain during the active video portion of the signal is not affected. Notice that the sync tip is no longer the highest portion of the envelope of the signal. The highest portion of the envelope now corresponds to the black portion of the scene. The black (really, blacker than black since the illustrated waveform ramps from 0 IRE, and black is 7.5 IRE) portion has a peak-to-peak amplitude in the example, of 1.5, which is -2.5 dB from the peak of Figure 16-27:

$$V_{scrambled} = 20\log\frac{1.5}{2} = -2.5\ dB_{unscrambled}$$

Thus, if the signal on which one is taking a reference for AGC, is scrambled, the AGC set point will drop by 2.5 dB, causing the amplitude of other signals on the plant to increase by 2.5 dB.

Since a sync suppressed scrambled signal has a peak-to-peak amplitude of less than the corresponding non-suppressed signal, some manufacturers take advantage of the reduced peak by amplifying the entire signal by 2.5 dB. This gets the peak signal level back to what it was without scrambling, so the load on the network is unchanged (assuming the operative signal level is the peak level rather than the average level).

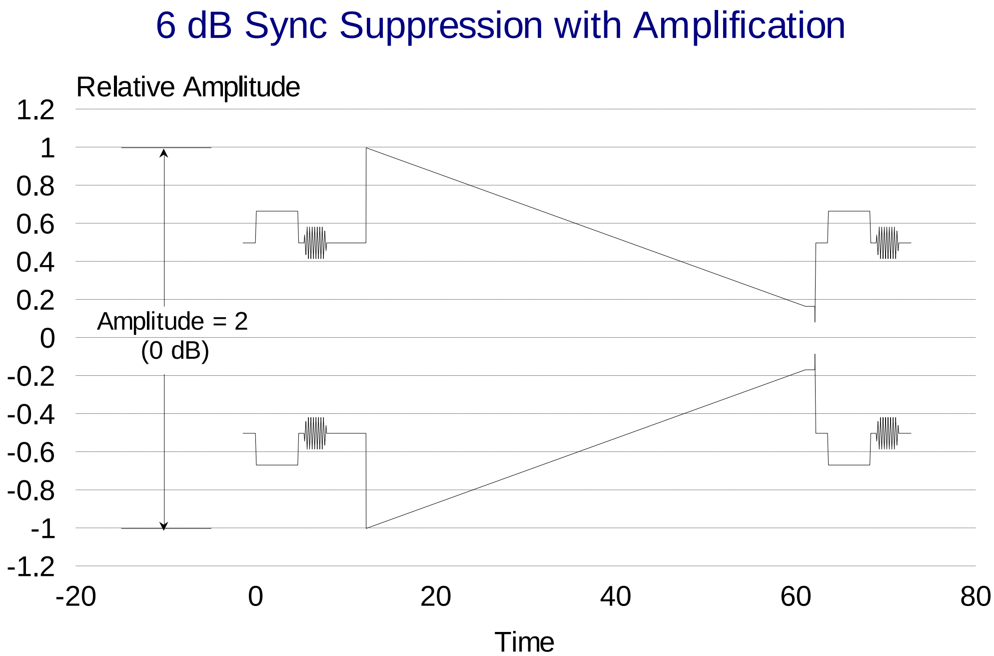


Figure 16-29: Same as Figure 16-28 Except Add Gain to Obtain the Same Peak Amplitude

Figure 16-29 is the same as Figure 16-28 except that enough gain (2.5 dB) has been added to the *entire* signal to bring the peak amplitude back to the same as with the unscrambled signal. Comparison between Figure 16-29 and Figure 16-27 show that the active video amplitude is 2.5 dB higher than it was in the unscrambled case, so the subscriber will see a 2.5 dB better carrier-to-noise ratio in the picture. The maximum envelope is a function of active video level rather than the sync tip, so if there is no black in the picture, the peak amplitude will again be less than that of the unscrambled signal. This amplification is frequently applied if the signal is to be sync suppressed all of the time. However, if the signal is to have a scrambling mode in which sync is not suppressed,[4] then the 2.5 dB gain cannot be added, because there will be some modes in which it is not possible to add the gain.

[4] This is common in modern systems employing changing scrambling states, which include video inversion and others. Sometimes the sync is suppressed, and sometimes it is not.

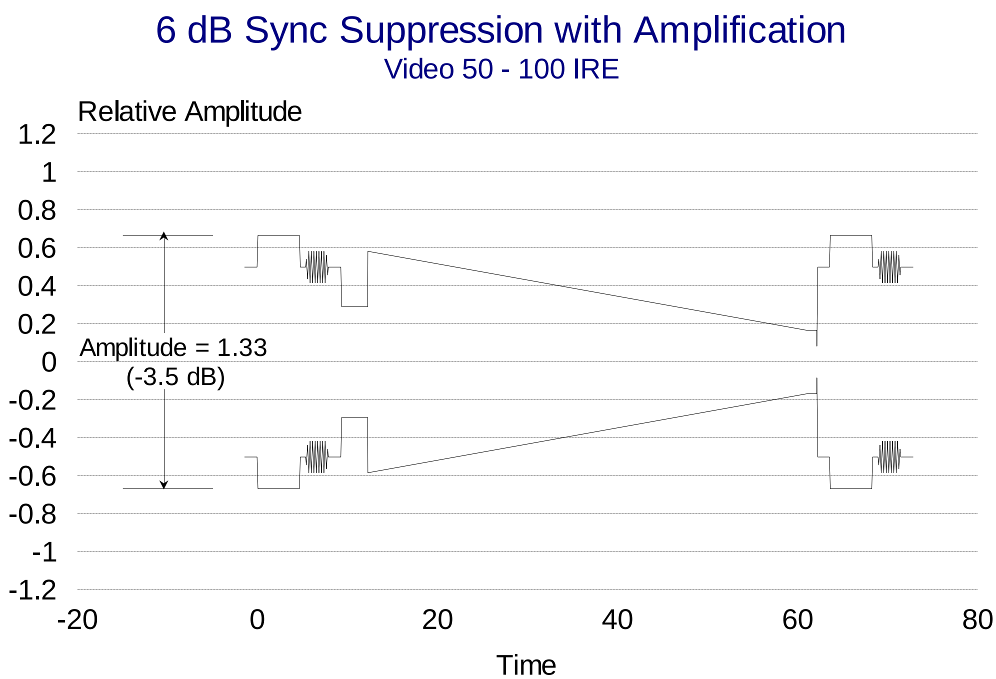


Figure 16-30: Same as Figure 16-29 Except Video Varies from 50 to 100 IRE Rather than 0 to 100 IRE

Figure 16-30 compared with Figure 16-29 illustrates the sensitivity of the peak level to the video waveform. The two figures are the same except in Figure 16-29 the video ramp varied from 0 to 100 IRE (blacker than black to white), while in Figure 16-30 it varied from 50 to 100 IRE (gray to white). Note that the sync tips and the white portion of the video are at the same amplitude as in Figure 16-29, but that the peak envelope of the signal is 3.5 dB lower, due to the fact that there is no black content in this picture. Also note that, if the picture contains no dark material, the amplitude of the signal is once again a function of the sync tip amplitude.

It is conceivable that the scrambled signal will have a lower amplitude for hours or months, if it is configured according to Figure 16-28, and for minutes (depending on the video content) if it is the situation is as depicted in Figure 16-30. The method illustrated in Figure 16-29 will be practiced in some instances but not in others.

We have shown several examples of scrambled signals which have differing peak amplitudes. Since a single channel level measurement responds to the peak amplitude of that channel, we have shown that use of a scrambled channel for signal measurement is unwise.

16.12 Power Supplies

History of Power Supplies

Historically, power has been supplied to amplifiers by superimposing a utility-frequency ac voltage onto the coaxial cable. Historically, at first 30 VAC was used, and then as amplifier bandwidths increased, with the corresponding increased demands for power, 60 VAC was used. Currently, 90 VAC is being implemented in HFC networks.

In early days of cable, the core technology for these power supplies was a step-down transformer reducing the voltage from the nominal utility supply of 117 VAC to the required cable operating voltage of 30 VAC or 60 VAC. To further enhance the supply of energy to the network, these step-down transformers are now commonly a self-regulating or ferroresonant type, which provides some protection against "brown-outs", voltage sags, and most importantly, transients. Self-regulating transformers are highly reliable and seldom, if ever, require maintenance.

Elements of a Self-Regulating or Ferroresonant Transformer

The self-regulating transformer uses two secondary windings to accomplish its task. As shown in Figure 16-31, there are two windings; a primary winding connected to the utility company's 117 VAC supply, a secondary output winding which has two taps: one secondary tap supplies the requisite voltage to the network (30, 60, 90 VAC) and another tap for the resonant capacitor essential to the operation of the system. The extra horizontal lines between the input and output windings represent the leakage flux path which is an integral feature of a self-regulating or ferroresonant transformer. The resonant capacitor and the leakage flux path create a resonant circuit with a resonant frequency near the primary supply frequency (in the U.S. – 60 Hz). Commonly, this ferroresonant transformer by itself is referred to as a non-standby power supply since any failure or significant disruption of the input voltage interrupts the power flow to the output; there is no standby function with this type of power supply.

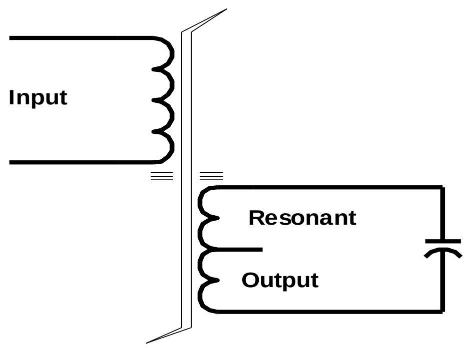


Figure 16-31: Self-Regulating or Ferroresonant Transformer Diagram.

This resonant circuit, leakage flux path, and a saturable-transformer core operate together to provide transient isolation and voltage regulation. Typical regulation is a nominal 1% or 2% output voltage change for a 15% input voltage change. Regulation is somewhat dependent on load: Greater regulation is possible with a lighter output load. Most network operators design for an 80% output load, which improves regulation and also allows the transformer to provide additional current as required during network startup. Efficiency approaches 90% at full output loading. Loading the ferroresonant transformer to a maximum of 80% of output rating allows the power supply to provide the extra current necessary during network startup when the constant-power amplifiers draw higher than normal currents while the network voltage is below its normal level.

Another benefit of the ferroresonant transformer is the output waveform distortion. The input waveform is typically a sine wave and the output is a quasi-square wave. A typical output waveform is shown in Figure 16-32. The waveform is modified by saturation of the transformer core for part of the sine wave. The waveform shape will vary depending upon load. The output waveform will gradually approach a sine wave as the output loading is increased to 110% load rating, as shown in Figure 16-33.

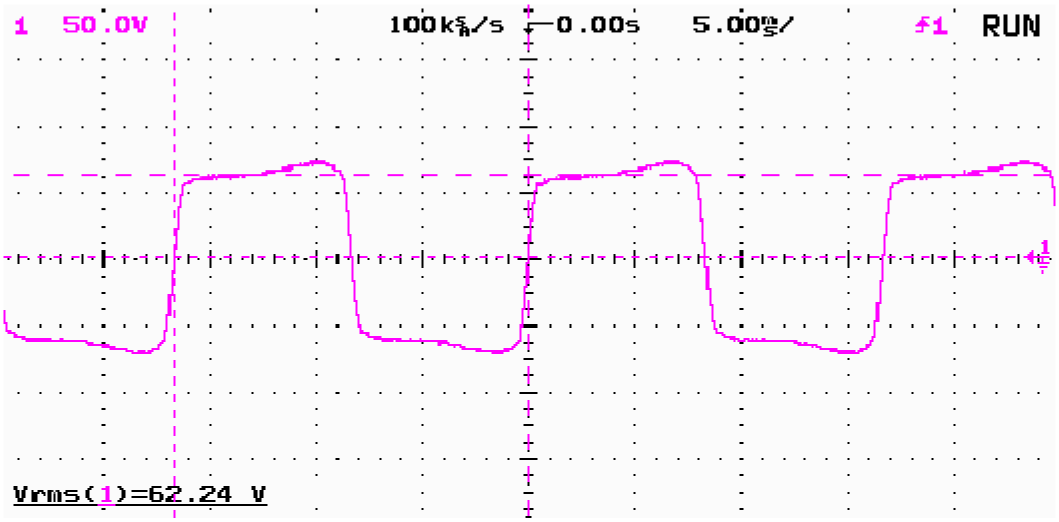


Figure 16-32: Typical Output Voltage Waveform from a Ferroresonant Transformer.

Because the waveshape is non-sinusoidal, for accurate measurements it is important to perform all voltage, current and power measurements using a true rms instrument. Many modern digital multimeters (DMM) read true rms, however it is an important specification to verify. Using a DMM which is not a true rms meter is probably the most frequent cause for unfounded concern for a power supply.

Because of its resonant nature, the self-regulating or ferroresonant transformer is not an effective performer when powered by an unregulated AC standby generator. Most small AC generators (< 10 KW) do not employ frequency and voltage regulation. Output voltage and voltage regulation cannot be guaranteed when the transformer is powered by a small AC gas generator.

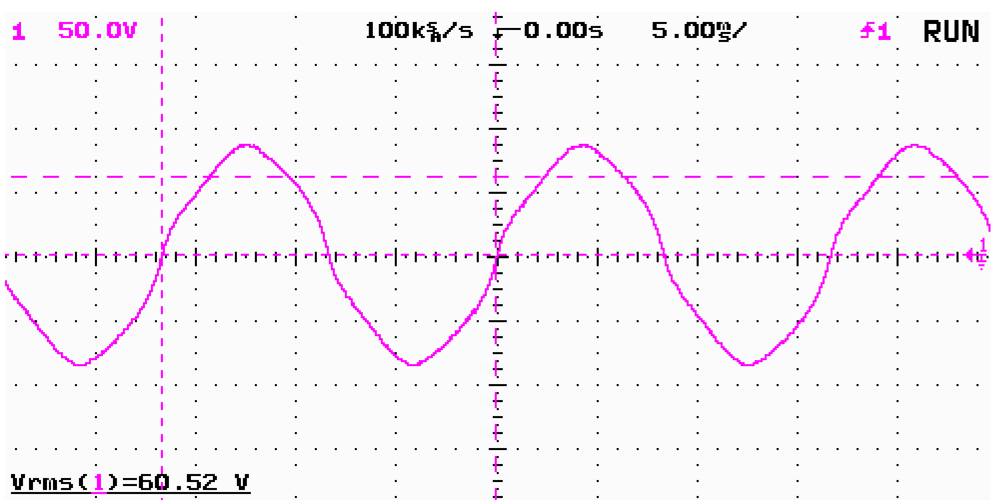


Figure 16-33: Typical Output Voltage Waveform from a Ferroresonant Transformer Under Overload Conditions.

Elements of a Ferroresonant-based UPS or Standby Power Supply

A standby power supply is created by the addition of a dc input to the self-regulating or ferroresonant system seen in Figure 16-31. Power during utility failures is obtained from batteries, typically, three 12 VDC batteries, or sometimes, for longer standby times, four 12-V batteries. Since all magnetic transformers require an ac input waveform for operation, a standby power supply must have an inverter which converts the dc into a suitable ac voltage. One common system involves adding another winding to produce the system seen in Figure 16-34, where the additional winding is referred to as the inverter winding.

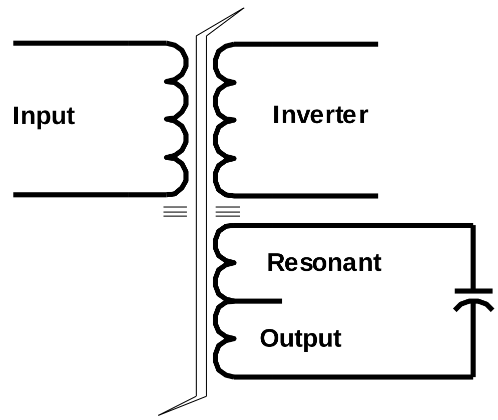


Figure 16-34: Ferroresonant-Based Standby Power Supply.

The inverter, which converts the dc power from the 36 VDC battery string or 48 VDC string, coordinates with a control system to provide ac power to the inverter winding during a utility outage. During times when the utility is functioning normally, the standby power supply must charge the battery string using power obtained from the utility. Typically, a charged, three-battery string will provide about 90 minutes standby time, and from a fully discharged state, require about 24 hours of normal utility power to fully charge.

Elements of a Switch-Mode Standby Power Supply or UPS

Other standby power supplies do not use a ferroresonant transformer, but instead use a high-frequency switching pattern to create the quasi-square wave output waveform. See Figure 16-35. These switch-mode supplies obtain power from a 36 VDC or 48 VDC battery string during times of utility outages. Batteries are charged during normal utility operation, and these batteries provide the power for standby operation when the utility fails.

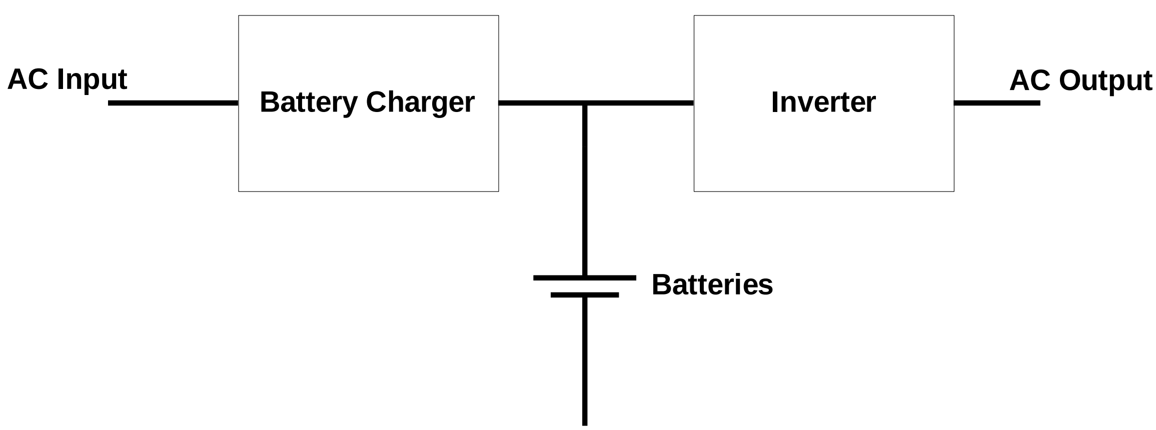


Figure 16-35: Example Block Diagram for a Switch-Mode Standby Power Supply

Grounding

All power supplies must be grounded to allow the protection circuits to function properly. Some power supplies have an isolated output and thus both the output and input must be grounded. Electrical codes must be considered when the ground configuration is designed.

Batteries for Cable Television Power Systems

The ideal battery for standby power systems is a maintenance-free battery which would not need replacement for over 20 years. Unfortunately, no such battery exists with presently known technology. In actual practice, the sealed valve regulated gelled lead-acid batteries are most frequently used in today's networks.

Battery life is affected by many things, including battery temperature, charging technique and discharge cycles. Temperature of the battery is often the most significant factor in battery life expectancy. Enclosures and power supply positioning are critical to producing a system with the lowest possible battery temperatures. Charging techniques vary among power supply manufactures and also among type of power supply. Most often, however, the charger has a initial, fast recharge cycle, followed by a time-limited cycle which helps equalize the charge among a string of three or four batteries, and a float charge. Float charging is the most important cycle since over 90% of the time, the power supply and batteries are in this state. Temperature compensation is an essential factor to preventing thermal runaway of the batteries. Thermal runaway occurs when the battery temperature rises and the charging current also rises. In this undesired situation, the battery temperature will continue to rise.

Batteries can also be abused by either chronically under or over-charging batteries which can lead to significant reduction in battery life expectancies. Manufacturers charging recommendations are key in preserving the life of a battery. It is critical that a charger be used that limits the highest voltage to no more than 14.1 volts and no less than 13.8 volts per battery at 68 °F (20 °C.) Constant current chargers should never be used on gel cell batteries. Discharge should never extend below 10.2 volts per battery, and if the battery voltage does fall below this level, irreversible damage to the batteries can occur.

16.13 Considerations for Digital Launch

This section presents a series of do’s and don'ts for introducing digital signals into an existing analog cable system.

Make certain that digital satellite signal levels and carrier/noise ratios are within the manufacturer’s specified limits for error free reception.

Check for correct terminations and grounding at the headend before putting the digital signals on the system.

Set the digital signal levels correctly. Digital signal levels should be set relative to peak analog carrier values.

When adding digital signals to an analog system, make sure that the additional signal power does not overdrive the lasers in fiber optic transmitters. If necessary, re-adjust input levels to the transmitters to maintain correct total power.

Make certain that a non-interfering sweep is used in the digital portion of the spectrum. Use of a high-level sweep will cause errors in the digital signal.

Do not operate a digital signal next to a trapped channel. The group delay in the trap can produce distortion in the adjacent channel. Although this distortion may not affect an adjacent analog signal, it can produce errors in a digital channel.

Do not operate digital signals in the rolloff portion of the spectrum. The rolloff is a part of the system where amplifier group delay and return loss performance is not guaranteed. Although satisfactory results may be obtained on some parts of the system, there may be locations where reflections and group delay prevent proper operation of digital signals.

16.14 Frequency Response and Digital Signal Processing Fundamentals

The Fourier Transform for Continuous-Time Systems

An analog or continuous time signal may be described by a function of time *f(t)*. Under general conditions met in most engineering practice, we may take the Fourier Transform of *f(t)* defined as

$$F(\omega) = \int_{-\infty}^{\infty} f(t)\, e^{-j\omega t}\, dt \qquad (1)$$

evaluated for all real values of ω (frequency in radians per second), which is often referred to as the spectrum of the signal *f(t)* and $|F(\omega)|^2$ is called the power spectrum of the signal. There is an inverse relationship to (1) given by

$$f(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} F(\omega)\, e^{-j\omega t}\, d\omega \qquad (2)$$

where *f(t)* and *F(ω)* form a Fourier Transform pair.

The output *y(t)* of a linear time-invariant system (e.g., a filter) is related to the input *x(t)* by the superposition integral

$$y(t) = \int_{-\infty}^{\infty} h(\tau)\, x(t-\tau)\, d\tau \qquad (3a)$$

where *h(t)* is the unit impulse response of the system. With a change in variable $\tau' = t - \tau$ this can also be expressed as

$$y(t) = \int_{-\infty}^{\infty} x(\tau')\, h(t-\tau')\, d\tau' \qquad (3b)$$

Equation (3) is commonly called the convolution of the functions *x(t)* and *h(t)* often denoted by $x(t) \circledast h(t)$.

By a well known theorem in Fourier integrals

$$Y(\omega) = H(\omega)\, X(\omega) \qquad (4)$$

where *Y(ω), H(ω),* and *X(ω)* are the Fourier Transforms of *y(t), h(t),* and *x(t)* respectively.

The Z Transform for Discrete-Time Systems

Analogously, a digital or discrete time (sampled) signal may be described by a sequence $\{x_n\}$. The output $\{y_n\}$ of a linear time-invariant discrete-time system (e.g., a digital filter) is related to the input $\{x_n\}$ by the convolution sum

$$y_n = \sum_{m=-\infty}^{\infty} x_m \, h_{n-m} \qquad (5a)$$

or equivalently by substituting m = n - k

$$y_n = \sum_{k=-\infty}^{\infty} h_k \, x_{n-k} \qquad (5b)$$

The Z Transform of a sequence $\{x_n\}$ is defined as

$$Z[x_n] = X(z) = \sum_{n=-\infty}^{\infty} x_n \, z^{-n} \qquad (6)$$

where z is a complex variable and plays a similar role to the variable s in the Laplace Transform. An elementary property of the Z Transform is the shift property

$$Z[x_{n+k}] = \sum_{n=-\infty}^{\infty} x_{n+k} \, z^{-n} = \sum_{m=-\infty}^{\infty} x_m \, z^{-(m-k)} = z^k X(z) \qquad (7)$$

A linear time invariant discrete time system described by the relationship

$$y_n = \sum_{k=0}^{M} a_k \, x_{n-k} - \sum_{k=1}^{L} b_k \, y_{n-k} \qquad (8)$$

is commonly called a digital filter, an example of which is depicted in Figure 16-7: Finite Impulse Response (FIR) Filter, which is also the basis of the DFE depicted in Figure 16-6. Such a digital filter is recursive (the output contains weighted previous outputs in addition to weighted previous inputs), which is characterized by an infinite impulse response and is known as an Infinite Impulse Response (IIR) filter. The impulse response is given by substituting the unit impulse $x_{n-k} = 1$ for n = k and 0 otherwise in (8). For the special case where the digital filter shown in Figure 16-6 has recursive output coefficients $b_1, \dots, b_L$ that are all zero, the filter is non-recursive (the output contains a finite number of weighted previous inputs only). As a result, the impulse response of the filter is finite and the filter is known as a Finite Impulse Response (FIR) filter as shown in Figure 16-7.

Taking the Z Transform of equation (8) yields

$$\sum_{n=-\infty}^{\infty} y_n \, z^{-n} = \sum_{n=-\infty}^{\infty} \left[\sum_{k=0}^{M} a_k \, x_{n-k} - \sum_{k=1}^{L} b_k \, y_{n-k} \right] z^{-n} \qquad (9)$$

The left side is simply $Y(z)$, the Z Transform of $\{y_n\}$. The right side can be evaluated using the shift property (7). The Z Transform of $\{x_{n-k}\}$ is $z^{-k} X(z)$ and that of $\{y_{n-k}\}$ is $z^{-k} Y(z)$. Thus (9) becomes

$$Y(z) = \left[\sum_{k=0}^{M} a_k z^{-k}\right] X(z) - \left[\sum_{k=1}^{L} b_k z^{-k}\right] Y(z) \qquad (10)$$

This can be more conveniently expressed as

$$Y(z) = H(z)\, X(z) \qquad (11)$$

where the system transfer function $H(z)$ is given by

$$H(z) = \frac{\sum_{k=0}^{M} a_k z^{-k}}{1 + \sum_{k=1}^{L} b_k z^{-k}} \qquad (12)$$

Frequency Response of Discrete-Time Systems

The frequency response of the digital filter is obtained by evaluating $H(z)$ along the unit circle in the z-plane by setting $z = e^{j\omega T}$ as

$$H\left(e^{j\omega T}\right) = \frac{\sum_{k=0}^{M} a_k e^{-j\omega kT}}{1 + \sum_{k=1}^{L} b_k e^{-j\omega kT}} \qquad (13)$$

where ω denotes the frequency in radians per second (equal to 2πf where f is the frequency in Hz) and T denotes the sampling period of the digital filter in seconds. If the filter $H(e^{j\omega T})$ equalizes the frequency response of the channel, then the channel frequency response $G(e^{j\omega T})$ is given as the inverse of the equalizer response as

$$G\left(e^{j\omega T}\right) = 1/H\left(e^{j\omega T}\right) \qquad (14)$$

The amplitude response of the channel is given by the magnitude of the frequency response as

$$Mag\left[G\left(e^{j\omega T}\right)\right] = \sqrt{(Re[G(e^{j\omega T})]^2 + Im[G(e^{j\omega T})]^2} \qquad (15)$$

and the phase is given by

$$Phase\left[G\left(e^{j\omega T}\right)\right] = \tan^{-1}\frac{Im\left[G\left(e^{j\omega T}\right)\right]}{Re\left[G\left(e^{j\omega T}\right)\right]} \qquad (16)$$

Finally, the group delay GD of the channel is given by

$$GD\left[G\left(e^{j\omega T}\right)\right] = -\frac{d}{d\omega} Phase\left[G\left(e^{j\omega T}\right)\right] \qquad (17)$$

In terms of the frequency f (in Hz), (15), (16), and (17) are given by

$$Mag\left[G\left(e^{j2\pi fT}\right)\right] = \sqrt{(Re[G(e^{j2\pi fT})]^2 + Im[G(e^{j2\pi fT})]^2} \qquad (18)$$

the phase is given by

$$Phase\left[G\left(e^{j2\pi fT}\right)\right] = \tan^{-1}\frac{Im\left[G\left(e^{j2\pi fT}\right)\right]}{Re\left[G\left(e^{j2\pi fT}\right)\right]} \qquad (19)$$

and the group delay GD of the channel is given by

$$GD\left[G\left(e^{j2\pi fT}\right)\right] = -\frac{1}{2\pi}\frac{d}{df}\, Phase\left[G\left(e^{j2\pi fT}\right)\right] \qquad (20)$$

Frequency Response and the Discrete Fourier Transform

Note that the frequency response of transfer function (13) was obtained by evaluating H(z), the Z Transform of h_n , around the unit circle in the z-plane by setting $z = e^{j\omega T}$. If the sequence $\{x_n\}$ in the Z Transform is of finite duration with N samples, then the Z Transform of the finite sequence is

$$Z[x_n] = X(z) = \sum_{n=0}^{N-1} x_n\, z^{-n} \qquad (21)$$

Evaluating (21) at the point $z = e^{j\left(\frac{2\pi}{N}\right)k}$ on the unit circle in the z-plane gives

$$X\left(e^{j\left(\frac{2\pi}{N}\right)k}\right) = \sum_{n=0}^{N-1} x_n\, e^{-j\left(\frac{2\pi}{N}\right)kn} \triangleq X_p(k) \qquad (22)$$

This is the definition of the N-point Discrete Fourier Transform (DFT) with coefficients $X_p(k)$ of the periodic sequence

$$x_p(n) = x_p(n + mN) = x_n\,;\; n = 0, 1, \dots, N-1;\; m = 0, \pm 1, \pm 2, \dots \qquad (23)$$

with a period of N samples.

Thus the Z Transform of an N-point sequence evaluated at N evenly spaced points around the unit circle equals the DFT of the N-point periodic sequence. Here the normalized frequency $\Omega_k = \frac{2\pi k}{N}$ is sampled in N points in a period of 2π, or equivalently, $F_k = \frac{k}{N}$ with a period of 1. Thus the number of points N in one period of the sequence $x_p(n)$ determines the normalized resolution in the DFT frequency domain $\Delta\Omega = \frac{2\pi}{N}$ or $\Delta F = \frac{1}{N}$.

By substituting nT for n and $\omega_k = \frac{2\pi k}{NT}$ for Ω_k in (22), we obtain the frequency response $X_p(k/NT)$ in a period of 2π/T of the periodic sampled sequence $x_p(nT)$ in (23) with a period of NT which scales the normalized frequency by the discrete-sampling period T. The frequency resolution becomes $\Delta\omega = \frac{2\pi}{NT}$ or $\Delta f = \frac{1}{NT}$.

The value of N should be chosen to be a large enough multiple of the digital filter sampling period T where the time sequence period NT spans a time encompassing the significant portion of the infinite impulse response of the digital filter so as to prevent aliasing in the DFT.

Consider the following example of a QAM signal transmitted through a channel that is input into an equalizer in a receiver of the form depicted in Figure 16-8. Consider the following example of the calculation of the channel amplitude and group delay response from the equalizer filter coefficients. The channel combines the main (desired) signal with four micro-reflections given in the following table:

Table 16-16: Amplitude and Delay of Main Signal with Four Micro-Reflections

Signal:	Main	Micro-Reflections			
Amplitude (dB):	0	-10	-15	-20	-30
Delay (µs):	0	0.5	1.0	1.5	4.5

A spectrum analyzer trace of the 256-QAM signal transmitted through this linearly distorted channel is shown below:

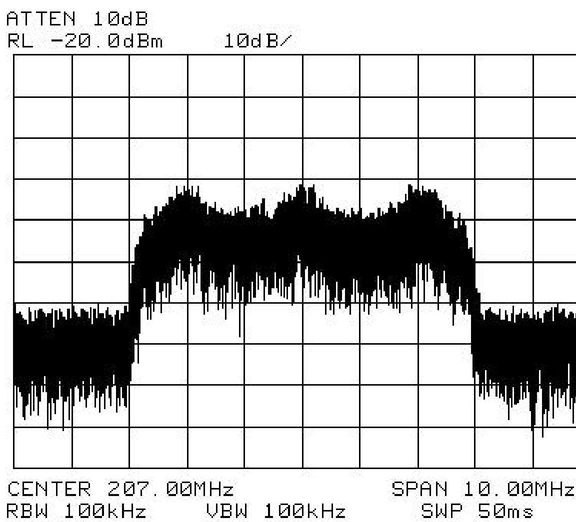


Figure 16-36: Spectrum of 256-QAM Signal Transmitted Through Linearly Distorted Channel

The receiver equalizer tap delays are equal to the QAM symbol duration T which is the inverse of the symbol rate or 1/T. The symbol rate for ITU-T J.83 Annex B 256-QAM is 5,360,537 Hz which yields a symbol duration of T = 0.186548 µs.

The receiver equalizer coefficients (normalized for unity DC gain) for this linearly distorted channel are:

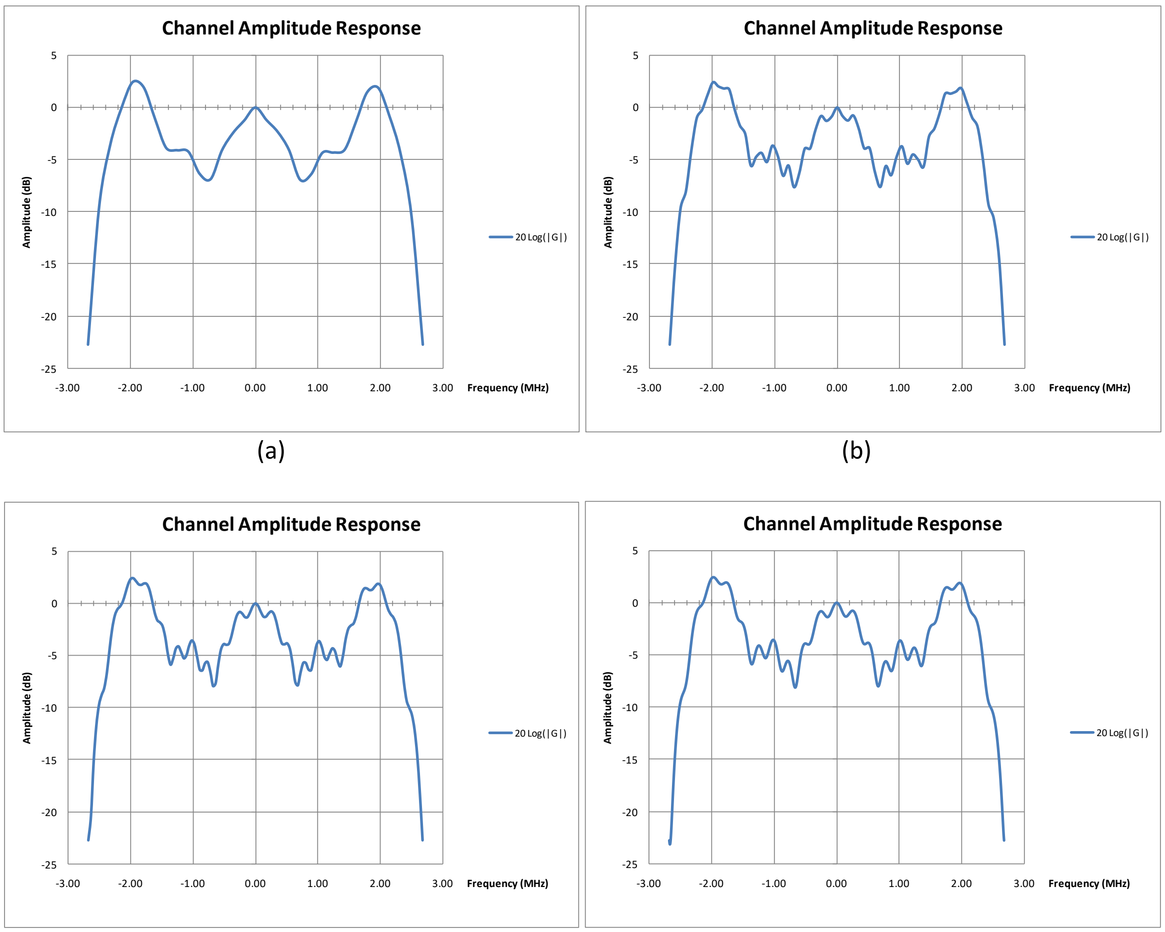
FFE Coefficients Position:	0	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20
Real:	-25	197	20	-25	374	-202	330	-335	728	-1088	1437	-1752	2455	-3125	4360	-5517	8169	-12027	22474	-60967	544248
Imaginary:	134	45	134	-43	134	1	178	-88	311	-176	488	488	665	-752	1241	-1284	1905	-2479	3721	-5978	1
DFE Coefficients Position:	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21
Real:	-25651	27231	45212	-20203	33431	6547	-2886	18948	-1823	613	-362	258	-317	170	-185	125	-185	170	-185	170	-229
Imaginary:	-1020	1106	-887	442	-134	-90	43	43	-134	87	-90	-90	-46	-1	-46	-1	-46	-1	-46	-1	-90
DFE Coefficients Position:	22	23	24	25	26	27	28	29	30	31	32	33	34	35	36						
Real:	303	-583	5573	170	-185	170	-140	125	-52	170	-52	37	37	37	-7						
Imaginary:	-1	43	43	-1	43	-46	-1	-1	43	43	-46	-46	-46	-1	-1						

The FFE and DFE coefficients at position n are equal to a_n and b_n respectively in (8) and Figure 16-8.

The frequency response of the equalizer $H(e^{j\omega T})$ is obtained by inserting these coefficient values in (13) and evaluating at the point $z = e^{j\left(\frac{2\pi}{N}\right)k}$ as done in (22) and then substituting nT for n and $\omega_k =$

$\frac{2\pi k}{NT}$ for Ω_k to obtain the DFT coefficients $H_p(k/NT)$ as previously discussed. The frequency response of the channel is simply the inverse given by (14). The results of this calculation for the channel frequency response and evaluating the magnitude (18) using an N-point DFT with values of N equal to 32, 64, 128, and 1024 are shown below:

The frequency response of the equalizer $H(e^{j\omega T})$ is obtained by inserting these coefficient values in (13) and evaluating at the point $z = e^{j\left(\frac{2\pi}{N}\right)k}$ as done in (22) and then substituting nT for n and $\omega_k = \frac{2\pi k}{NT}$ for Ω_k to obtain the DFT coefficients $H_p(k/NT)$ as previously discussed. The frequency response of the channel is simply the inverse given by (14). The results of this calculation for the channel frequency response and evaluating the magnitude (18) using an N-point DFT with values of N equal to 32, 64, 128, and 1024 are shown below:



(c) (d)

Figure 16-37: Channel Amplitude Response using an N-point DFT for (a) N=32, (b) N=64, (c) N=128, and (d) N=1024.

Note that using a DFT size of less than 128 results in aliasing that distorts the response and using 128 points or more results in no difference in the response. Thus choosing a 128 point DFT is sufficient for evaluating this equalizer filter to calculate the channel frequency response shown in Figure 16-38:

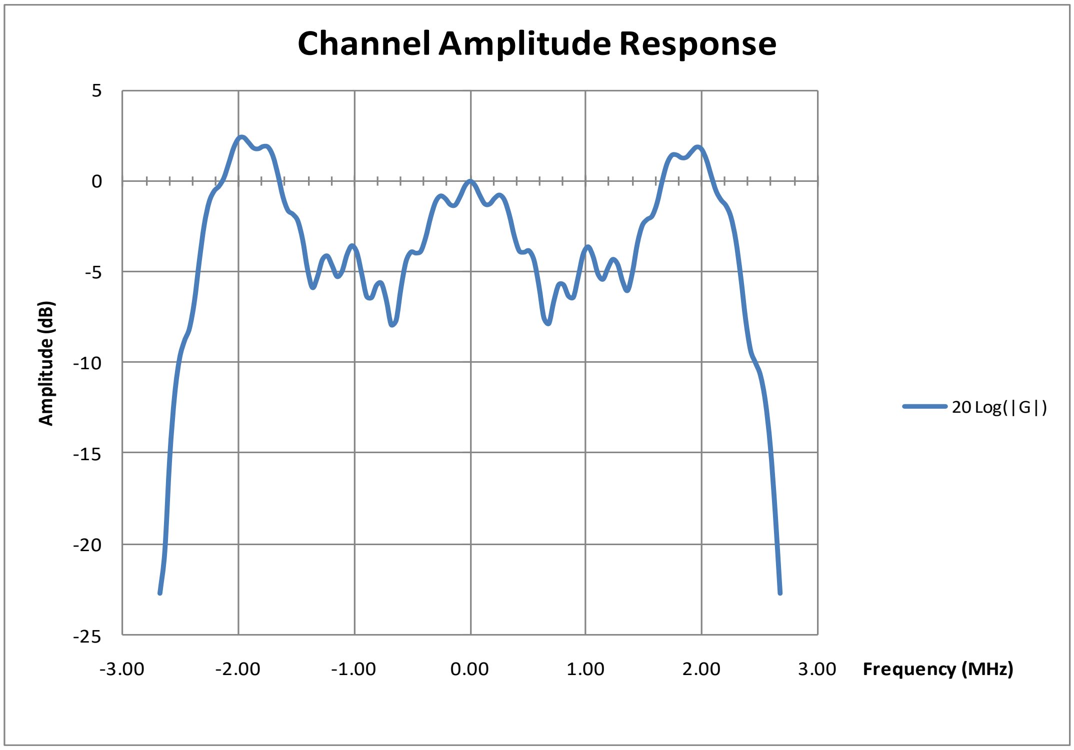


Figure 16-38: Channel Amplitude Response using 128 point DFT

Similarly, the Group Delay Response using (19) and (20) is shown in Figure 16-39:

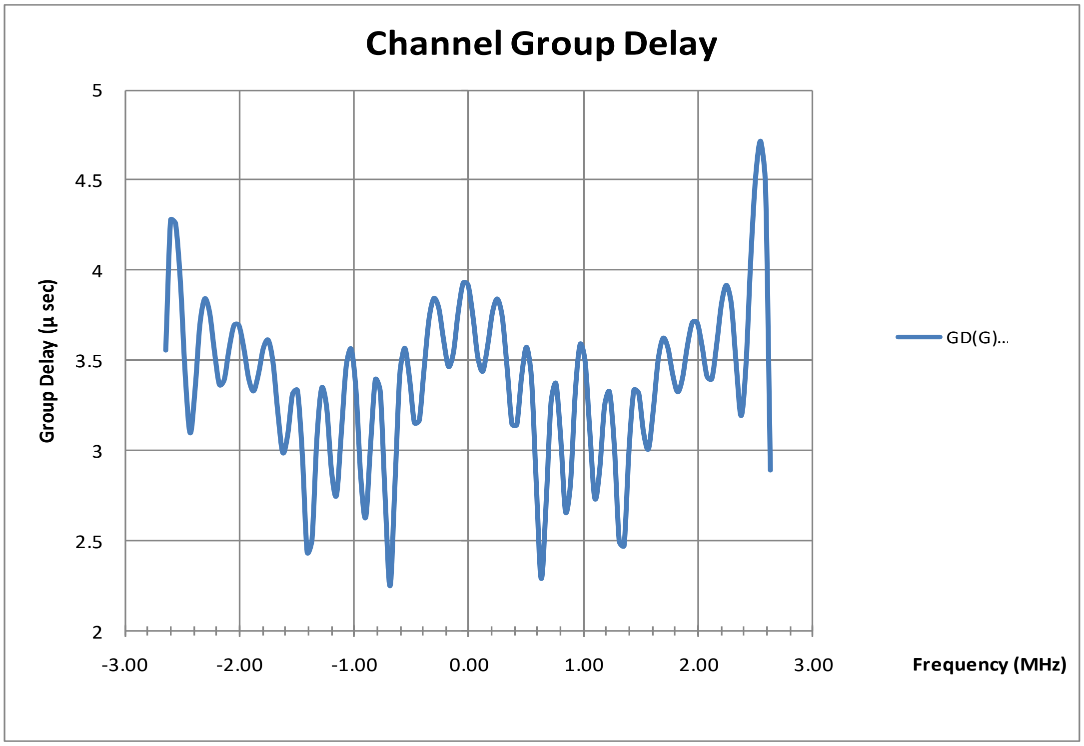


Figure 16-39: Channel Group Delay Response using 128 point DFT

16.15 OFDM

The DOCSIS 3.1 specifications, first published in 2013, introduced then-new physical layer (PHY) technology for cable networks: orthogonal frequency division multiplexing (OFDM) in the downstream, and its upstream counterpart, orthogonal frequency division multiple access (OFDMA). The combination of OFDM/A, a sophisticated forward error correction (FEC) known as low density parity check (LDPC)[5], and much higher modulation orders, allows DOCSIS 3.1 technology to scale to multi-gigabit data rates in HFC networks.

A closer look

Cable networks have for decades used frequency division multiplexing (FDM) to allow the transmission of multiple RF signals through the same length of coaxial cable at the same time. Each RF signal is on a separate frequency, or more specifically, assigned to its own channel (see Figure 16-40). National Television System Committee (NTSC) analog TV signals each occupy 6 MHz of bandwidth, and each 6 MHz-wide chunk of spectrum is a channel. For instance, what is called CTA channel 2 occupies 54 MHz to 60 MHz. Within each channel used for NTSC analog TV transmission, one will find an amplitude modulated (more specifically, vestigial sideband amplitude modulation or VSB-AM) visual carrier located 1.25 MHz above the lower channel edge, and a frequency modulated aural carrier 4.5 MHz above the visual carrier. A color subcarrier is located in between the visual and aural carriers, approximately 3.58 MHz above the visual carrier.

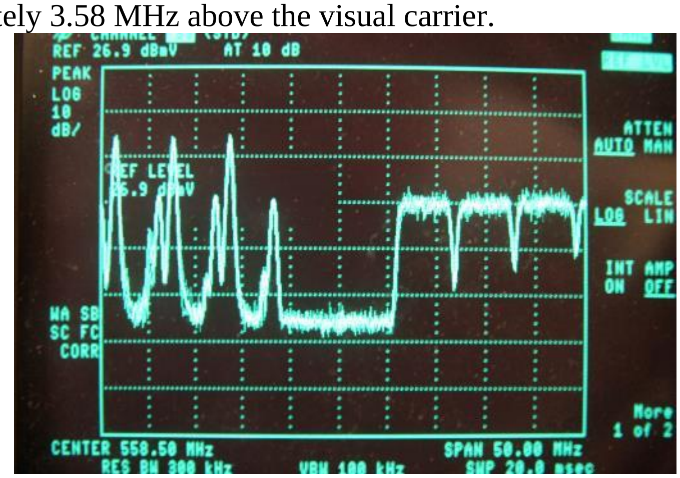


Figure 16-40: Example of FDM in a Cable Network

[5] The concept of LDPC was introduced by Robert G. Gallager in his 1960 doctoral dissertation, but because of encoder and decoder complexity wasn't practical to implement until much later. DOCSIS 3.1 uses a combination of LDPC and BCH (Bose-Chaudhuri-Hocquenghem) coding in the downstream, and LDPC in the upstream.

Three 6 MHz-wide analog NTSC signals can be seen on the left, and three 6 MHz-wide SC-QAM signals on the right.

When the cable industry made the transition to digital transmission, the modulation of choice was double-sideband suppressed carrier (DSB-SC), where the carrier uses quadrature amplitude modulation (QAM).[6] Each downstream DSB-SC signal occupies a 6 MHz bandwidth, the same as an analog NTSC TV signal.[7] This method of transmission has come to be known in the cable industry in recent years as single carrier QAM (SC-QAM), to distinguish from OFDM (which has multiple carriers within a channel).[8] The term SC-QAM is applied even when DOCSIS 3.0 channel bonding is used. Each channel slot carries only one modulated carrier – a QAM signal – hence the SC-QAM moniker. The entire data payload transmitted in the channel modulates just that one QAM signal.

Now imagine transmitting a large number of very narrow bandwidth QAM signals – hundreds or even thousands – within a given channel. A 6 MHz-wide channel could, for example, contain up to 240 narrow QAM signals that are spaced only 25 kHz apart. Each of the narrow QAM signals, called a subcarrier, subchannel, or tone (the term subcarrier is used in the remainder of this section), carries a small percentage of the total payload at a very low data rate. The aggregate of all of the subcarriers' data rates comprises the total data payload. This variation of FDM is known as OFDM.

For improved spectral efficiency, the subcarriers actually overlap one another.[9] This sounds counterintuitive, because one would be inclined to think that if signals overlap each other, interference will occur. With OFDM, the subcarriers are mathematically orthogonal to – that is, distinguishable from – one another, which takes care of the interference concern. "Orthogonal" in this case means that filtering one subcarrier through a filter tuned for another subcarrier will result in zero voltage at the properly timed sampling instant.[10] At other times, between the proper sampling instant, the contribution from adjacent subcarriers is non-zero, because of the frequency overlap of the signals. The concept is analogous (equivalent) to

---

[6] Double-sideband suppressed-carrier modulation, abbreviated DSB-SC, is described in many references, such as [see Bell Telephone Laboratories reference]. DSB-SC can carry analog or digital modulation. In cable digital modulation, the carrier is both amplitude and phase modulated, which is also called quadrature amplitude modulation, denoted as QAM.

[7] Euro-DOCSIS SC-QAM signals each occupy 8 MHz of RF bandwidth.

[8] It is a possible source of confusion that the traditional double-sideband suppressed carrier modulation, DSB-SC, and single-carrier QAM, SC-QAM, both employ the abbreviation "SC," but it has different meaning in the two cases.

[9] The individual carriers "overlap" in the sense that two or more of the subcarriers have a portion of their energy at the same frequencies.

[10] The bank of receiver filters tuned for each subcarrier are implemented digitally. The number of multiplications and additions is reduced by using the FFT algorithm compared to an implementation where each tuned filter is implemented separately (without shared mathematical operations) from each other.

having zero inter-symbol interference (ISI) in the time domain, after passing through the receiving filters.

Orthogonality is achieved by assigning the center frequency of each of the subcarriers to exactly an integer number of cycles in a selected interval, T, which is called “the FFT duration.” The spacing of the subcarrier center frequencies is the reciprocal of the interval T. This spacing results in the sinc (sin x/x) frequency response curves of the subcarriers lining up so that the peak of one subcarrier’s response curve falls on the first nulls of the lower and upper adjacent subcarriers’ response curves (see Figure 16-41).

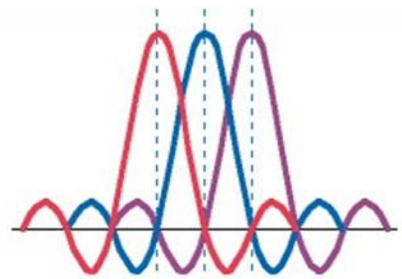


Figure 16-41: The Sinc (Sin X/X) Response Of OFDM Subcarriers; The Horizontal Axis Is Frequency

Note the overlap of the subcarriers and the alignment of the peaks and nulls. The subcarrier-to-subcarrier spacing is 1/T.

With OFDM, the concept of a 6 MHz- or 8 MHz-wide channel is no longer necessary. DOCSIS 3.1 downstream OFDM signals can be up to 192 MHz wide,[11] comprising up to 7,600 active subcarriers spaced 25 kHz apart or 3,800 subcarriers with 50 kHz spacing. Among the subcarriers are pilot tones for synchronization and other purposes. There are guard bands at each end of the 192 MHz-wide channel (minimum of 1 MHz each, resulting in an encompassed spectrum of 190 MHz). With the 25 kHz subcarrier spacing, the FFT duration (sometimes called useful symbol time) is 40 microseconds (µs), the reciprocal of which is the previously noted subcarrier spacing: 1/0.000040 second = 25,000 Hz. Note: The total symbol duration is equal to the FFT duration plus the length of the cyclic prefix, the latter configurable from 0.9375 µs to 5.0 µs.[12] Assuming 4096-QAM on each data subcarrier, the maximum data rate for a 192 MHz-wide downstream OFDM channel is in the range of about 1.8 Gbps to 1.89 Gbps, depending on subcarrier spacing and other factors.

[11] The minimum downstream DOCSIS 3.1 OFDM channel bandwidth is 24 MHz (22 MHz encompassed spectrum).

[12] The FFT duration for 25 kHz subcarrier spacing is 40 µs, and for 50 kHz subcarrier spacing is 20 µs. The cyclic prefix is a copy of a portion of the end of the symbol added to the beginning of the symbol. Assuming the FFT duration is 40 µs and the cyclic prefix is 2.5 µs, the total symbol duration would be 42.5 µs.

Upstream OFDMA, as the name suggests, supports multiple access. Multiple access, meaning multiple transmitters operating into a single receiver or sharing a single channel, can be accomplished by having different subsets of subcarriers assigned to different transmitters. In this case, for the cable industry, different cable modems can transmit simultaneously within the same channel, using different subcarriers, to a single OFDMA upstream receiver. OFDMA also can be used in combination with other multiple access schemes such as time division multiple access (TDMA). In strict TDMA, the full channel would be assigned to one user at a time, and the multiple access achieved via time division. The DOCSIS 3.1 upstream OFDMA actually combines FDMA and TDMA, with both the frequency domain (different subcarriers) and time domain (different time-slots) available for transmission among the multiplicity of cable modems transmitting to a single upstream receiver. Because of this use of TDMA in both SC-QAM and OFDMA upstream, cable modems are generally not transmitting continuously, but rather in assigned or granted bursts. The TDMA-capable upstream receiver is often termed a "burst receiver" for both SC-QAM and OFDMA upstream operations.

For more information on OFDMA, see reference U.S. Patent Number: 5,815,488 .

Advantages

Advantages of OFDM include the ability to adapt to degraded channel conditions such as the presence of severe micro-reflections, without the need for computationally intensive adaptive equalization algorithms. One reason for this that a very narrow bandwidth subcarrier typically experiences what is known as flat fading when micro-reflections affect channel response. This is in contrast to an SC-QAM signal that occupies the full channel bandwidth (e.g., 6 MHz), and is susceptible to amplitude ripple (standing waves) across that full signal. Each OFDM subcarrier "sees" just a tiny portion of the ripple, which for the most part affects the amplitude and phase of the narrow subcarrier with (essentially) the same value across the entire subcarrier's spectrum. The equalization is simplified to just applying an individual amplitude and phase adjustment to each subcarrier, a "single-tap" equalizer per subcarrier, unlike with SC-QAM, where shaping an equalizing filter's frequency response across the channel is required. The use of a "single-tap" equalizer for each of the subcarriers simplifies the adaptive equalization process for OFDM compared to having to shape a frequency response across a wider channel, even though there are many such "single-tap" adjustments required (one for each subcarrier) in OFDM. As a general rule, determination of the required "single-tap" adjustments for OFDM receivers requires fewer computations than required for a traditional SC-QAM time-domain adaptive equalizer in a highly distorted channel.[13]

[13] Research into frequency-domain adaptive equalizers for SC-QAM has developed solutions that claim to mitigate the computational advantage of OFDM equalization. This equalization is sometimes called single-carrier frequency domain equalization (SC-FDE) [see Pancaldi et al]. OFDM has demonstrated usefulness in channels with rapidly varying, non-flat channel response, such as in mobile phone applications, with a complexity and robustness advantage over single-carrier practices at the time of DOCSIS 3.1 specification development, through the beneficial use of pilot tones, which come at the expense of data throughput. In the cable environment, with a medium that is less dynamic than a mobile wireless application, a more-sparing use of pilot tones allows an increase in data throughput compared to conventional wireless OFDM.

Further, the composite OFDM signal can be more robust than SC-QAM in the presence of interference. For example, an ingressor such as a long term evolution (LTE) tower's transmitted signal may affect up to a few hundred subcarriers rather than taking out the full channel. Depending on the severity of the interference, FEC may be able to compensate in the OFDM system. Alternatively, the cable operator can change the modulation order on the affected subcarriers (lower) to increase robustness. In severe cases, the operator can create an exclusion band within the OFDM signal (in effect, turning off affected subcarriers) to avoid interference on problem frequencies.

Inter-symbol interference is generally less of a problem with OFDM because of the low data rate per subcarrier.

As discussed earlier, the overlapping nature of OFDM's subcarrier transmission provides high spectral efficiency in terms of spectral occupancy.[14]

If information about the channel's condition is sent back to the transmitter by the receiver, then multiple modulation profiles can used to optimize data throughput on all subcarriers, blocks of subcarriers, or even individual subcarriers. In other words, some subcarriers in the channel can use lower, more robust orders of modulation, while others can use higher modulation orders, on an as-conditions-warrant basis.

Disadvantages
OFDM does have a few disadvantages: It is susceptible to phase noise, frequency and clock errors, although the pilot carriers that accompany the subcarriers help to mitigate this by providing the receiver a means of synchronization with the transmitter. OFDM has a high peak-to-average power ratio (PAPR), but a spectrum full of SC-QAM signals does, too. Some of OFDM's overall efficiency, which benefits from high spectral efficiency, is reduced by the use of cyclic prefixes, which help to maintain subcarrier orthogonality and compensate for echoes, but introduce significant time-domain inefficiency compared to SC-QAM.

Why bother?
Some may wonder why the cable industry even considered a new PHY for DOCSIS 3.1. After all, SC-QAM works well, and channel bonding can be used to significantly increase data throughput. What about increasing the symbol rate and consequently also the SC-QAM per-channel RF bandwidth? Doing so would have been impractical in part because of the very large number of taps needed for time-domain adaptive equalization, and the sheer computational complexity.

The good news is that OFDM isn't a new technology without a proven history. It is used in Wi-Fi networks, worldwide interoperability for microwave access (WiMAX), LTE, digital audio broadcasting (DAB), ultra wideband (UWB), and Europe's digital video broadcasting

[14] "Spectral efficiency" is one aspect of overall "efficiency." See Appendix L in SCTE Operational Practice SCTE 270 2021 Mathematics of Cable for a thorough discussion of efficiency.

(DVB). A variation of ODFM also is used in asymmetric digital subscriber line (ADSL) and very high-speed digital subscriber line (VDSL).

The previously discussed advantages bring a lot of signal transmission flexibility to the table. When OFDM is combined with more powerful FEC, higher orders of modulation can be used – within the limits of the channel conditions, of course.

16.16 DOCSIS 3.1 OFDM transmission and reception

Figure 16-42 illustrates a high-level overview of transmission and reception of OFDM in a cable network. Following the block diagram in the lower part of the figure, data enters the coding and interleaving stage, followed by bit-to-symbol mapping. After modulation, serial to parallel conversion, and inverse fast Fourier transform (IFFT), the cyclic prefix is added to the beginning of each symbol. In the receiver, essentially the opposite processes occur, producing the output data.

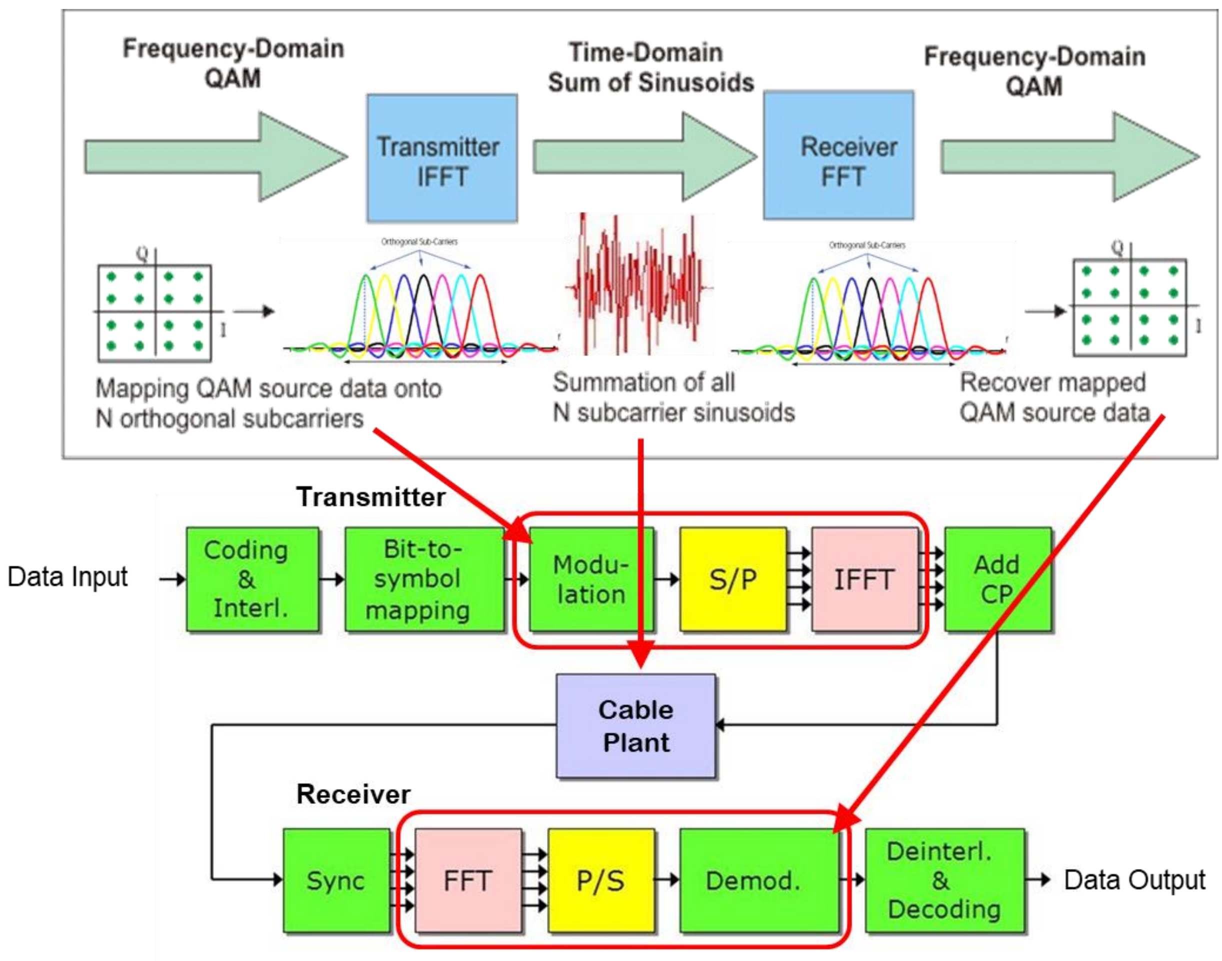


Figure 16-42: Diagram Illustrating Downstream OFDM Transmission and Reception (courtesy of Rich Prodan, Comcast).

16.17 Anatomy of an OFDM signal

The following information and figures highlight some of the major components of a 192 MHz-wide DOCSIS 3.1 downstream OFDM signal, as seen on a spectrum analyzer. The analyzer configuration was optimized to make it easier to see various parts of the signal, such as the pilots.

Figure 16-43 shows the OFDM channel bandwidth, which includes spectral guard bands (taper regions) on the ends of the signal. The minimum width of the guard bands is about 1 MHz on each bandedge, but can be greater.

Figure 16-43: Spectrum Analyzer Screen Shot Of A 192 Mhz-Wide Downstream OFDM Signal.

Figure 16-44 shows the encompassed spectrum, 190 MHz in this example, of the OFDM signal in the previous figure. Encompassed spectrum is defined in the DOCSIS 3.1 Physical Layer Specification as "…the range of frequencies from the center frequency of the channel's lowest active subcarrier minus half the subcarrier spacing, to the center frequency of the channel's highest active subcarrier plus half the subcarrier spacing."

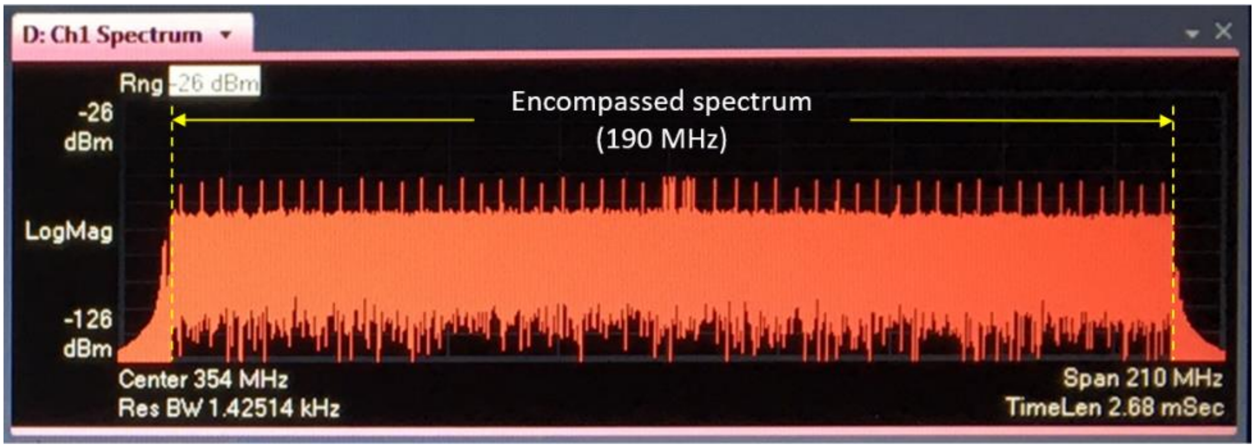


Figure 16-44: 190 MHz-Wide Encompassed Spectrum of the OFDM Signal In The Previous Figure

Figure 16-45 shows the location of the OFDM signal's data subcarriers.

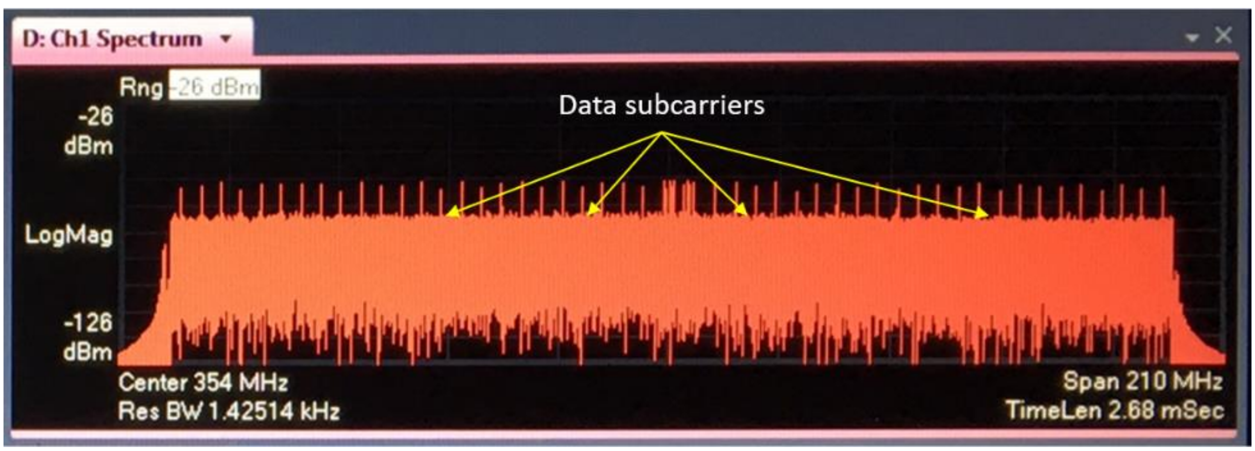


Figure 16-45: Data Subcarriers in the 192 MHz-Wide OFDM Signal

Figure 16-46 shows the OFDM signal's continuous pilots, boosted by 6 dB relative to the data subcarriers. Continuous pilots always remain on the same frequencies, and are configured to be spread uniformly throughout the OFDM signal. Another type of pilot called scattered pilots – which are also boosted by 6 dB – are not shown here (they change frequency periodically throughout the OFDM signal).

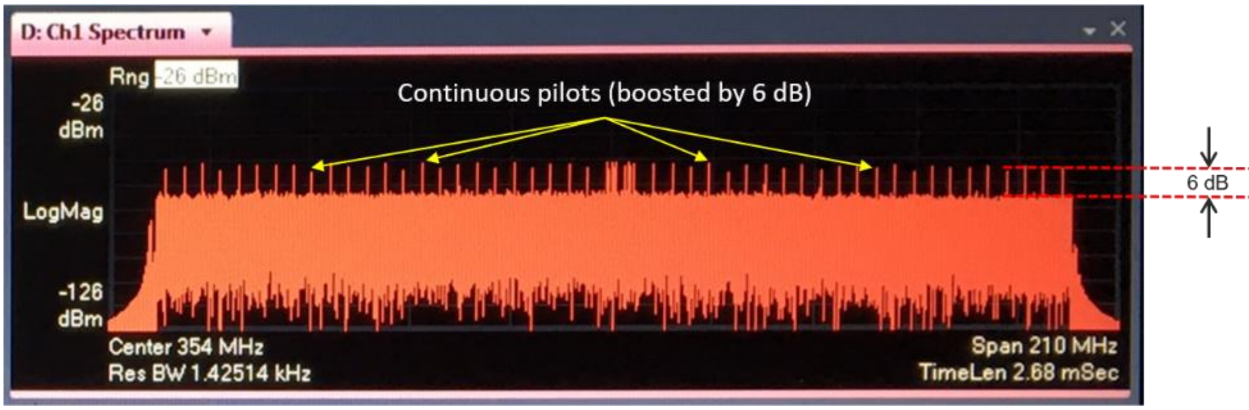


Figure 16-46:The "spikes" Sticking Up Above The Data Subcarriers Are The Continuous Pilots, Which Are Used For Frequency And Phase Synchronization

The physical layer link channel or PHY link channel (PLC) band is a 6 MHz-wide portion of the OFDM signal within which the PLC is centered. The PLC band cannot have any excluded subcarriers or exclusion bands. See Figure 16-47

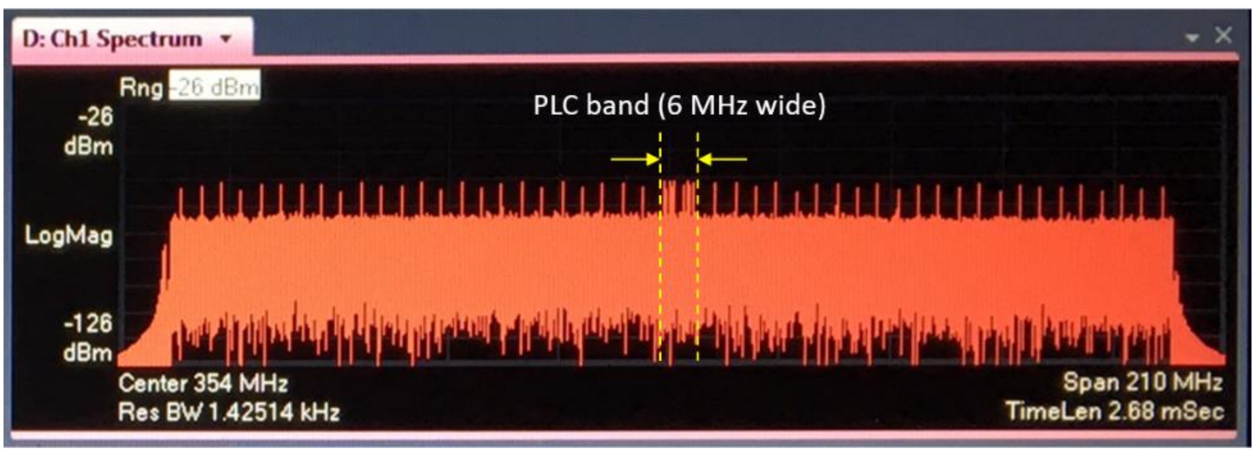


Figure 16-47: Location of the 6 MHz-wide PLC Band

Figure 16-48 shows the continuous pilots within the PLC band (circled). The continuous pilots in the PLC band have a unique pattern that tells the cable modem where the PLC is located. The PLC band's continuous pilot pattern is not user-adjustable; it is a fixed pattern defined in the DOCSIS 3.1 Physical Layer Specification.

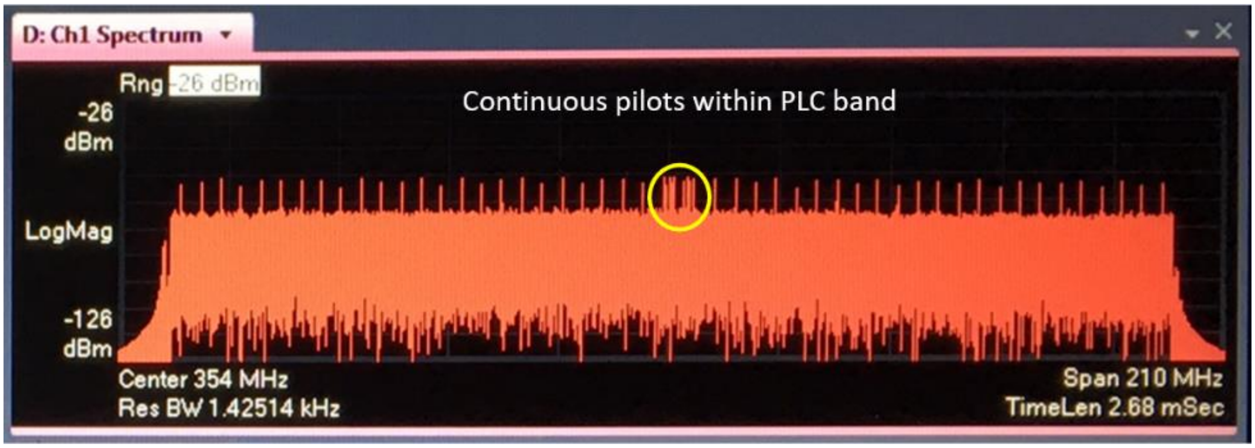


Figure 16-48: PLC Band's Continuous Pilots (circled)

Figure 16-49 shows the location of the 400 kHz-wide PLC, centered within the 6 MHz-wide PLC band. The PLC conveys physical layer parameters from the CMTS to the cable modems. An important point: The cable operator chooses where in the OFDM channel to place the PLC band and PLC (it is not necessary to use the CMTS's default setting). Ideally, the PLC should be located in a known clean part of the OFDM channel that is not susceptible to ingress, direct pickup, and other types of interference.

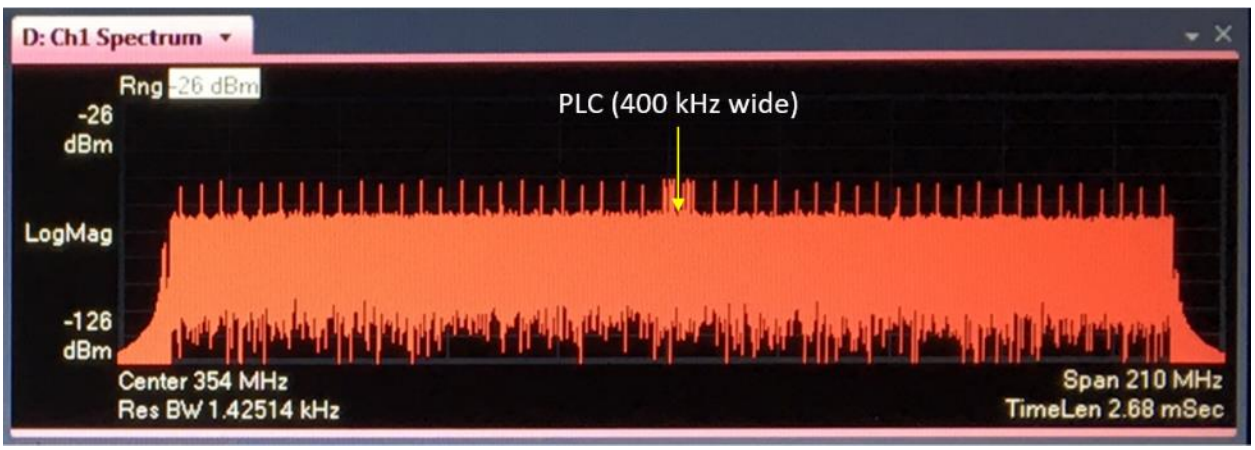


Figure 16-49: The 400 kHz-wide PLC, centered within the PLC band, comprises eight subcarriers (50 kHz subcarrier spacing) or 16 subcarriers (25 kHz subcarrier spacing), and uses 16-QAM

16.18 Bandwidth of DOCSIS OFDM channels

As mentioned earlier, with DOCSIS 3.1 OFDM signals, channels are no longer restricted to just 6 MHz or 8 MHz width. Downstream OFDM channel bandwidth can vary from a minimum of 24 MHz to a maximum of 192 MHz. However, "bandwidth" as used here is a bit more complicated than one might initially assume. From SCTE 270 2021, *bandwidth* is defined as "1) The amount of spectrum, measured in units of hertz, that an electromagnetic signal significantly occupies. 2) The operating passband of a device or system, typically expressed in units of hertz."

Terms such as *encompassed spectrum* (discussed previously), *occupied bandwidth*, *modulated spectrum*, and *taper region* are important characteristics to understand when dealing with DOCSIS OFDM signals. The DOCSIS 3.1 PHY specification glossary defines most of those terms; those definitions are included here.

> **Channel –** A portion of the electromagnetic spectrum used to convey one or more RF signals between a transmitter and receiver. May be specified by parameters such as center frequency, bandwidth, or [CTA][15] channel number.
>
> **Encompassed spectrum –** 1) For an OFDM or OFDMA channel, the range of frequencies from the center frequency of the channel's lowest active subcarrier minus half the subcarrier spacing, to the center frequency of the channel's highest active subcarrier plus half the subcarrier spacing. 2) For an SC-QAM channel, the encompassed spectrum is the signal bandwidth (i.e., 6 MHz or 8 MHz in the downstream; 1.6 MHz, 3.2 MHz, and 6.4 MHz in the upstream). 3) For the RF output of a downstream or upstream port including multiple OFDM, OFDMA, and/or SC-QAM channels, the range of frequencies from the lowest frequency of the encompassed spectrum of the lowest frequency channel to the highest frequency of the encompassed spectrum of the highest frequency channel.
>
> **Modulated spectrum –** 1) Downstream modulated spectrum - Encompassed spectrum minus the excluded subcarriers within the encompassed spectrum, where excluded subcarriers include all the individually excluded subcarriers and all the subcarriers comprising excluded sub-bands. This also is the spectrum comprising all active subcarriers. Note: For this definition, the width of an active or excluded subcarrier is equal to the subcarrier spacing. 2) Upstream modulated spectrum - The spectrum comprising all non-zero-valued subcarriers of a cable modem's OFDMA transmission, resulting from the exercised transmit opportunities. Note: For this definition, the width of a transmitted subcarrier is equal to the subcarrier spacing.
>
> **Occupied bandwidth:** 1) Downstream - The sum of the bandwidth in all standard channel frequency allocations (e.g., 6 MHz spaced [CTA] channels) that are occupied by the OFDM channel. The [CTA] channels which are occupied by the OFDM signal are those which contain any of the Modulated Spectrum and/or taper region shaped by the OFDM channels' transmit windowing, where the values for the taper regions are defined in Appendix V as a function of the Roll-Off Period. It is possible, but not problematic, for a [CTA] channel to be "occupied" by two OFDM channels. 2) Upstream - a) For a single OFDMA channel, the sum of the bandwidth in all the subcarriers of that OFDMA channel which are not excluded. The upstream occupied bandwidth is calculated as the number of subcarriers which are not excluded, multiplied by the subcarrier

[15] The DOCSIS 3.1 PHY specification glossary uses "CEA" rather than the more current "CTA."

spacing. b) For the transmit channel set, the sum of the occupied bandwidth of all OFDMA channels plus the bandwidth of the legacy channels (counted as 1.25 times the modulation rate for each legacy channel) in a cable modem's transmit channel set. The combined bandwidth of all the minislots in the channel is normally smaller than the upstream occupied bandwidth due to the existence of unused subcarriers. The bandwidth occupied by an OFDMA probe with a skip value of zero is equal to the upstream occupied bandwidth.

Taper region is defined in **Appendix V CMTS Proposed Configuration Parameters (Informative)** of the DOCSIS 3.1 PHY spec as follows: "The taper region consists of the spectrum extending from a) the spectral edge of the encompassed spectrum, to b) the frequency above (or below) the edge of the encompassed spectrum by the amount (MHz) as given in the third column of Table 55."

Consider the graphic in Figure 1, which shows a 192 MHz-wide OFDM signal that has no exclusion bands, and an encompassed spectrum of 190 MHz. In this example, the taper regions on the lower and upper band edges of the OFDM signal are each about 1 MHz wide (the actual width of each taper region in this example would be 0.975 MHz).

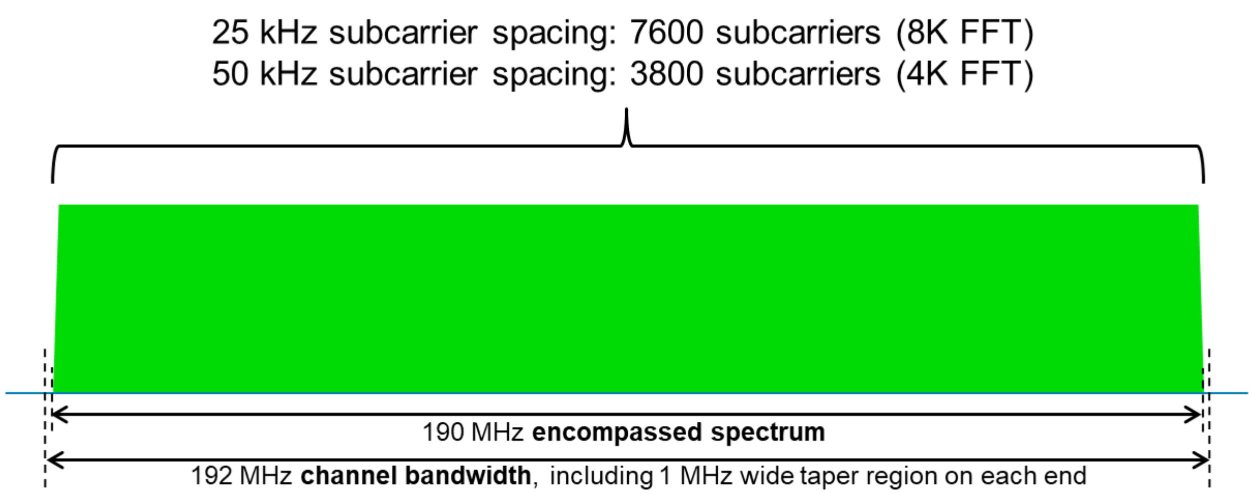


Figure 16-50: 192 MHz-wide OFDM Signal with 190 MHz Encompassed Spectrum

The graphic in Figure 2 highlights the approximately 1 MHz-wide taper regions at the lower and upper band edges of the OFDM signal.

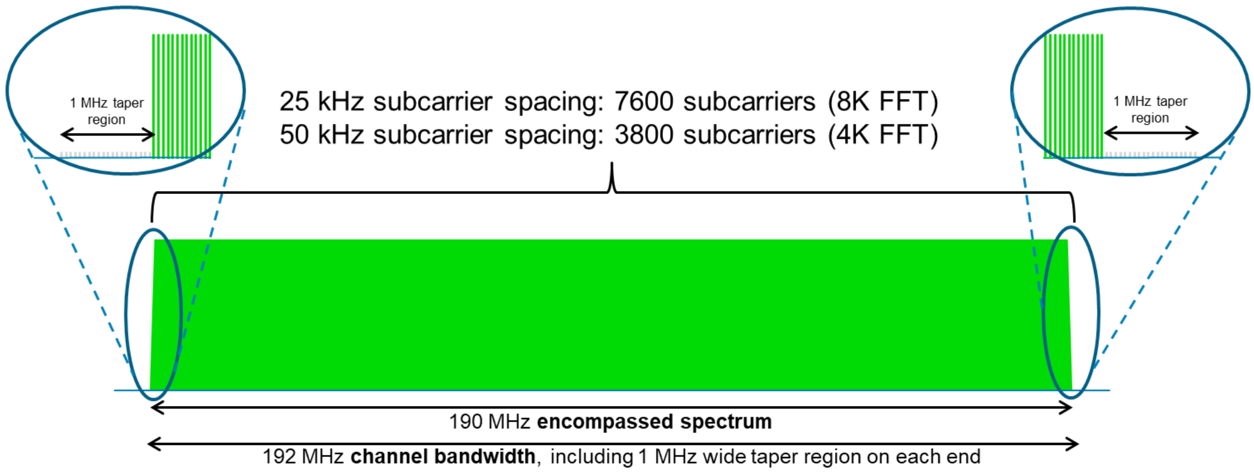


Figure 16-51: Closer Look At Taper Regions At Lower And Upper Edges of OFDM Signal.

Table 55 in the DOCSIS 3.1 PHY spec summarizes (defines) the taper region widths that occur with supported roll-off period (N_{rp}) sample configurations. Figure 3 shows what a 0.975 MHz-wide taper region looks like for a 96 MHz-wide OFDM channel when N_{rp} = 256.[16]

[16] Note: The three "impact" parameters in the figure's legend ("Impact to 750 kHz = 0.52", etc.) refer to the increase (in dB) in combined theoretical OFDM sidelobes and allowed spurious emissions compared to just the allowed spurious emissions by themselves in DOCSIS 3.1 downstream modulator requirements for a 96 MHz-wide OFDM channel. The shaded green shows the PSD of the theoretical OFDM sidelobe in the 750 kHz adjacent spectrum.

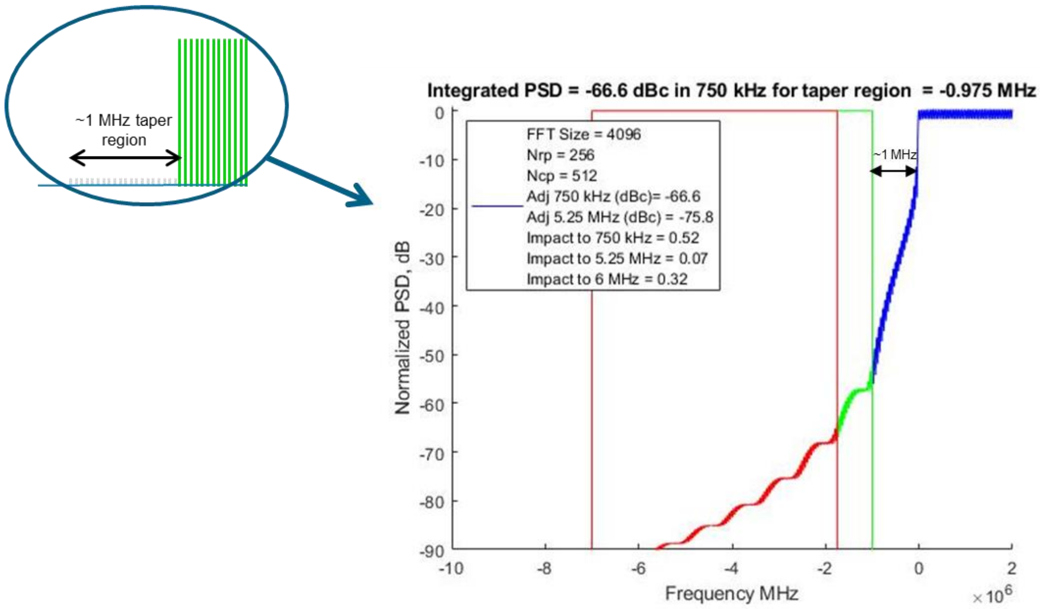


Figure 16-52: The Approximately 1 MHz-Wide Taper Region (Actual Width Is 0.975 MHz When N_{rp} = 256) At The OFDM Signal Band Edges Looks More Like The Simulation Shown Here.

Note: 0 dBc in the figure is the channel power in 6 MHz. Graphic courtesy of Broadcom.

Figure 4 shows what a 3.575 MHz-wide taper region looks like for a 96 MHz-wide OFDM channel when N_{rp} = 64.[17]

[17] Note: The three “impact” parameters in the figure’s legend (“Impact to 750 kHz = 0.44”, etc.) refer to the increase (in dB) in combined theoretical OFDM sidelobes and allowed spurious emissions compared to just the allowed spurious emissions by themselves in DOCSIS 3.1 downstream modulator requirements for a 96 MHz-wide OFDM channel. The shaded green shows the PSD of the theoretical OFDM sidelobe in the 750 kHz adjacent spectrum.

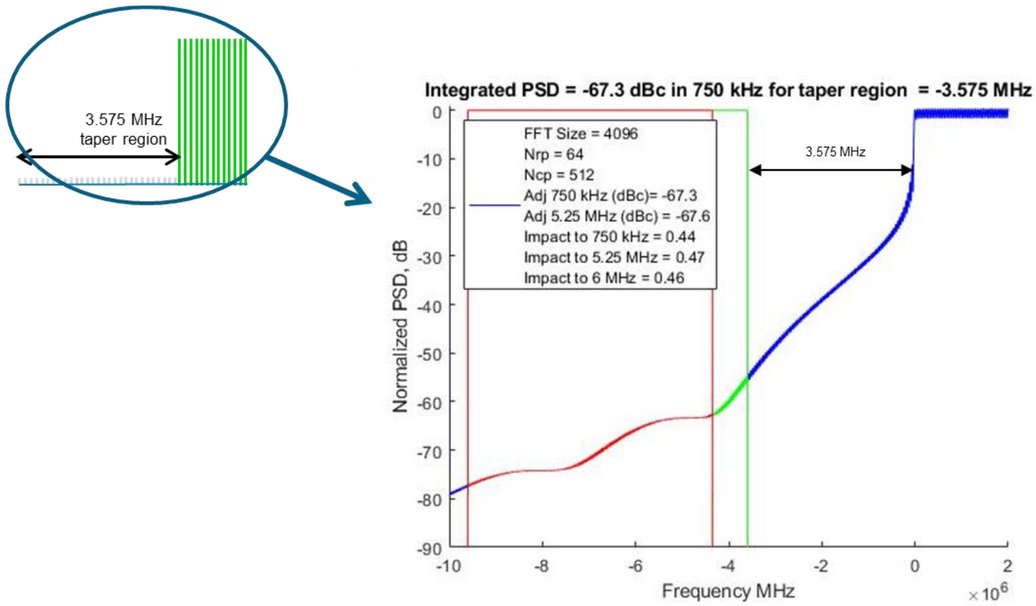


Figure 16-53: The 3.575 MHz-wide Taper Region at the OFDM Signal Band Edges Looks More Like The Simulation Shown Here When N_{rp} = 64. Note: 0 dBc in The Figure Is Channel Power In 6 MHz. Graphic courtesy of Broadcom.

The graphic in Figure 5 illustrates the occupied bandwidth of a 192 MHz-wide downstream OFDM signal (190 MHz encompassed spectrum) that is aligned with the CTA channel grid. In the figure, the OFDM signal's occupied bandwidth is 32 CTA channels x 6 MHz = 192 MHz.

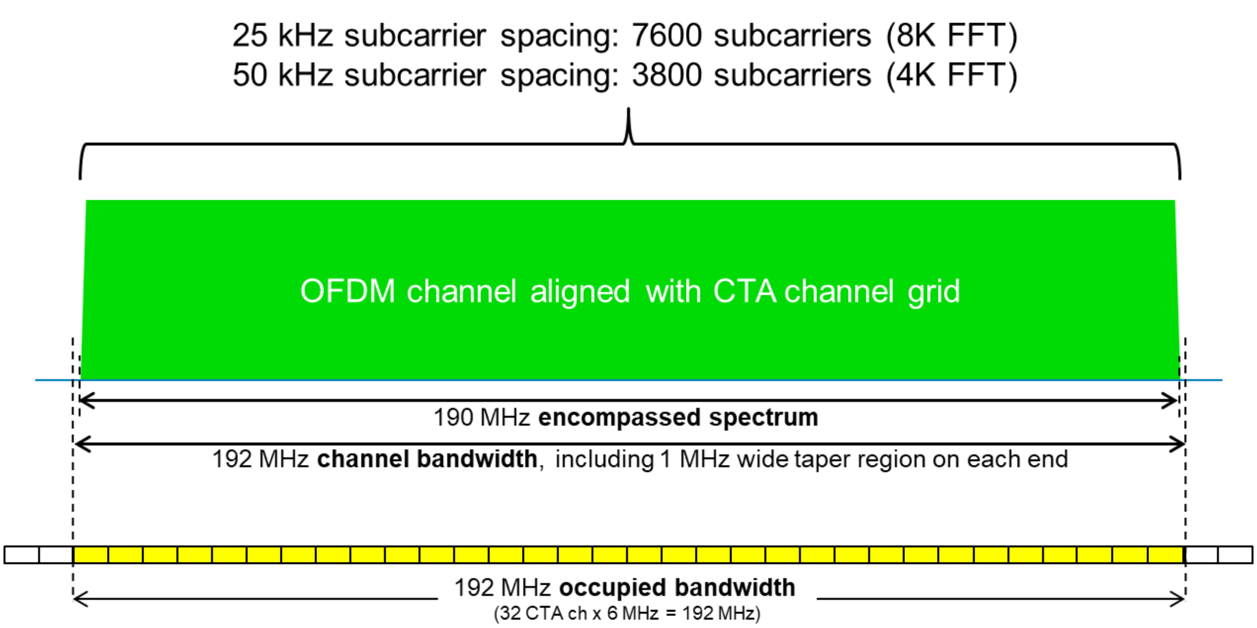


Figure 16-54: Occupied Bandwidth of a 192 MHz-wide OFDM Signal (190 MHz Encompassed Spectrum).

The graphic in Figure 6 shows the modulated spectrum for a 192 MHz-wide OFDM signal whose encompassed spectrum is 190 MHz, and that has no exclusion bands. In this example the modulated spectrum is 190 MHz.

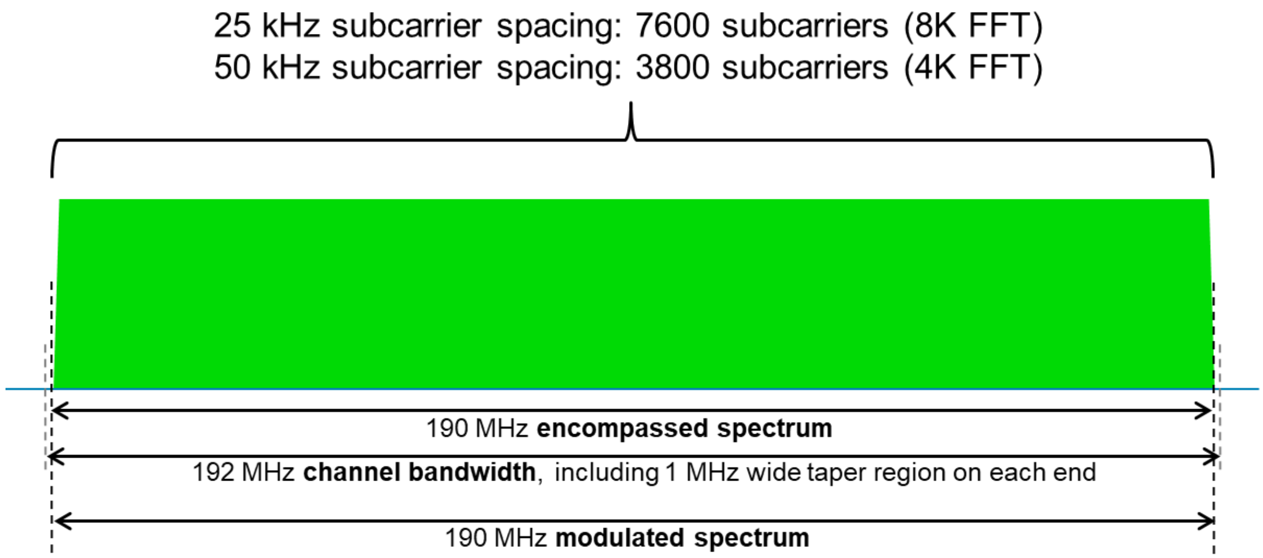


Figure 16-55: Modulated spectrum of a 192 MHz-wide OFDM Signal With no Exclusion Bands and 190 MHz Encompassed Spectrum.

When one or more exclusion bands are configured in an OFDM signal, determining modulated spectrum is a bit more difficult than the previous example. The graphic in Figure 7 shows a 192 MHz-wide OFDM signal (190 MHz encompassed spectrum) with a 20 MHz-wide exclusion band. In this example the modulated spectrum is 170 MHz.

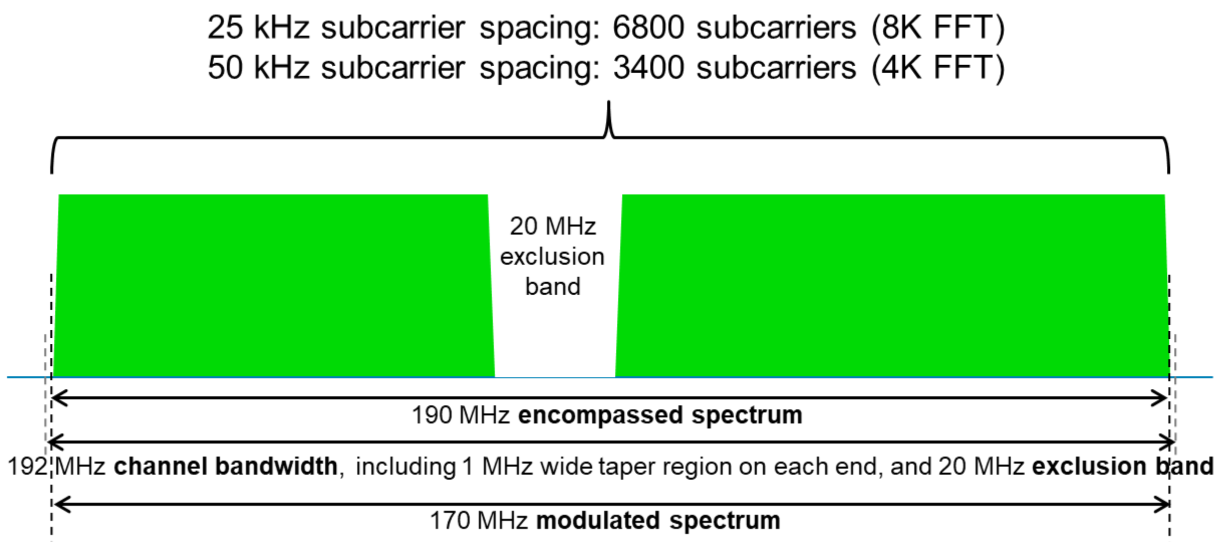


Figure 16-56: Modulated Spectrum of a 192 MHz-Wide OFDM Signal with a 20 MHz-Wide Exclusion Band and 190 MHz Encompassed Spectrum.

As shown in Figure 8, the occupied bandwidth for the previous example is 29 CTA channels x 6 MHz = 174 MHz. The channel bandwidth is still 192 MHz and the encompassed spectrum is 190 MHz.

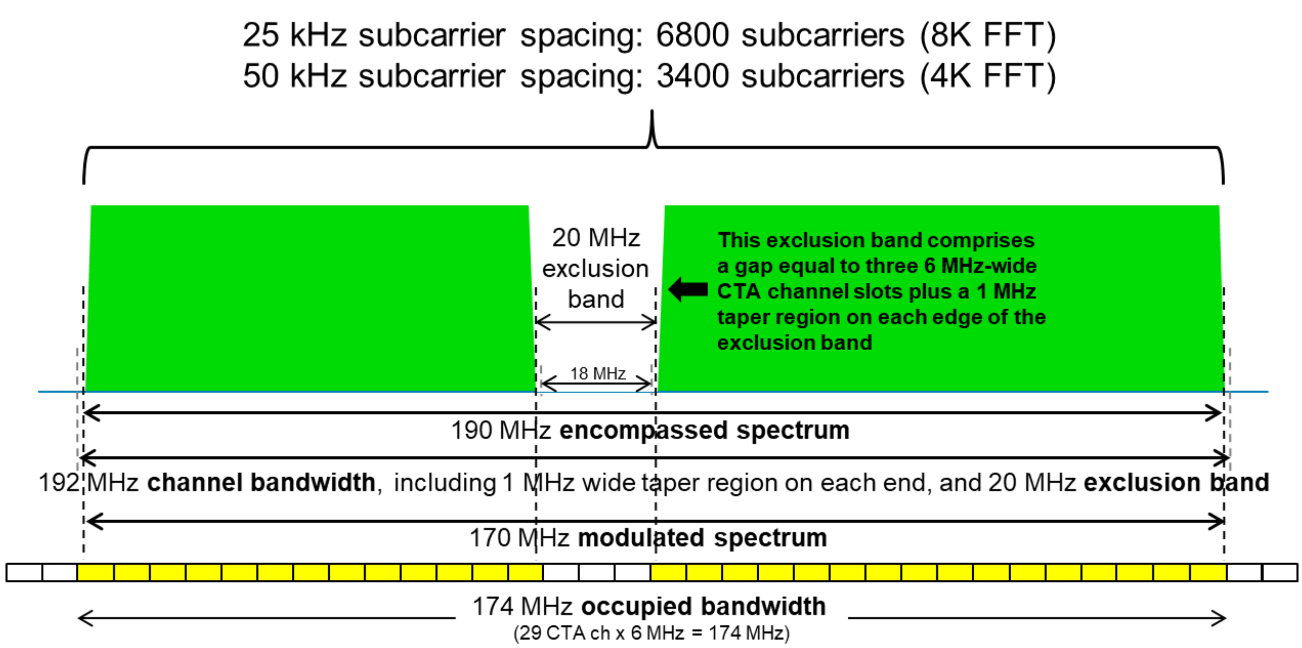


Figure 16-57: Occupied Bandwidth of a 192 MHz-wide OFDM Signal with a 20 MHz-wide Exclusion Band and 190 MHz Encompassed Spectrum.

The graphic in Figure 16-58 shows a 24 MHz-wide OFDM signal with approximately 1 MHz-wide taper regions on the lower and upper band edges of the signal. The encompassed spectrum is 22 MHz, and the occupied bandwidth is 4 CTA channels x 6 MHz = 24 MHz.

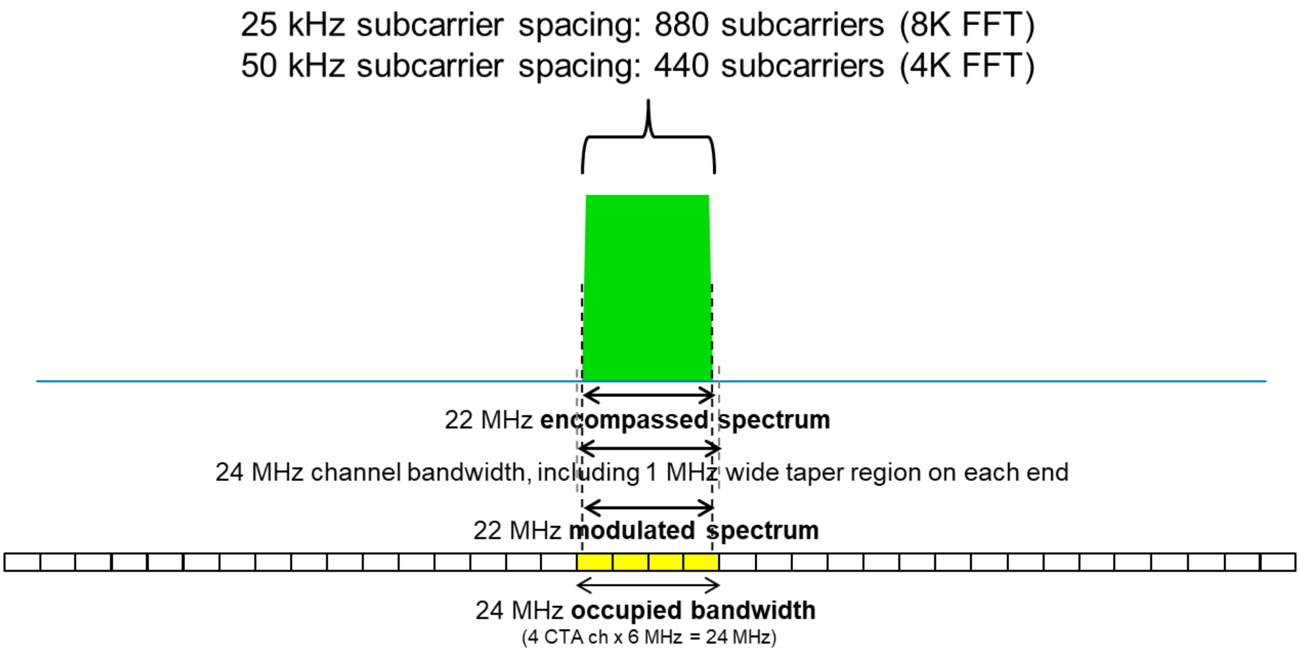


Figure 16-58: Encompassed spectrum, modulated spectrum, and occupied bandwidth of a 24 MHz-wide OFDM signal.

Characterizing DOCSIS 3.1 downstream OFDM performance

Many digital signal analyzers available to the cable industry include embedded DOCSIS 3.1 cable modem silicon, which facilitates testing and measuring various parameters of a downstream OFDM signal. Some of the parameters of interest include: RF signal level, error correction performance, and receive modulation error ratio (RxMER, often called just MER or sometimes SNR).

The information included here is intended to be general in nature. Different makes/models of test equipment may support different measurement capabilities. In all cases, follow the test equipment manufacturer's instructions for setup and operation. During operation, ensure that the instrument is tuned to the OFDM signal to be measured, and at that it is locked to the PLC and boot profile.[18]

[18] All DOCSIS 3.1 cable modems, including the test equipment's embedded modem, must be able to receive profile A, the boot profile. This profile typically uses a lower, more robust modulation order such as 256-QAM.

RF signal level

OFDM channel power is the RF power per CTA channel – that is, the power per 6 MHz – which provides signal level information comparable to SC-QAM digital channel power. Figure 16-59 shows an example of channel power per 6 MHz for a 96 MHz-wide OFDM signal. The short blue horizontal lines represent the power in 6 MHz segments across the bandwidth of the OFDM signal. Here, the average channel power is –3.3 dBmV, but note that the power per 6 MHz varies somewhat across the channel, from a minimum of –5.3 dBmV to a maximum of –2.7 dBmV. Per the DOCSIS 3.1 Physical Layer Specification, the OFDM signal level (power per 6 MHz) at the cable modem input is supposed to be in the –15 dBmV to +15 dBmV range.

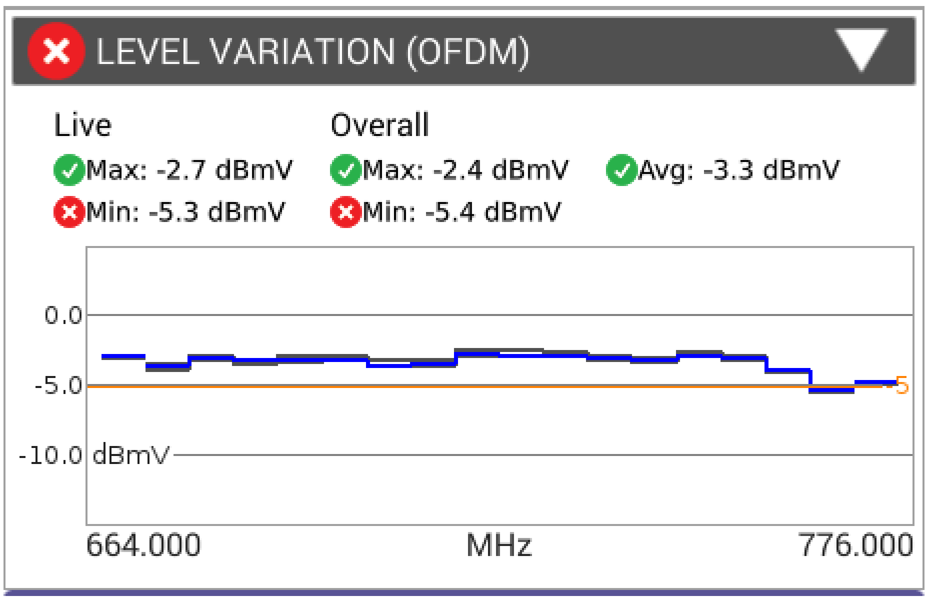


Figure 16-59: OFDM Channel Power (That Is, Power Per 6 MHz) for a 96 MHz-wide OFDM Signal

In most cases, OFDM channel power should be set to the same power as the cable network's SC-QAM signals: the power spectral density (PSD) of both OFDM and SC-QAM vary across their respective channels, but averaged over 6 MHz (e.g., the CTA channel grid), the averages are set to be the same (recommended practice).[19] In other words, the heights of the OFDM and SC-QAM "haystacks" as seen on a spectrum analyzer should be the same (when averaged over 6 MHz). Figure 16-60 shows a spectrum analyzer display in which the OFDM signal on the right half of the screen has been set to the same PSD as the SC-QAM signals on the left half of the screen. The OFDM pilots are included in the averaging, as is the OFDM taper region (bandedges) and any excluded subcarriers, and similarly, the SC-QAM bandedge spectral rolloff is included in the averaging, when matching the PSDs; these factors

[19] SCTE-258-2020, Section 6.2.1.

are visible in Figure 16-60, where the 6 MHz averages are set equal. The SC-QAM rolloff at the channel edges can be seen, as can the OFDM continuous pilots. Also, some small amount of amplitude variation across the full spectrum seems to have occurred in generating Figure 16-60

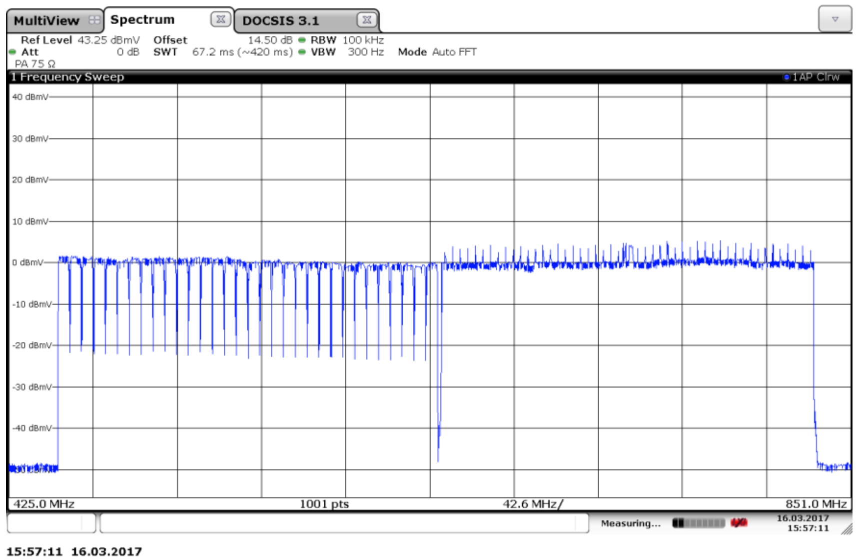
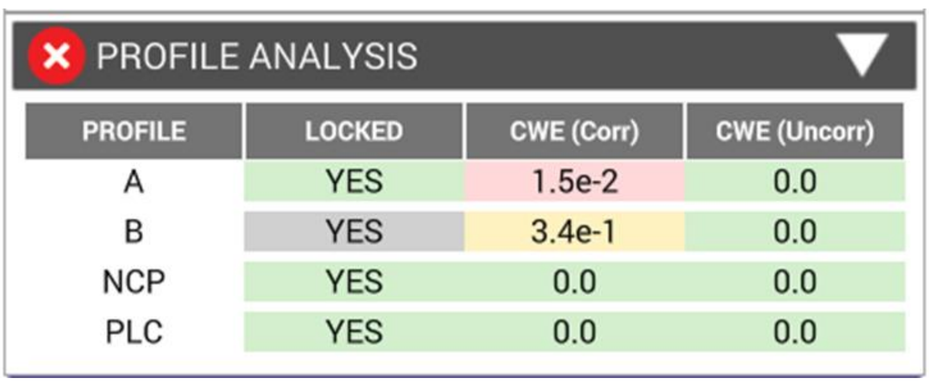


Figure 16-60: In most cases the PSD of the cable network's SC-QAM signals and OFDM signal(s) should be the same (see text).

Error correction



PROFILE ANALYSIS

PROFILE	LOCKED	CWE (Corr)	CWE (Uncorr)
A	YES	1.5e-2	0.0
B	YES	3.4e-1	0.0
NCP	YES	0.0	0.0
PLC	YES	0.0	0.0

Figure 16-61: This test equipment display shows codeword error statistics for modulation profiles A and B, NCP, and the PLC.

Most DOCSIS 3.1-capable digital signal analyzers report FEC performance for the OFDM signal, typically as correctable and uncorrectable codeword error statistics. Depending on the make/model of test equipment, codeword error information can be reported for the active modulation profile(s), next codeword pointer (NCP), and PLC, as shown in the example in Figure 16-61. Correctable codeword errors can usually be ignored. However, the goal should be few or no uncorrectable codeword errors (the uncorrectable codeword error performance, or post-FEC errors, is discussed later in this section). Note that the display in Figure 16-61also indicates that the instrument is locked to the active profiles, NCP, and PLC.

RxMER

DOCSIS 3.1 cable modems can report the RxMER per subcarrier, a capability that provides a powerful troubleshooting tool. However, it would be impractical to scroll through a list of thousands of per-subcarrier values for a 192 MHz-wide OFDM signal with up to 7,600 active subcarriers. A better approach is a graph of RxMER per subcarrier, as shown in Figure 16-62. In this example, the vertical axis is RxMER in dB, and the horizontal axis is frequency (sometimes subcarrier number). From such a graph, the user can quickly discern what the OFDM signal's RxMER per subcarrier looks like across the bandwidth of the signal, and identify problems such as frequency-specific ingress.

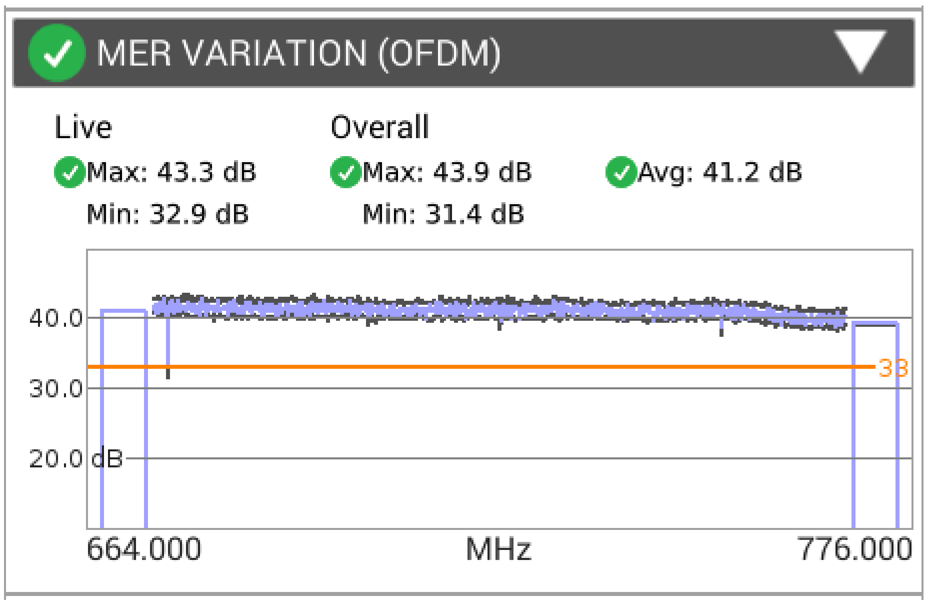


Figure 16-62: Graph of RxMER per subcarrier for a 96 MHz-wide OFDM signal. This particular display also includes the average, minimum, and maximum RxMER per subcarrier values.

Some digital signal analyzers also report RxMER for the PLC, and additional parameters such as RxMER standard deviation[20] and RxMER 2nd percentile.[21] See Figure 16-63.

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[20] Good engineering practice suggests that the standard deviation should be less than 2 dB, but 1 dB or less is even better.

[21] "RxMER 2nd percentile" means that 98% of the subcarriers have higher RxMER than what is shown.

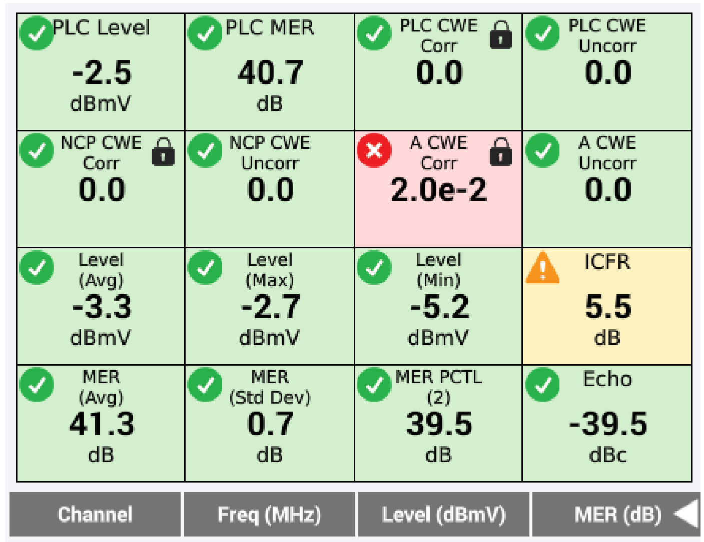

Figure 16-63: Some Digital Signal Analyzers Can Provide A Variety Of OFDM performance Metrics, Including Several RxMER Parameters.

Cable operators occasionally ask about the network performance necessary to support reliable downstream OFDM operation. From the DOCSIS 3.1 Physical Layer Specification: "The required level for CM downstream post-FEC error ratio is defined as less than or equal to 10^{-6} PER (packet error ratio) with 1500 byte Ethernet packets." The values shown in Figure 16-64 "…describe the conditions at which the cable modem is required to meet this error ratio."

Figure 16-64 : Copy of Table 46 from the DOCSIS 3.1 Physical Layer Specification, Summarizing Cable Modem Minimum CNR Performance in an AWGN Channel.

Constellation	CNR[1,2] (dB) Up to 1002 MHz	CNR[1,2] (dB) 1002 MHz to 1218 MHz	Min P_{6AVG} dBmV
4096	41.0	41.5	-6
2048	37.0	37.5	-9
1024	34.0	34.0	-12
512	30.5	30.5	-12
256	27.0	27.0	-15
128	24.0	24.0	-15
64	21.0	21.0	-15
16	15.0	15.0	-15

Table Notes:

Note 1. CNR is defined here as total signal power in occupied bandwidth divided by total noise in occupied bandwidth.

Note 2. Channel CNR is adjusted to the required level by measuring the source inband noise including phase noise component along with transmitter noise and distortion and adding the required delta noise from an external AWGN generator to achieve the desired CNR at the CM F-connector.

Note 3. Applicable to an OFDM channel with 192 MHz of occupied bandwidth.