Understanding Jitter and Wander Measurements and Standards

Second Edition

Agilent Technologies

Contents

Welcome to the second edition of Agilent Technologies *Understanding Jitter and Wander Measurements and Standards* booklet, the distillation of over 20 years' experience and know-how in the field of telecommunications jitter testing.

The Telecommunications Networks Test Division of Agilent Technologies (formerly Hewlett-Packard) in Scotland introduced the first jitter measurement instrument in 1982 for PDH rates up to E3 and DS3, followed by one of the first 140 Mb/s jitter testers in 1984. SONET/SDH jitter test capability followed in the 1990s, and recently Agilent introduced one of the first 10 Gb/s Optical Channel jitter test sets for measurements on the new ITU-T G.709 frame structure. Over the years, Agilent has been a significant participant in the development of jitter industry standards, with many contributions to ITU-T O.172/O.173, the standards for jitter test equipment, and Telcordia GR-253/ ITU-T G.783, the standards for operational SONET/SDH network equipment.

Agilent's long-term commitment to leadership in jitter testing has gained us a reputation for accuracy and repeatability of measurement. We believe Agilent has the highest-performing instruments in the industry and the best overall understanding of the various nuances of jitter testing, many of which are discussed in this booklet. Our goal is to provide instruments with very low intrinsics and high accuracy so that measurements made on network equipment are a true reflection of their performance without compromise.

In this second edition, our goal remains the same: to bring Agilent's jitter measurement knowledge and experience together in one place, providing a valuable reference for engineers in the telecommunications industry. We have added new papers on Jitter Testing in the Optical Transport Network (OTN) and An Overview of Wander Measurements. Two additional papers also explore the performance of jitter test sets and the validity of calibration schemes, particularly for exacting jitter generation measurements.

Jitter is a complicated topic and there is always ongoing debate and argument about the integrity of measurements between industry players, whether equipment manufacturers, network operators or test equipment suppliers. Sometimes measurements made in good faith cause disagreement between parties, due to misunderstandings or invalid assumptions about the measurement set-up, execution and compliance with standards. I hope that the papers in this booklet will help to prevent and/or resolve these disagreements, while also stimulating industry debate over some controversial measurement issues on accuracy and the compatibility of various standards.

Ronnie Neil
Product Marketing Manager
Functional Test for Optical Transport

1. Introduction to Jitter and Wander

Abstract: This paper sets the scene for the following papers on more detailed topics. It describes why jitter measurements are important and the fundamental classes of jitter and wander measurements.

Why jitter is important

Error free communications is something every user would like to enjoy. Digital transmission, with its ability to completely avoid cumulative noise-induced degradation, should provide this. One reason for the digital reality not meeting expectations is mis-timing inside transmission equipment when data is regenerated. When mis-timing becomes large, errors are produced and the system can become unusable. Even at low values of mis-timing, sensitivity to amplitude and phase variations is increased and performance suffers.

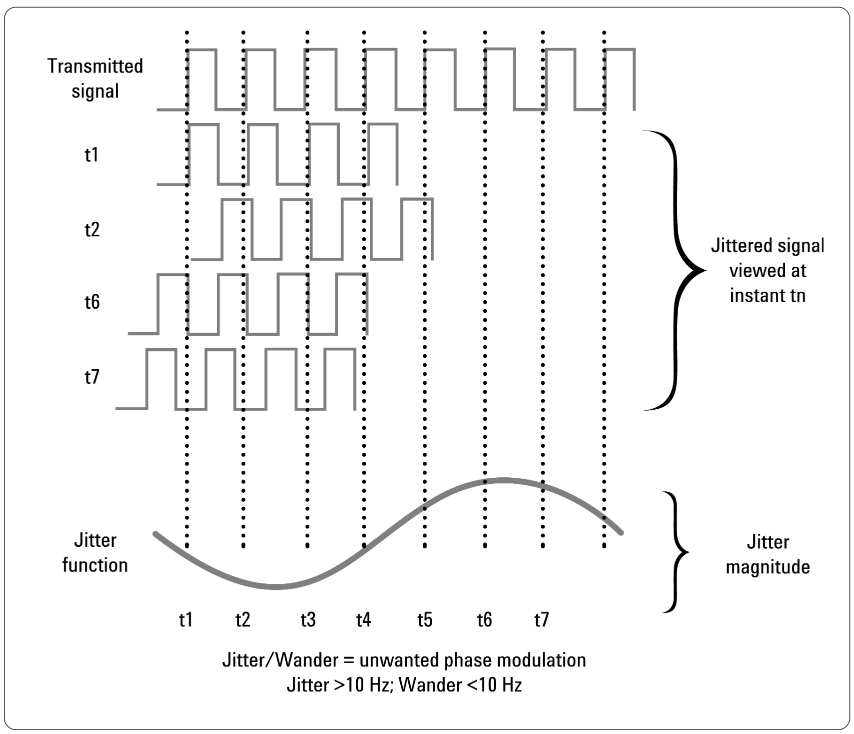


Figure 1.1 Jitter – unwanted phase modulation of a digital signal

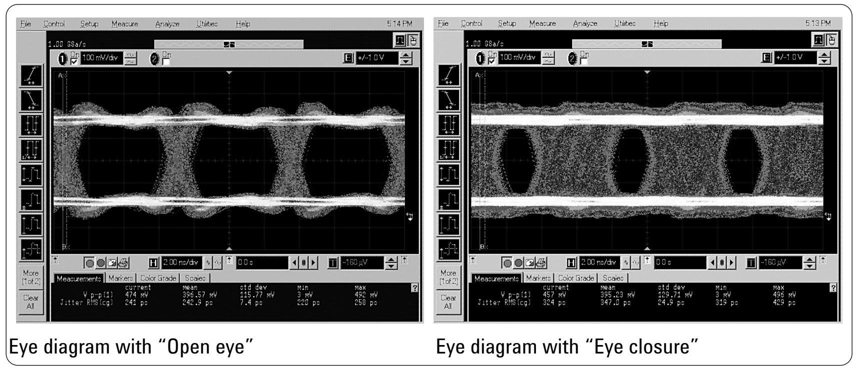


Figure 1.2 The effects of jitter (as viewed on an oscilloscope)

As shown in Figure 1.2, jitter causes eye-closure in the horizontal axis and this prevents correct sampling and ultimately results in bit errors. Even if the jitter doesn't cause errors itself, it reduces the noise margin of the system and makes it more prone to errors.

Mis-timing may be referred to as skew, wander or jitter depending on its frequency band. It may be the result of pattern dependency or due to noise sources such as thermal noise or crosstalk. It can also be inherent in the system design and caused by de-multiplexing (justification) in PDH systems or pointer movements in SDH/SONET systems. However it is caused, every system will generate some degree of mis-timing, and therefore has to operate in its presence.

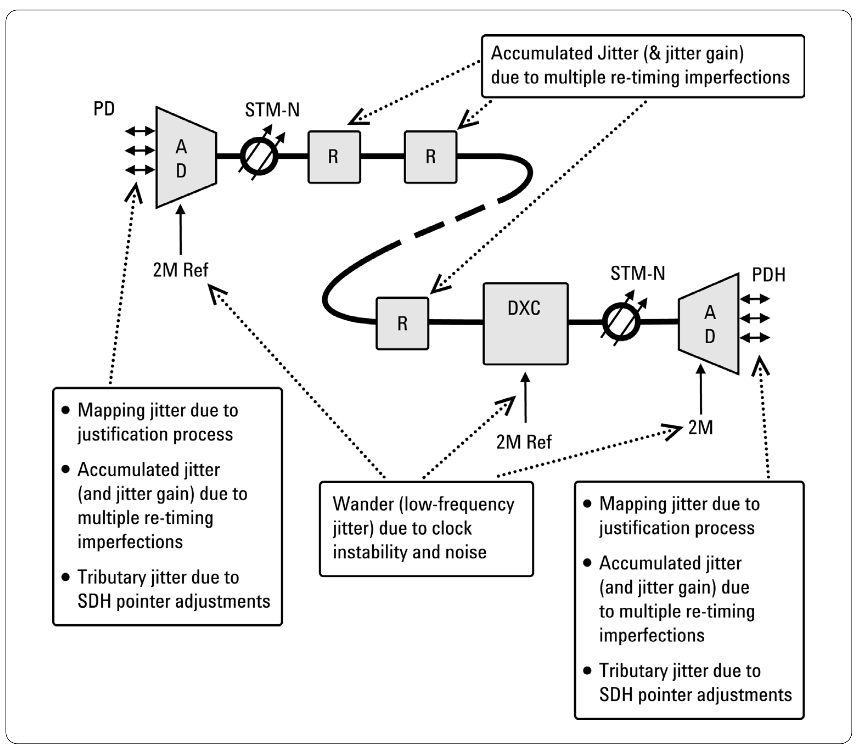


Figure 1.3 Sources of jitter and wander in transmission networks

Static mis-timing in a system isn't normally measurable because it requires access to internal signals to see it. It is detectable through its effect on the equipment's sensitivity to noise and phase variations, and a jitter tolerance test (see later) will highlight that. Timing variation or *skew* across the multiple channels of a bus structured transmission scheme such as Gigabit Ethernet or some VSR (Very Short Reach) optical links can be viewed using a multi-channel oscilloscope. The difficulty in interpreting the results on an oscilloscope, when combined with the effect of internal de-skewing circuitry, still makes a jitter tolerance test a useful measure of timing margin and system performance.

Slow variations in signal timing through a system are called wander. Higher speed variations are termed *jitter*. The division between the two is taken at 10 Hz. Wander is measured using a single pole low-pass filter with its –3 dB point at 10 Hz while jitter uses a high-pass filter with the same –3 dB frequency.

A fundamental operation in every digital transmission system is to receive a degraded signal and regenerate it. All high capacity systems transmit only a suitably coded data signal, and the first task of a regenerator is to produce a local clock signal from that data. There are two contradictory requirements. First, the local clock should be stable for onward transmission and easier aggregation with other data sources. Second, the local clock should track incoming phase variations of the data signal so that as the optimum sampling point for the input data varies, the clock tracks it. This leads to the danger of phase variations building up as a signal traverses a network and each regenerator in turn attempts to track incoming phase variations. In practice the design of a clock recovery circuit is a compromise, and different compromises are struck for different systems. Systems with low-bandwidth clock recovery circuits where little of the incoming phase variation is transferred to the output are referred to as high-Q. Systems with wider clock recovery bandwidths are referred to as low-Q. Either can provide satisfactory performance, but mixing them together in the same system will cause problems.

Jitter measurements

Jitter performance of transmission systems is mandated by standards written by bodies such as the ITU-T and Telcordia (formerly Bellcore). These are then referenced by equipment specifications and usually taken as a minimum requirement. There are three measurements that define the jitter performance of a transmission system, as shown in Figure 1.4, and specifications and standards can be expected to refer to all three:

- Output jitter
- Jitter tolerance
- Jitter transfer

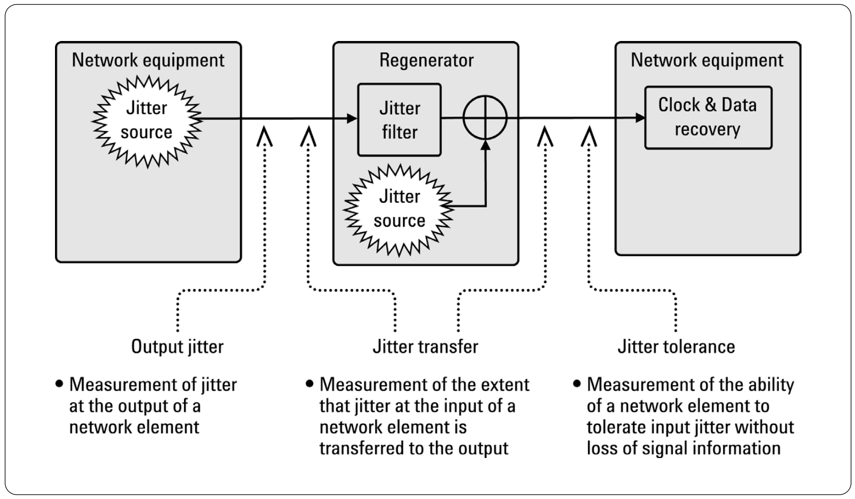


Figure 1.4 Jitter measurement categories

Output jitter is a measurement of the jitter present on an output from a system. This could have been generated within a single piece of equipment (jitter generation or intrinsic jitter), or may have built up as the signal traversed a large network (network jitter). It is specified in unit intervals and the result expressed as a Root Mean Square (RMS), or peak-to-peak value. RMS values give information about the total amount of average jitter present, while peak-to-peak results tell more about the effect on performance, as it is the extremes that can cause errors. While jitter is defined as any phase variations above 10 Hz, most measurements use additional high-pass and low-pass filters, and some systems define more than one set.

Output jitter results are also strongly influenced by the data being carried. Test results can vary widely between those for a simple repetitive pattern such as 1010, and those for a complex PRBS (Pseudo Random Binary Sequence). Similarly, a SDH/SONET system with its 8 kHz frame structure and block of framing bytes can create yet another set of results. It is crucial when measuring and specifying jitter generation to define the data being transmitted, and especially so when comparing specifications and results.

Jitter tolerance is a measurement to check the resilience of equipment to input jitter. A signal is generated with added sinusoidal jitter and applied to the DUT (Device Under Test). At each jitter frequency, the amplitude of the jitter is increased until transmission errors are detected. Alternatively, a specified level of input jitter is generated and error-free operation checked. In the real world, jitter is unlikely to be sinusoidal, but it is easy to generate and gives repeatable results. It allows results for different systems to be compared and for system specifications to be written, usually in the form of a jitter tolerance mask.

Jitter transfer is a measure of how much jitter is transferred between input and output of network equipment. As mentioned earlier, this is a function of jitter frequency and the type of clock recovery used. As a signal traverses a network, the jitter generated by each piece of equipment becomes the input jitter to the following equipment. If this jitter is amplified as it passes through the network, then it could exceed the jitter tolerance of subsequent equipment. To avoid this, a jitter transfer function is specified for equipment, typically allowing a maximum of 0.1 dB jitter gain.

Wander measurements

A different set of measurements is used to characterize wander, the longer-term phase variations ranging from 10 Hz down to micro-Hertz and below. While jitter is normally measured with reference to a clock extracted from the data signal, wander is measured against an external reference clock . The fundamental measurement is of Time Interval Error (TIE). This represents the time deviation of the clock signal under test relative to the reference source.

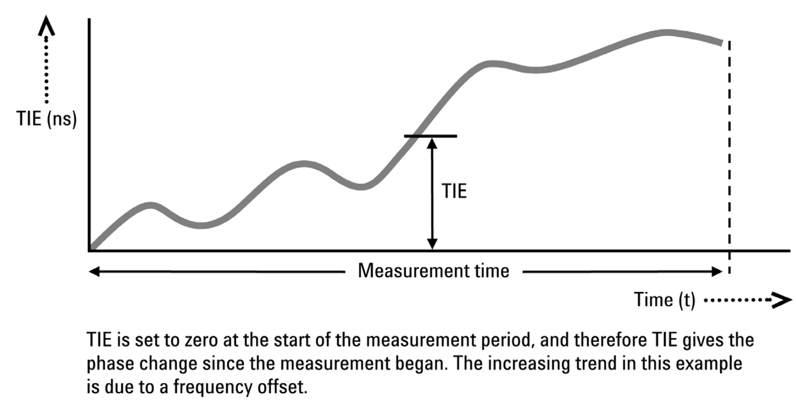


Figure 1.5 Example TIE wander measurement

Several results requiring intensive computation can be calculated from TIE according to the ITU-T G-series recommendations:

- MTIE (Maximum Time Interval Error): The peak-to-peak variation of TIE within a defined observation interval τ.
- TDEV (Time Deviation): A measure of the spectral content of wander and again is a function of the observation interval τ.
- Frequency Offset: A measure of the degree to which the clock frequency deviates from its ideal value.
- Frequency Drift Rate: A measure of how the frequency offset changes with time (i.e. frequency stability).

These wander results are usually obtained by processing the TIE samples from the measurement equipment by a program running on a PC. This can be done in real time or accumulated results post-processed at a later date and/or in a different location.

Jitter test equipment

It is possible to measure or estimate jitter using general-purpose test equipment such as oscilloscopes and spectrum analyzers. While these may give useful information, conforming to the test conditions required by the jitter standards generally isn't possible. To do that, a purpose designed jitter measurement system, which provides wide-band phase demodulation of the input jittered signal, is required. The demodulated signal is then filtered by high-pass and low-pass filters, and processed to give peak-to-peak and RMS jitter measurements.

Test equipment standards

The ITU-T, as well as writing recommendations for operational equipment (e.g. the G-series recommendations for digital transmission equipment), writes specifications for measuring equipment - the O-series recommendations. There are two relevant recommendations for jitter and wander measurements, O.171 and O.172. The main differences between them are that O.171 covers PDH systems whereas O.172 is primarily concerned with SDH systems. O.172 also addresses requirements for SDH tributaries and covers new test applications required in synchronous systems (e.g. pointer jitter). O.172 also, in some cases, has tighter requirements for measurement accuracy.

Conclusions

With the introduction of SONET/SDH and dense high-capacity networks, the sources and consequences of jitter have become more complex. Specifications, particularly for intrinsic jitter are more exacting and the measurement pitfalls more prevalent. The subsequent papers in this booklet discuss these issues.

Further reading

For more information on the basics of jitter and wander, refer to the following publications:

- "Timing and Delay Jitter" by David Robertson, Chapter 23, pp. 489-519, Communications Network Test and Measurement Handbook, Coombs and Coombs, McGraw-Hill 1998, ISBN 0-07-012617-8.

- Agilent application note AN 1432 (Pub No. 5988-8425EN), "Jitter Analysis Techniques for High Data Rates".

2. Jitter Standards and Applications

Abstract: Agreed international standards for jitter and wander are essential for equipment and network interoperability as well as maintaining quality of service. This paper provides an overview of jitter standards in North America, Europe and through the ITU-T, including the standards applicable to test equipment. The objective is to guide the reader on which standards are relevant to a particular equipment or application.

Introduction

Jitter is always present within devices, systems and networks to a certain degree. In order to ensure interoperability between devices and minimize signal degradation due to jitter accumulation across long distances, it is important that there are limits set on the maximum level of jitter present at an output interface and the minimum level that can be tolerated at an input. Adherence to these limits will ensure interworking between different vendor equipment and networks, as well as providing the basis for demarcation.

These limits have been determined by both Telcordia and ITU-T and have been implemented in several standards for different types of device and interface. The very fact that there are so many interface and device types, means that finding the relevant jitter standard for a specific application can be a bewildering process! The objective of this paper is to highlight which standards apply to which types of device.

Standards bodies

There are four main standards bodies concerned with the efficient transportation of communications traffic. These are Telcordia (formerly Bellcore) and the American National Standards Institute (ANSI) in North America, and the International Telecommunication Union (ITU) and the European Telecommunications Standards Institute (ETSI) in Europe. Within the ITU, standards for wireline communications are laid down by ITU Telecommunication Standardization Sector (ITU-T).

ITU-T

ITU-T have two distinct series of recommendations that govern all jitter requirements, namely the G.xxx and O.xxx series.

ITU-T G-series recommendations form the basis for all network, system and Network Element (NE) standards, but are not limited to simply recommending jitter performance levels. The scope of the G-series runs from definitions used in telephone connections to setting transmission performance characteristics, and covers both synchronous and asynchronous line rates.

This series of standards exists to enable Network Equipment Manufacturers (NEMs) to produce equipment for transmission systems and networks which will interoperate with equipment from other vendors, thus providing an open interface for exchanging traffic.

ITU-T O-series recommendations define the required specification of test instruments to enable NEMs to test and verify the performance of their products. For test equipment, it is common to see jitter performance specified as being "compliant with ITU-T O.172".
This simply illustrates that the test equipment in question meets the recommendations laid down by ITU-T. NEMs and service providers do not need to test to any recommendation in the O-series as it does not apply to NEs or networks.

Telcordia

Telcordia maintain fewer standards than the ITU-T, but the information is no less relevant. In terms of jitter within networks and NEs, the two primary publications from Telcordia are GR-253-CORE "Synchronous Optical Network (SONET) Transport Systems: Common Generic Criteria", and GR-499-CORE "Transport Systems Generic Requirements (TSGR): Common Requirements". Up until recently GR-1377-CORE provided information for networks and equipment running at OC-192, but this has since been superceded by the latest revision of GR-253-CORE, which now includes these requirements.

These standards define jitter requirements in terms of SONET equipment interfaces. Every NE interface falls into one of two categories:

- Category I – Asynchronous DSn interfaces to SONET.
- Category II – OC-n, STS-n electrical and synchronous DS1 interfaces to SONET NEs.

GR-253-CORE details jitter performance requirements for "Category I" and "Category II" interfaces for SONET NEs. Non-SONET (T-carrier/DSn) NEs are covered by GR-499-CORE.

European Telecommunications Standards Institute

ETSI provides a forum for service providers, NEMs, research bodies or any other relevant parties to discuss and lay down criteria for all aspects of telecommunications. Topics for standardization are raised by the Institute's members, rather than being directly required by ETSI itself.

In terms of jitter, the two primary specifications developed by ETSI are EN 300-462-3-1, which deals with jitter and wander in synchronization networks, and EN 302-084, which deals with jitter and wander in transport networks. There is very little difference between the limits specified in EN 302-084 and the limits in ITU-T G-series recommendations. Due to the coverage of ITU-T standards, it is more often these that are specified. On the other hand, ETSI EN 300-462-3-1 is more comprehensive on synchronization than ITU-T recommendations, and so tends to be singled out as the benchmark recommendation for this type of network.

American National Standards Institute

ANSI is similar in a way to ETSI as it does not generate standards itself, but encourages development by working to establish agreements between qualified working groups such as the Institute of Electrical and Electronics Engineers (IEEE), the Telecommunications Industry Association (TIA) etc. Although recommendations from Telcordia are quoted more often than ANSI, Telcordia standards are often seen to change to reflect alterations and updates made by ANSI. The key ANSI standard for SONET is T1.105.03 (1994) with later supplements A and B. The standard is currently undergoing revision in the T1X1.3 committee to combine these supplements and include jitter specifications for OC-192. T1X1.3 is also making contributions to the new ITU-T OTN jitter standard (G.8251 or G.otnjit) in Study Group 15, Question 13.

ITU-T standards summary

The ITU-T follows a set process during the development of its recommendations. The procedure takes recommendations from “to be determined”, at which point the specific criteria are established and put into draft form. The newly formed proposals go through an approval process and are then put together into a “pre-published version”. By the time a standard reaches this stage it is often made available to the general marketplace. Although by this stage the limits and criteria are almost completely set, there still remains some scope for change. Once the pre-published version has been approved by ITU-T members, it is made into a full standard, published and made openly available. It will also be noted if any new standard supercedes or replaces any existing standards.

Confusion often arises when a change is made between the release of the pre-published version and the release of the fully approved version. Due to the fact that changes can happen, we always recommend that the latest edition of any standard be referred to during design, development and testing. The current status of any ITU-T standard can be accessed using the URL given at the end of this paper. The website also provides a download of the current standard for a small fee.

Operational standards vs equipment standards

The G-series can be roughly split into two categories:

- Line equipment standards
- Network standards

Line equipment standards specify the requirements for equipment in a network running at any synchronous or asynchronous rate. Network standards tend to take the jitter characteristics of whole networks into account and often specify long-term objectives for cumulative jitter. Many of the tests required for network standards compliance can be performed as in-service tests over a duration of time which would not be appropriate in the production line test environment.

ITU-T recommendations are briefly summarized below, however for full details please refer to the most recent edition of the full specification. While the standards have been categorized as “Line Equipment” or “Network” based, please bear in mind that many of them contain sections that may apply equally to either scenario.

Line equipment standards

ITU-T Rec.	Title	Description
G.735	Characteristics of Primary PCM Multiplex Equipment Operating at 2048 kb/s and Offering Synchronous Digital Access at 384 kb/s and/or 64 kb/s.	G.735 offers criteria for use when testing equipment, which processes PDH signals operating at 2 Mb/s and has tributaries at 64 or 384 kb/s. G.742 and G.751 also contain similar PDH information but at different rates.
G.783	Characteristics of SDH Equipment Functional Blocks	The latest edition of G.783 replaces the 1994 versions of ITU-T G.781, G.782, G.783 and G.958. G.783 specifies the required jitter and wander criteria for all SDH and PDH interfaces on an SDH network element. It makes reference to jitter tolerance, jitter transfer, jitter generation, jitter resulting from pointer adjustments and jitter resulting from tributary mapping.
G.813	Timing Characteristics of SDH Equipment Slave Clocks	This publication specifies the performance requirements of clocks used inside SDH NEs. While it is not strictly speaking a NE itself, its performance is inextricably linked with the performance of a NE. It specifies parameters for clocks used in SDH/2 Mb/s systems and clocks used in SONET/ 1.5 Mb/s systems.

Table 2.1 Line equipment standards

Network standards

ITU-T Rec.	Title	Description
G.822	Controlled Slip Rate Objectives on an International Digital Connection.	This publication sets the requirements for the overall service performance level at 2 Mb/s. This then allows the wander limit and clock requirements within the network to be determined.
G.823	The Control of Jitter and Wander within Digital Networks which are based on the 2048 kb/s hierarchy.	This standard sets network performance expectations for PDH interfaces outside N. America. The parameters covered in this specification include output jitter and input jitter/wander tolerance.
G.824	The control of Jitter and Wander within Digital Networks which are based on the 1544 kb/s hierarchy.	This is almost identical in nature to G.823, but specifies the network perfor-mance parameters for PDH interfaces in N. America.
G.825	The Control of Jitter and Wander within Digital Networks which are based on the Synchronous Digital Hierarchy.	G.825 is the equivalent of G.823 and G.824 for any SDH interface.
G.8251 removed G.OTNJIT ITU-T SG15 Question 13	The Control of Jitter and Wander within the Optical Transport Network (OTN).	This recommendation is still under develop-ment by ITU-T, but is currently available as a pre-published version. It specifies jitter/ wander parameters for networks based on the ITU-T G.709 Digital Wrapper.

Table 2.2 Network standards

Test equipment standards

While it is never necessary for any NEMs or service providers to test to limits laid down by test equipment standards, it is worthwhile knowing what these documents contain.

ITU-T Rec.	Title	Description
O.171	Timing jitter and wander measuring equipment for digital systems which are based on the Plesiochronous Digital Hierarchy (PDH).	This publication details the minimum requirements for a test instrument in order to test and measure jitter and wander in PDH signals. This specification also contains appendices containing guidelines for the measurement of jitter and wander.
O.172	Jitter and wander measuring equipment for digital systems which are based on the Synchronous Digital Hierarchy (SDH).	This document essentially provides information for SDH jitter and wander testing, equivalent to O.171 for PDH signals.
O.173 (draft)	Jitter measuring equipment for digital systems which are based on the Optical Transport Network (OTN)	Still only at the draft stage, this standard will form the minimum requirements for test equipment designed to test jitter in OTN.

Table 2.3 Test equipment standards

While O.172 is the benchmark jitter and wander standard for today's test equipment, it does have limitations. Some of the limits recommended for network element jitter are very difficult to measure if the test equipment only meets O.172. In certain cases it is preferable that test equipment exceeds O.172 in order to guarantee accuracy[1].

[1] See "What O.172 Doesn't Tell You" and "Measuring 100 mUI p-p Jitter Generation with an O.172 Tester" in this booklet.

The Optical Transport Network (OTN) is developing rapidly, with future application in the all-optical network. The OTN frame structure and layered architecture is similar to SONET/SDH, but has significant differences which affect jitter performance and measurements. The control of jitter and wander within the OTN has thr=erefore been split out into G.8251 for performance limits and O.173 for test equipment[2].

Telcordia standards summary

It is inherently more difficult to split Telcordia's standards by any specific criteria, since there are only two publications which are of interest when testing jitter and wander. These are GR-253-CORE and GR-499-CORE.

GR-253-CORE: SONET Transport Systems: Common Criteria

This defines parameters for SONET network interface jitter criteria and SONET NE jitter criteria. In order to present some consistency, GR-253 refers to different regenerator types as defined by ITU-T's G.958 for 2.5 Gb/s. GR-253 also sets out 10 Gb/s criteria, however only one type of regenerator is assumed.

GR-253 specifies the same jitter criteria for SONET NEs that ITU-T specifies for SDH NEs in G.825:

- Input jitter tolerance
- Jitter transfer
- Jitter generation
- Jitter as a result of tributary mapping
- Jitter as a result of pointer adjustments

In some cases the recommended jitter limits in GR-253 are different to those specified by ITU-T.

GR-499-CORE: Transport Systems Generic Requirements (TSGR): Common Requirements

This provides recommendations for T-carrier/DSn NEs in the same way GR-253 does for SONET NEs. On top of this, it offers explanations of each of the jitter measurements, provides some measurement methodology information and gives guidelines on the cumulative effect of jitter transferring through long lines of regenerators. These explanations are applicable to either SONET or T-carrier/DSn systems and provide a good background for understanding jitter tests.

[2] See "Jitter Testing in the Optical Transport Network (OTN)" in this booklet.

Application of the standards

The above information provides a background of the current standards being enforced. Table 2.4 illustrates the types of test scenarios in which each standard applies. Note that blank sections are either not currently specified, or are not applicable. Each of the standards illustrated contains the relevant limits and masks in order to allow jitter and/or wander tests to be performed. GR-253-CORE provides similar information for SONET devices and networks.

		Output Jitter	Input Jitter Tolerance	Jitter Transfer Function	Pointer/ Mapping Jitter	Output Wander	Input Wander Tolerance	Wander Transfer Function	Transient Response
Network Equipment	SDH (DXC,ADM)	G.813	G.813 G.825		G.783	G.813	G.813 G.825	G.813	G.813
	SDH Regenerator	G.783	G.783	G.783					
	PDH	G.735 G.742 G.751 GR-499	G.832	G.735 G.742 G.751			G.823		
	PRC Clock	G.811				G.811			
	SSU Clock	G.812	G.812			G.812	G.812	G.812	G.812
Network Interface	PDH Transport	G.823	G.823				G.823		
	SONET Transport	GR-253	GR-253	GR-253 GR-499	GR-253	GR-253	GR-253	GR-253	
	SDH Transport	G.825	G.825						
	OTN Transport	G.8251				G.8251			

Table 2.4 Application of standards per jitter test and measurement application

Table 2.5 and Table 2.6 illustrate some of the maximum jitter limits recommended by the above specifications. Please note that while this represents a summary of the standards currently available and in force, these documents are continuously being updated to reflect the latest trends and technologies. Any of these may be superceded by a revised or completely new standard with little or no notice.

Network Interface	Standard	Bit Rate	Jitter Limits	
			Wideband (UI p-p)	High-band (UI p-p)
SDH Transport	ITU-T G.825	STM-1e	1.5	0.075
		STM-1	1.5	0.15
		STM-4	1.5	0.15
		STM-16	1.5	0.15
		STM-64	1.5	0.15
SONET Transport	Telcordia GR-253	OC-1	1.5	0.15
		OC-3	1.5	0.15
		OC-12	1.5	0.15
		OC-48	1.5	0.15
		OC-192	1.5	0.15
PDH Transport	ITU-T G.823	2048 kb/s	1.5	0.2
		8448 kb/s	1.5	0.2
		34368 kb/s	1.5	0.15
		139264 kb/s	1.5	0.075
	ITU-T G.824 Telcordia GR-499	1544 kb/s	5	0.1
		6312 kb/s	3	0.1
		44736 kb/s	5	0.1
OTN Transport	ITU G.8251	OTU1	1.5	0.15
		OTU2	1.5	0.15
		OTU3	6	0.15
Synchronization	ITU-T G.823 ETSI 300 462-3-1	2048 kb/s PRC	0.05	N/A
		2048 kb/s SSU	0.05	N/A
		2048 kb/s SEC	0.5	0.2
		2048 kb/s PDH	1.5	0.2

Table 2.5 Network jitter recommendations for network interfaces

Network Equipment	Standard	Bit Rate	Jitter Limits	
			Wideband (UI p-p)	High-band (UI p-p)
OTN	ITU-T G.8251	OTU1	0.3	0.1
(ODCr)		OTU-2	0.3	0.1
		OTU-3	1.2	0.1
SDH	ITU-T G.783	STM-1	0.5	0.1
(ADM, DXC etc.)	ITU-T G.813	STM-4	0.5	0.1
		STM-16	0.5	0.1
		STM-64	-	0.1
SONET	Telcordia GR-253	OC-1	0.1 (0.01 rms)	-
(ADM, DXC etc.)		OC-3	0.1 (0.01 rms)	-
		OC-12	0.1 (0.01 rms)	-
		OC-48	0.1 (0.01 rms)	-
		OC-192	0.1 (0.01 rms)	-
SDH Regenerators	ITU-T G.783	STM-1	0.3	0.1
		STM-4	0.3	0.1
		STM-16	0.3	0.1
		STM-64	0.3	0.1
PDH	ITU-T G.735	2048 kb/s	0.05	-
	ITU-T G.742	8448 kb/s	0.05	-
	ITU-T G.751	34368 kb/s	0.05	-
		139264 kb/s	0.05	-
	Telcordia GR-499	1544 kb/s	1.0 (0.3 rms)	-
		6312 kb/s	1.0 (0.3 rms)	-
		44736 kb/s	1.0 (0.3 rms)	-

Table 2.6 Jitter generation levels recommended for network equipment.

Conclusions

There is a great potential for the accumulation of jitter to degrade network performance. It is therefore imperative that components within a network and networks as a whole are tested and screened for jitter to ensure that optimum levels of quality can be maintained. Adherence to the standards currently set by ITU-T and Telcordia will help keep jitter levels low enough to ensure signal integrity under normal operation. With this all-encompassing approach, if everyone adheres to the appropriate standards, quality of service need never be degraded by jitter.

Further Information

See the following web sites for further standards information.

ANSI	http://www.ansi.org/
ETSI	http://www.etsi.org/
ITU-T	http://www.itu.int/
Telcordia	http://www.telcordia.com/

3. Jitter Testing in the Optical Transport Network (OTN)

Abstract: This paper gives an overview of the emerging OTN standard and highlights the differences from existing SONET/SDH frame structures. This has implications for jitter performance and measurements. Framing, payload and scrambling all have effects on the jitter performance, which is defined in ITU-T G.8251. The paper discusses these performance limits and shows why it is important to use an OTN jitter test set rather than a SONET/SDH tester. These measurement requirements are being defined in the draft ITU-T Recommendation O.173.

Why develop a new optical transport network?

The SONET/SDH standard has been the bedrock of the world's telecommunications networks for more than 10 years, and continues to be the key technology for multiplexing and transport of TDM traffic, particularly for rates of 2.5 Gb/s (STM-16/OC-48) and below. However, much of the new traffic growth is in wideband data applications for IP routers and ATM/Ethernet switches, that typically interconnect at 2.5 Gb/s or 10 Gb/s using non-channelized payloads over optical transport systems.

These emerging requirements have led to the development of the new ITU-T Recommendation G.709 "Interface for the optical transport network (OTN)" [Reference 1]. It has been developed to cater for the transmission needs of today's wide range of digital services, and to assist network evolution to higher bandwidths and improved network performance. Furthermore, it takes another step towards the all-optical network, and thus opens the door to potentially extensive cost savings. ITU-T G.709 is an international standard, so, unlike SONET and SDH, the ITU-T G.709 optical transport network is truly global. This paper gives a brief introduction to the OTN to provide a background for jitter measurements. For more in-depth information, please refer to Agilent Application Note 1379 "An overview of ITU-T G.709" [Reference 2].

ITU-T G.709 builds on the experience and benefits gained with SONET and SDH, and many of the concepts in ITU-T G.709 have their roots in SONET/SDH. For example, all the standards have a similar layered structure, in-service performance monitoring, protection and other management functions. However, some key elements have been added to continue the cycle of improved performance and reduced cost. These include:

- Management of optical channels in the optical domain without the need to convert into the electrical domain.

- Forward error correction (FEC) to improve error performance and enable longer optical spans without regeneration.

OTN frame format

The ITU-T G.709 frame (Figure 3.1) has three distinct parts, with two that are broadly similar to a SONET/SDH frame:

- Overhead area for operation, administration and maintenance functions
- Payload area for customer data

In addition, the ITU-T G.709 frame also includes a Forward Error Correction (FEC) block.

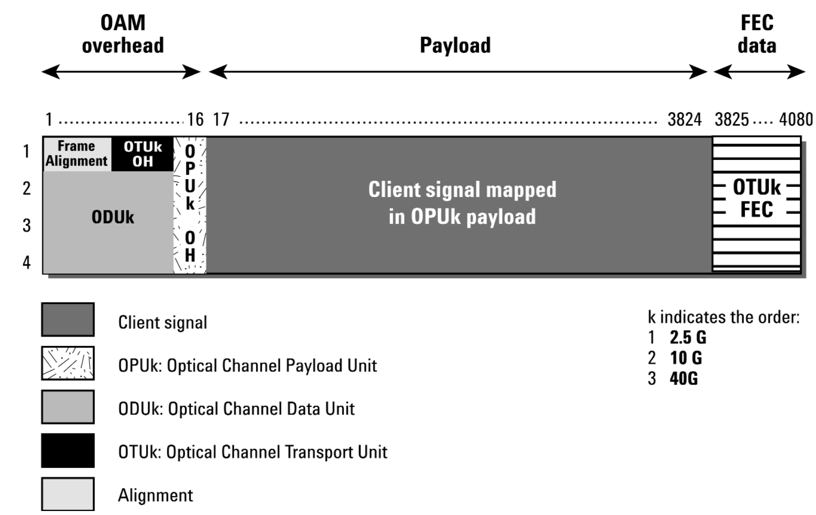


Figure 3.1 OTN frame format

The size of the frame is four rows of 4080 bytes.

There are three line rates currently defined in ITU-T G.709:

- 2.66 Gb/s – Optical channel Transport Unit 1 (OTU1)
- 10.7 Gb/s – Optical channel Transport Unit 2 (OTU2)
- 43 Gb/s – Optical channel Transport Unit 3 (OTU3)

Unlike SONET/SDH (with a fixed frame rate of 8 kHz) , the ITU-T G.709 frame size (4 × 4080) remains the same at each line rate, but the frame rate increases.

The three frame rates are:

- 20.420 kHz for OTU1
- 82.027 kHz for OTU2
- 329.489 kHz for OTU3

This means that to carry one SONET/SDH 10 Gb/s frame, for example, requires approximately eleven OTU2 optical channel frames.

OTN overhead structure

The optical transport module overhead consists of:

- **Optical channel Payload Unit (OPU) overhead**, which is added to the client signal. This overhead describes the payload type and enables positive and negative justification depending on any difference between the clock rates of the client signal and the OTN signal.

- **Optical channel Data Unit (ODU) overhead**, which provides information on tandem connection monitoring (TCM), end-to-end path supervision, and client signal adaptation via the OPU.

- **Optical Transport Unit (OTU) overhead**, which provides supervisory functions and conditions the signal for transport between 3R (reamplification, reshaping and retiming) regeneration points in the OTN. It is at this level that forward error correction is calculated and added to the overhead.

The Optical Channel (OCh) is simply the conversion of the OTU to light. The OCh will typically form one of the wavelengths of a DWDM signal.

Figure 3.2 illustrates this layered structure. The OPU, ODU and OTU layers are electrical, the OCh optical.

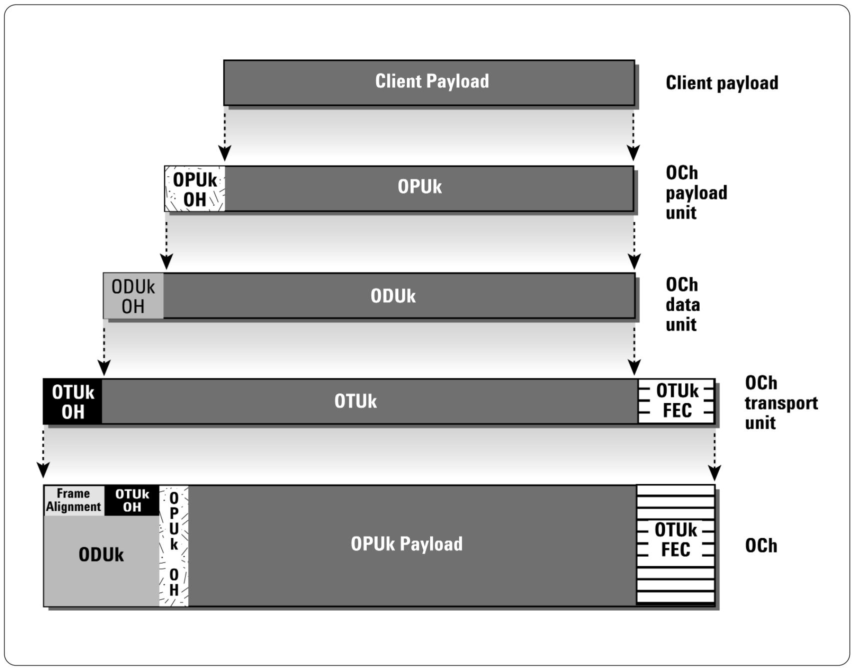


Figure 3.2 OTN layers

Payload mappings

The optical transport hierarchy has been designed to transport a range of payloads. Today, there are defined mappings for SONET/SDH signals, ATM cells, generic framing procedure (GFP) frames, and a Pseudo Random Bit Sequence (PRBS) test pattern.

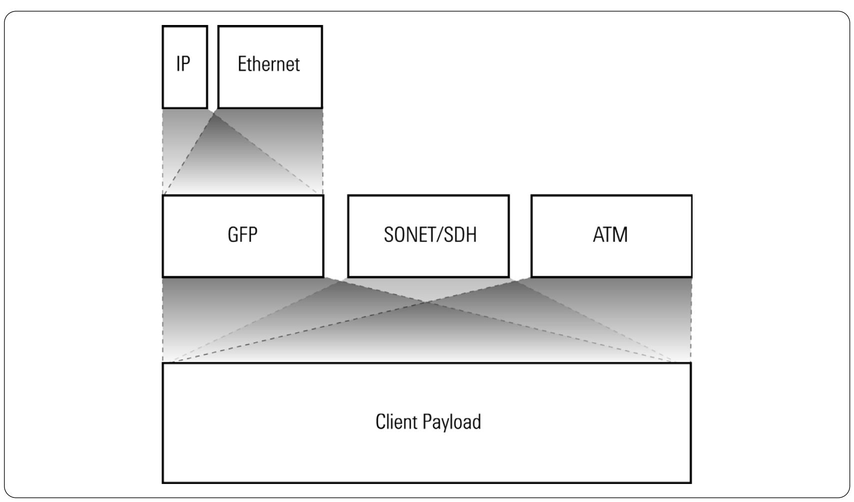


Figure 3.3 Client signal mappings into OPU payload

Forward Error Correction (FEC)

As transmission bit rates increase to 10 Gb/s and beyond, physical parameters of the optical fiber play a more significant part in degrading transmitted pulses of light. At 10 Gb/s, there is less margin in terms of Bit Error Ratio (BER) compared to the same optical power at 2.5 Gb/s. This can be compensated by increasing the transmit power or by reducing the span distance. Another approach is to use Forward Error Correction (FEC) for 10 Gb/s transmission rates and above.

FEC can be used to provide more system margin if the span length remains constant, or to increase the span length with a given BER objective and optical power. FEC detects and corrects errors, effectively delivering a 7 to 8 dB improvement in signal-to-noise ratio or power budget.

The FEC scheme used in the ITU-T G.709 standard is a Reed-Solomon RS(255,239) code. This means that for every 239 bytes of data, an additional 16 bytes (255 – 239 = 16) of data is added for error correction (raising the gross bit rate by around 6.7 percent).

Jitter and wander testing requirements

As explained earlier, the OTN frame format shares many similarities with SONET/SDH, but there are also some differences. This section picks up on some of the key differences in order to explain why specific standards have been developed to address OTN jitter and wander testing requirements.

The paper on jitter generation measurement (Paper 6 in this booklet) describes how jitter measurements are strongly influenced by the data being carried. Framing, scrambling, mapping and pattern all affect the spectral content of jitter. A comparison of these data influences on jitter between SONET/SDH and OTN, is shown in Table 3.1.

	SONET/SDH	**OTN**
Framing	n × A1, n × A2	3 × A1, 3 × A2
Payload	$2^{23} - 1$ PRBS	$2^{31} - 1$ PRBS
Scrambling	$1 + x^6 + x^7$	$1 + x + x^3 + x^{12} + x^{16}$

Table 3.1 Comparison of data influences on jitter between SONET/SDH and OTN

Framing effects

One difference from SONET/SDH is the overhead, which for OTN does not vary in size with line rate, but remains constant.

- The frame alignment word of an OC-n/STM-n frame is composed of $n \times A1$ bytes followed by $n \times A2$ bytes. For OC-192/STM-64, this represents a 384 byte framing word.
- For OTN, the framing word is a fixed 6-byte word ($3 \times A1$ $3 \times A2$).

In both cases, the frame alignment word is unscrambled. For SONET/SDH, the Z0 bytes are also unscrambled.

The framing word for OTN is much smaller, and therefore much less susceptible to jitter experienced in SONET/SDH signals caused by the DC imbalance of the header ($n \times A1$, $n \times A2$, $n \times Z0$).

Payload effects

Digital circuits have pattern-dependent behavior, which is why PRBS sequences with long runs of “ones” and “zeros” are used to create the most stressful scenario for testing.

- For OC-n/STM-n signals, the test pattern defined by ITU-T O.172 [Reference 5] is a bulk-filled concatenated signal, with a $2^{23} - 1$ PRBS.
- ITU-T G.709 defines two test signal mappings into an OPUk:
  - *$2^{31} - 1$ PRBS* test signal
  - *NULL client* which is an OPUk payload signal where the entire payload is filled with an “all-zeros” pattern. The OPUk payload for the NULL mapping consists of 4×3808 bytes.

For OTN application, the $2^{31} - 1$ PRBS test signal will generate a longer “ones”/“zeros” run length (31 “ones”) than the SONET/SDH $2^{23} - 1$ PRBS signal (23 “ones”). The measured peak-to-peak jitter may therefore be higher.

The “ones” or “zeros” run length is also affected by scrambling.

Scrambling effects

Scrambling is used to ensure the signal has sufficient bit timing, by preventing long sequences of “ones”/“zeros”. Different scrambling patterns are used for SONET/SDH and OTN:

- OC-n/STM-n uses a scrambler of sequence length 127 operating at the line rate, with a generating polynomial $1 + x^6 + x^7$. The first row of the OC-n/STM-n (including the A1 and A2 framing bytes) are not scrambled.

For an OC-n/STM-n bulk filled, concatenated signal with a $2^{23} - 1$ PRBS filling the container, scrambling potentially gives a worst-case run of 30 consecutive “ones”/“zeros”.

- The OTN uses a scrambler of sequence length 65535 operating at the OTUk rate, with generating polynomial $1 + x + x^3 + x^{12} + x^{16}$. All bytes of the OTUk frame are scrambled, with the exception of the framing bytes.

For an OTN signal with a $2^{31} - 1$ PRBS filling the OPUk payload, scrambling gives a potential worst-case run of 47 consecutive “ones”/“zeros”.

Clearly, the different scrambling pattern may also result in higher jitter in the case of OTN. In practice this worst-case run may not occur during jitter measurements as the repetition rate of a $2^{47} - 1$ PRBS (the result of a $2^{31} - 1$ test pattern combined with the $2^{16} - 1$ scrambler) at 10.7 Gb/s is around 4 hours.

ITU-T recommendations

This section outlines the relevant standards that contain OTN jitter and wander specifications.

ITU-T G.8251 “The control of jitter and wander within the Optical Transport Network (OTN)”

This standard [Reference 3] specifies network limits for jitter and wander at any OTN interface. Its counterpart in the SDH world is ITU-T G.825 “The control of jitter and wander within digital networks which are based on the Synchronous Digital Hierarchy (SDH)”.

ITU-T O.173 (Draft) "Jitter measuring equipment for digital systems which are based on the Optical Transport Network (OTN)"

This is a new recommendation [Reference 4] dealing with OTN jitter test equipment, currently in the early stages of development. Its counterpart in the SDH world is ITU-T O.172 "Jitter and wander measuring equipment for digital systems which are based on the Synchronous Digital Hierarchy (SDH)" [Reference 5].

ITU-T O.173 covers measurement requirements for OTN line interfaces. Client interfaces, such as SDH line interfaces, are addressed by ITU-T O.172. For a comparison between ITU-T O.172 and O.173, see the paper on jitter standards (Paper 2 in this booklet).

Network interface and network equipment requirements

This section covers the jitter/wander requirements for OTN compliant networks and network equipment.

In an OTN, jitter and wander accumulate on transmission paths according to the jitter and wander generation and transfer characteristics of the network equipment being interconnected. This equipment may be, for example, 3R regenerators, client mappers, and client demappers/ desynchronizers (Figure 3.4).

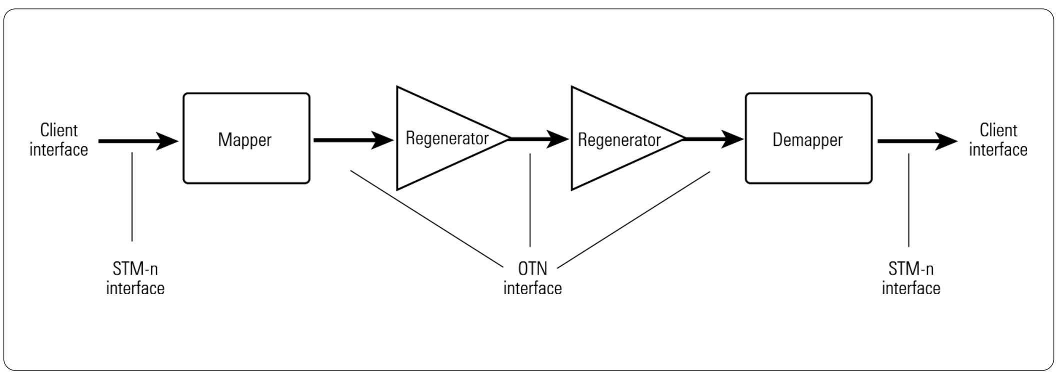


Figure 3.4 OTN network equipment

Network interface limits for jitter and wander

The maximum permissible jitter at any OTUk interface within an OTN is shown in Table 3.2. These figures apply regardless of the amount of equipment preceding the interface.

Interface	Wide-band		High-band	
	Measurement bandwidth (Hz)	Jitter limits (UI p-p)	Measurement bandwidth (Hz)	Jitter limits (UI p-p)
OTU1	5k to 20M	1.5	1M to 20M	0.15
OTU2	20k to 80M	1.5	4M to 80M	0.15
OTU3	20k to 320M	6.0	16M to 320M	0.15

Table 3.2 ITU-T G.8251 network jitter requirements for network interfaces

The wander generation of OTN equipment is expected to be negligible, so there is no specification of network wander limits.

Jitter and wander tolerance for network interfaces

Jitter and wander tolerance for OTN interfaces

The OTUk input ports of all equipment shall tolerate, as a minimum, the input jitter applied according to the mask in Figure 3.5, with values specified in Table 3.3. This ensures that any equipment can be connected to any appropriate interface within a network without risk of errors due to jitter.

OTN interfaces must tolerate jitter and wander at frequencies below f_0, but for practical reasons, it is not required to test below this region. This is because the phase-locked loops which tolerate the sloped region of the mask between f_1 and f_0, will also, by design, tolerate an extension of this mask to frequencies below f_0.

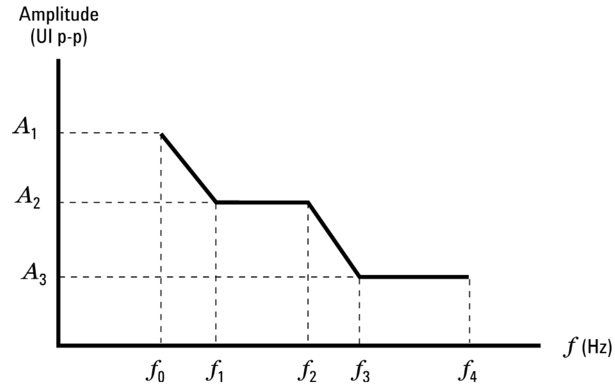


Figure 3.5 ITU-T G.8251 input jitter tolerance limit

Interface	Amplitude (UI p-p)			Frequency (Hz)				
	A_1	A_2	A_3	f_0	f_1	f_2	f_3	f_4
OTU1	15	1.5	0.15	500	5k	100k	1M	20M
OTU2	15	1.5	0.15	2k	20k	400k	4M	80M
OTU3	15	6.0	0.15	8k	20k	400k	16M	320M

Table 3.3 ITU-T G.8251 input jitter tolerance limit

Jitter and wander tolerance for client interfaces

Jitter and wander tolerance requirements and network limits for client interfaces (2.5G, 10G, and 40G) are derived from the corresponding requirements for STM-16 and STM-64 signals, respectively, given in ITU-T G.825. (Jitter and wander tolerance requirements and network limits for STM-256 signals are not currently defined in ITU-T G.825).

Network equipment requirements

This section covers the requirements for OTN compliant network equipment. ITU-T G.8251 Annex A defines network equipment specifications for the ODUk Clock (ODC). The ODC generates timing for the signals produced by the OTN network equipment types. The jitter performance of the ODUk clock can be tested at the OTUk optical interface.

Four ODC types are defined:

- ODCa for asynchronous mapping of clients into ODUk,
- ODCb for bit-synchronous mapping of clients into ODUk,
- ODCr for 3R regeneration
- ODCp for demapping of constant bit rate (CBR) clients.

As ODUk interfaces are not accessible for testing, only the requirements for testing at the OTUk optical interface are considered. (Note: the input and output interfaces depend on clock type.)

Jitter generation (ODCa, ODCb, ODCr, ODCp)

In the absence of input jitter, the intrinsic jitter at output interfaces should not exceed the limits given in Table 3.4 (ODCr) and Table 3.5 (ODCp). There are no jitter generation requirements for ODCa and ODCb, as the ODUk interface is not accessible for testing.

Interface	Wide-band		High-band	
	Measurement bandwidth (Hz)	Jitter limits (UI p-p)	Measurement bandwidth (Hz)	Jitter limits (UI p-p)
OTU1	5k to 20M	0.3	1M to 20M	0.1
OTU2	20k to 80M	0.3	4M to 80M	0.1
OTU3	20k to 320M	1.2	16M to 320M	0.1

Table 3.4 ODCr jitter generation requirements

Interface	Wide-band		High-band	
	Measurement bandwidth (Hz)	Jitter limits (UI p-p)	Measurement bandwidth (Hz)	Jitter limits (UI p-p)
STM-16	5k to 20M	1.0	1M to 20M	0.1
STM-64	20k to 80M	1.0	4M to 80M	0.1
STM-256	20k to 320M	1.0	16M to 320M	0.1

Table 3.5 ODCp jitter generation requirements

Jitter tolerance (ODCa, ODCb, ODCr, ODCp)

Jitter tolerance requirements for ODC types are as follows:

- ODCa - no jitter tolerance requirements, as ODCa is free running.
- ODCb - must satisfy requirements as per STM-16/64/256 client interfaces (see above).
- ODCr - must satisfy requirements as per OTUk interface within an OTN (see above).
- ODCp - must satisfy requirements as per OTUk interface within an OTN (see above).

Jitter transfer (ODCa, ODCb, ODCr, ODCp)

The jitter transfer function shall be under the curve given in Figure 3.6 when input sinusoidal jitter up to the respective masks (defined by jitter tolerance requirements) is applied. The parameters of Figure 3.6 are given in Table 3.6. There are no jitter transfer requirements for ODCa, ODCb, and ODCp, as the ODUk interface is not accessible for testing. This means that jitter transfer testing is only applicable for network equipment where both input and output line interfaces are OTUk.

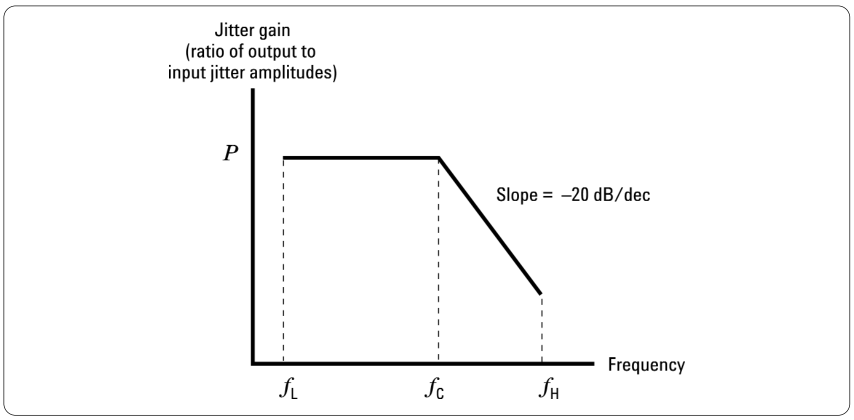


Figure 3.6 ITU-T G.8251 jitter transfer pass mask

OTUk level	f_L (Hz)	f_C (kHz)	f_H (kHz)	P (dB)
OTU1	2.5	250	20	0.1
OTU2	10	1000	80	0.1
OTU3	40	4000	320	0.1

Table 3.6 ODCr jitter transfer requirements

Network synchronization

This section covers some of the possible effects on network synchronization associated with the introduction of the OTN layer.

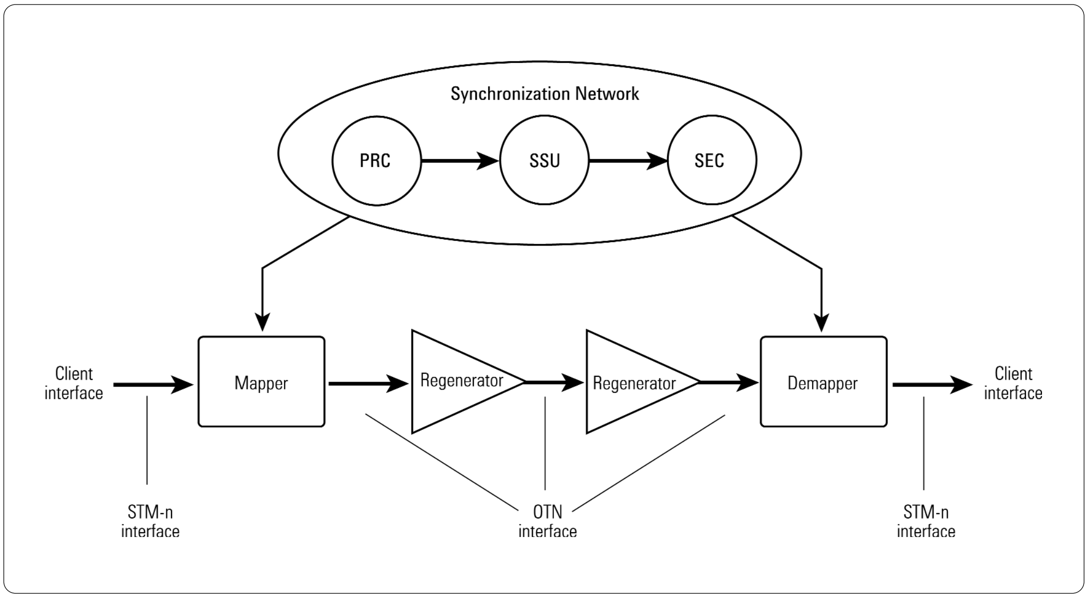


Figure 3.7 OTN/SDH network model

A simple OTN/SDH network model is shown in Figure 3.7. The OTN physical layer will not be used to transport synchronization. Network synchronization distribution is a function of the client layer, for example SDH. As the timing of OTN signals is determined from a separate synchronization reference signal, the OTN signals can have slow phase movements (wander) relative to the reference. In most cases, the wander generation of OTN equipment is expected to be negligible, so there is no specification of wander limits in the published standards.

Conclusions

OTN is the emerging network standard developed for high capacity data over optics with the future application in the all-optical network. OTN frame structure and layered architecture is similar to SONET/SDH, but has significant differences which affect jitter performance and measurements. The performance limits are defined in ITU-T G.8251 and the test equipment specification will be defined in ITU-T O.173, currently in draft form.

References

1. ITU-T Recommendation G.709 (02/01) “Interface for the optical transport network (OTN)”

2. Agilent Application Note 1379 (09/01) 5988-3655EN “An overview of ITU-T G.709”

3. ITU-T Recommendation G.8251 (10/02) “The control of jitter and wander within the Optical Transport Network (OTN)”

4. ITUT Recommendation O.173 (Draft, 10/02) “Jitter measuring equipment for digital systems which are based on the Optical Transport Network (OTN)”

5. ITU-T Recommendation O.172 (03/01) “Jitter and wander measuring equipment for digital systems which are based on the Synchronous Digital Hierarchy (SDH)”

4. Jitter Tolerance Measurements

Abstract: This paper examines the important topic of jitter tolerance, Maximum Tolerable Jitter (MTJ) and the various jitter tolerance masks referenced in the standards. Different measurement methods including "Power Penalty" and "Onset of Errors" are discussed along with advice on accuracy and measurement time. The paper concludes with case studies and advice on the causes of degraded performance.

Introduction

Jitter tolerance measurements are required to confirm that Network Elements (NEs) in the transmission system can operate error-free in the presence of the worst-case jitter from preceding sections. These tests can be a simple confirmation that the minimum requirements are met or may be more comprehensive tests where the additional margin and Maximum Tolerable Jitter (MTJ) are measured. For convenience in testing, the recommendations are based on the use of sinusoidal jitter to simulate the phase variations of a preceding interface. In practice, the jitter in transmission systems carrying real traffic is more like random noise.

Checking a DUT meets tolerance mask

To confirm the Device Under Test (DUT) meets the minimum or lower limit of the jitter tolerance requirement, a sweep of the jitter mask (jitter amplitude in UI versus jitter frequency) is applied at the input while checking there are no transmission errors at the output. The transmission signal is generated with jitter at the mask levels (typically defined in the appropriate standard) and is swept in a controlled phase-transient-free sequence. No transmission errors should be recorded at the output, thus confirming the DUT meets the applied jitter mask.

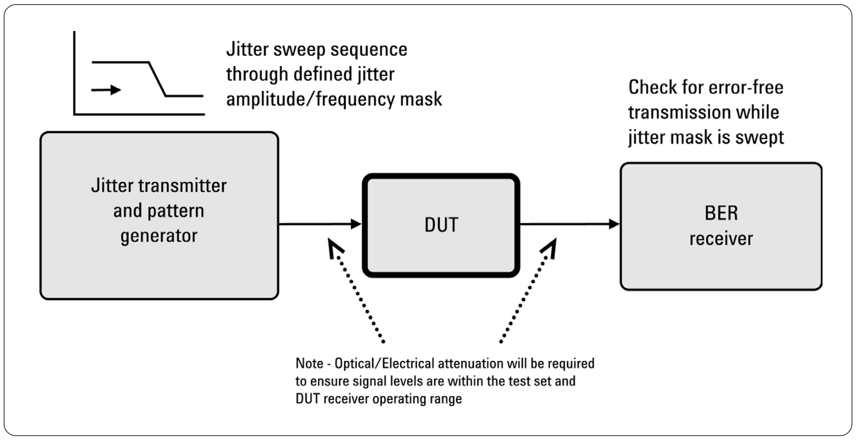


Figure 4.1 Sample DUT swept mask check

The swept-mask technique can also be used to confirm a specific margin on the standards recommendation by generating, for example, a 10%, or 20% high version of the regulation mask. Alternatively, a user-defined mask can be tailored to provide an appropriate DUT benchmark test based on typical device margins. Manual, spot jitter-mask frequency tests can be useful for identifying any swept-mask region that needs further analysis.

The swept-mask method can also be applied between network interfaces with the test set in “Through Mode” operation, while using the NE itself or an additional BER receiver to check for errors, as shown in Figure 4.2.

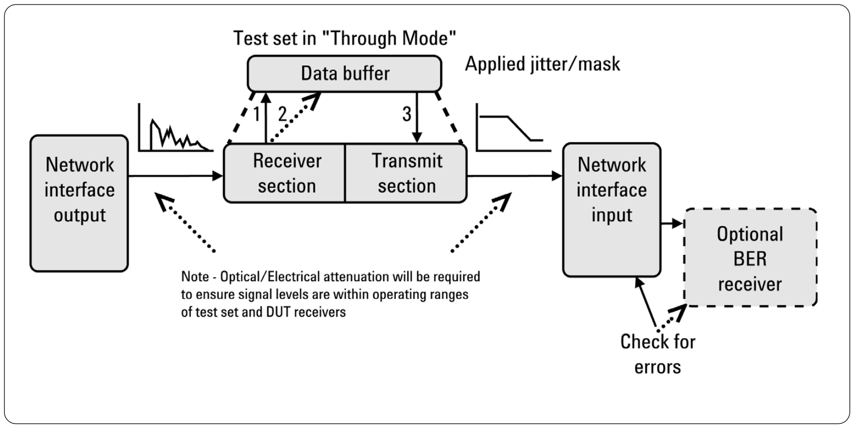


Figure 4.2 “Through Mode” DUT swept-mask check

In “Through Mode”, the test set in effect acts as a de-jitterizer. The data from an output interface of the NE is clocked into a through-mode data buffer in the test set (arrow 1), and is clocked-out using a de-jittered recovered clock (arrow 3). To ensure the buffer does not overflow/ underflow it must be large enough to accommodate the jitter on the received signal and the maximum jitter to be applied. The data is centered in the store initially, when the test set locks onto incoming signal, to ensure sufficient buffer margin (arrow 2). Note, very low frequency phase variations (wander) are not removed as this would be within the through-mode bandwidth.

Jitter, or a fixed jitter-mask sweep, can be applied to the de-jittered signal by the transmit section of the test set to stress/check the NE input interface. Note that the test set’s jitter receiver is also operational in “Through Mode”, and therefore the output interface jitter level can also be measured.

Measuring maximum tolerable jitter

Where actual measurement of the DUT Maximum Tolerable Jitter (MTJ) and margin on the minimum tolerance is required, an Automatic Jitter Tolerance test mode is useful.

A standard jitter tolerance mask is used as a comparison/reference during the test. Jitter is applied sequentially at a number of frequency points in the mask range. The receiver checks for the onset of transmission errors and increment/decrements the jitter amplitude using a control algorithm, re-testing for errors at one frequency point until the MTJ threshold is found. This is repeated for each subsequent jitter frequency test point to provide a plot of the DUT tolerance.

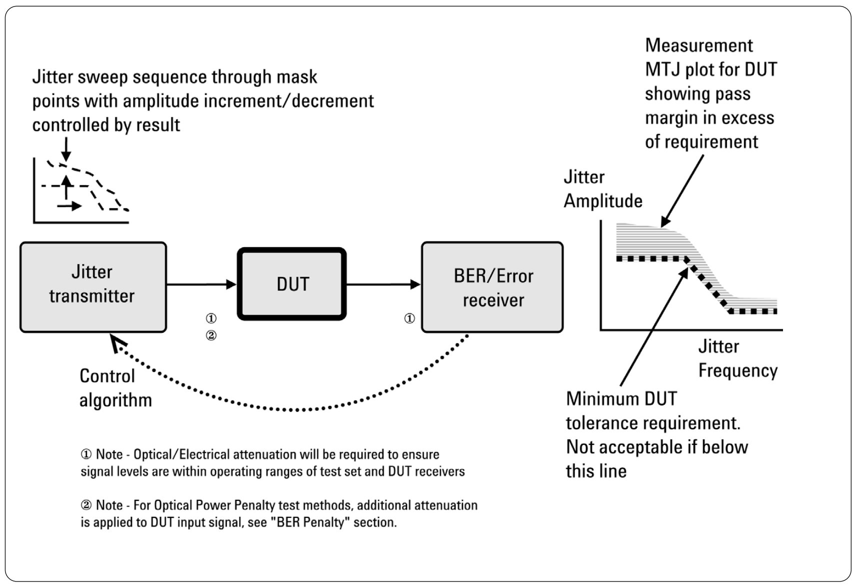


Figure 4.3 Maximum Tolerable Jitter Test

The control algorithm used to detect the maximum tolerance employs successive approximation to home-in on the value. After any change to jitter amplitude in the search, or when moving to the next frequency point, a measurement settling time is applied. This is followed by a gate time during which the error measurement takes place. The purpose of the settling time is to allow the DUT to settle after changes, and before making the measurement. This is important, as some DUTs can take time to recover from error conditions caused when jitter exceeds the margin during the search for actual threshold.

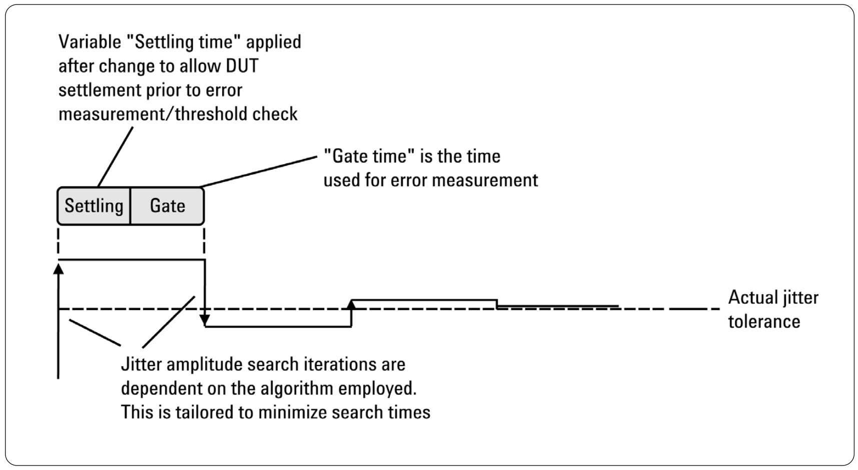


Figure 4.4 Jitter amplitude search process

The transmission error measurement is made during the gate time period, and the threshold used for detection of jitter tolerance may simply be the instance of any transmission errors (or alarms), or alternatively a defined number of bit errors. The gate time and number of bit errors used can be varied, allowing a derived bit error ratio to be used. Alarm detection suits situations where the DUT can report alarms like Remote Error Indication (REI), but access or return loop of the bit-errored signal is not possible.

The different detection threshold methods allow the jitter tolerance of a device to be measured and expressed in either of two ways, using “Onset of Errors” or “Bit Error Ratio (BER) Penalty” methods. The reason that these two techniques are used is to explore the different circuit characteristics in DUTs that primarily limit jitter tolerance:

- Clock recovery performance and jitter tolerance in the presence of noise.
- Size of buffer stores, and the capacity of justification/bit-stuffing process.

While either method is likely to give similar tolerance result plots, for optical interfaces the “BER Penalty” method is recommended by most network equipment standards, as it is considered best for testing the clock recovery tolerance. The “Onset of Errors” technique is best where buffer-store size and justification/bit-stuff capacity are concerned.

"Onset of Errors" method

The "Onset of Errors" method is usually performed with the signal level into the DUT at nominal levels. Recommendations in ITU-T O.171 suggest using a tolerance detection threshold of two Errored Seconds (ES) in a 30 second measurement period for each incremental jitter amplitude change when searching for the actual tolerance at each frequency. This can be somewhat time consuming in practice. In general, using a single error in successive one-second measurement gate periods, gives sufficient accuracy, assuming reasonable device margins.

"BER Penalty" method

The "BER Penalty" method tests the jitter tolerance under worst-case DUT sensitivity conditions, by using a degraded Signal-to-Noise Ratio (SNR). With no jitter applied, the DUT input (optical) power level is attenuated until a particular measured BER is reached. Recommendations often suggest some very low error ratios for this purpose, but derived ratios of $4E^{-10}$ to $4E^{-8}$ (approximately 1 to 100 errors per second, for OC-48/STM-16) are normally used in practice.

Once the error ratio is set, the SNR is improved 1 dB by reducing attenuation (usually a 1 dB power penalty is used) with a resultant reduction in BER. Jitter is then applied using a successive approximation amplitude controlled technique, as previously described, to find the level of jitter that causes the same BER as the 1 dB power penalty, and this is repeated at other jitter frequencies to complete the plot.

Measurement time considerations

The time required for a MTJ measurement plot can be difficult to assess because it is dependent on several factors:

- Number of points required in plot – reduce where possible depending on the plot accuracy required.
- Errors/error ratio being used as a threshold – ensure a realistic threshold is used.
- Payload size/mapping over which errors are monitored – where possible, to reduce time to error threshold detection, use maximum bulk payloads and longest Pseudo Random Binary Sequence (PRBS) loading available.
- Settling period – keep settling time to minimum, only increase as required for DUT settlement if test results are being impacted. (See example plots later in this paper).
- Gate period, payload size and threshold error count are interrelated and therefore their combination affects the relative error ratio used in jitter tolerance detection.
- Number of jitter amplitude iterations used in search of threshold is variable and can be largely influenced by DUT to DUT differences.
- The "BER Penalty" method performed at minimal signal levels is likely to take longer than "Onset of Errors" method at nominal signal level.

The above points need to be taken into account, as well as the purpose of the testing. For example, maximum plot resolution and statistical accuracy would be considered more important than measurement time for a design/development application.

Where measurement time/completion and overall DUT test times are important, as in production test, a remote end-of-measurement status flag can be useful for programmed control.

As with the simple swept fixed-mask method, it is also possible to create user-defined masks for MTJ, for example to reduce test frequency range or focus on an area of interest.

Other measurement consideration and advice

When using the “BER Power Penalty” method, the optical signal will almost certainly be attenuated to a level below the DUT minimum specified sensitivity. This could cause some DUTs to invoke Signal Loss/AIS conditions etc. and therefore may restrict the choice of BER level that can be used in the test.

It is important that the DUT output also meets any jitter transfer and jitter generation (output jitter) requirements applicable to the DUT, as any excessive phase noise could affect the accuracy of the jitter tolerance measurements.

The signal applied to the test set input should be set at nominal input levels. This ensures that tolerance measurement being performed on the DUT is not compromised or restricted by the test set being operated under marginal signal conditions.

If necessary, the test set’s own jitter tolerance margin and measurement headroom can be checked by performing a back-to-back check without the DUT. The margin available will depend on the maximum jitter generation capability and MTJ of the receiver.

Jitter tolerance masks

A variety of standard jitter tolerance masks is shown in Figure 4.5, with an explanation of the steps in the jitter mask characteristics. While there is general agreement between the various standards, there are differences and the standards are always under review and subject to change, as explained elsewhere in this booklet.

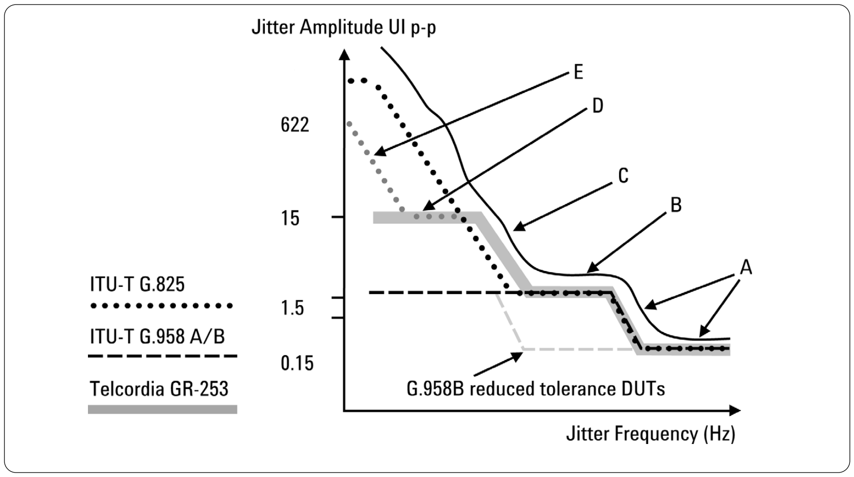


Figure 4.5 Typical SDH/SONET regulation mask(s) and tolerance plot for OC-48/STM-16

Typical SONET/SDH interface jitter tolerance masks and MTJ plot characteristics

Tolerance masks/plots have a characteristic slope/stepped form, which relates to the bandwidth limitations of DUT circuitry sections.

A This section shows the input clock recovery eye width limitations where jitter frequency is higher (flat portion) than the clock-recovery bandwidth. The sloped section is within the clock-recovery bandwidth, therefore jitter at lower-frequencies can be accommodated/tracked up to the peak frequency deviation of the clock-recovery circuit.

B Shows the buffer eye width of some regenerators where the output timing uses a Phase-Locked Loop (PLL). This portion is above the PLL bandwidth where jitter is absorbed in the buffer.

C This slope is within the bandwidth of a regenerator PLL output and therefore jitter at lower frequency can be accommodated/tracked up to the clock-recovery's peak frequency deviation.

D Shows the Telcordia GR-253 requirement low-frequency jitter limits

E Shows the different ITU-T G.825 low frequency jitter and wander limits for systems based on 1544 kb/s hierarchy.

Results requiring explanation/interpretation

This section gives some examples of results that may require explanation. Unexpected results, failure to meet specification or abnormal measurements can occur depending on the type of DUT, the bandwidth limits and measurement set-up. The following examples include an explanation of the plot characteristics and interpretation of the likely situation with advice on the cause.

Case 1 – Clock recovery bandwidth limitation or incorrect mask

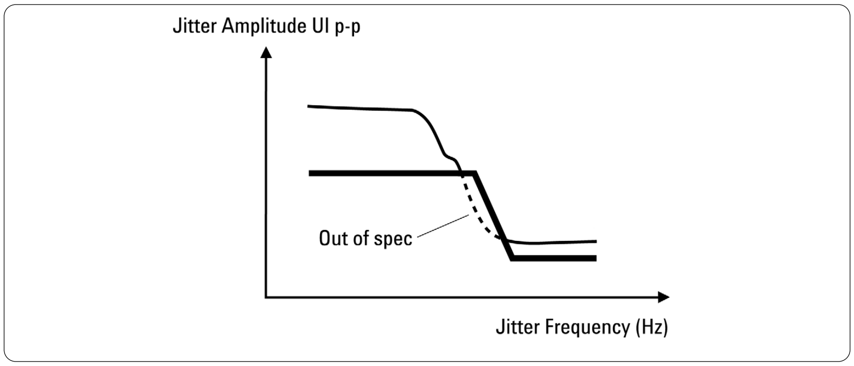


Figure 4.6 Clock recovery bandwidth limitation or incorrect test mask

In this example, the jitter tolerance measurement fails the mask at the corner frequency (Figure 4.6). This may be due to insufficient clock recovery bandwidth to meet the mask, or an incorrect mask being used in DUT test. The slope shows the peak frequency deviation limit of the re-timing circuit.

Case 2 - Clock recovery appears to have excessive bandwidth margin

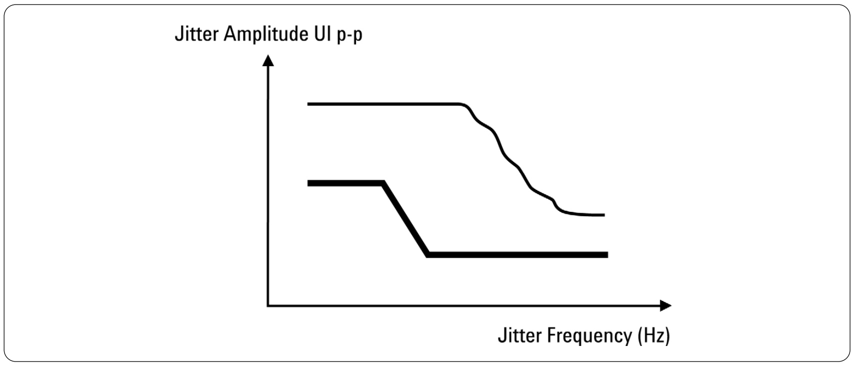


Figure 4.7 Large margin on requirement

If the DUT appears to have a very large margin (Figure 4.7) on the required jitter tolerance mask, it is possible that the wrong mask or test pattern is being used. Some devices such as a test instrument receiver may exhibit very high tolerance.

Case 3 - Unexpected/occasional fail at some mask frequencies

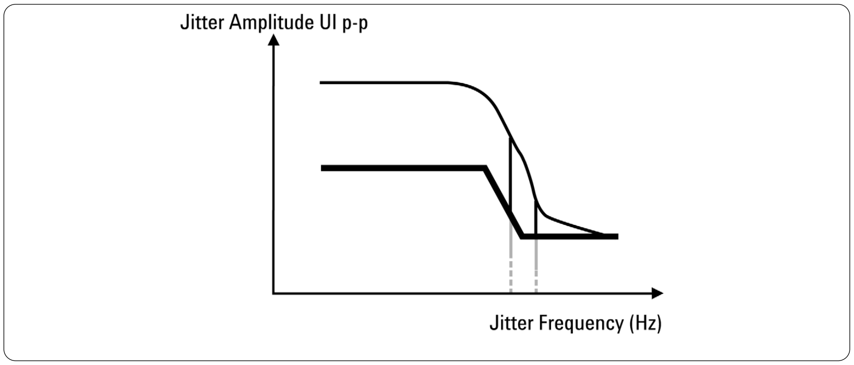


Figure 4.8 Fails mask occasionally at some point

This can happen if the DUT requires extra settlement time after jitter amplitude changes (for example some DUTs may lose frame and have a long re-frame time). Try repeating the measurement with additional settling time and if this does not resolve problem then use a user-defined mask or spot jitter test to investigate.

If a "Power Penalty" method is being used, check the statistical error rate/measurement time threshold is not compromised by background errors. Repeat the measurement with nominal level signals.

Case 4 – Failing the mask at high frequencies above the clock recovery bandwidth

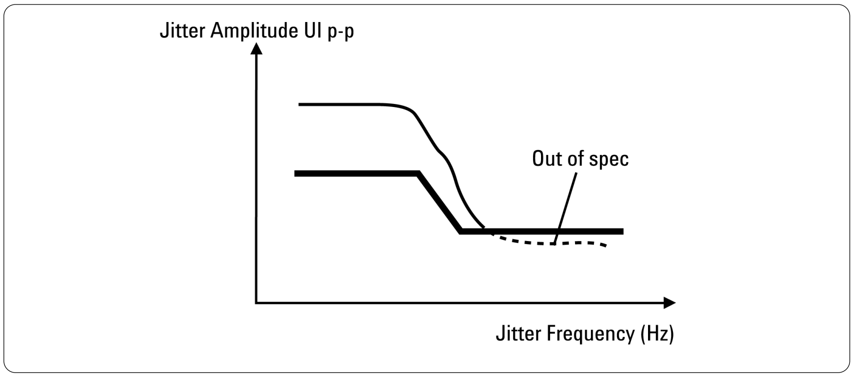


Figure 4.9 Fails mask at upper frequencies

In this case (Figure 4.9), the most likely cause is limited eye-width alignment or noise.

Case 5 - Failing to start measurement or taking excessive time

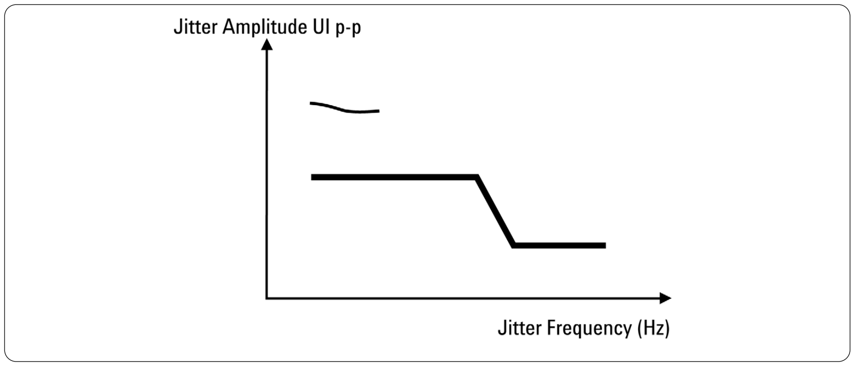


Figure 4.10 Fails to start measurement run or takes excessive time

Check that there are no errors being counted before starting the measurement and applying jitter. Note, in the case of the "Power Penalty" technique, there may be a very low background error rate but this should be significantly lower than the threshold BER being used.

Where measurement time is excessive, check the BER threshold selection, and ensure that gate time and settling time and error count are kept to a realistic minimum. Also check that the maximum available payload bulk size/concatenation is used where possible (see the earlier section on measurement time considerations).

Case 6 – Failing the mask at low jitter frequencies

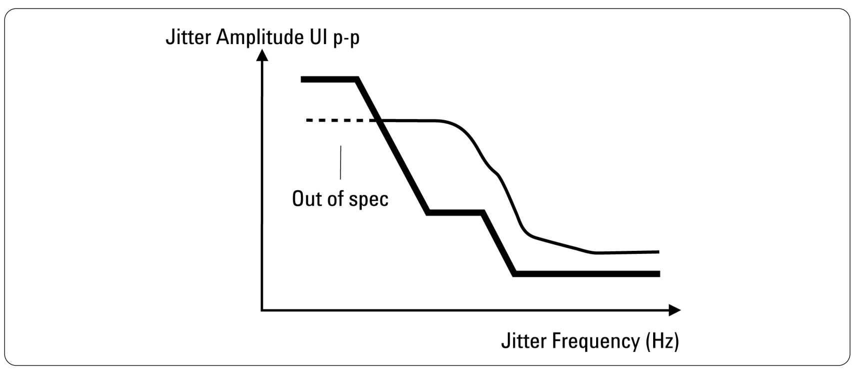


Figure 4.11 Fails to meet a buffer store limit

Although the DUT meets the demanding higher frequency portion of the mask, it appears to have limited buffer store and fails to meet the minimum size required for the low-frequency plateau (Figure 4.11). However, check the correct mask/regulation applicable to the DUT is being used.

Conclusions

Jitter tolerance is one of the most important characteristics for the clock recovery and input circuitry of network equipment. To test the full jitter specification defined by the jitter mask, requires a large number of measurements so automation and the selection of the correct test conditions are essential to minimize the test time. A full plot of the jitter tolerance can reveal features of the design and clock recovery bandwidth, which can be used diagnostically, and for evaluating design margins.

5. Jitter Transfer Measurement

Abstract: This paper explains the need for jitter transfer measurements and the standard ways of making measurements and ensuring high accuracy. The difference between high-Q and low-Q clock recovery is explained, and the paper concludes with a review of unexpected measurement results with advice on how to interpret the jitter transfer plots.

Introduction

Jitter transfer measurements are employed as part of network element/component design, in manufacturing test and perhaps in procurement. They may be used in installation, commissioning or maintenance of transmission systems, but this is less likely as transfer characteristics and performance are largely controlled by the element/component design factors, and also because these tests are done intrusively.

This article provides information on some of the practical application aspects of jitter transfer measurements and test requirements. Included are examples of results, what to expect, how to interpret the measurement information and how to avoid common problems.

Network element bandwidth and jitter transfer requirements

Jitter transfer measurement is required to confirm there is no amplification of jitter by Network Elements (NEs) in the transmission system. This is particularly important in line systems, where regenerators are cascaded, and any systematic jitter gain could accumulate and cause transmission errors. Line systems have specific bandwidth limits relating to the clock/data recovery, which help to control the build-up of jitter. There is typically a trade-off between jitter tolerance at the input of the network equipment and jitter transfer to the re-clocked output.

Systems may specify narrow or wide recovery bandwidth and this affects the jitter transfer and tolerance characteristics. For example, at OC-48 there are Type A and Type B line systems, where A has low-Q (wider bandwidth) and B has high-Q (narrower bandwidth).

Line System Bandwidth Terminology and Characteristics				
Type A	Low-Q	Wide bandwidth	Typically employ Surface Acoustic Wave (SAW) or dielectric resonator filters, and sometimes configurable Phase-Locked Loops (PLL).	Higher jitter tolerance with wider -band jitter transfer and more high-frequency line jitter.
Type B	High-Q	Narrower bandwidth	Typically employ PLLs.	Reduced jitter tolerance to high-frequency jitter, with narrower jitter transfer.

The bandwidth limiting of the recovery circuits attenuates higher jitter frequencies giving a characteristic roll-off to the jitter transfer response. Type A line-system regenerators are generally characterized by greater jitter tolerance to higher jitter frequencies, and by more high frequency jitter accumulation due to their relatively broad jitter transfer characteristics. Conversely, Type B line-system regenerators are typically characterized by less tolerance to high frequency jitter due to the high-Q factors[1] associated with Phase-Locked Loops (PLLs), usually employed in Type B regenerators, and by less high frequency jitter accumulation due to their narrow phase transfer characteristics.

In higher bit-rate line systems, like OC-192/STM-64, the regenerators may use a hybrid arrangement of the Type A/Type B characteristics by employing a highly tolerant timing recovery circuit at the input, followed by a buffer and PLL at the output. This combination results in OC-192/STM-64 network-interface jitter, and jitter tolerance specifications, that are comparable to Type A, and a jitter transfer specification that is comparable to Type B. The purpose of this is to optimize the OC-192/STM-64 line system robustness by providing greater margin between input tolerance and output transfer.

Jitter transfer tests are mainly performed on the line regenerators, and give a classic clock-recovery bandwidth transfer response. More complex transmission sections/elements like Add Drop Multiplexers (ADMs), de-jitterizers, transponders etc., will include very narrow PLLs or re-timed buffer stores. When tested for jitter transfer, these may not exhibit a typical recovery response and some examples of these are shown later in this paper.

[1] Q-factor of a bandpass filter is the ratio of the center frequency to the –3 dB bandwidth ($f_c/BW_{-3\,dB}$). High-Q implies a relatively narrow filter.

In general, jitter transfer measurement is only required where the device output is loop-timed from the input. Devices which are re-timed from reference clocks usually do not require jitter transfer measurement, as the external synchronization requirements are normally stringent enough to ensure transfer compliance. However, this depends on the synchronization referencing method and implementation.

The measurement process

As jitter transfer is a relative gain/loss measurement, it is performed by comparing the absolute level of jitter being transmitted into the Device Under Test (DUT) with an absolute measure of the jitter output from the DUT. A gain/loss measurement could be performed using a broadband measurement technique. However in practice, to achievethe necessary accuracy a very narrow-band tracking filter method is employed.

The standard jitter tolerance mask is usually used to set the input jitter level during the jitter transfer test. Sinusoidal jitter, at a number of spot frequency and amplitude points in the mask, is applied sequentially during the measurement process. The test set receiver tracks the applied jitter frequency and makes a narrow-band measurement of the level to plot the gain/loss at each point. To ensure optimum accuracy, the transmitter/receiver of the test set can be normalized by performing a loop-back calibration run, prior to making DUT measurements.

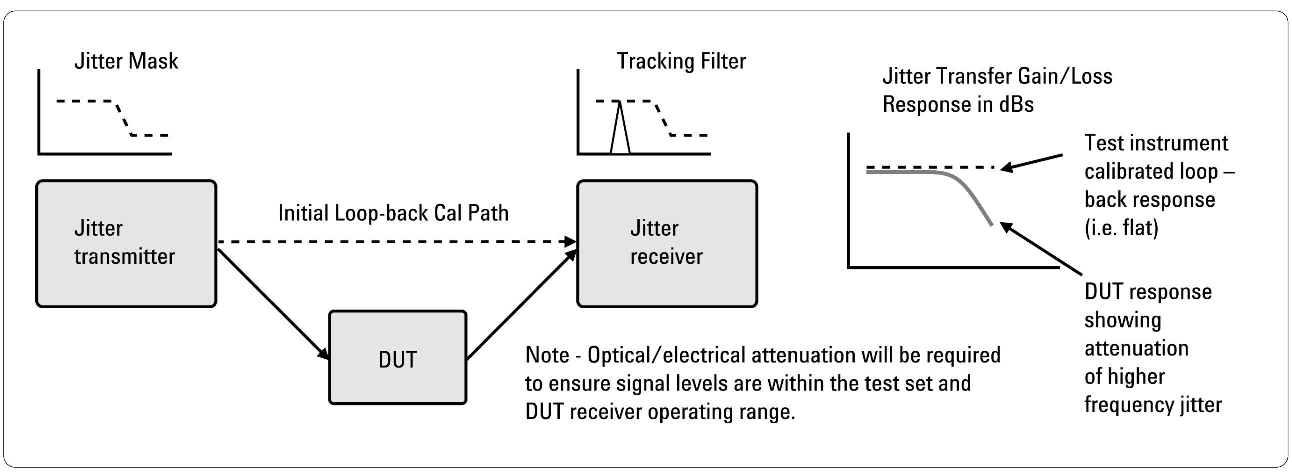


Figure 5.1 Simplified measurement process

The calibration sequence also takes the temperature environment into account to achieve the best possible accuracy. Measurement resolution and repeatability can be improved by extending the *gate time* (time spent in measurement at each point). *Settling time* (device settlement time before measurement at each point), can also be altered. However, there is a trade-off between extended settling/gate times versus environment/temperature changes, and of course the consequential increase in overall measurement time. In practice, a default of 5 seconds settling/gate time at each point is used, with a recommended increase to 10 seconds settling and 20 seconds gate time if required to optimize measurement accuracy versus time.

When testing transfer at very low jitter frequencies, below 15 Hz with low level jitter input masks (1.5 UI p-p), some improvement in accuracy may be achieved by increasing the mask level for low frequency points. User-defined jitter masks can be created/edited and allow the flexibility to perform transfer over specific range of jitter frequencies and also with increased jitter amplitude at low frequencies (within maximum tolerable jitter of DUT) to further improve measurement noise floor margin if necessary.

Ideally, before performing jitter transfer measurements, the *jitter generation* and *jitter tolerance* requirements for the DUT should have been checked and met. It is important that the DUT meets the jitter generation (output jitter) requirements, as any excessive spurious phase noise which correlates to a jitter transfer test point frequency could affect the accuracy of the measurement.

Where tests are being performed at DUT binary interfaces, great care must be taken to ensure good impedance matching and clock/data phasing of the test set-up to avoid reflections and crosstalk that could impair measurement accuracy.

Input jitter mask for jitter transfer testing

A typical input masks used to provide the jitter amplitude and frequency points for sinusoidal jitter generation used during jitter transfer measurements is shown in Figure 5.2. These are usually the standard tolerance masks for the DUT as this ensures the jitter input applied is within the input jitter tolerance. Note the different high frequency jitter tolerance masks depending on DUT type. SONET masks may use additional amplitude at lower frequencies (solid grey line in Figure 5.2).

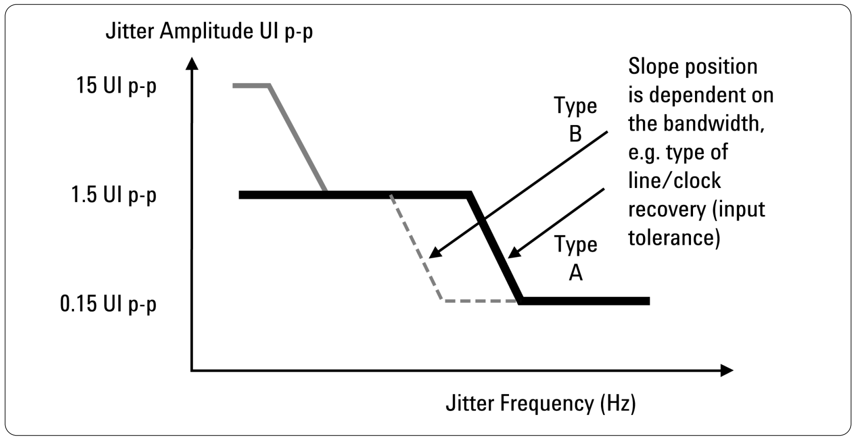


Figure 5.2 Typical jitter input masks for jitter transfer

Normal/expected jitter transfer plot of a DUT clock recovery

A typical jitter transfer plot for a DUT with a wider, Type A, characteristic is shown in Figure 5.3. Typically, there will be a flat region where the jitter frequency is within the bandwidth of the DUT clock recovery with approximately 20 dB/decade roll-off at higher frequencies. The gain must not exceed the +0.1 dB limit in the flat region and should be below the 20 dB/decade jitter attenuation slope. The recovery circuit –3 dB bandwidth is designed to be less than the transfer-pass mask corner frequency. However, it must still be high enough to meet the jitter input tolerance requirements. Therefore, for simple regenerators, this is usually a compromise between meeting the tolerance and transfer requirements

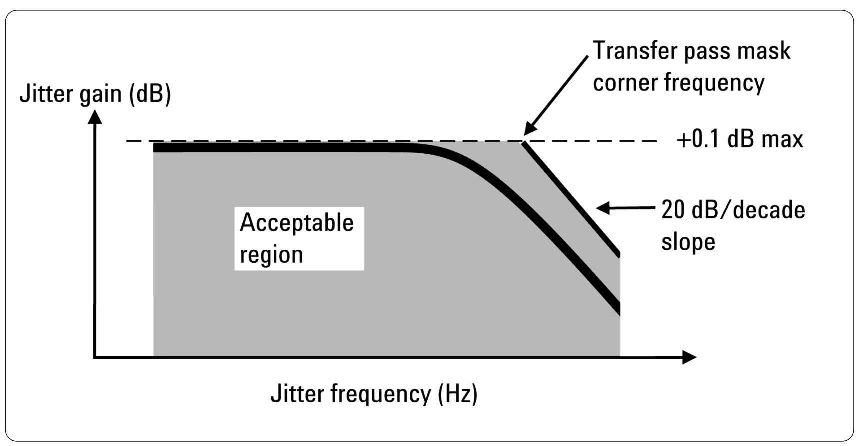


Figure 5.3 Typical jitter transfer plot

Results requiring explanation/interpretation

This section gives some examples of results that could occur depending on the type of DUT, the bandwidth and measurement set-up. These case studies include an explanation of the plot characteristics, and interpretation of the likely situation with advice on the cause.

Case 1 - Using the wrong pass mask intended for narrower bandwidth (high-Q) Type B clock recovery.

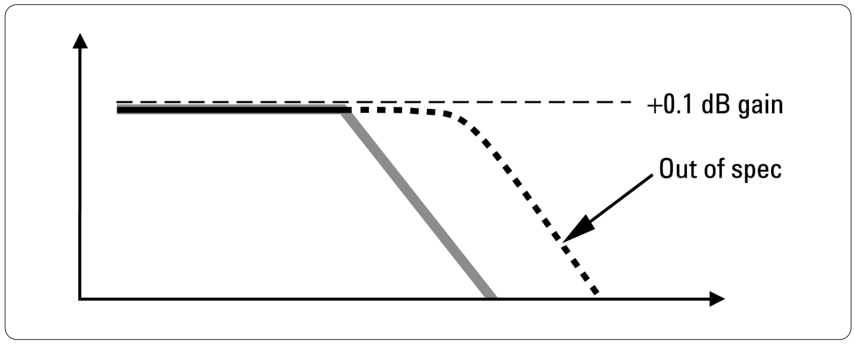


Figure 5.4 Using the wrong pass mask

Because of the different requirements in Type A and Type B clock recovery, a failure might be due to selecting the wrong mask. Check if the DUT is a Type A (low-Q) or Type B (high-Q) device and then check the correct transfer pass mask is selected. Check the DUT design bandwidth.

Case 2 - Testing beyond the test set-up noise floor into an undefined region

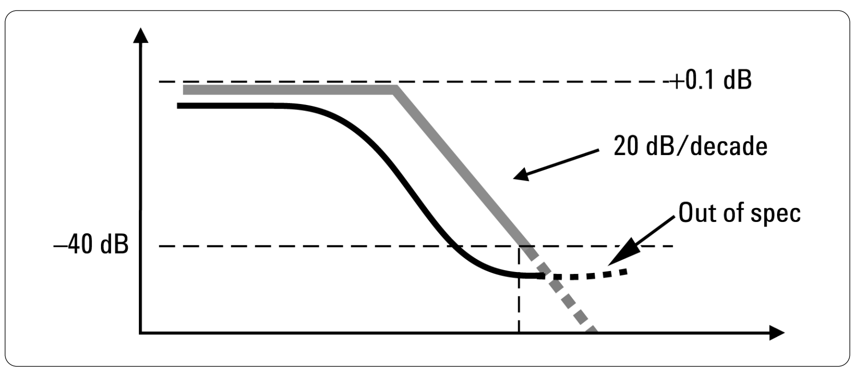


Figure 5.5 Testing below the noise floor

This problem can occur with the narrower-band Type B characteristic DUT if there is a conflict between the pass mask attenuation and measurement set-up noise floor at higher frequencies. Usually it is sufficient to check to two decades above pass mask corner (e.g. –40 dB) if the network/standard requirement does not quote an upper frequency limit. Ensure the test upper frequency is limited so the combination of pass mask and attenuation does not exceed the two decades or –40 dB level.

Case 3 - DUT showing gain at the low-frequency end of clock recovery response

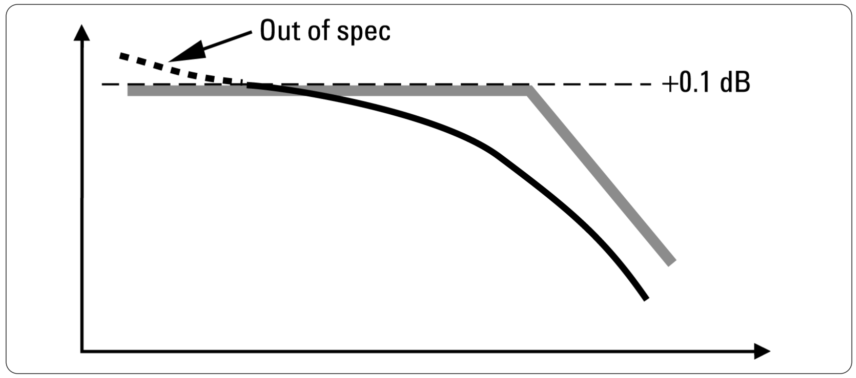


Figure 5.6 Peaking at low frequencies

In this case the most likely cause is additive low frequency noise affecting the measurement. Check for excessive low-frequency phase noise/power-line crosstalk in the DUT or measurement setup.

Increase measurement time and then re-run the calibration and measurement.

Improve measurement noise margin at low frequencies by using modified jitter input mask with additional low-frequency amplitude. Note most devices will accept more than the standard 1.5 UI p-p at low jitter frequencies.

Case 4 - DUT shows peaked response and gain >0.1 dB close to roll-off

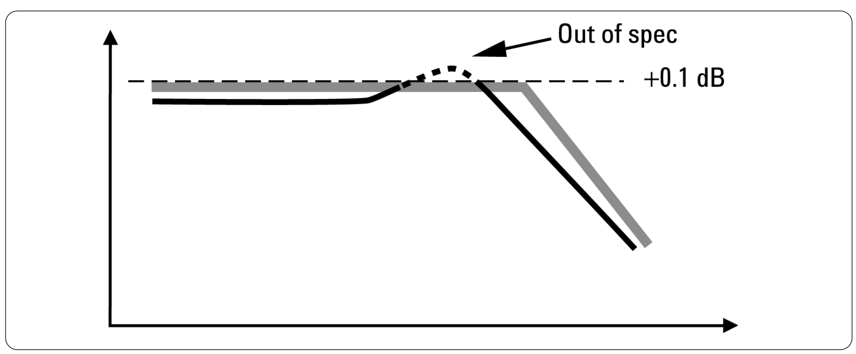


Figure 5.7 Peaking near clock-recovery bandwidth limit

Most likely caused by incorrect DUT recovery filter gain characteristics. There may be too much peaking or the roll-off bandwidth too high.

This result can also be due to DUT or set-up crosstalk or spurious signals affecting measurement.

Case 5 - DUT response is flat beyond expected roll-off

The most likely cause is that the DUT is jitter transparent or does not have a band-limiting clock recovery circuitry.

Note: A measurement run of the test instrument (without the DUT in path) should give a response similar to this confirming the flat characteristic of the test instrument.

Case 6 - DUT response shows attenuated response across all frequencies

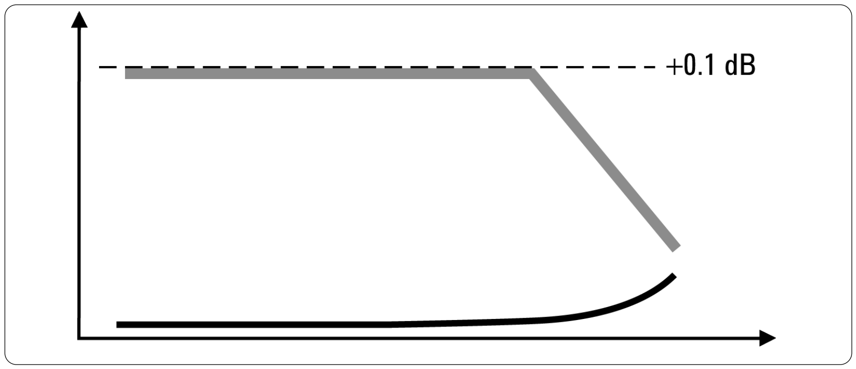


Figure 5.8 Jitter attenuation at all frequencies

This device is being re-timed, removing the jitter applied by test set, and therefore the plot shows a set-up noise-floor. Jitter transfer testing may not be required on such DUTs, as any jitter is adequately removed by re-timing.

Case 7 – DUT has very low-frequency clock recovery bandwidth rejecting all but the very lowest jitter frequencies

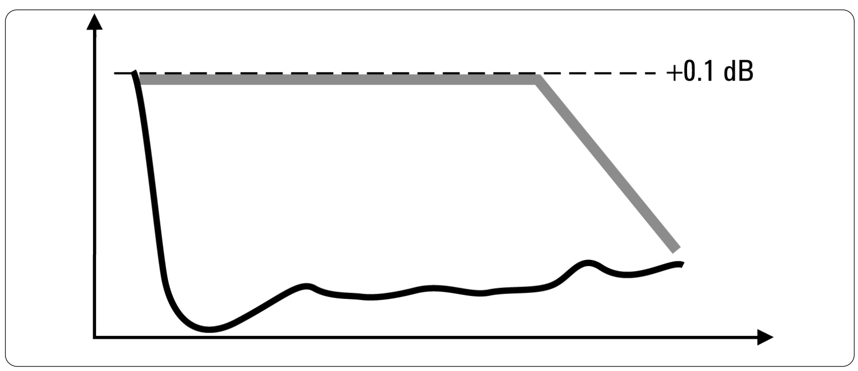


Figure 5.9 Jitter Attenuation, except very low frequency

De-jitterizers use buffer stores and narrow-band loop-timing to reduce jitter and may exhibit this type of transfer response.

Note: Check if a device is loop-timed, or being re-timed from another source, by applying a small frequency offset and checking if this passes through DUT.

Conclusions

Jitter transfer measurements are important for cascaded clock recovery circuits as might be found in long-distance transmission systems with regenerators and line terminals. If jitter transfer gain is not controlled, excessive jitter may accumulate and cause errors in subsequent equipment. Jitter transfer masks are defined by bit-rate and whether high-Q or low-Q clock recovery is used. As with other jitter measurements, there is a variety of unexpected and anomalous test results which may arise depending on the test conditions.

6. Accurate Measurement of Jitter Generation

Abstract: This paper discusses techniques and conditions that are important for accurate measurement of output jitter or jitter generation. The effects of test set intrinsics, test set filtering, pattern dependency and the choice of test pattern are discussed. Consideration of these points will lead to a better understanding of this key measurement, and help reduce measurement uncertainty.

Introduction

A certain amount of jitter will appear at the OC-n/STM-n output port of any Network Element (NE), even if an entirely jitter-free digital signal or clock is applied to its input. This is known as *jitter generation*. The NE itself produces this “intrinsic” jitter, due to:

- Thermal noise and drift in clock oscillators
- Spurious emissions by crystals in clock oscillators
- Influences from other system modules on the clock supply (cross-talk)
- Pattern-dependent delay in scramblers and encoders
- Clock data recovery circuits

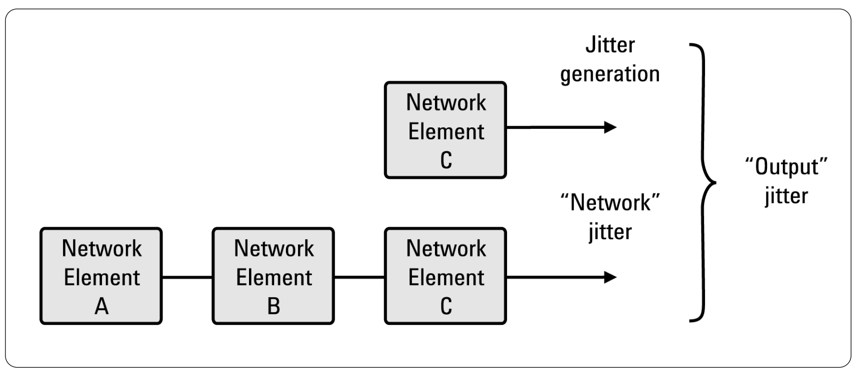


Figure 6.1 Output jitter, jitter generation and network jitter

Closely related is *network jitter*, which can be measured on any interface in the network. Network jitter accounts for the accumulation of jitter across multiple network elements. For completeness, both *jitter generation* and *network jitter* are also referred to as *output jitter*. This paper will refer to *jitter generation*, but it could equally apply to network jitter measurement also.

Measuring jitter generation

Jitter generation measurements use a peak detector to record the maximum amplitude of jitter that occurs during a specified period. As jitter is statistical in nature and can exhibit bursts of high amplitude (jitter transients), 'peak-to-peak' measurements will give a good indication of performance, as it is the extremes that often cause errors. The results should be cumulative and retrieved in real time so that *all* jitter events are caught during the observation interval. Root Mean Square (RMS) measurements can also be made to provide a statistical indication of the average jitter noise power, however peak values may be missed.

Standards specify the maximum amounts of jitter generation (within a defined bandwidth) permitted in the telecommunications network. This is to ensure that the amount of jitter never exceeds the specified lower limit of Maximum Tolerable Input Jitter (MTIJ) specified for the NE's input ports. In other words, if the jitter level is excessive, the NE's input circuitry (clock recovery circuits etc.) may not have been designed or qualified to work error-free under such conditions.

A demanding maximum limit of 100 mUIp-p is defined by both Telcordia[1] and ITU-T standards. To measure jitter accurately at such low levels requires careful consideration of the measurement techniques and conditions.

Alternative measurement techniques

Phase noise technique

Phase noise measurements analyze the spectrum of a clock signal. The spectrum can be filtered using software, to calculate the RMS and estimate the peak-to-peak jitter. While the technique has limitations, it is useful in measuring quasi-static intrinsic jitter on clock sources and reference oscillators, which may not run at standard telecom rates.

The limitations of this technique are:

- Not real time, so the results cannot be correlated with other "events".
- Peak-to-peak 'predictions' are not reliable, as real-time data and jitter Probability Density Function (PDF) are unpredictable.
- Software results are post processed so jitter transient peaks may be missed.
- Requires an external clock recovery circuit (frequency specific) to measure on data interfaces.

[1] Telcordia GR-253 Category 2 interfaces for OC-n, STS-n electrical and synchronous DS1 interfaces to SONET NEs. See the paper on jitter standards in this booklet.

Eye diagram and histogram technique

The quality of a digital signal can be evaluated using an oscilloscope, as shown in Figure 6.2.

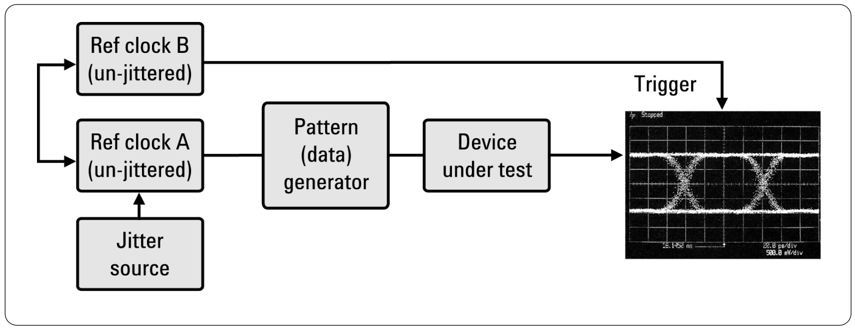


Figure 65.2 The eye-diagram technique, using an oscilloscope

With a suitable jitter free reference clock, data symbols are superimposed on each other. The resultant eye diagram can be inspected to deduce a peak-to-peak (p-p) or RMS result. Any jitter on the trigger signal, or in the oscilloscope trigger circuit, will affect the validity of the results. An eye diagram gives a qualitative indication of the signal only.

The limitations of this technique are:

- Not real time.
- No frequency-band information – only the sum of all spectral components is measured, so the result cannot be compared with ITU-T and Telcordia standards.
- Limited to measuring 1 UIp-p. Above this level, the eye diagram is totally closed.
- Transient peaks may be missed (sampling oscilloscope).

Use a dedicated output jitter amplitude tester

To obtain quantitative measurements that are traceable to international standards, a dedicated jitter test is required.

The advantages of this are:

- Telecom data rates with clock recovery and built-in measurement filters.
- Specifically designed peak-to-peak and RMS detectors.
- Accumulates and displays output jitter amplitude statistics in real time.
- Measures data – specifically designed clock-recovery circuits.
- Compliant with O.172 (the ITU-T recommendation for jitter and wander test equipment applied to SDH networks).
- Demodulated jitter output is made available for further analysis, for example using a spectrum analyzer.

Measurement block diagram for jitter and wander test set

The block diagram of a typical jitter and wander measurement test set, taken from ITU-T O.172, is shown in Figure 6.3.

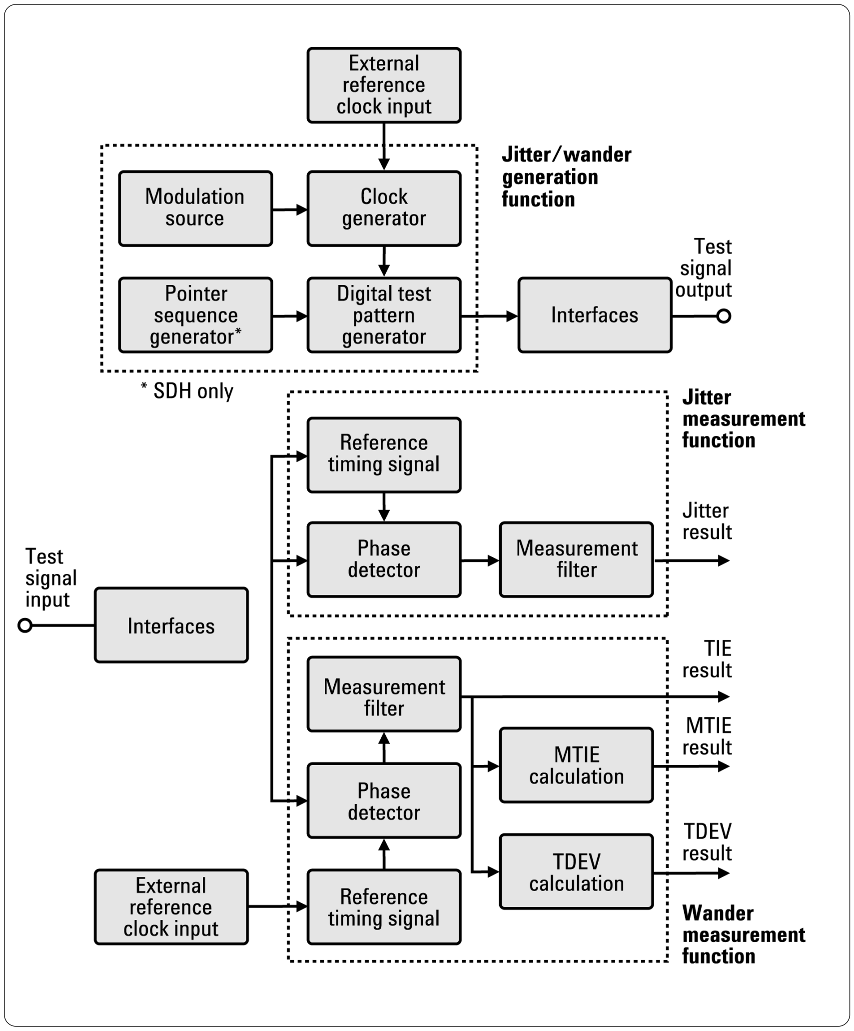


Figure 6.3 Block diagram of a jitter and wander test set

The test signal input is typically an optical data signal. A jittered clock signal applied to the jitter measurement function must first be derived from the SONET/SDH interface (jitter measurements are always made on a recovered clock signal). This requires an optical/electrical (O/E) conversion, and wide-band clock data recovery, specifically designed to ensure the jitter content of the optical signal is completely preserved in the recovered clock signal. The clock recovery bandwidth is dependent on both jitter frequency and jitter amplitude.

The phase detector produces a value that is proportional to the phase difference between the jittered clock signal and the un-jittered internal reference timing signal. The output voltage (after a low-pass filter) can be made available as the demodulated jitter waveform (0.172 specifies this as optional). This demodulated waveform can act as a useful diagnostic tool – for example, where a DUT fails to meet specification. It allows further analysis in the time or frequency domain using external equipment to help identify the most significant jitter components contributing to the jitter generation. The rest of the demodulated jitter waveform is selected by a high-pass filter and made available for data processing to produce the normal measurement results (peak-to-peak, RMS jitter, etc.).

Measurement considerations

Calibration

The jitter measurement method described above, requires the clock to be recovered transparently from the SONET/SDH signal to include all the jitter, followed by removal of jitter from the recovered clock to create the reference for jitter measurement in the phase detector. A common source of intrinsic jitter is in the clock recovery process itself! To ensure optimum performance and measurement accuracy, a test set must be designed and calibrated to minimize intrinsic noise in the measurement circuitry.

The calibration applied to the test set's receiver intrinsic jitter should be minimal, and compensates mostly for wide-band noise components. Correction for a more complex signal (for example a SONET/SDH-framed Pseudo Random Binary Sequence (PRBS) $2^{23} - 1$ pattern) leads to a much greater potential for error, due to greater correction factors. The problem with this approach is readily apparent if you were then to measure a simpler signal (such as an "all-zeros" payload) - the greater correction factor would cause the measurement results to be over-compensated and excessively low. The danger is that an apparently more favorable measured result would appear, and the output jitter amplitude measurement *under reports* the actual value of output jitter from the DUT.

The fundamental point is that to optimize the accuracy of any overall measured jitter result, the test set itself has to have absolute minimum noise levels before correction.

Finally, it is worth noting that simply connecting the test set's transmitter and receiver back-to-back and measuring the noise floor – a common way to "measure the test set's intrinsic jitter" – can be misleading. It will usually yield a result that is a combination of the transmitter and receiver jitter, far in excess of the true intrinsic jitter of the receiver. To then subtract that value from the overall measured jitter result will consistently under-estimate the true levels of output jitter amplitude of the DUT.

Measurement range

The jitter receiver accuracy is specified over the measurement bandwidth, which is further defined by the addition of filters. The test set's measurement bandwidth is dependent on both the jitter amplitude and the jitter frequency. For intrinsic jitter measurements the jitter range of the peak detector should be at its minimum so that the resolution/accuracy is maximized (the smallest range has the highest resolution and therefore greatest accuracy).

Measurement time

The output jitter measurement aims to record the peak-to-peak (p-p) value in the selected measurement period, therefore any transient event should be captured and recorded accurately as intrinsic noise. The measurement period should be cumulative and be set to some realistic gating period, with the latest standards recommending a 60-second observation interval.

Optical power

For optimum output jitter measurement, always ensure the correct level is being applied to the receiver by making a power measurement at the fiber end which is to be connected to the "optical in" port. If the power level is incorrect (too high power levels may damage optics!), perhaps due to an unclean receiver connector/port or dirty cables, then intrinsic jitter results may well be a lot higher than expected. An attenuator may be required to achieve an input power level in the operating range.

Frequency offset

Digital input ports of SONET/SDH equipment are specified to tolerate a line frequency offset within the range ±20 ppm from the nominal bit-rate.

Clock data recovery and the accuracy of the output jitter measurement should NOT be affected by the presence of such a frequency offset. Should the test set lose lock, it is recommended that a "frequency offset" measurement be made to check for excessive offset.

Measurement bandwidth

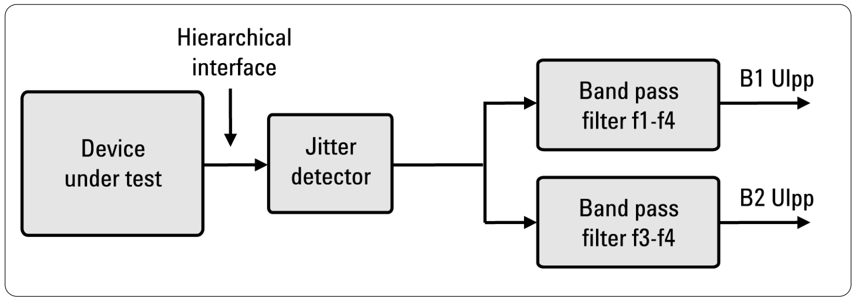


Fig 6.4 Jitter test set filtering

Jitter has noise-like characteristics (particularly in the presence of real-life traffic or a PRBS test signal) and exists across the full band of frequencies. Not all jitter frequencies will affect the digital signal. For example, consider the Phase-Locked Loop (PLL) of a DUT, which provides the timing for the output signal. Very-high jitter frequencies (higher than the bandwidth of the PLL) could be absorbed by the PLL's buffer. Jitter at frequencies lower than this bandwidth can pass through without affecting transmission performance.

Consequently, the 'measurement' bandwidth should be limited, depending on the requirements of the interface being tested, in order to examine only the jitter frequencies of interest. This eliminates non-problematic jitter components, which would impair the actual measurement.

For SONET/SDH rates, different measurement bandwidths are specified in recommendation ITU-T O.172, which assist in partitioning the main jitter sources. These are shown in Figure 6.5.

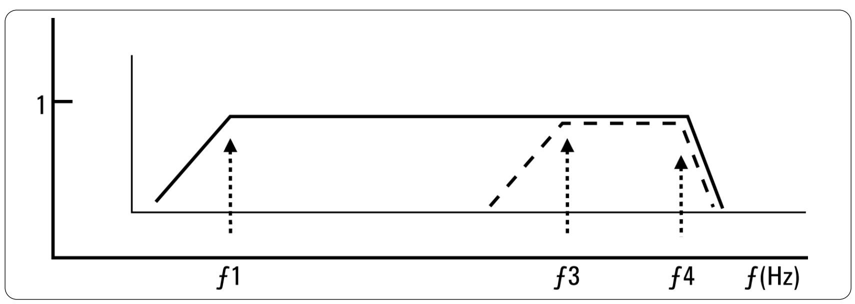


Fig 6.5 Jitter measurement bandwidth cut-off frequencies (O.172)

Signal	Jitter measurement bandwidth (–3 dB cut-off frequencies)		
	f_1 (Hz) High Pass 1	f_3 (Hz) High Pass 2	f_4 (Hz) Low Pass
STM-0	100	20k	400k
STM-1	500	65k	1.3M
STM-4	1k	250k	5M
STM-16	5k	1M	20M
STM-64	20k	4M	80M
The value of f_1 reflects the narrowest timing circuit cutoff frequency expected in a line system. The value of f_3 is related to the bandwidth of input timing acquisition circuitry. The value of f_4 establishes minimum measurement bandwidth requirements and is chosen to include all expected, significant alignment jitter.			

Note that a 12 kHz high pass filter is also specified up to OC-48 rates by GR-253. Similarly, for OC-192, a 50 kHz filter is added.

Test signal considerations

Jitter measurements are strongly influenced by the data being carried. Test results can vary widely between those for a simple repetitive pattern such as a fixed word (no pattern dependency, plenty of transitions), e.g. 1010 or 11001100, and those for a complex PRBS or for a SDH/SONET system (with its 8 kHz frame structure and block of framing bytes). It is crucial when measuring and specifying jitter generation to define the data being transmitted, especially when comparing specifications and results.

Pattern dependency

Pattern-dependent jitter (systematic jitter) results from a distorted digital signal. Basically, the timing of the signal edges (rising/falling) and peak amplitude are affected by the pattern content.

This results in *Inter-Symbol Interference (ISI)*, where preceding data affects the timing edges of the data that follows. Usually this is due to bandwidth limitations and AC coupling and it can happen in the generation or receiving sections. Figure 6.6 illustrates a simplified example.

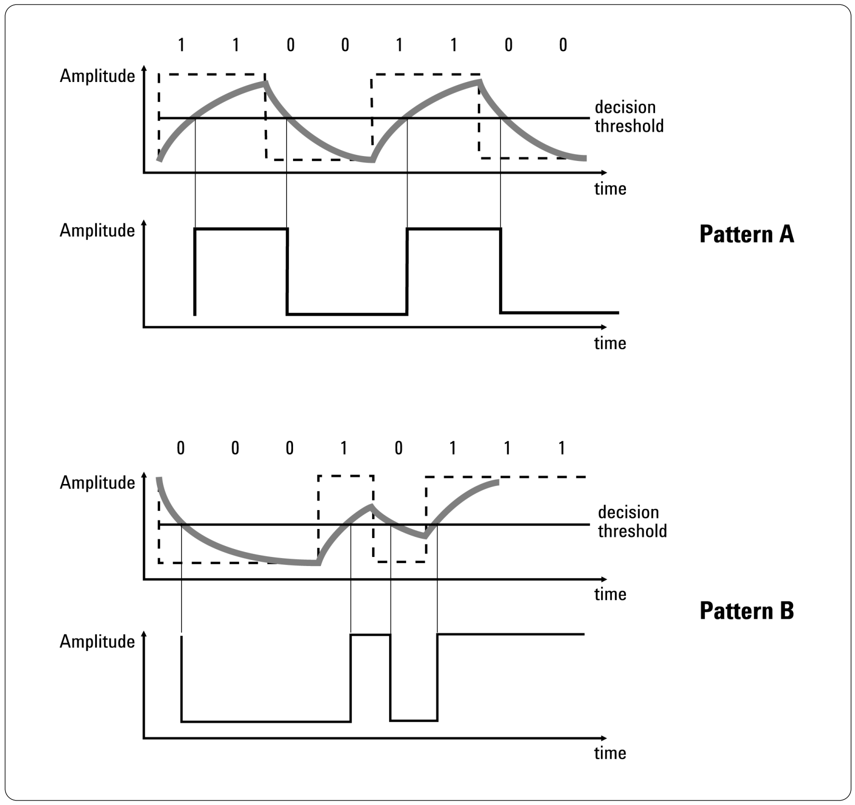


Fig 6.6 DC pedestal affects the timing of the recovered clock

With Pattern A (11001100) the bandwidth limitations slow down the timing edges uniformly. The resulting output signal is phase shifted from the original, the relationship between edges is maintained and there is no jitter.

With a different Pattern B (00010111), the DC pedestal effects of previous data combine with the bandwidth limitations to slow down the timing edges irregularly. In this case the resulting output signal is not only phase shifted, but the relationship between edges has changed and there *is* jitter.

Long runs of zeros (where there are no timing edges to detect) will affect a data recovery circuit's ability to maintain timing in the network equipment. It cannot cope with the excessive amount of jitter induced by the *pattern dependency*, and ultimately ISI becomes the limiting factor for signaling rate and transmission distance.

This can be equally true of a test set's ability to measure jitter, depending on the method used and quality of the clock recovery or sampling used. With Agilent test equipment, the jitter measurement clock recovery is specially designed to meet two apparently conflicting requirements that traditional recovery circuits cannot:

- *Wide bandwidth* to faithfully recover the jitter modulation up to the highest frequency without attenuation – (Low-Q clock recovery).
- *Narrow Bandwidth* to maintain timing without drift during long zero/one runs – (High-Q clock recovery).

Normally both cannot be met and there is a compromise, but the recovery pre-processor for Agilent test equipment is a design specific to the task of recovering the jitter with minimal effects of pattern dependency.

ISI due to bandwidth limitations can be minimized by careful design, especially at higher data rates.

Scrambling and SONET/SDH overhead byte conditions

The SONET/SDH signal (with binary line coding) must have sufficient bit timing content at the network interface. The requirement of preventing a long sequence of "ones" or "zeros" is met by using a scrambler. The SONET/SDH signal is scrambled with a frame synchronous scrambler of sequence length 127 (PRBS $2^7 – 1$), operating at the line rate. The scrambling system does *not guarantee* that a long run of "ones" or "zeros" would not occur. There are two main reasons for this:

Overhead bytes

The standards (ITU-T G.707 and Telcordia GR-253) state that the first row of the SONET/SDH frame shall *not be scrambled* (this maintains integrity of the A1 and A2 framing bytes).

Care should be taken in selecting the binary content of the remaining J0 and Z0 bytes and of the bytes reserved for national use (which are excluded from the scrambling process of the SONET/SDH signal), to ensure that long sequences of "ones" and "zeros" do not occur. For SDH, the recommended content of these bytes is AAhex (10101010), to minimize the pattern dependent jitter contribution. In the SONET world, the value CChex (11001100) is used. The content of these bytes can expose pattern dependent jitter problems in network equipment.

Data content

There is a remote possibility that the actual data content, even after scrambling, could result in a lengthy sequence with no transitions.

For example, if the traffic emulates the scrambling pattern, then many bytes of all “ones” or “zeros” will appear in the coded line signal. As the extreme cases of this will be a rare occurrence, and as it would be difficult for a jitter test set to continue to perform accurate measurements under these conditions, it is important that a representative worst-case signal is defined for the purposes of test set specification.

Recommended test signal

As discussed earlier, the spectral content of the jitter can change with pattern, mapping, scrambling and framing. This is why a *bulk payload with the longest PRBS* is usually considered more stressful and usually generates the largest peak-to-peak jitter. *Concatenated* payloads provide the worst-case scenario for OC-n/STM-n signals. For bulk-filled concatenated signals with a $2^{23} - 1$ PRBS filling the container, the result of scrambling this data is a worst case run of 30 consecutive “ones”/”zeros” (i.e. there will be a maximum of 30 clock periods with no transitions on the line signal).

For non-concatenated payloads generated by the SONET/SDH test set, the byte interleaving of the STS/VC containers reduces the maximum length of runs produced, consequently reducing the pattern dependent jitter. In this case, payload containers, which do not contain the test signal, may be unequipped or should contain an “all-ones” or “all-zeros” fixed byte pattern. All jitter accuracy requirements should be met with the signals described. If any other structured signals, pseudo-random or random signals are used, larger measurement errors could occur. For example, signals with longer runs of “ones” or “zeros” (say 50 or 60 rather than 30) could even infringe the sampling theorem, rendering it impossible to make the jitter measurement because the clock recovery circuit cannot sample the data correctly.

Conclusions

Accurate measurement of output jitter or jitter generation requires the use of a high-performance jitter test set with very low receiver intrinsic jitter to resolve the 100 mUIpp specified in the current standards. Use of “calibration” to reverse out intrinsic jitter in the test set, is only reliable if the background residual jitter is already below the 100 mUIp-p level. It is also extremely important to select the right test pattern for reliable and repeatable results, particularly with SONET/SDH framed/scrambled signals. Long PRBS bulk-filled payloads are the most exacting test signals, and it is essential that the test set intrinsic remain well below 100 mUIp-p under these conditions for reliable results on the DUT.

7. How Tester Intrinsics and Transients affect your Jitter Measurement

Abstract: This paper discusses the design of a tester capable of making accurate measurement of jitter generation and jitter transients. It shows that large errors can occur if the performance of an inherently noisy jitter receiver is "improved" using software calibration. Many jitter generation waveforms are noise like and contain isolated transient spikes that create jitter peaks which must be accurately recorded. Jitter is composed of random and deterministic components and both must be accurately measured. The paper concludes with a block diagram description for a jitter test set and a practical method of evaluating the intrinsic jitter performance.

Introduction

A common problem when making jitter measurements is the variability and lack of consistency between different test sets, particularly for low-level jitter generation measurements. This is most likely to be caused by significant performance aspects of the test equipment, not addressed by the current ITU-T standards.

The measurement of jitter generation on high-speed optical signals is arguably the most demanding challenge for jitter test equipment. Currently ITU-T O.172 [Reference 1] sets fixed intrinsic error limits for the jitter receiver, however, no definition is given which validates the various calibration schemes employed by different test set vendors. Without some common understanding or control, there is inevitably inconsistency between test sets, and at worst the measurement may be completely invalid.

This paper will highlight some potential performance issues in the areas of test equipment calibration and peak detection, and importantly will show a conformance test to ensure consistency between test sets in jitter generation measurements.

Calibration or too much correction?

ITU-T Recommendation O.172 has limits to govern the maximum amount of fixed error that should exist in compliant test equipment. Intuitively one might expect that to measure a low-noise signal, it would be necessary to use low-noise measurement equipment. Indeed, the premise of ITU-T O.172 started with the concept of a test equipment residual of *one-third* of the Device Under Test (DUT) performance limit. Subsequently this requirement has been relaxed somewhat as for example in ITU-T G.783[1] where a test equipment requirement of *one-half* the DUT performance limit is realised. Nevertheless, it is important that the measurement equipment's own noise is significantly less than the limit being tested, to guarantee accuracy.

[1] ITU-T G.783 (10/00) "Characteristics of synchronous digital hierarchy (SDH) equipment functional blocks"

The test equipment fixed error term or receiver intrinsic jitter is a limit put on the performance of the test equipment itself. If the receiver noise floor exceeds this amount then the ability to give a truly representative reading is impaired. While it is generally agreed that calibration of a jitter receiver is useful, it should also be recognised that a receiver unable to make inherently low-noise optical data pattern jitter measurements, cannot be made standards-compliant by software correction to give artificially low readings. Because the statistics of jitter being measured are unknown, any wholesale subtraction of calibration terms doesn't give reliable readings. In fact there is a high probability of misreading the DUT jitter.

To illustrate this, a demodulated jitter intrinsic waveform of an SDH signal at 10 Gb/s is shown in Figure 7.1, displaying a random noise-like component and a systematic or transient event. The test equipment has a poor parametric performance and a high fixed-error or intrinsic term so the overall measured value may be too high. If software correction is employed, the test equipment may display compliant values but will measure incorrectly as shown by Figure 7.2.

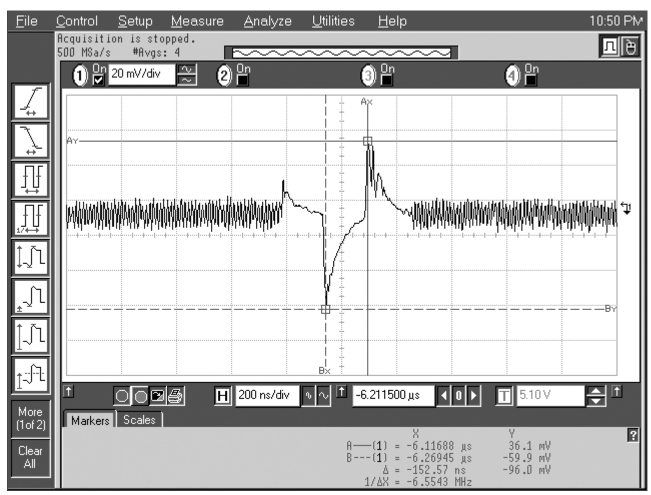


Figure 7.1 Test equipment jitter intrinsic waveform of an SDH signal

A similar demodulated jitter waveform is shown in Figure 7.2, but this time 'real' jitter is present, generated by the DUT. This is, however, at the same level as the instruments own intrinsic or fixed error term and therefore will not be measured accurately since it is indistinguishable from the instrument intrinsics.

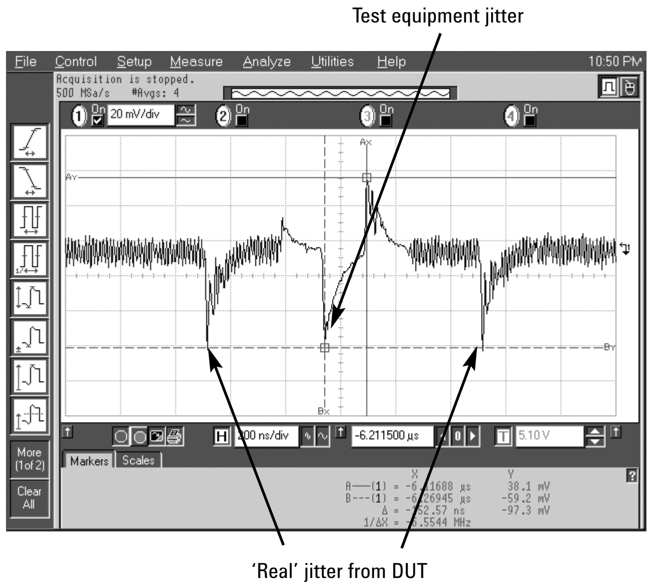


Figure 7.2 Demodulated jitter waveform with DUT components superimposed on instrument intrinsic jitter

To make accurate measurements, the instrument must inherently meet or preferably exceed the ITU-T-O.172 specification. It is clear that if calibration and nulling schemes are applied arbitrarily to test equipment of poor quality, then measurement errors far greater than expected may occur.

The example in Figure 7.3 illustrates a tester which has a fixed error that is well within the limits of ITU-T O.172. In this case both the random jitter and the systematic jitter components in the instrument residual are well below the ITU-T O.172 50 mUI specification, thus allowing the DUT jitter to be clearly discerned and accurately measured.

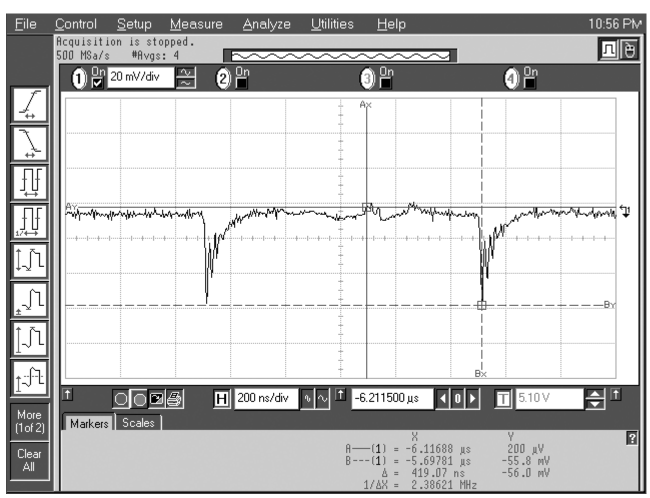


Figure 7.3 DUT jitter accurately measured in the absence of significant instrument intrinsics

Comparing Figures 7.2 and 7.3 shows the difficulties of measuring peak-to-peak jitter if the measurement system has a high noise floor – in some cases the jitter to be measured becomes invisible to the tester.

Where jitter noise is truly random some assumptions can be made about jitter addition (similar to noise power addition). It may be valid to perform subtraction and calibration, but care is needed to understand the potential measurement errors.

Typically, the jitter characteristics and statistics being measured by the tester are unknown. Jitter may be systematic, random or "bursty" in nature, with causes varying from pattern dependence, oscillator noise effects or any interference within the system. It may be possible to use effective calibration schemes within the tester which do not compromise jitter measurement accuracy. However, it must be emphasized that if the test equipment is inherently noisy or has itself significant pattern dependence issues, calibration or correction schemes may result in an erroneous measurement of the DUT jitter.

Incorrect jitter peak detection

The ITU-T definition of *jitter* is "short-term non-cumulative variations of the significant instants of a digital signal from their ideal position in time". This means that jitter is an (unwanted) phase modulation of the digital signal. The frequency of the phase variations is called *jitter frequency*.

The nature of jitter can be described by terms such as systematic, random or bursty. Again the test equipment should be able to detect any and all jitter components accurately within the measurement bandwidth. This may indeed be only one jitter transient type event or hit, but this still accounts for the peak reading during the measurement period. The jitter transient may in some cases be enough to cause a bit error in the DUT, so the jitter tester must be able to report and measure this type of event. Ability to measure only continuous sine-wave modulation accurately is insufficient for true jitter measurements.

The examples below show the performance of two different DUTs.

The high-quality DUT has low measured jitter (Figure 7.4) and a good optical eye (Figure 7.5).

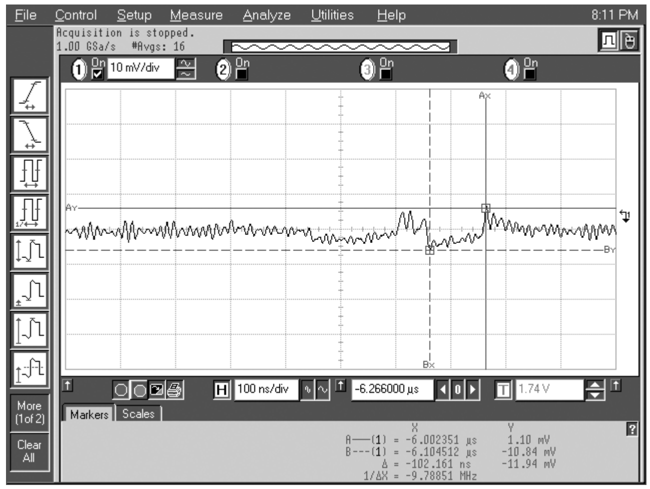


Figure 7.4 Residual jitter waveform from a high-quality DUT

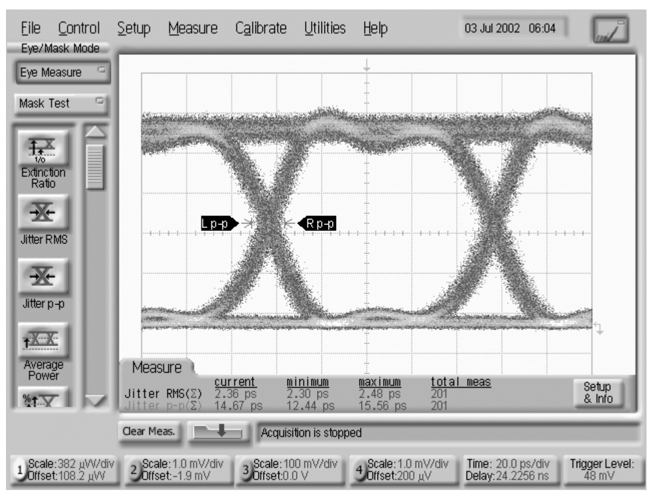


Figure 7.5 Eye-diagram for high-quality DUT

The lower-quality DUT has poorer performance and higher intrinsic jitter. The demodulated jitter (Figure 7.6) indicates a peak jitter event which is transient in nature. This is a severe pattern-dependence caused by the unscrambled bytes of the SDH header, due in this case to poor return loss between the data multiplexer and the Electrical/Optical (E/O) converter. The poorer resultant eye-pattern is shown in Figure 7.7.

In both Figures 7.5 and 7.7, waveforms may pass the optical eye mask and therefore the only way to detect the problem is by correctly measuring the peak jitter.

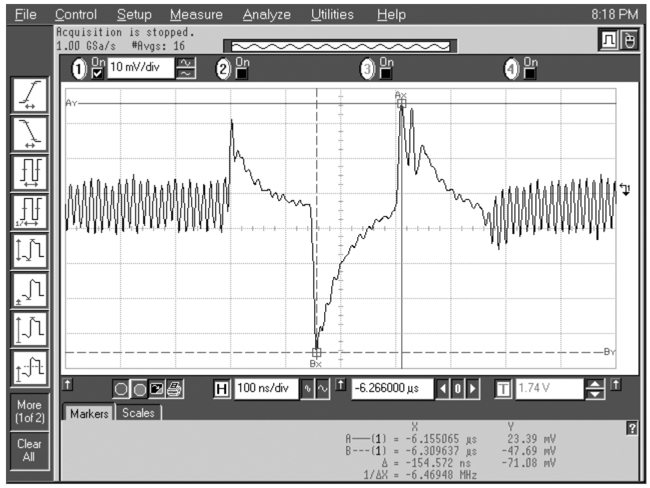


Figure 7.6 Residual jitter waveform from a lower-quality DUT

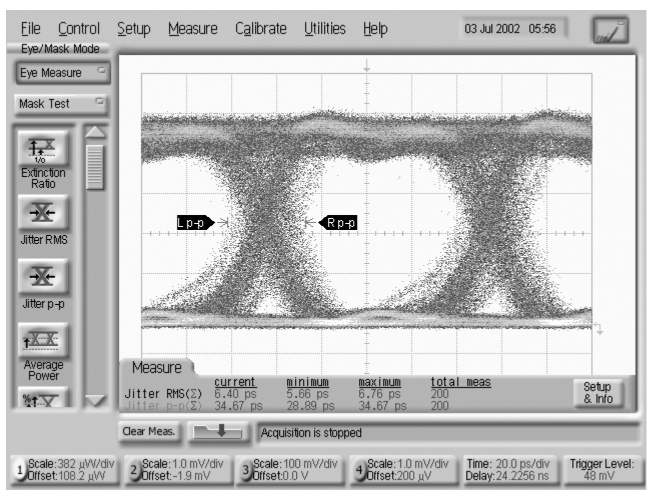


Figure 7.7 Eye-diagram for lower-quality DUT

To accurately assess the jitter performance of these two DUTs, the jitter tester must have low intrinsics and also read the peak values correctly.

In general, jitter transients may arise from a wide range of sources, but all these effects will degrade performance within the network and must be detected and measured accurately, even if they occur only once or very infrequently.

An example of an isolated jitter transient is shown in Figure 7.8. This single jitter hit might be caused by a timing slip in a multiplexer or a possible network synchronizing clock glitch. The jitter tester should detect and record this isolated jitter transient.

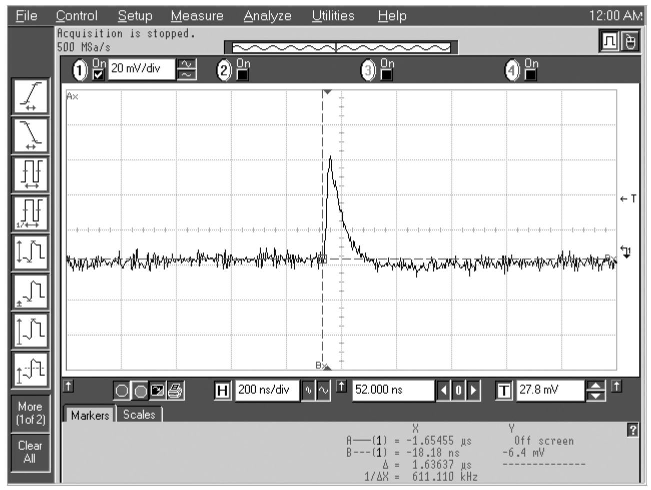


Figure 7.8 Demodulated jitter waveform of an isolated jitter transient

Other transient-type jitter events are shown in Figures 7.9 and 7.10, respectively a burst of jitter possibly due to interference from a digital data bus, and series of repetitive jitter hits caused for example by possible switch-mode power supply breakthrough.

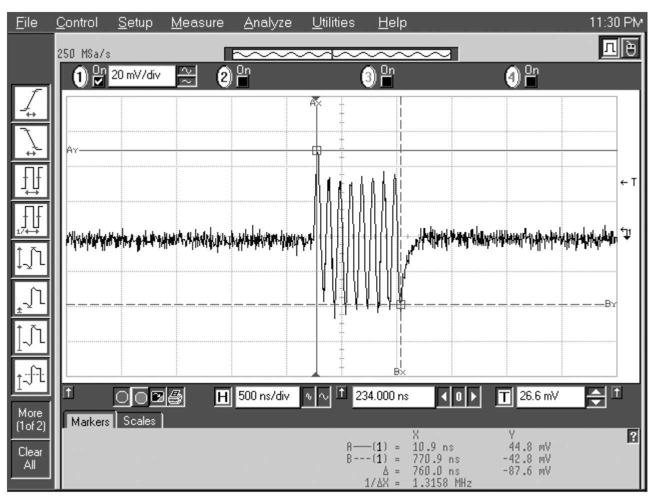


Figure 7.9 Jitter burst due to possible data bus interference

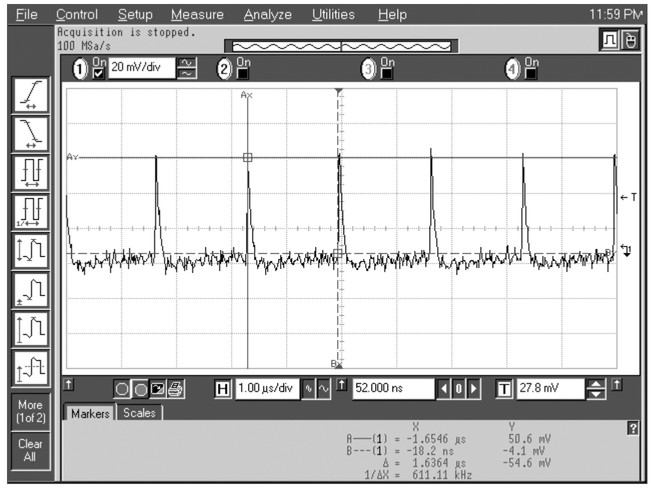


Figure 7.10 Repetitive jitter hits possibly due to switched-mode power supply breakthrough

It is important to note that the above baseband jitter plots show some jitter events that are potentially infrequent and unpredictable. Jitter measurements are sometimes made by statistical analysis of the eye-diagram displayed on a digital sampling oscilloscope. However, the sub-sampling nature of such an eye-diagram plot means some of the isolated jitter hits will be missed in the sampling process or be impossible to see on the displayed eye-mask. These jitter transients may only be possible to detect with a jitter tester using real-time data processing.

Intrinsic jitter calibration principles

Basic block diagram and design considerations

The block diagram shown in Figure 7.11, is typical of most jitter test sets. The diagram shows the elements involved in a jitter generation measurement.

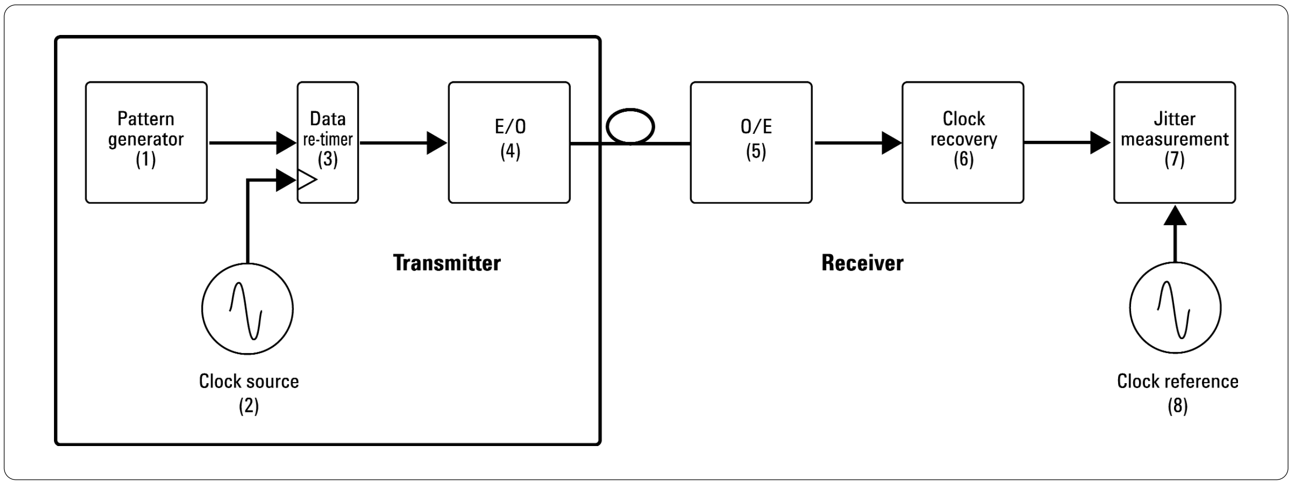


Figure 7.11 Back-to-back configuration for instrument jitter intrinsic measurement

The elements of the 10 Gb/s system of Figure 7.11 have been numbered and will be discussed in turn below:

(1) *Pattern generator* – is important as the properties of the digital pattern generated here will influence pattern-dependence of the system. In general, all the components will exhibit increased intrinsics with increasing Pseudo Random Binary Sequence (PRBS) test pattern complexity, plus unscrambled pattern areas (such as the SONET/SDH header) may cause a DC imbalance within the components. The system should be specified to operate with a representative pattern, such as a bulk-filled 2^{23} - 1 or 2^{31} - 1 PRBS, in SONET or SDH, giving stressful and realistic conditions for the components.

(2) *Generator clock source* – is the line-rate clock that times the digital electrical data. This clock will have some noise present, and a common way to specify the performance of a clock is to talk about its phase noise. In general, the higher the quality of clock, the lower the phase noise and the lower the peak-to-peak jitter. The quality and jitter contribution of this clock depends on how it is generated, but usually the peak-to-peak jitter contribution increases the closer it is measured to the carrier frequency. This means the peak-to-peak jitter measured is likely to increase as the jitter measurement high-pass filter is reduced in frequency, for example from 4 MHz to 50 kHz.

(3) *Electrical re-timer* – is the final electrical re-timing element before application to the optical modulator. This component may be a multiplexer output stage or a component such as a clocked modulator driver IC. In any case, the performance of the re-timer/optical modulator chain is key to minimizing pattern dependence and the consequential pattern-dependant jitter components. At the output of the electrical re-timer the electrical data signal may have jitter present due to pattern dependence within the device itself as well as additive noise due to the generator clock source.

(4) *Electrical/Optical (E/O) converter* – where the electrical data signal is converted to optical. The modulator may vary in type, but in general will add some pattern-dependent jitter to the transmitted optical data.

(5) *Optical/Electrical converter* (O/E) – which converts the optical data into electrical data and is usually a diode type detector. It can either be a PIN or APD device but must be selected for properties which don't add intrinsic jitter to, or degrade the data signal being measured.

(6) *Clock recovery* – is the processe whereby a clock is recovered from the scrambled data signal. The recovered clock must contain the jitter information in the data signal and therefore has to have sufficient bandwidth so that this is retained. Low intrinsic jitter and wide bandwidth are conflicting parameters here.

(7) *Jitter measurement* – which, in this example, is achieved by phase-detection on the recovered clock. The demodulated phase information is filtered in the required bandwidth, and peak-to-peak and RMS jitter values measured. Noise addition here is likely to some extent due to sampling processes or inherent component noise.

(8) *Clock reference* – is the jitter measurement timing reference. It can be phase locked to the recovered clock or be free running depending on the phase demodulation scheme. Similar to the clock source (2) above, the quality of this reference can make a difference to the intrinsic noise floor of the measurement.

Random jitter and deterministic jitter

Its worth now making some distinctions between the types of jitter talked about above. In terms of intrinsic jitter (or in ITU-T O.172 terminology, fixed error) these can be thought of as two distinct components: *Random Jitter* (RJ) and *Deterministic Jitter* (DJ).

Random jitter – is the accumulation of jitter through the system due to processes such as additive thermal and phase noise. This is generally Gaussian in nature and therefore assumptions can be made about its effect on the overall result in a given time period.

Deterministic jitter – is generated by various systematic effects, and the case examples described earlier show some potential causes of this. Pattern dependence, such as the unscrambled portion of the SDH/SONET header being one of those highlighted.

The jitter to be measured by a SONET/SDH tester is therefore the combination of both DJ and RJ components. Elements (5) - (8) in the block diagram of Figure 7.11 illustrate a possible implementation of a jitter measurement solution. Any practical implementation will have its own fixed error or intrinsic jitter residual that ultimately determines the accuracy of the measurement and hence the tester's ability to differentiate DUT jitter from its own intrinsics. The goal of the test equipment design must be minimize DJ and RJ components in the jitter measurement chain. Broadly speaking, elements (5) and (6) dominate with DJ and elements (7) and (8) with RJ.

Calibration guidelines

In order to qualify the fundamental intrinsic jitter performance of a jitter test set, some observations may be helpful. The same guidelines apply if one decides to "improve" the residual performance using a software calibration algorithm

- It may be obvious, but in a back-to-back measurement, the tester must measure its own generated jitter: the test set transmitter is the DUT for the jitter generation test. A jitter measurement receiver with no intrinsic jitter will display the jitter generation of the transmitter elements (1)-(4) in Figure 7.11.

- Calibrating out the intrinsic components is only valid if it does not impact the accuracy of the measurement. The example of "over-calibration" described earlier shows that the measurement may be meaningless.

- Calibration of DJ in the receive path must ensure measurement accuracy is not affected. DJ does not add linearly and therefore a noisy design cannot be reliably compensated for in calibration. Reliable calibration for DJ effects may be impossible.

- Calibration of RJ in the receive path can be achieved making assumptions about the additive nature of Gaussian noise to the result. This in no way means a large RJ component can be compensated without a detrimental effect on accuracy.

In summary calibration can be applied but only to enhance the overall accuracy of an already good measurement. Calibration applied to make the instrument appear to have low back-to-back intrinsics, when its system is in fact noisy, will render the measurement inaccurate or at worst useless.

Jitter conformance test

The conformance test described here will help the equipment user verify and quantify jitter measurement performance and accuracy. This is recommended as a method to evaluate the test set low-level intrinsic sensitivity and calibration integrity.

As shown in Figure 7.12, the user injects transient jitter onto an optical data signal on the transmit side and evaluates the jitter measurement receiver performance. The baseband demodulated jitter output should be used for a visual comparison on an oscilloscope of the demodulated jitter waveform versus the displayed jitter measurements. The results should be comparable.

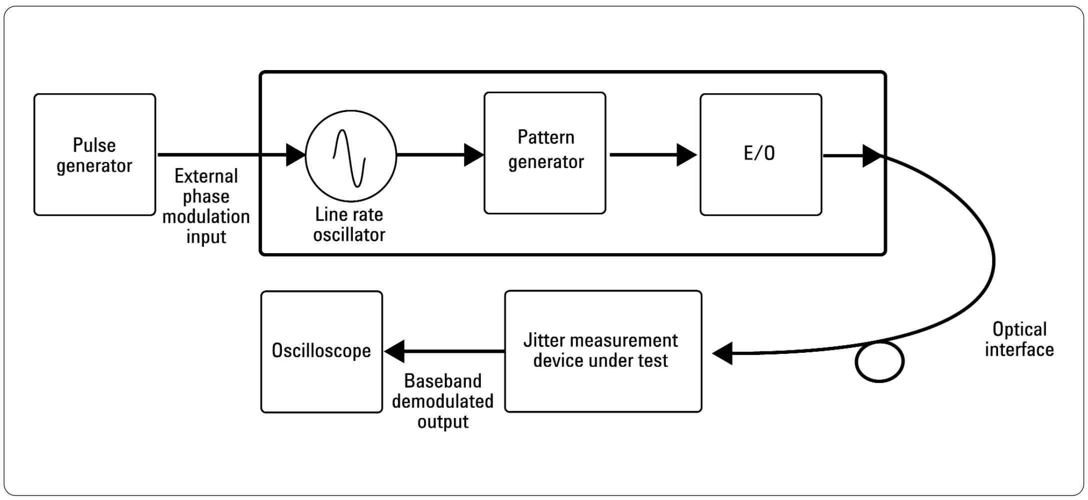


Figure 7.12 Test set up for evaluating residual jitter and peak detection performance

If the test equipment is exhibiting poor intrinsic performance, poor peak detection, or "over-calibration" then the method shown in Figure 7.12 can expose this phenomenon.

Conclusions

Test equipment may be deficient in two areas. Firstly a jitter receiver may not respond accurately to transients. Secondly an inherently poor intrinsic performance may be wrongly suppressed by software correction or calibration. In either case it may render the tester unfit for purpose.

It has been shown above that there may be significant issues measuring jitter generation with some current jitter testers that appear to conform to ITU-T O.172. The examples shown in this paper demonstrate the need for a jitter measurement system to be low-noise as well as possess the ability to measure and accurately characterise all types of jitter waveform.

References

1. ITU-T Recommendation O.172 (03/01) “Jitter and wander measuring equipment for digital systems which are based on the Synchronous Digital Hierarchy (SDH)”

2. ITU-T Recommendation G.783 (10/00) “Characteristics of synchronous digital hierarchy (SDH) equipment functional blocks”

8. What O.172 Doesn't Tell You

Abstract: This paper explores some of the less well-understood aspects of jitter measurement based on ITU-T O.172. It shows that genuine errors arise due to the statistics of different jitter modulation such as sinusoidal versus noise-like intrinsics. Measurement filters also perform differently depending on their noise bandwidth. Problem-free measurements require adherence to standards, good instruments and a good understanding of the problems.

Introduction

From other papers in this booklet, the reader will have realised that jitter is a difficult topic with many ramifications governed by a multitude of standards. The experienced practitioner in jitter measurements may also realise that these standards are subject to a range of interpretation which yields different measurement results depending on the test conditions such as observation period, jitter waveform, filter response and so on. One of the most important standards for measurement is ITU-T O.172, the specification for the jitter test equipment used to verify networks and network elements. This standard in particular has some areas of uncertainty in its current form (revision 03-2001), which the reader needs to be conscious of when comparing measurement results and different vendor's test equipment.

ITU-T O.172 is a good example of a specification, which is essential in itself and has brought many benefits to its users, but has some problems hidden away in the low-level details. Amongst other things it specifies and limits jitter intrinsics for measurement instruments, essential for evaluating a piece of network equipment. Unfortunately, a test set could pass O.172, yet compromise a network related intrinsic jitter test (see the companion paper on O.172 in this booklet). This paper examines some of the issues with O.172, and looks at the things people forget to include in intrinsic specifications.

How to get an intrinsic jitter answer, any answer!

The essential problem with O.172 is that it never properly considers the jitter modulation waveforms that occur in the real world. For jitter measurements, O.172 refers to either intrinsic measurements or sinusoidal amplitude measurements, and pure sinusoidal jitter is not the most common waveform on a real network.

One critical aspect of noise-like jitter intrinsic measurements is that the observation time is important for real measurements. This appears in some network specifications, but not in O.172. Jitter is normally checked as a peak-to-peak measurement, and occasionally as an average RMS value. RMS jitter is easy to think about: for a constant jitter input, the measured result converges on a constant value the longer the measurement is made.

Peak-to-peak jitter is different, especially for intrinsic measurements. There is no steady final answer, the measured value depends on the observation period. The basic issue can be understood by assuming that jitter intrinsics have a gaussian (normal) distribution, in other words like gaussian noise. A normal distribution has a mean value and a standard deviation. We can assume a zero mean value for jitter as the jitter waveform has been through a high-pass filter (to remove any DC offset) and for normally-distributed noise, the standard deviation is the same as the RMS value of the waveform. The standard deviation and the RMS value will be constant. What can be said about the peak-to-peak values of noise?

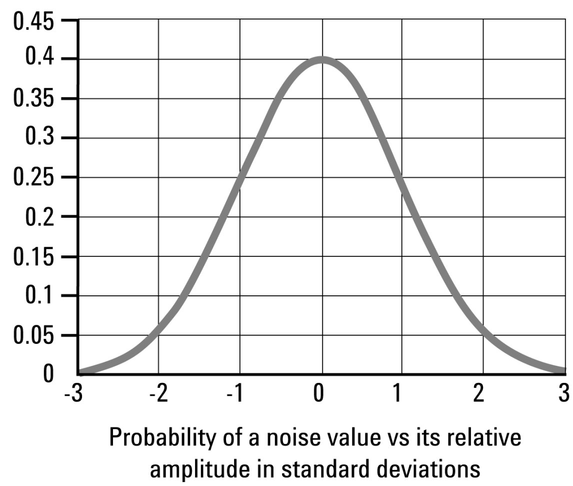


Figure 8.1 Probability density function (PDF) of additive white gaussian noise

The normal distribution curve is shown in Figure 8.1, in which the horizontal axis is in multiples of the standard deviation (sigma) equivalent to the RMS value, and the vertical axis is probability. It illustrates that the probability of a spike of a particular value is determined by a density function, the more extreme the value the lower its probability. The probability of a noise spike being the same as the RMS value (one sigma) is about 0.25, the probability of being twice that value is about 0.05. The probability of seeing a particular value during a measurement will depend on how many noise spikes are observed in the measurement period, and the RMS value of the noise. Larger jitter impulses become more probable, as more impulses are examined. Clearly time is involved, more time equals more impulses, but the measurement bandwidth is also a factor since more bandwidth equals more impulses per unit time. So it is actually the time-bandwidth product of the measurement that links the peak-to-peak jitter amplitude to the RMS value. The relationship is shown in Figure 8.2.

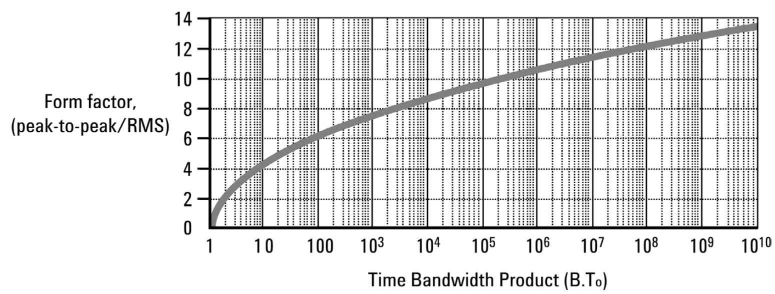


Figure 8.2 J_{p-p} / J_{RMS} form factor versus time-bandwidth product

For typical jitter measurement time-bandwidth products, the form factor between peak-to-peak jitter and RMS jitter is in the range of 10:1 to 14:1. There are a number of assumptions that we will come back to later, but the basic relationship is :

$$J_{p-p} = J_{RMS}\sqrt{(8\log_e(B.T_o))}$$

So measuring at STM-16 with a 20 MHz bandwidth for 60 seconds, we would expect the ratio of peak-to-peak jitter to RMS jitter to be greater than for a 60 second measurement at E1 with a 100 kHz bandwidth.
In the real world, it may be difficult to verify this due to the resolution of the RMS measurement and other factors discussed later.

It's obvious now, that the peak-to-peak jitter result is related to the observation period, the longer the gating time, the higher the result. The point is that while O.172 specifies the peak-to-peak intrinsic jitter levels, it does not define how long the instrument should be gated to get that answer. Instruments with very different intrinsic performance can claim compliance with O.172. It is possible however to get instruments that guarantee the intrinsic specifications in real-world situations, such as 60 second gating times, to ensure the repeatability and compatibility of different measurements.

How to save on company time

Jitter testing of network equipment usually has a specified observation interval for jitter measurement, often 60 seconds. Unfortunately this can become very expensive in test time and occasionally jitter testing is skipped, which raises the prospect of expensive problems once customers start their testing.

Take the simple example of a unit that generates 100 mUIp-p when observed over a 60 second interval. What results will be measured over shorter intervals? The theoretical results can be tabulated for a 20 MHz measurement bandwidth, assuming gaussian noise:

Observation Time (seconds)	Jitter (mUIp-p)
60	100.0
50	99.6
30	98.3
20	97.3
10	95.6
5	93.9
3	92.6
2	91.5
1	89.7

These results suggest that by carefully reducing the test-pass threshold, a unit can be tested over a much shorter interval. The standard deviation of the results will increase slightly as the test time is reduced so the drop-out rate will increase slightly if the threshold is not corrected for this effect. An alternative approach would be to automatically re-run the measurement over a longer time (or the full specification time) if a unit fails, thus allowing deterministic intrinsic components to be dealt with efficiently.

Alternative measurement tools

O.172 suggests the structure of a typical measurement instrument, but it does not indicate what assumptions are made. This becomes very important when trying to make measurements with an instrument that does not measure jitter in the same manner, such as a spectrum analyzer. By integrating the phase noise plot on the spectrum analyzer, a value for RMS jitter can be estimated.

The peak-to-peak jitter can than be calculated if one assumes a PDF for say gaussian noise. The spectrum analyzer is unable to detect the jitter peaks in real time.

Most telecom jitter waveforms contain quasi-random noise as well as deterministic signals that are structure/framing related such as 8 kHz spikes. The measurement of the composite jitter envelope is far more accurate with a broadband jitter measurement tool than an indirect calculation approach such as a spectrum analyzer. The O.172 standard has a number of built-in assumptions: they may not be clearly stated but they can impact measurement reproducibility when alternative measurements are attempted.

Why O.172 filter specification affects the measured result

Another consideration is the measurement bandwidth. O.172 only specifies the frequency response for sinusoidal waveforms, while for intrinsic jitter the noise-bandwidth of the filter is important and this is greater than the sinusoidal bandwidth by almost 5%. The filter corner-frequency specification allows a significant variation in the corner frequencies, and this will impact the noise bandwidth thus adding more uncertainty to the intrinsic measurement.

Clearly the O.172 specification has some interesting features for the unwary, and trying to perform an apparently equivalent measurement, for example with a spectrum analyzer, can lead to significant errors. Going back to the spectrum analyzer phase noise integration feature, it is possible to specify the limits of the frequency integration, but probably not the O.172 first-order high-pass filter at the lower cut-off and the third-order Butterworth low-pass filter at the high-frequency end.

A similar set of issues exist when using a spectrum analyzer to estimate RMS and by inference, peak-to-peak jitter. It probably uses a gaussian or pseudo-rectangular response in its IF bandpass filter[1], but what is its impulse response, what is its noise bandwidth, and can it really perform an O.172 measurement?

O.172 is universally used as a fundamental specification for jitter measurements and it is used to define test equipment used by both network equipment suppliers and purchasers. If there are potential errors caused by the standard it often does not affect commercial transactions since both the network equipment supplier and purchaser will have made the same measurement and got a similar result. If one party tries an alternative jitter measurement that is not O.172 compliant, then problems can arise – the two parties think they are performing a similar measurement but get different results.

How much wander gets into the jitter measurement?

Wander goes from very low frequencies up to 10 Hz, while jitter goes from 10 Hz upwards, usually to a maximum at some frequency below 5% of the line rate. The crossover between wander and jitter measurements is defined by a high-pass filter (HPF).

One issue is that the HPFs are essentially first-order in response, they don’t stop low frequency signals absolutely. Wander will always make some contribution to a jitter measurement, the only issue is how much and is it significant?

[1] The Intermediate Frequency (IF) filters in the spectrum analyzer provide the resolution bandwidth for the displayed measurement.

The nice thing about first-order filters is the simplicity making first-order approximations to the frequency response. A 200 Hz HPF will be down by a factor of 100 at 2 Hz (–40 dB). So, 1 UIp-p of wander at this frequency could contribute 10 mUI to a jitter measurement, it rather depends on the waveforms. This problem is reduced at the higher SDH rates due to the higher frequency of the HPF responses but can cause real problems at tributary rates such as DS3 with its 10 Hz filter. Never forget that a jitter intrinsic problem may have nothing to do with jitter, wander can be the problem. The clue can be that the short-term jitter results can be seen to move up and down in a slow, gentle manner.

Temperature effects

Jitter measurements are affected by temperature through two basic mechanisms. Firstly the phase-noise performance of the reference source is temperature-dependent, and secondly the jitter measurement hardware such as clock recovery can be temperature sensitive. It can be shown that an oscillator's phase noise will increase by about 8% when its temperature is raised from 25 °C to 50 °C. Correspondingly, its jitter envelope will also increase by about 4%. It does not matter how a jitter measurement system is implemented, whether it is analog or entirely digital, it must use a time/frequency reference that will have temperature-related intrinsic jitter.

Another concern in jitter measurement instruments is that the intrinsic jitter performance of the measurement hardware will be temperature dependent. The O.172 standard does not give any guidance in this area. This dependence can be partially compensated for by using good design practice, but the result can be that the compensated temperature-dependency might have either a positive or negative coefficient. Practical instruments include software compensation to partly correct the predicted variations in the internal intrinsic-jitter temperature performance.

This compensation will normally be a single algorithm that is the best fit for typical instruments. There is a better alternative available for some instruments where the test equipment manufacturer characterizes each individual instrument for effects such as temperature, before programming it with a unique correction algorithm. This is normally available as an option when consistent low intrinsics are needed. Without this option, instruments may have low intrinsics in most respects and be within the O.172 specification, but how well they stay within specification may vary with temperature.

Conclusions

This paper has reviewed a number of the less well-known issues of jitter measurement using the O.172 standard.

The O.172 specification has areas of vagueness, so compliant instruments may perform differently. Jitter measurements can be very difficult to justify when using a non-standard instrument such as a spectrum analyzer or high-speed oscilloscope. Results consistency is dependent on consistency of the measurement environment (including measurement filters) and the observation/gating time. Jitter testing can be speeded up in a justified manner by making assumptions regarding the PDF of intrinsic jitter. In general, problem-free measurements require adherence to standards, good instruments and a good understanding of the problems.

9. Measuring 100 mUIp-p Jitter Generation with an O.172 Tester?

Abstract: This paper examines one of the most difficult measurements to make accurately in the field of jitter testing. The paper concludes that the only practical solution is to use a jitter receiver with very low intrinsics, below that specified in ITU-T O.172, to reliably measure jitter generation of less than 100 mUIp-p.

Introduction

Increasingly, Network Equipment Manufacturers (NEMs) are being confronted with the need to design equipment having less than 100 mUIp-p intrinsic jitter. *Jitter generation*, to give this characteristic its correct name, should not be confused with *network jitter*, which can be measured at any interface in a network. (See the paper on Jitter Generation in this booklet). To account for the accumulation of jitter throughout the network, *network jitter* requirements are less stringent than *jitter generation* requirements, which apply to a single piece of equipment. Network jitter requirement is typically 150 mUIp-p compared with 100 mUIp-p for jitter generation.

The 100 mUIp-p jitter generation requirement is most clearly stated in Telcordia standard GR-253[1], the definitive SONET standard. The 100 mUIp-p requirement can be found in the SDH world too, such as in ITU-T recommendation G.813[2], but here it is somewhat hidden in a tangle of different requirements that few are brave enough to navigate with authority. So for the sake of simplicity, this paper will focus on the relatively straightforward SONET world of GR-253.

Test conditions for the 100 mUIp-p jitter generation requirement

Jitter Generation refers to the intrinsic jitter generated within a Network Element (NE). It is measured at the output of the NE, with no jitter or wander applied at the input.

The 100 mUIp-p jitter requirement applies to Category II interfaces[1], which include all OC-n optical and STS-n electrical interfaces to SONET NEs.

Jitter is a noise process, so it is statistical in nature. Maximum peak-to-peak jitter is accumulated and measured over a 60 second interval[3]. Low-pass and high-pass filters are also used, to limit the measurement to the jitter frequency range of interest[4]. Referring to the accompanying paper in this booklet on jitter generation measurement, the recommended test signal is a *bulk payload with the*

[1] Telcordia Technologies GR-253-CORE Issue 3 September 2000, section 5.6.2.3.6 Category II Jitter Generation, requirement R5-258.
[2] ITU-T Recommendation G.813 (08/96), section 7.3 Jitter, part b) Option 2.
[3] See GR-253-CORE section 5.6.1 Network Interface Jitter Criteria.
[4] See GR-253-CORE section 5.6.2.3 Jitter Generation.

longest Pseudo Random Binary Sequence (PRBS) as this is considered more stressful and usually generates the largest peak-to-peak jitter. *Concatenated* payloads provide the worst-case scenario for OC-n/STM-n signals. For bulk-filled concatenated signals with a 2^{23}-1 PRBS filling the container, the result of scrambling this data is a worst case run of 30 consecutive “ones”/”zeros” (i.e. there will be a maximum of 30 clock periods with no transitions on the line signal). In general, a very long-run PRBS will create worst case noise-like jitter generation.

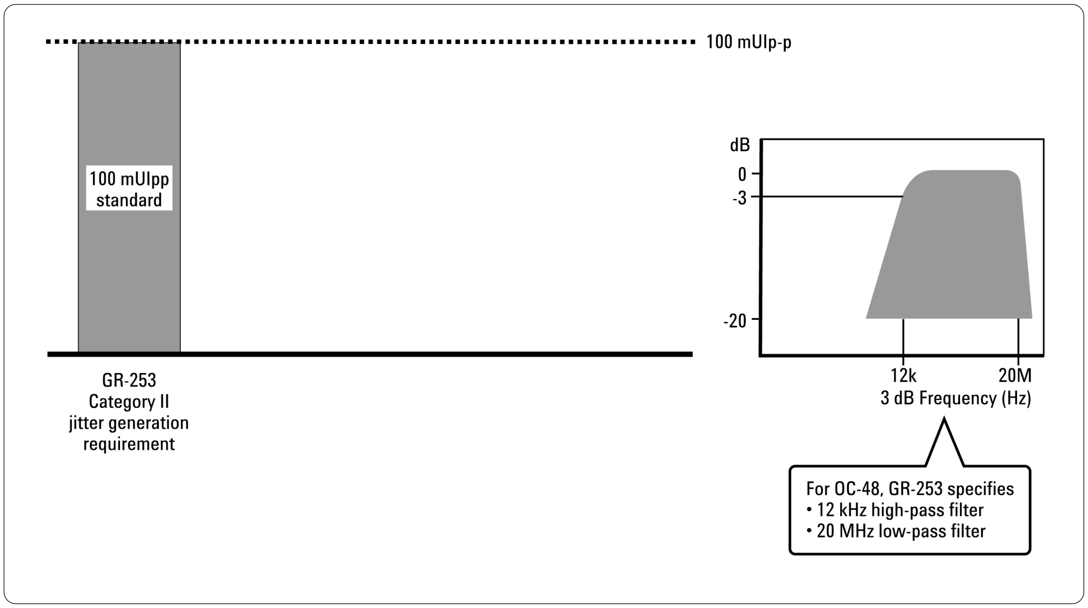


Figure 9.1 Telcordia GR-253 Cat. II jitter generation requirement

Measurement problems with an O.172 specified test set

Most SONET jitter test sets available today specify the accuracy of their jitter measurement according to ITU-T recommendation O.172.

O.172 specifies jitter measurement accuracy in a number of ways, all in terms of maximum peak-to-peak jitter using a concatenated SDH (SONET) signal with a $2^{23} - 1$ PRBS payload[5]. The total measurement error allowed comprises fixed (W) and variable (R) components:

Error = $\pm R\%$ *of reading* $\pm W$

- R is the variable component, expressed as a percentage of the reading. R varies according to the frequency of jitter present in the signal. At STM–16 (OC–48), R ranges from ±7% for low jitter frequencies to ±20% for high jitter frequencies.

- W is the fixed error, expressed in UI. This is the intrinsic jitter of the tester’s receiver.

[5] ITU-T O.172 (03/2001), section 9.4 measurement accuracy.

The potential problem with using an O.172 specified test set is illustrated in Figure 9.2. The GR-253 specification calls for a 100 mUIp-p maximum jitter generation for the Device Under Test (DUT), however the O.172 specification also calls for a maximum tester receiver residual or intrinsic jitter (W) of 100 mUIp-p, making it impossible to measure accurately the jitter generation performance of the DUT. In addition, there is the variable component (R). Over the narrower band of 1 MHz to 20 MHz, the O.172 residual specification of 50 mUIp-p is more realistic for measurement of 100 mUIp-p jitter generation in the DUT. Here, however, the bandwidth is completely inadequate since it excludes all of the lower frequency jitter.

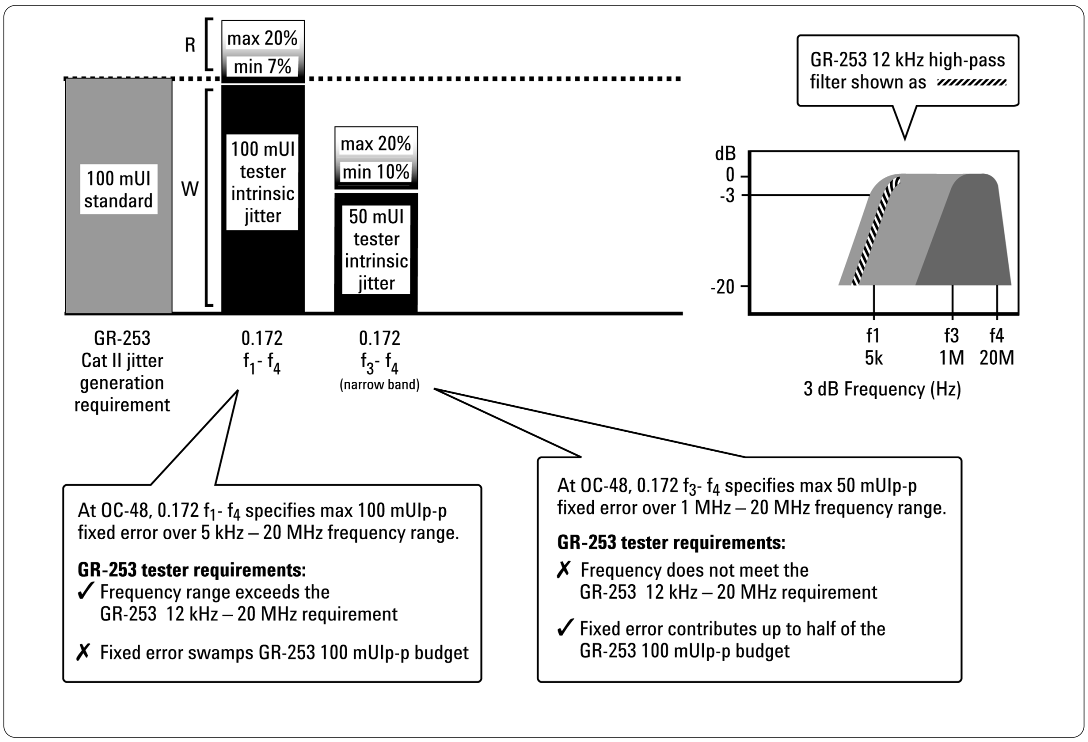


Figure 9.2: ITU-T 0.172 jitter measurement accuracy cannot guarantee compliance with Telcordia GR-253

Finding a solution

The stringent jitter generation level specified in GR-253 and ITU-T recommendations for SONET/SDH equipment has created a challenge for test equipment manufacturers. From the above analysis, it is clear that simply meeting the O.172 specification is insufficient to give reasonable accuracy with a DUT performing at or better than the 100 mUIp-p specification.

One proposed solution is to normalize out the intrinsics of the test set by subtracting its residual value from the total measured result. This in itself presents a problem as to what level of correlation exists between the jitter generation in the DUT and the intrinsics of the test set. Both may be caused by similar pattern dependency or may be uncorrelated random noise. How do you quantify the residual jitter of the measurement receiver? Connecting back-to-back with the measurement transmitter may not be accurate depending on the jitter generation in the transmitter. Certainly, subtracting a residual or intrinsic jitter that is of similar magnitude to the expected measured result, will be prone to very large errors and the DUT test will be meaningless.

At Agilent Technologies, we believe the only satisfactory solution to this problem is to build a jitter receiver with much lower intrinsics than required by O.172. For some time now, the Agilent OmniBER 718 has delivered the industry's best jitter intrinsic performance in the shape of option 200. Typically, this allows designers to work to a 30 mUIp-p safe design margin for jitter generation measurements, as shown in Figure 9.3.

Now, the OmniBER also offers an optimized version – option 210 – that guarantees receiver jitter intrinsics, 30% lower than the current best-in-class levels of the option 200[6]. You can now work, with even more confidence, to a 45 mUI safe design margin for peak-to-peak jitter, and prove your DUT meets GR-253.

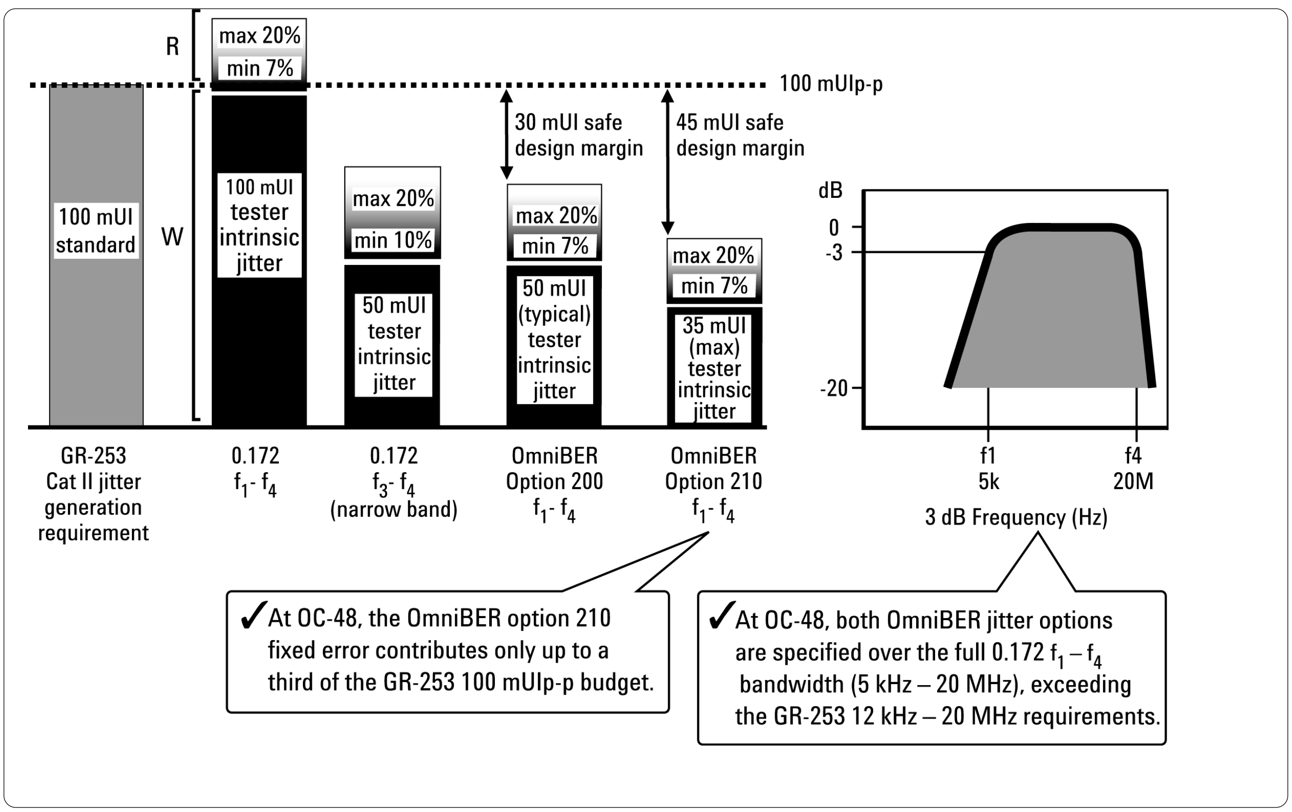


Figure 9.3: New options on the OmniBER 718 guarantee low intrinsics and accurate measurement of jitter generation.

6 30% reduction in receiver jitter intrinsics applies at OC-48 and OC-12 data rates.

Conclusions

Testing to the 100 mUIp-p jitter generation requirement and achieving compliance with standards such as GR-253 is a difficult task. In fact, when using a test set that specifies its jitter measurement accuracy according to O.172, it's impossible.

OmniBER offers a way out of this situation by specifying its 2.5 Gb/s jitter measurement accuracy across the full 5 kHz to 20 MHz bandwidth. There are two optional levels of receiver jitter intrinsics, both well within the 100 mUIp-p budget. OmniBER 718A opt 210 guarantees receiver jitter intrinsics to be no more than 35 mUI, allowing measurements to 100 mUI with full traceability.

Appendix

Performance comparison between Telcordia GR-253, ITU-T O.172 and OmniBER 718 options.

Bandwidth comparison

	GR-253[7]		O.172 f1-f4[8]			O.172 f3-f4[8]			OmniBER 718 options 200 and 210		
	HP	LP	f1 HP	f4 LP	Meets GR-253?	f3 HP	f4 LP	Meets GR-253?	HP	LP	Meets GR-253?
OC-48	12 kHz	20 MHz	5 kHz	20 MHz	✓	1 MHz	20 MHz	✗	5 kHz	20 MHz	✓
OC-12	12 kHz	5 MHz	1 kHz	5 MHz	✓	250 kHz	5 MHz	✗	1 kHz	5 MHz	✓
OC-3	12 kHz	1.3 MHz	500 Hz	1.3 MHz	✓	65 kHz	1.3 MHz	✗	500 Hz	1.3 MHz	✓
STS-3	12 kHz	1.3 MHz	500 Hz	1.3 MHz	✓	65 kHz	1.3 MHz	✗	500 Hz	1.3 MHz	✓
OC-1	12 kHz	400 kHz	100 Hz	400 kHz	✓	20 kHz	400 kHz	✗	100 Hz	400 kHz	✓
STS-1	12 kHz	400 kHz	100 Hz	400 kHz	✓	20 kHz	400 kHz	✗	100 Hz	400 kHz	✓

In addition to the high-pass filters listed above, OmniBER also offers a 12 kHz high-pass filter at all rates, convenient for GR-253 peak-to-peak and RMS jitter measurements.

Peak-peak jitter comparison

	GR-253	0.172 f1-f4			OmniBER 718 opt 200			OmniBER 718 opt 210		
	Max p-p[9]	W[10]	Max R[11]	Meets GR-253?	W (typical)	Max R	Meets GR-253?	W	Max R	Meets GR-253?
OC-48	100 mUI	100 mUI	±20%	✗	50 mUI	±20%	✓	35 mUI	±20%	✓
OC-12	100 mUI	100 mUI	±15%	✗	50 mUI	±15%	✓	35 mUI	±15%	✓
OC-3	100 mUI	70 mUI	±10%	✓	35 mUI	±10%	✓	35 mUI	±10%	✓
STS-3	100 mUI	70 mUI	±10%	✓	35 mUI	±10%	✓	35 mUI	±10%	✓
OC-1	100 mUI	70 mUI	n/a	?	35 mUI	±8%	✓	35 mUI	±8%	✓
STS-1	100 mUI	n/a	n/a	?	35 mUI	±8%	✓	35 mUI	±8%	✓

RMS jitter comparison

	GR-253	0.172 f1-f4	OmniBER 718 options 200 and 210		
	Max rms[9]	W	W (typical)	Max R	Meets GR-253?
OC-48	10 mUI	n/a	4 mUI	±20%	✓
OC-12	10 mUI	n/a	4 mUI	±15%	✓
OC-3	10 mUI	n/a	4 mUI	±10%	✓
STS-3	10 mUI	n/a	4 mUI	±10%	✓
OC-1	10 mUI	n/a	4 mUI	±8%	✓
STS-1	10 mUI	n/a	4 mUI	±8%	✓

O.172 treats RMS jitter measurement as optional[12], so it contains no accuracy specification for RMS jitter measurement.

7 Figures from GR-253-CORE section 5.6.2.3 and Table 5-9.
8 Figures from ITU-T O.172 Table 7/O.172.
9 Figures from GR-253-CORE section 5.6.2.3.6, requirement R5-258.
10 Figures from ITU-T O.172 Table 8/O.172.
11 Figures from ITU-T O.172 Table 10/O.172.
12 See ITU-T O.172 section 9.2.3 Measurement of RMS jitter.

10. Faster Jitter Testing with Simultaneous Filters

Abstract: This paper describes the advantages of a parallel-filter configuration in the jitter measurement receiver, which greatly reduces overall test time for jitter generation measurements. The parallel architecture is also more reliable in capturing transient events.

Introduction

Jitter specifications for network equipment recommended by the standards bodies require detailed testing as described elsewhere in this booklet. These tests can be time-consuming and costly, but they are not easily bypassed or abbreviated if reliable measurements are needed. Up until now, test and manufacturing plans have had to take into account the extended test time required to ensure compliance with both ITU-T and Telcordia standards.

Unfortunately, implementation of the required test routines by test equipment vendors has sometimes led to slow measurements and inconsistent results. This is now all set to change with the development of parallel testing, as introduced into the Agilent Technologies OmniBER OTN jitter analyzer. This ensures that the OmniBER OTN provides all the necessary tools for fast and accurate jitter characterization and analysis in development and production.

Parallel jitter measurements

Two arrangements for measuring output jitter are shown in Figure 10.1. The top diagram (Figure 10.1(a)) is the most common method of jitter measurement, whereby high-pass and low-pass filters are selected and jitter measured, before moving sequentially to the next appropriate filter combination and measuring that result. While this is a perfectly valid method of measuring jitter, it can be very time-consuming, since the standards bodies recommend a gating time of 60 seconds each time a new filter is selected.

The lower diagram (Figure 10.1(b)) depicts a recently developed method for jitter test, in which all of the standard filter bandwidths are continuously monitored and results displayed for all filter bandwidths simultaneously. This presents the user with several key advantages, while not affecting the measurement accuracy:

- Significantly faster testing, ultimately reducing cost and time to market.
- The ability to identify and analyze any intermittent “spikes” of jitter which may otherwise be missed.
- Fewer button pushes required to set-up measurements and view results, thus reducing the chance of errors.

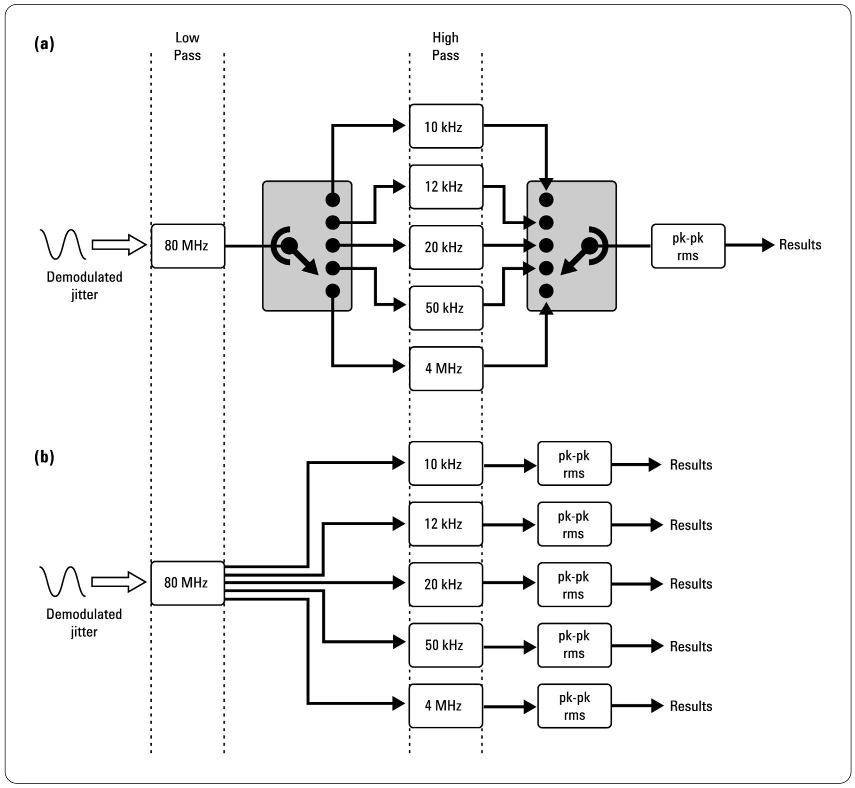


Figure 10.1(a)/10.1(b) Sequential and parallel filter configurations for 10 Gb/s measurements

An immediate advantage of parallel filters is the reduced test time required because all results for all filters are obtained with one measurement period. By using sequential filters, users must switch between filters, only recording one set of results at a time. This means that tests must be repeated if more than one filter combination is required.

The filters illustrated above are relevant to both SONET and SDH. For OC-192 at 10 Gb/s, Telcordia GR-253 states a requirement for 20 kHz and 4 MHz high-pass and 80 MHz low-pass filters at network interfaces, and 50kHz high-pass and 80MHz low-pass for jitter generation measurements. For STM-64, ITU-T specifies 20 kHz and 4 MHz high-pass and 80 MHz low-pass filters for network interfaces.

Since the standards bodies recommend a test time of 60 seconds every time the test is performed, the time saving can be very significant in a high-throughput production line or verification lab. Test time will also be dramatically reduced for any network equipment destined to cover both SONET and SDH, since all SONET and SDH jitter filter options will be tested simultaneously. A manufacturer currently testing the complete set of high-pass and low-pass filter combinations for SONET and SDH, will need to run jitter tests at least three times (assuming only one line-rate is used). Obtaining the same results now in a third of that time, implies that there will be significant savings not only in time, but also in cost-per-test. These savings will be multiplied considerably if the device under test (DUT) requires testing at more than one line-rate.

The ability to measure and analyze any jitter transient events is also improved by using parallel filters. Standards bodies recognize jitter can be systematic, random or bursty in nature so it is vital that jitter testers are capable of detecting and accurately measuring non-sinusoidal events. In parallel-filter architectures, the very fact that all results for all filters are displayed simultaneously means that there is an extremely low risk of any transient anomalies being missed by not having the appropriate filter bandwidth activated. Engineers using the parallel filter configuration can be more confident that there are no hidden issues waiting to cause problems further down the line.

Another benefit in terms of reliability is that there will be no chance of erroneously selecting an inappropriate filter, since all filters are simultaneously active. Logging results will always capture all data from all filters, increasing traceability and allowing a full spread of results to be saved after only one test period.

Understanding measurement results

One feature of parallel-filter results, which often causes confusion, is that some narrowband-filter jitter results can appear to be higher than wideband-filter results. At first glance this appears slightly odd, but in fact this is not an error in the test equipment, nor is it an incorrect result due to any problem with the device under test. Figure 10.2 depicts two jitter waveforms that might be expected from a wideband filter measurement, 50 kHz – 80 MHz (a) and a narrowband filter measurement, 4 MHz – 80 MHz (b).

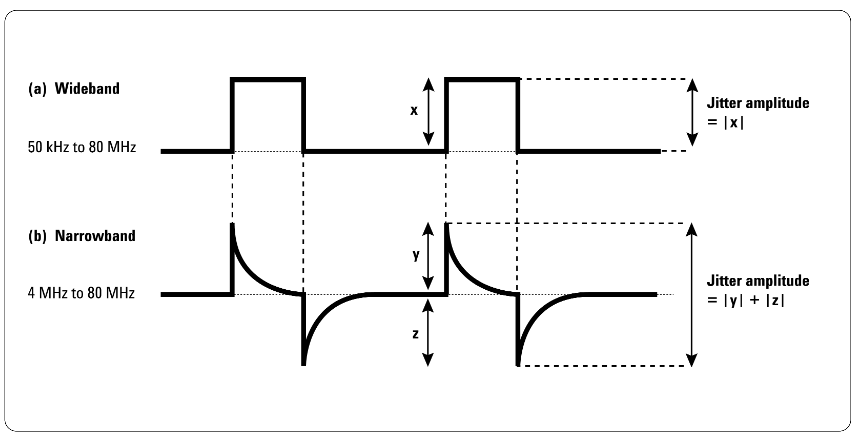


Figure 10.2 Jitter waveform for wideband (a) and narrowband (b) filters

These show the effects of the 50 kHz and 4 MHz high-pass filters on the peak-to-peak jitter waveform. It is interesting to note that for this demodulated jitter waveform the narrower bandwidth has a higher peak-to-peak reading than the wideband reading. This is not intuitively what might be expected. The 4 MHz high-pass filter acts as a differentiator and increases the absolute peak-to-peak amplitude of the waveform.

The RMS jitter levels on the other hand would typically be lower for the narrowband example than for the initial square wave, because the waveform contains less energy. This effect is not any fault in the test equipment, but is a natural consequence of decreasing the measurement bandwidth. It should be evident when measured on any tester, but is made all the more apparent with the parallel configuration allowing a direct comparison between different bandwidths during the same measurement period.

The waveform in Figure 10.3 shows the demodulated jitter output created through pattern-dependence in the DUT caused by the SONET/SDH header. This shows that the 4 MHz – 80 MHz waveform has a higher peak-to-peak value than the 50 kHz – 80 MHz measurement, for this particular example.

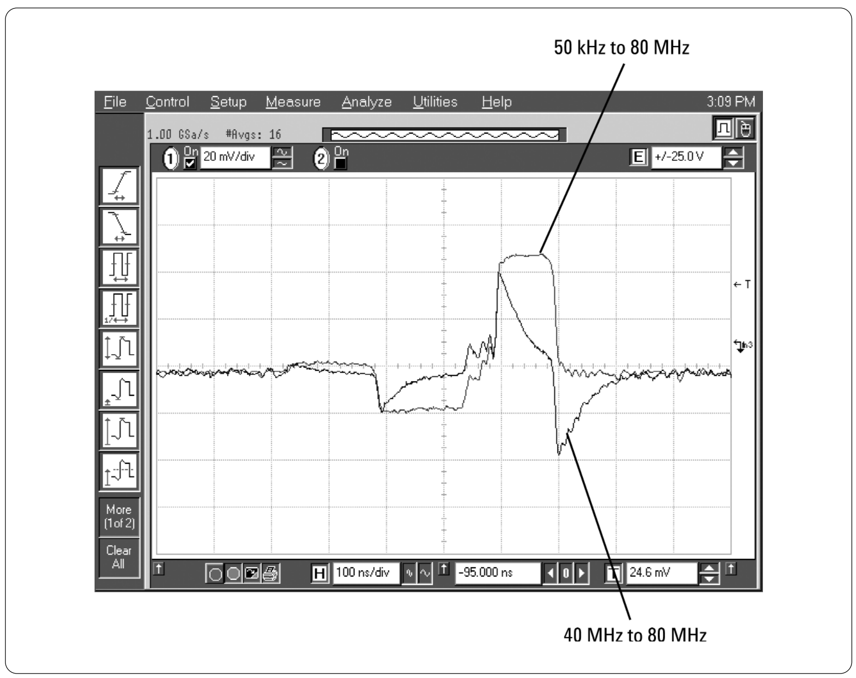


Figure 10.3 Demodulated jitter waveform caused by patter-dependence due to SONET/SDH header, as measured with wideband and narrowband filters

Conclusions

Accurate measurement of output jitter and jitter generation is time consuming as several filter combinations are specified in the standards, each requiring a minimum gating period for reliable results. Parallel filter configuration in the test set substantially reduces this overall measurement time, provides a comprehensive test record for proving equipment compliance with standards and is more dependable when capturing transient events.

11. Verifying Jitter Generator Performance

Abstract: This paper describes techniques based on spectrum analysis to independently verify the calibration of a jitter test transmitter. Accuracy of the jitter generator is important as it has a direct bearing on the reliability of jitter tolerance measurements. Modulation theory and the application of Bessel functions lead to three measurement methods, applicable to different levels of jitter. The paper concludes with test results and error analysis using the Agilent 718 communications performance analyzer as an example.

Introduction

The Agilent OmniBER 718 communications performance analyzer has a jitter option, which allows the generation of SONET and SDH signals, with accurately defined levels of sinusoidal jitter. The transmitter part of this instrument is show in Figure 11.1 below.

The jitter is generated in the modulated clock source, which is frequency modulated by an internally generated sinusoidal signal. Since the clock source electrical output is at the full line-rate, and there are no phase-locked loops (PLLs) or multipliers within the data generator, the clock source defines the jitter of the optical line rate signal.

During production of the OmniBER 718 analyzer, the accuracy of the clock source jitter is automatically tested by measurements using a spectrum analyzer. This paper outlines the methods used to verify the jitter amplitude, and some typical results are included.

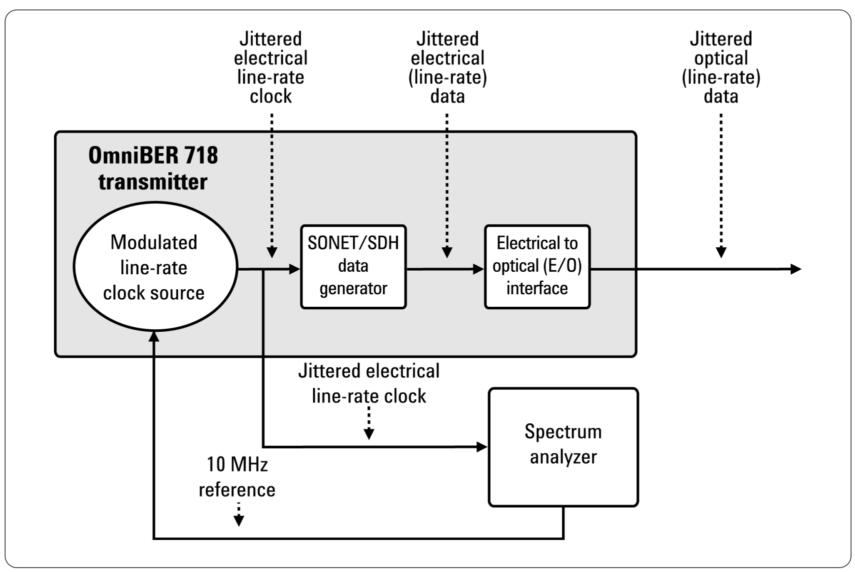


Figure 11.1 Transmitter jitter verification block diagram

Theory

In this section the background theory of jitter is discussed. Jitter is effectively phase modulation of a carrier signal, so the normal mathematical equations for phase/frequency modulation apply.

For SONET/SDH signals, the jitter modulation is described by:

$v(t) = V\sin(2\pi L_r t - \pi(UIp\text{-}p)\cos(2\pi J_r t))$ .... equation 1

L_r = line rate in Hz
J_r = jitter rate in Hz
UIp-p = unit intervals of jitter measured peak-to-peak
where 1 UIp-p = 1 unit interval = $1/L_r$ = 2 π radians peak-to-peak

Jitter is also described by Bessel functions which use modulation-index M_f to indicate the amount of jitter. The equivalence between modulation-index M_f and UIp-p jitter is:

Peak-to-peak phase deviation (UIp-p) = M_f/π .... equation 2

Hence Bessel functions can be used to calculate the spectrum of a clock signal for any jitter amplitude.

The spectrum of a jittered clock contains a carrier and symmetrically placed sidebands, the usual notation is:

J0 = amplitude at the line-rate, or carrier frequency
J1 = amplitude at (the line rate ± jitter rate)

Several important deductions can be made for measurement purposes from the amplitude of the carrier (J0) and first sidebands (J1):

- The ratio J1/J0 can be used to measure jitter amplitude.
- The ratio method is useful for low levels of jitter (0 to 2 UIp-p).
- The amplitude of J0 becomes zero (or null) at defined jitter amplitudes, thus making these nulls a useful calibration point for specific jitter amplitudes.
- The nulls are useful for intermediate levels of jitter up to about 50 UIp-p.

Equation 1 can also be used as the basis for measuring much larger levels of jitter. The instantaneous frequency deviation can be derived by differentiating equation 1, (frequency = differential of phase, dø/dt).

Frequency deviation = $d(\pi(UIp\text{-}p)\cos(2\pi J_r t))/dt$ .... equation 3

Peak frequency deviation = $\pm\, \pi(UIp\text{-}p)J_r$ .... equation 4

For signals with large levels of sinusoidal jitter, it is found that the displayed spectrum has peaks located very near to:

Line-rate – π(UIp-p)Jr, and line rate + π(UIp-p)Jr

Two observations follow from this:

- The frequencies of the peaks allow a measurement of jitter amplitude for signals with large levels of jitter. These signals are wide-band frequency modulation signals (wide-band FM).
- This method is useful for high levels of jitter (20 UIp-p to beyond 800 UIp-p).

Three Measurement Techniques

From the above theory, three measurement techniques can be derived for verifying the levels of injected sinusoidal jitter.

For low jitter amplitudes: J1/J0 ratio method

Following measurement, the J1 to J0 ratios can be calculated along with the corresponding jitter amplitude, using the mathematical formulae. For example:

J1/J0 = 1.200895 corresponds to 0.5 UIp-p.

The ratio of J1 to J0 is useful for measuring jitter in the ranges:

- 50 mUIp-p to 600 mUIp-p
- 900 mUIp-p to 1100 mUIp-p .... for a test at 1 UIp-p
- 1300 mUIp-p to 1600 mUIp-p .... for a test at 1.5 UIp-p

These ranges avoid the regions where either J0 or J1 are at or near a null, which would push the ratio towards zero or infinity, making it unusable.

Note that J1 is the amplitude of the first-order sidebands, and are located at the line-rate ± jitter modulation rate.

For intermediate jitter amplitudes: Bessel “nulls” method

The nulls in the amplitude of J0 (the line rate) correspond to particular jitter amplitudes of:

0.7655 UIp-p, 1.7571 UIp-p, 2.7546 UIp-p, ... in general, approximately N.75 UIp-p, where N is an integer.

The J0 nulls allow the jitter amplitude to be verified at specific values of UIp-p.

Note that the nulls are defined by the jitter amplitude only, and are independent of the jitter modulation rate.

For high jitter amplitudes: wide-band FM peaks method

For very large amplitudes of jitter, for example 800 UIp-p, the spectrum has two amplitude peaks near to the frequencies:

Line rate $\pm$ (UIp-p) πJ_r

Where J_r = jitter modulation rate

For practical measurements, a more accurate result is obtained by using the –3 dB frequencies of the peaks, which are related to the jitter amplitude and modulation rate.

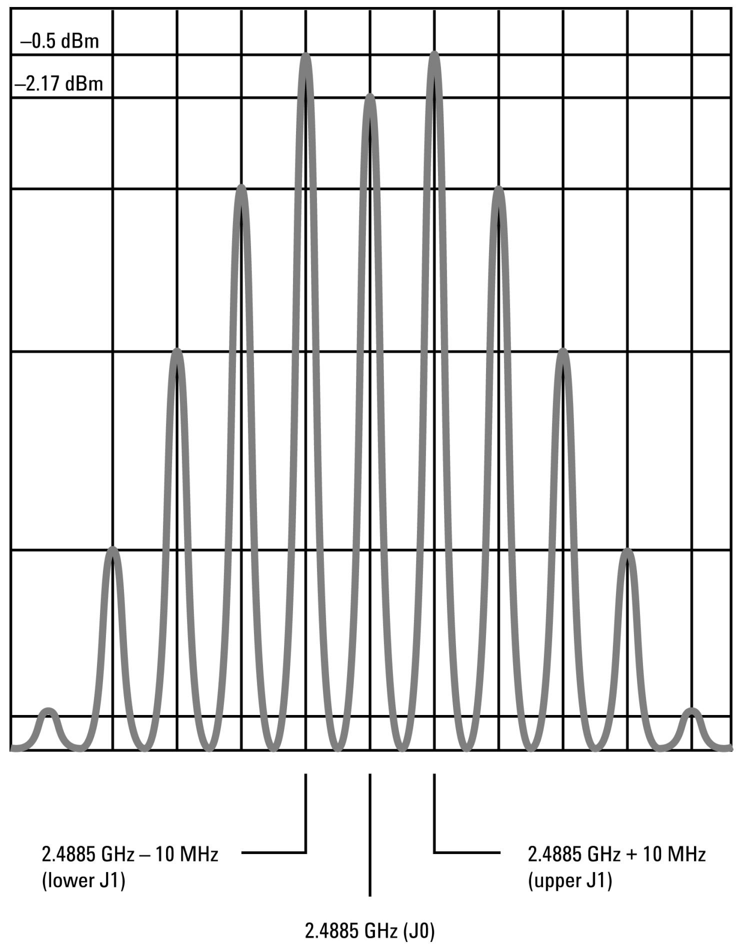
The jitter amplitude can be measured using:

$$\text{UIp-p} = (F_{\text{upper} - 3\text{ dB}} - F_{\text{lower} - 3\text{ dB}})/(2\pi\, J_r)$$

Having reviewed the theory and application of these spectrum-based jitter measurements, the following sections give methods and examples of practical measurements with an analysis of the measurement errors.

J1/J0 ratio method

- The jitter amplitude and rate are set on the instrument transmitter under test, and the amplitudes of J0 (line rate) and J1 (upper and lower sidebands) are measured using the spectrum analyzer.
- All measurements are performed without amplitude-range changes to eliminate range-change errors.
- The spectrum analyzer detector type is set to LOG for jitter amplitudes ≤ 150 mUIp-p and LINEAR for jitter amplitudes >150 mUIp-p. This maximizes the accuracy of the ratio measurement.
- The ratio J1/J0 is located in a look up table to find the corresponding jitter amplitude.
- Accuracy: ~5% for jitter amplitudes <200 mUIp-p, ~3% for jitter amplitudes ≥ 200 mUIp-p
- Measurement integrity is monitored by measuring both upper and lower sidebands, and checking the amplitudes are within 0.5 dB. Additional check point: 0.460 mUIp-p corresponds to a ratio = 1.013719



Jitter measurement:

J1 = –0.5 dBm = 0.211098 V_{rms}
J0 = –2.17 dBm = 0.174175 V_{rms}
J1/J0 = 1.21199 = 0.502 UIp-p

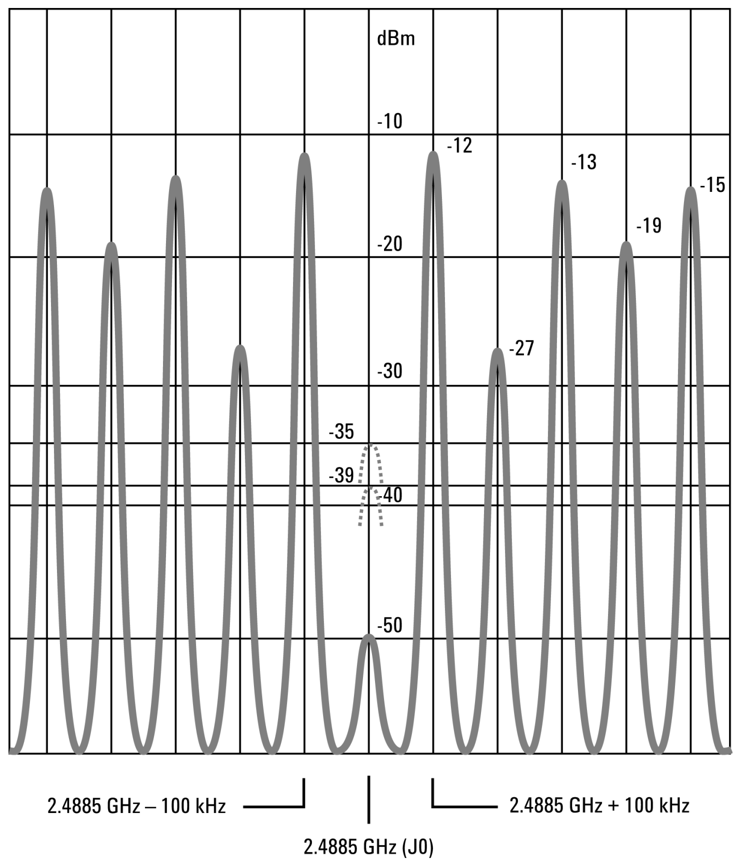
Jitter Tx settings:

Line rate = 2.4885 GHz (b/s)
Jitter amplitude = 0.5 UIp-p
Jitter rate = 10 MHz

Figure 11.2 J1/J0 ratio method

Bessel "null" method

- The jitter amplitude is incremented from 0 UIp-p to 20 UIp-p (or the range maximum) on the transmitter under test, and the magnitude of J0 measured on the spectrum analyzer. The results are stored in a table.
- The table of UIp-p setting and J0 amplitude is searched to locate the position of the nulls in J0.
- Interpolation allows for accurate null determination, avoiding any limitations due to the jitter amplitude resolution of the transmitter under test.
- Accuracy: better than 10 mUIp-p for any null.



Jitter measurement:

J0 amplitude is –35 dBm at 9.73 UIp-p (near the null)
J0 amplitude is –50 dBm at 9.75 UIp-p (at the null)
J0 amplitude is –39 dBm at 9.77 UIp-p (near the null)

Jitter Tx settings:

Line rate = 2.4885 GHz (b/s)
Jitter amplitude = 9.75 UIp-p
Jitter rate = 100 kHz

Figure 11.3 Bessel "null" method

Wide-band Frequency Modulation (WBFM) method

- The spectrum analyzer is used to measure the frequency of the –3 dB points of the peaks, see Figure 11.4.
- The –3 dB point frequencies are used to calculate the jitter.
- This measurement requires a slow sweep-rate and low resolution and video bandwidths.
- Accuracy: ~ 1%.
- Measurement integrity is monitored by measuring both peaks and checking the peaks are symmetrical about the line rate (to better than ±1%).

Specifically:

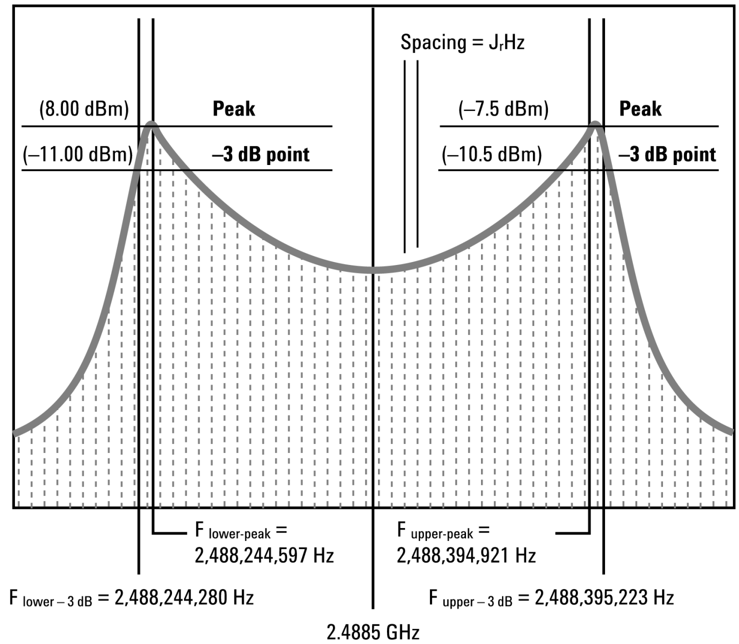
$$\frac{\text{Line rate} - \text{Negative peak frequency}}{\text{Positive peak frequency} - \text{Line rate}} < 1.02\text{, and}$$

$$\frac{\text{Positive peak frequency} - \text{Line rate}}{\text{Line rate} - \text{Negative peak frequency}} < 1.02$$

Example:
Line rate = 2,500, $\Delta f = \pm 100$, Peaks offset by +1
Negative peak = 2,500 – 100 + 1 = 2,401
Positive peak = 2,500 + 100 + 1 = 2,601

$$\frac{2{,}601 - 2{,}500}{2{,}500 - 2{,}401} = 101/99 = 1.0202 \text{ detects excessive offset}$$

- The method can be cross-checked against Bessel “nulls”.



Jitter measurement:

Delta F = 2,488,395,223 - 2,488,244,280 = 150,943 Hz
UI = Delta F/$(2\pi J_r)$
UI = 150,943/$(2\pi 30)$ = 800.78 UIp-p (error = $\pm$0.1%)

Note: the upper peak and lower peak amplitudes are typically slightly different.

Note: the individual lines are spaced at J_r (30 Hz) and are not resolved.

Jitter Tx settings:

Line rate = 2.4885 GHz (b/s)
Jitter amplitude = 800 UIp-p
Jitter rate (J_r) = 30 Hz

Figure 11.4 Wide-band frequency modulation (WBFM) method

Some typical results and example data

Some typical results are shown below. Four test points (0.1 UIp-p, 0.5 UIp-p, 18.8 UIp-p, 800 UIp-p) were repeated 12 times, corresponding to several hours of test time. The instrument tested was an Agilent OmniBER 718 communications performance analyzer. The spectrum analyzer used was an Agilent 8562EC (30 Hz to 13.2 GHz).

J1/J0 ratio method (log detector) at 0.1 UIp-p

Line Rate (b/s)	Jitter Rate (Hz)	Transmitter Jitter Setting (UIp-p)	Spectrum Analyzer Measurement (UIp-p)	Error
2.48 Gb/s	20,000,000	0.100	0.099	–1.3%
2.48 Gb/s	20,000,000	0.100	0.099	–1.3%
2.48 Gb/s	20,000,000	0.100	0.098	–2.2%
2.48 Gb/s	20,000,000	0.100	0.098	–2.2%
2.48 Gb/s	20,000,000	0.100	0.100	–0.3%
2.48 Gb/s	20,000,000	0.100	0.098	–2.2%
2.48 Gb/s	20,000,000	0.100	0.099	–1.3%
2.48 Gb/s	20,000,000	0.100	0.099	–1.3%
2.48 Gb/s	20,000,000	0.100	0.100	–0.3%
2.48 Gb/s	20,000,000	0.100	0.098	–2.2%
2.48 Gb/s	20,000,000	0.100	0.095	–4.9%
2.48 Gb/s	20,000,000	0.100	0.096	–4.0%

Spectrum Analyzer measurement Mean = 0.0983 UIp-p
Sigma = 0.00142 UIp-p
Three-sigma = 0.00427UIp-p (~4.3%)

Verification test accuracy = ±5% = ± 0.005 UIp-p
Verification test repeatability = ± 0.00427 UIp-p
Verification test total error = ± 0.00927 UIp-p

Instrument specification = 0.1 UIp-p ± (10% + 0.05 UIp-p) = 0.1 ± 0.06 UIp-p.

The verification test is about six times better than the instrument specification.

J1/J0 ratio method (linear detector) at 0.5 UIp-p

Line Rate (b/s)	Jitter Rate (Hz)	Transmitter Jitter Setting (UIp-p)	Spectrum Analyzer Measurement (UIp-p)	Error
2.48 Gb/s	10,000,000	0.500	0.501	+0.3%
2.48 Gb/s	10,000,000	0.500	0.497	–0.6%
2.48 Gb/s	10,000,000	0.500	0.500	+0.0%
2.48 Gb/s	10,000,000	0.500	0.498	–0.5%
2.48 Gb/s	10,000,000	0.500	0.498	–0.4%
2.48 Gb/s	10,000,000	0.500	0.499	–0.2%
2.48 Gb/s	10,000,000	0.500	0.496	–0.8%
2.48 Gb/s	10,000,000	0.500	0.498	–0.4%
2.48 Gb/s	10,000,000	0.500	0.497	–0.6%
2.48 Gb/s	10,000,000	0.500	0.501	+0.2%
2.48 Gb/s	10,000,000	0.500	0.498	–0.5%
2.48 Gb/s	10,000,000	0.500	0.499	–0.1%

Spectrum Analyzer measurement Mean = 0.4985 UIp-p
Sigma = 0.0015 UIp-p
Three-sigma = 0.0045 UIp-p (~ 0.9%)

Verification test accuracy = ± 3% = ± 0.015 UIp-p
Verification test repeatability = ± 0.0045 UIp-p
Verification test total error = ± 0.0195 UIp-p

Instrument specification = 0.5 UIp-p ± (7% + 0.05 UIp-p) =
0.1 ± 0.085 UIp-p.

The verification test is about four times better than the instrument specification.

Bessel “null” method at 18.800 UIp-p

Line Rate (b/s)	Jitter Rate (Hz)	Transmitter Jitter Setting (UIp-p)	Spectrum Analyzer Measurement (UIp-p)	Error
2.48 Gb/s	100,000	18.800	18.748	–0.3%
2.48 Gb/s	100,000	18.800	18.746	–0.3%
2.48 Gb/s	100,000	18.800	18.745	–0.3%
2.48 Gb/s	100,000	18.800	18.746	–0.3%
2.48 Gb/s	100,000	18.800	18.745	–0.3%
2.48 Gb/s	100,000	18.800	18.745	–0.3%
2.48 Gb/s	100,000	18.800	18.745	–0.3%
2.48 Gb/s	100,000	18.800	18.745	–0.3%
2.48 Gb/s	100,000	18.800	18.745	–0.3%
2.48 Gb/s	100,000	18.800	18.745	–0.3%
2.48 Gb/s	100,000	18.800	18.746	–0.3%
2.48 Gb/s	100,000	18.800	18.747	–0.3%

Spectrum Analyzer measurement Mean = 18.746 UIp-p
Sigma = 0.00089 UIp-p
Three-sigma = 0.00267 UIp-p (~ 0.01%)

Verification test accuracy = ± 0.01 UIp-p
Verification test repeatability = ± 0.00267 UIp-p
Verification test total error = ± 0.01267 UIp-p

Instrument specification = 18.8 UIp-p ± (5% + 0.05 UIp-p) = 18.8 ± 0.99 UIp-p.

The verification test is about 70 times better than the instrument specification.

(Bessel “nulls” are an extremely accurate method)

Wide-band FM method at 800 UIp-p

Line Rate (b/s)	Jitter Rate (Hz)	Transmitter Jitter Setting (UIp-p)	Spectrum Analyzer Measurement (UIp-p)	Error
2.48 Gb/s	30	800.000	800.455	+0.1%
2.48 Gb/s	30	800.000	800.993	+0.1%
2.48 Gb/s	30	800.000	800.455	+0.1%
2.48 Gb/s	30	800.000	800.455	+0.1%
2.48 Gb/s	30	800.000	800.455	+0.1%
2.48 Gb/s	30	800.000	800.993	+0.1%
2.48 Gb/s	30	800.000	800.455	+0.1%
2.48 Gb/s	30	800.000	800.993	+0.1%
2.48 Gb/s	30	800.000	800.993	+0.1%
2.48 Gb/s	30	800.000	800.993	+0.1%
2.48 Gb/s	30	800.000	799.918	–0.0%
2.48 Gb/s	30	800.000	799.918	–0.0%

Spectrum Analyzer measurement Mean = 800.59 UIp-p
Sigma = 0.39 UIp-p
Three-sigma = 1.16 UIp-p (~ 0.145%)

Verification test accuracy = ±1% = ± 8.0 UIp-p
Verification test repeatability = ± 1.16 UIp-p
Verification test total error = ± 9.16 UIp-p

Instrument specification = 800 UIp-p ± (5% + 2 UIp-p) =
800 ± 42 UIp-p.

The verification test is about 4.5 times better than the instrument specification.

Conclusions

The above results show the verification tests using a spectrum analyzer are significantly more accurate than the required instrument specification limits, and allow reliable precise verification of the instrument performance. This ensures the OmniBER family delivers superior accuracy when used to perform jitter tolerance tests.

References

Frequency modulation theory:

1. “Modulation” by F. R. Connor, pp 26-28, pp 114-116, Edward Arnold ISBN 0-7131-3457-7

Bessel function theory and equations to calculate Bessel “null” positions and J1/J0 ratios:

2. “Advanced Engineering Mathematics” by Erwin Kreyszic, pp 205-213, John Wiley & Sons ISBN 0-471-85824-2

12. An Overview of Wander Measurements

Abstract: This paper describes the sources of wander in synchronization networks and defines the various measurements of wander with reference to the ITU-T standards. TIE, MTIE, TDEV and other wander computations are all discussed in detail, along with the performance limits for various network reference clocks. Examples are given throughout the paper using the Agilent Wander Analysis Package which greatly simplifies the complex calculation and long-term measurements required for wander analysis.

Introduction

Wander is the term given to low-frequency phase variations in a clock or digital signal, ranging from 10 Hz down to microhertz (μHz). Wander may be caused by very slight (parts per million or less) differences in reference clocks between two networks, or by slow changes in the relative phase of two clock signals due to climatic temperature changes. An example is the 2 Mb/s reference clock distribution to SDH equipment which might be subject to *diurnal wander* (period of a day) or *annual wander* due to environmental changes. Yet another source of wander is very low-frequency phase noise in a clock oscillator. Whereas jitter (phase variations greater than 10 Hz) is measured with reference to a recovered internal clock, wander must be measured against an external reference clock.

This paper overviews the techniques of wander measurement, covering these three objectives:

- Set out basic definitions for wander and its associated computations.
- Explore some of the causes and effects of wander.
- Review automated computation of wander metrics.

Basic definitions and discussion of key wander metrics

Wander

Wander is defined in ITU-T G.810 (Definitions and terminology for synchronization networks) as “The long-term variations of the significant instants of a digital signal from their ideal position in time (where “long-term” implies that these variations are of frequency less than 10Hz).” It is possible to express wander in terms of the unit interval (UI) or in terms of time. Given that wander frequency may extend into the μHz region, the time representation can be more convenient.

Time Interval Error (TIE)

This is defined in ITU-T G.810 as “The difference between the measure of a time interval as provided by a clock (as recovered from a data signal) and the measure of that same time as provided by a reference clock.”

This basic data (i.e. a graph of TIE versus time) may be used to represent the wander behavior of a given system, but inspection and analysis of that data is often necessary to capture specific characteristics.

A graph of TIE data versus time is shown in Figure 12.1, where a constant frequency offset is present in the device under test. It is easy to relate the slope of the graph to the divergence of time as perceived by the device when compared to that referred to a stable reference clock.

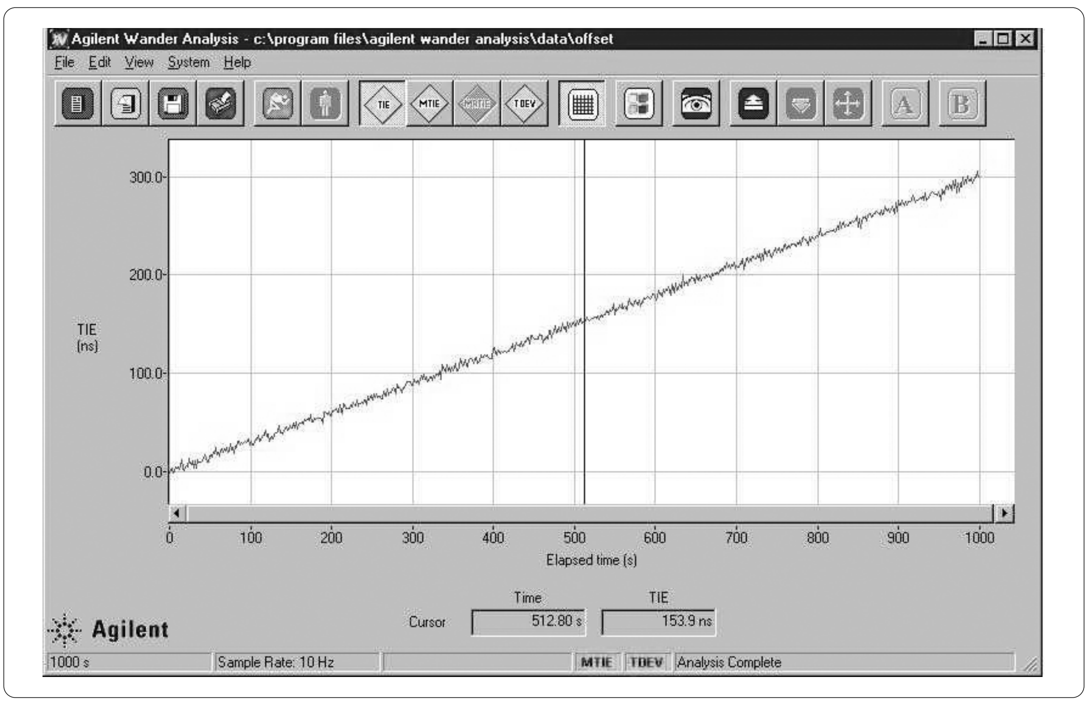


Figure 12.1 TIE versus time (constant frequency offset) as displayed on the Agilent E4547A Wander Analysis Package

Maximum Time Interval Error (MTIE)

MTIE is defined in ITU-T G.810 as “The maximum peak-to-peak delay variation of a given timing signal with respect to an ideal timing signal within an observation time ($\tau = n\tau_0$) for all observation times of that length within the measurement period (T).”

The observation time is defined as $\tau = n\tau_0$, where:
τ_0 = the time-error sampling interval
τ = the integration time
n = the number of sampling intervals within the integration time τ

MTIE is of particular use in identifying transients in phase and offsets in frequency.

Whereas it is quite easy to see what is going on in a TIE plot, the information represented by MTIE requires a little deeper thought. Here, values of MTIE are recorded for *a range of observation intervals.*

For example, a TIE data set could be analyzed by running an observation window of 0.1 seconds along the data. The maximum value of TIE is recorded, and this becomes the first point on the MTIE graph. The process is then repeated with a sliding observation window of 0.2 seconds, again recording the maximum value of TIE (which becomes the second point on the graph). This process is repeated with increasing length of observation window until the graph is built up.

- A graph of MTIE commencing at zero and increasing linearly with observation interval is indicative of a frequency offset in the signal under test, with respect to a reference source.

- An MTIE graph commencing with a non-zero value indicates that a transient change in TIE data has occurred during the measurement period.

- A graph of MTIE which begins with a ramp from zero and then which flattens off as observation interval increases is indicative of wander with sinusoidal content. This characteristic is intuitive - once the observation interval has increased beyond the period of the sinusoid, the MTIE value cannot increase beyond the peak-to-peak value of that wander.

A graph of MTIE data, where there has been a transient, is shown in Figure 12.2 (the plot commences with a non-zero value of MTIE). Although the graph indicates that a transient has occurred, inspection of the TIE data is required to identify when the event took place. (Note that in this example, the MTIE plot also reflects a possible frequency offset too).

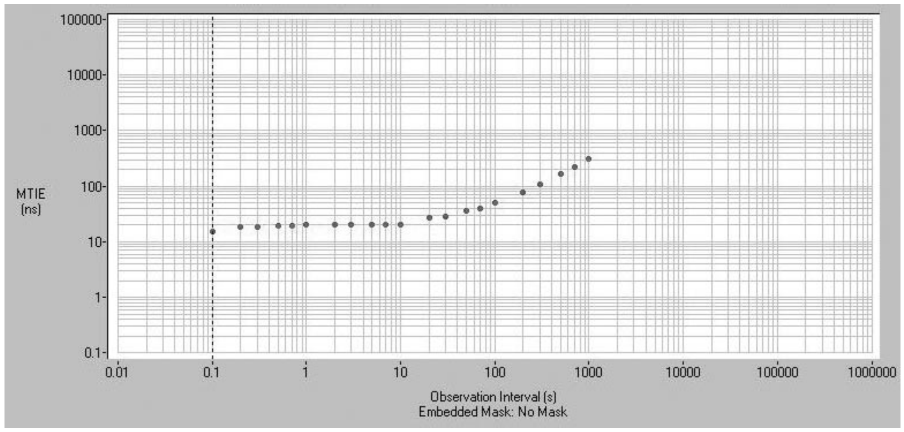


Figure 12.2 MTIE versus observation interval (phase-transient)

Because the MTIE computation is sensitive to phase-transients, it is less useful in capturing low-frequency noise characteristics. Hence another parameter is required as described below.

Time Deviation (TDEV)

Time Deviation is described in ITU-T G.810 as "A measure of the expected time variation of a signal as a function of integration time." TDEV is particularly useful in revealing the presence of a number of noise processes commonly found in clocks and oscillators.

It is usual to express instability of a system in terms of some statistical representation of its performance. In many cases, the computation of variance or standard deviation would give an adequate view. However in the area of clocks and oscillators, some of the typical noise sources do not lend themselves to this analysis, and it has been shown that the usual variance calculation does not converge.

The idea of convergence may be seen as follows. A population with Gaussian distribution will have a certain mean and variance, and as one computes these parameters using larger and larger samples of that population, the confidence in those parameters increases. Hence we observe that the variance converges towards the true value for that population. In the case of particular oscillator noise processes (flicker noise frequency modulation and random walk frequency modulation), no matter how large the sample, there will be no evidence of the variance computation iterating towards a settled value. In this situation, the variance function does not converge. Typical long-term frequency behaviour of clocks is characterized by a "walk-off" phenomenon, so it is understandable that the standard deviation of such a characteristic appears unbounded. In other words, it is not a stationary process.

To overcome this deficiency, three alternative measures have been developed - Time Variation (TVAR) being the most recent. This parameter does converge for oscillator noise processes. Time deviation (TDEV) is the square root of TVAR. TDEV was developed to fulfil several specific needs - one being to provide a method of assessing synchronization stability in telecom networks. It has distinct advantages over the other two statistical measures (Allan Deviation and Modified Allan Deviation) discussed later.

The computation of TDEV is given in the discrete time form below. The key to the understanding of the basic mechanism may be had from close inspection of the terms in the bold square brackets.

The classic variance involves computing the deviations from a mean value. In the case of TDEV, the double difference of three adjacent time error averages (each of duration τ) are computed, and it is these which are squared and summed (as the double difference computation "slides" along the TIE data).

As the observation interval increases, the number of TIE samples forming the time error averages increases accordingly. Hence then, from ITU-T G.810:

$$\mathrm{TDEV}\,(n\tau_0) \cong \sqrt{\frac{1}{6n^2\,(N-3n+1)} \sum_{j=1}^{N-3n+1} \left[\sum_{i=j}^{n+j-1} (x_{i+2n} - 2x_{i+n} + x_i) \right]^2}$$

$$n = 1, 2, \ldots, \text{integer part} \left(\frac{N}{3}\right)$$

where:

x_i are time error sample values
N = the total number of samples
τ_0 = the time error sampling interval
τ = the integration time, the independent variable of TDEV
n = the number of sampling intervals within the integration time τ.

Note that the double difference is given by:

((value 2 – value 1) – (value 1 – value 0))
= (value 2 – (2 x value 1) + value 0).

One may readily observe the form on the right hand side as being present in generalized expression in the square brackets of the TDEV equation.

It is also quite easy to see from the above, that for a given maximum observation interval, the data capture must run for at least a period of three times the observation interval in order to generate results in which there is any degree of confidence. This point may be observed by noting that the calculation is run from n = 1 to n= N/3. In some cases, three times is not considered enough, as for example in ITU-T G.811 which calls for TDEV tests to run for *twelve* times the maximum observation interval.

It was stated earlier, however, that TDEV has a use in characterizing the low frequency noise content of a signal. ITU-T G.810 ("Definitions and Terminology for Synchronization Networks") and ITU-T G.812 ("Timing Requirements of Slave Clocks suitable for use as Node Clocks in Synchronization Networks") go on to show that TDEV is approximately related to the power spectral density of time interval error by:

$$S_x(f) \approx \frac{0.75}{f}\left(TDEV\left(\frac{0.3}{f}\right)\right)^2$$

where:

$S_x(f)$ = the spectral density of the time interval error as a function of frequency offset from the carrier.
f = the offset frequency from the carrier.

The derivation of this equation is given in ITU-T G.812 Appendix 1. Indeed, to give further insight, ITU-T G.812 shows that TDEV may be regarded as the RMS value of phase (TIE), having been passed though a band-pass filter centred at $0.42/\tau$ Hz (see References at the end of this paper, in particular Annex D of ANSI T1.101, which adds clarity to the filter ideas).

The slope of the TDEV plot can be useful in revealing the characteristics of various noise types, which are summarised in Table 12.1.

Noise Process	**Slope of of TDEV (τ)**	**Equivalent Power Law Model**
White phase modulation	$\tau^{-1/2}$	f^0
Flicker phase modulation	τ^0	1/f
White frequency modulation	$\tau^{1/2}$	$1/f^2$
Flicker frequency modulation	τ	$1/f^3$
Random walk frequency modulation	$\tau^{3/2}$	$1/f^4$

Table 12.1 Relationship between noise process and TDEV slope

The applicability of these five noise types to various areas in a network is shown in Table 12.2.

Noise Process	**Applicability**			
	Caesium-based clocks	**Rubidium-based clocks**	**Quartz-based clocks**	**Telecom network**
White PM			●	●
Flicker PM			●	●
White FM	●	●	●	●
Flicker FM	●	●	●	
Random Walk FM	●	●	●	

Table 12.2 Relevance of noise processes to different areas of the network

An example of a TDEV plot is shown in Figure 12.3.

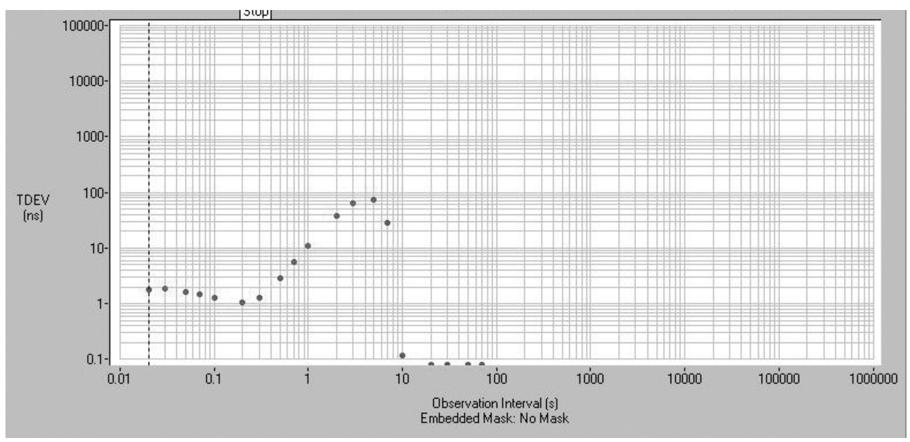


Figure 12.3 TDEV versus observation interval

This shows the simple case of the response of the TDEV function to sinusoidal wander input. The TDEV plot peaks when $\tau = 4.2$, indicating that the underlying wander has an approximate frequency of 0.1 Hz. This observation correlates with the simple band-pass filter idea given above.

Maximum Relative Time Interval Error (MRTIE)

This is defined in ITU-T G.810 as "The maximum peak-to-peak delay variation of an output timing signal with respect to a given input timing signal within an observation time ($\tau = n\tau_0$), for all observation times of that length within the measurement period (T)."

Essentially this parameter removes the effect of a frequency offset, leaving only phase transient and random information. MRTIE is indicated in ITU-T G.823 as being useful in specifying wander on a payload output relative to the clock phase of an input buffer, thus facilitating the design of the buffer size.

Additional parameters

ITU-T G.810 describes two additional statistical parameters related to TDEV, which may be used to identify noise types.

Allan Deviation (ADEV)

This is the oldest of the three statistical measures discussed in this paper. It was developed to provide a means of quantifying mid- to long-term stability of oscillators and clocks. In the context of telecommunications networks, this parameter would appear to be less useful than TDEV in that it cannot be used to distinguish between *white phase modulation and flicker phase modulation*. In both cases, the slope of the ADEV function varies according to τ^{-1}.

Modified Allan Deviation (MDEV)

MDEV was developed to remove the limitations of the Allan Deviation. This was done by introducing the computation of double differences of three adjacent time *averages*, as opposed to the computation of double differences of three adjacent time error *values*.

This parameter has significant dependence on the sample period in cases where White Phase Noise is dominant. G.810 reveals a direct relationship between MDEV and TDEV. Hence the dependence on sampling period, as indicated above, holds true for TDEV as well.

The relationship between TDEV and MDEV is given below:

$$\mathrm{TDEV}\,(n\tau_0) \cong \sqrt{\frac{1}{6n^2}\left\langle\left[\sum_{i=1}^{n}(x_{i+2n} - 2x_{i+n} + x_i)\right]^2\right\rangle} = \frac{n\tau_0}{\sqrt{3}}\,\mathrm{MDEV}\,(n\tau_0)$$

x_i are time error sample values
τ_0 = the time error sampling interval
n = the number of sampling intervals within the integration time τ.

(ITU-T G.810 Appendix II describes the derivation of this equation)

MDEV differs from TDEV in terms of the slopes which correspond to the five characteristic noise types. MDEV is less capable of discriminating between white phase modulation, flicker phase modulation and white frequency modulation noise types, which, as shown in Table 12.2, are particularly relevant in telecommunications networks.

Causes of wander

Wander may arise at many points in a network, and numerous standards have been developed to specify wander generation, wander transfer and wander tolerance limits. Synchronization of networks requires careful attention to ensure that the reference clocks operate within specified limits and that the build up of wander through the network is not excessive. Table 12.3 summarises the ITU-T and Telcordia standards governing network wander performance.

Equipment	**Input Wander Tolerance**	**Output Wander**	**Wander Transfer**	**Phase Transients**
SDH (DXC or ADM)	G.813, G.825	G.813	G.813	G.813
PDH	G.823			
PRC clock		G.811		
SSU clock	G.812	G.812	G.812	G.812
PDH transport	G.823			
SONET transport	Telcordia GR-253	Telcordia GR-253	Telcordia GR-253	
SDH transport	G.825			

Table 12.3 Summary of wander standards

A number of mechanisms for wander generation are now considered. Each of these areas has been addressed within the relevant standards above, and appropriate bounds have been set out for the performance of devices.

Oscillator noise processes

As mentioned above, the TDEV estimator can prove useful in identifying particular noise types. Five categories of noise are given below. Each noise process gives rise to a different slope on the TDEV plot, as summarised in Table 12.1.

- White phase modulation, commonly associated with amplification stages.
- Flicker phase modulation, commonly associated with amplification stages.
- White frequency modulation, commonly found in passive resonators, for example Caesium Standard.
- Flicker frequency modulation, which may be related to physical resonance.
- Random walk frequency modulation, which may be associated with shock, vibration or temperature.

Phase transients

Phase transients may occur when, for example, an SDH Equipment Slave Clock (SEC) loses its input reference and a back-up only becomes available after some period of time. During the switchover period, an accumulation of phase error might occur.

Frequency drift or offset

These may be caused when a slave clock loses all external reference sources and enters a holdover mode. During holdover mode, the clock continues to operate using calibration data built up during normal locked operation. Over a period of time, the slave clock may eventually develop frequency drift. Equally well, other factors such as ageing, temperature and power supply variations may contribute.

Effects of wander

The effects of wander may best be illustrated by considering pointer movements and slips, as shown in Figure 12.4.

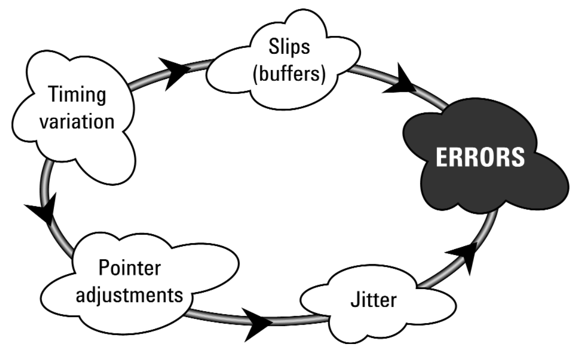


Figure 12.4 Chain of events leading from timing variation to errors

De-mapping jitter

Consider the case of synchronous networks which carry PDH payloads. Suppose that an SDH network element is required to cope with incoming data with high wander. This situation is handled with payload pointer adjustments which in turn may give rise to high jitter content once the payload is de-mapped.

Slips

In a similar vein, a fixed frequency offset on incoming data may well cause a network element's buffer to overflow or underflow. This may introduce frame slip which can in turn cause severe disruption to the payload.

Data arriving at a network element is typically clocked into a buffer using a clock recovered from the incoming data stream. In contrast, the information stored in that buffer is read out under control of a clock generated by the receiving equipment itself.

Should the incoming data arrive at a rate faster than the receiver is clocking it out, the slip buffer may overflow and data is lost. In the opposite situation, when the receiver clocks data out at faster rate than the incoming stream, the slip buffer may underflow leading to a block of data being repeated.

Slips in a network ultimately impact the end-user, whether through loss of colour or frame-freeze in video, or audible clicks in voice traffic. Poor synchronization means poor quality of service.

The buffer principle is illustrated in Figure 12.5.

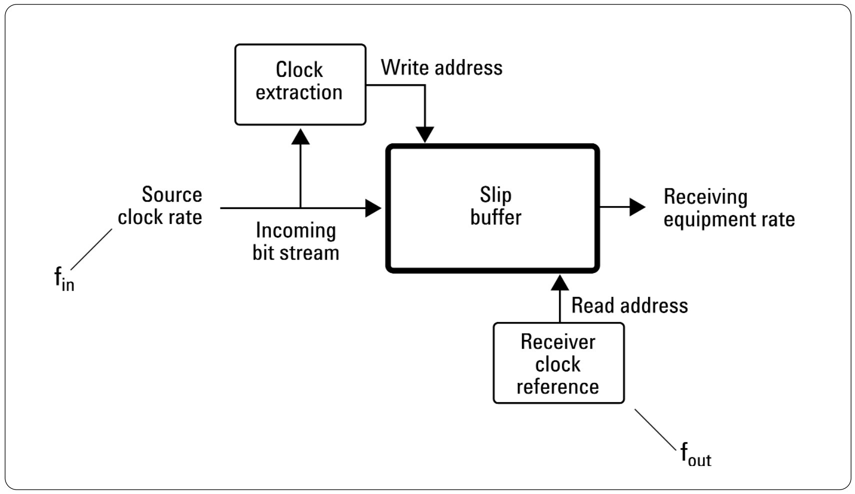


Figure 12.5 Slip buffer principle

Accuracy requirements of network clock sources

Given that good synchronization is key to successful performance of a network, the sources of time reference need to be well controlled. The following section gives an introduction to some of the key areas of network reference specification, and in so doing gives a feel for the typical values of MTIE and TDEV required.

ITU-T G.811, G.812 and G.813 define the requirements for stability and performance of the various sources of synchronization in a network. There follows a brief overview of some of the key elements of these specifications, but the reader is encouraged to source further details from the standards.

Primary reference clock/source (PRC/PRS)

The PRC is the most stable time reference source in a network. The PRC is typically based on caesium-beam technology and whilst fully operational will contribute of the order of 3 slips/year or better between networks.

An example of a PRC is the Agilent 5071A, whose frequency stability is at least an order of magnitude better than the ITU-T requirement. A survey of performance showed that the standard deviation for the 5071A frequency stability is seven times better than the manufacturer's accuracy specification of 1×10^{-12}, and the mean value of frequency offset only differs by 5×10^{-14}.

For observation times of greater than one week, the maximum allowable offset in frequency is 1 part in 10^{11}, as given in ITU-T G.811.

MTIE limits for the PRC as defined by ITU-T G.811 is shown in Table 12.4 and displayed graphically in Figure 12.6.

MTIE	Integration time (τ seconds)
$0.275 \times 10^{-3}\tau + 0.025\ \mu s$	$0.1\ s < \tau \leq 1000\ s$
$10^{-5}\tau + 0.29\ \mu s$	$\tau > 1000\ s$

Table 12.4 MTIE limits for a Primary Reference Clock (PRC)

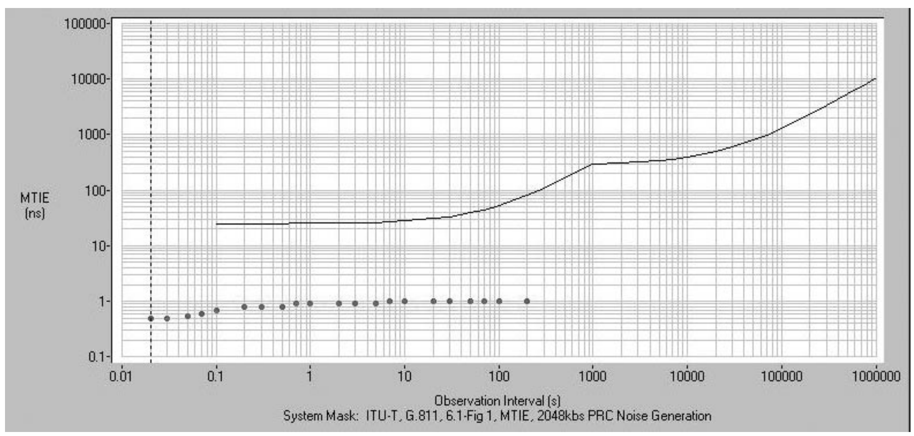


Figure 12.6 MTIE limits for a PRC, showing the ITU-T mask and example measurement points from an atomic clock

TDEV limits for a PRC are shown in Table 12.5 and graphically in Figure 12.7.

TDEV	Integration time (τ seconds)
3 ns	$0.1\ s < \tau \leq 1000\ s$
$0.03\ \tau$ ns	$100\ s < \tau \leq 1000\ s$
30 ns	$1000\ s < \tau < 10000\ s$

Table 12.5 TDEV limits for a PRC

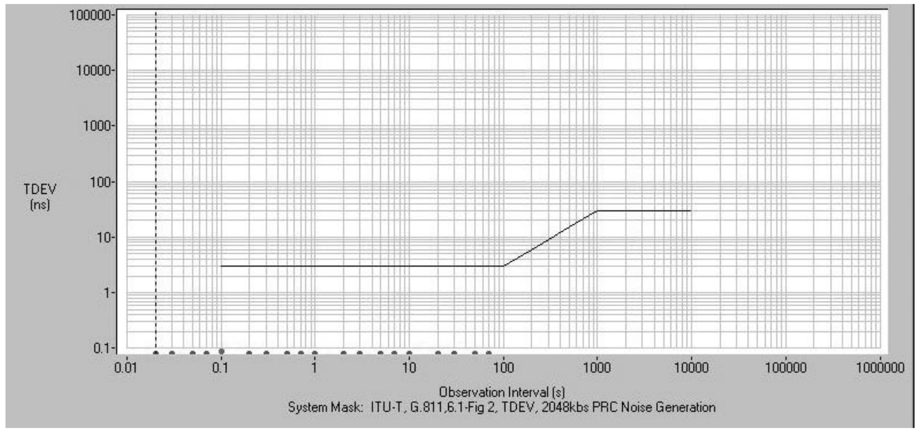


Figure 12.7 TDEV limits for a PRC, showing the ITU-T mask and example measurement points from an atomic clock

Synchronization Supply Unit/Building Integrated Timing Source (SSU/BITS) (Type II Example)

A Synchronization Supply Unit (SSU) may well be based upon Rubidium/quartz technology and has an accuracy requirement of 1 part in 1.6 x 10^{-8} over one year as defined in ITU-T G812.

MTIE limits (locked mode) are defined in Table 12.6 and illustrated graphically in Figure 12.8.

MTIE	Integration time (τ seconds)
40 ns	0.1 s < $\tau \leq$ 1 s
$40 \times \tau^{0.4}$ ns	1 s < $\tau \leq$ 10 s
100 ns	τ > 10 s

Table 12.6 MTIE limits for a SSU/BITS type II

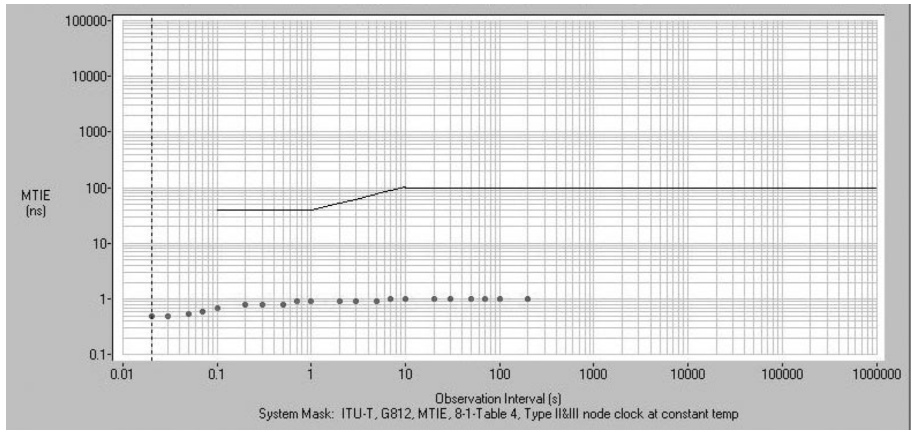


Figure 12.8 MTIE limits for a SSU/BITS type II, showing the ITU-T mask and example measurement points from a typical source

TDEV limits (locked mode) are defined in Table 12.7 and illustrated graphically in Figure 12.9.

TDEV	Integration time (τ seconds)
$3.2 \times \tau^{-0.5}$ ns	0.1 s < $\tau \leq$ 2.5 s
2 ns	2.5 s < $\tau \leq$ 40 s
$0.32 \times \tau^{0.5}$ ns	40 s < $\tau \leq$ 1000 s
10 ns	τ > 1000 s

Table 12.7 TDEV limits for a SSU/BITS

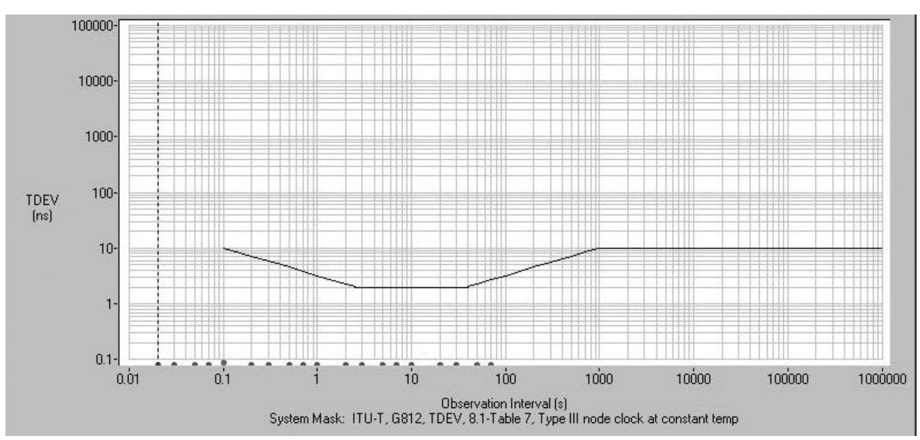


Figure 12.9 TDEV limits for a SSU/BITS, showing the ITU-T mask and example measurement points from a typical source

SDH Equipment Slave Clocks/SONET Minimum Clock (SEC/SMC) (option 1 constant temperature example)

An SDH Equipment Slave Clock (SEC) could be synthesized using an oven-controlled crystal oscillator. For observation times of greater than one month, the maximum allowable offset in frequency is 4.6 parts per million.

MTIE limits (locked mode) are defined in Table 12.8 and illustrated graphically in Figure 12.10.

MTIE	Integration time (τ seconds)
40 ns	$0.1\ s < \tau \leq 1\ s$
$40 \times \tau^{0.1}$ ns	$1\ s < \tau \leq 100\ s$
$20.25 \times \tau^{0.2}$ ns	$100\ s < \tau < 10000\ s$

Table 12.8 MTIE limits for an option 1 SEC

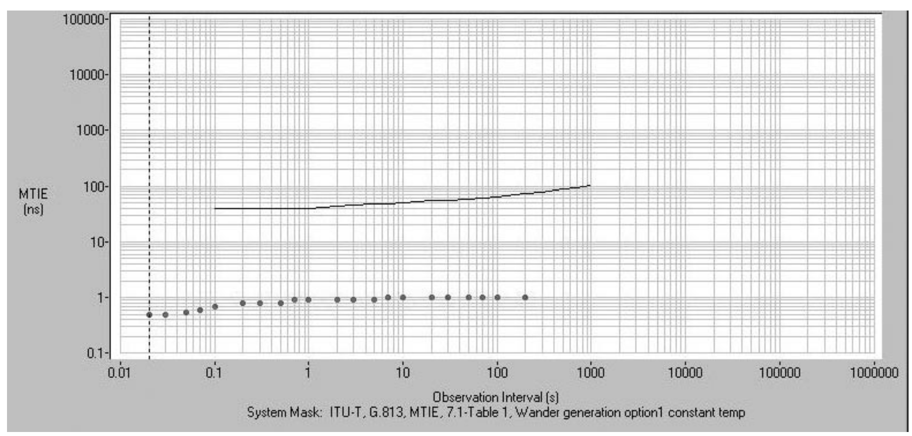


Figure 12.10 MTIE limits for option I SEC

TDEV limits (locked mode) are given in Table 12.9 and displayed graphically in Figure 12.11.

TDEV	Integration time (τ seconds)
3.2 ns	0.1 s < $\tau \leq$ 2.5 s
$0.64 \times \tau^{0.5}$ ns	25 s < $\tau \leq$ 100 s
6.4 ns	100 s < $\tau \leq$ 1000 s

Table 12.9 TDEV limits for SEC in locked mode defined in ITU-T G.813

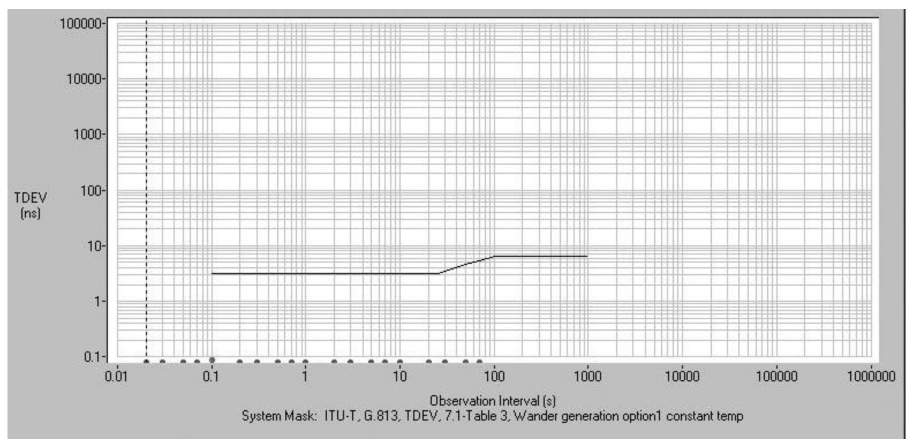


Figure 12.11 TDEV limits for option I SEC in locked mode

In addition, the standards for the SSU and SEC cover other aspects, summarised in Table 12.10:

Specification	**Attribute**
Wander in non-locked mode	Basic clock performance in the absence of reference input
Wander noise tolerance	The degree of immunity to phase noise on the clock input
Noise transfer	Noise observed at the output as a result of input phase noise
Short term transient response	Performance after switch over of reference
Long term transient response	Performance following loss of reference (holdover)
Input signal interruptions	Performance during interruptions where reference switching has not occurred
Hold-in range	Slave clock's ability to remain locked to a varying reference
Pull-in range	Slave clock's ability to lock to a reference

Table 12.10 Additional parameters for SSU and SEC references

Computation of Wander Metrics

Earlier sections have shown that computation of the key wander metrics is not trivial. Testing can last for many days, and the amount of data gathered can be enormous. Hence a practical approach to this situation is to use a PC-based analysis tool in conjunction with a tester such as the Agilent OmniBER 718/OTN communications performance analyzer.

Agilent E4547A Wander Analysis Package

Agilent Technologies E4547A Wander Analysis Software facilitates the real-time calculation of wander parameters. TIE data from a communications analyzer is transmitted to a PC over an RS232 link. This data is processed and displayed in graphical form. Four graph types are available:

- TIE vs time
- MTIE vs observation interval
- MRTIE vs observation interval
- TDEV vs observation interval

The software includes a number of pre-defined masks which facilitate network testing to various standards. The user may also generate mask files to suit any unique testing requirements. The format for user mask files is simple in structure and new files may be created in a text editor. Masks provide immediate visual feedback as to whether a device under test has met the appropriate performance criteria defined in the standards.

Full control is provided for test time and TIE data sampling rate. In some cases it may be of advantage to restrict the size of data files, however trade-off in the accuracy of the computation needs to be considered.

TIE data may be stored for post-processing purposes, and alternatively, TIE data from a source other than Agilent OmniBER 718/OTN may be imported (in comma-separated variable (CSV) format) into the software package for analysis.

E4547A includes a useful facility wherein the number of observation intervals used to compute MTIE, MRTIE or TDEV is restricted. In this way the wander computations are constrained to using relatively recent data. This feature may be useful in revealing underlying trends in MTIE data, where for example, a transient effect has obscured the plot.

Throughout this note, the figures used to illustrate the characteristics of wander have been taken from screen shots of the E4547A package. For the sake of completeness, the Figure 12.12 gives a brief description of the various controls and menus available through the graphics user interface (GUI).

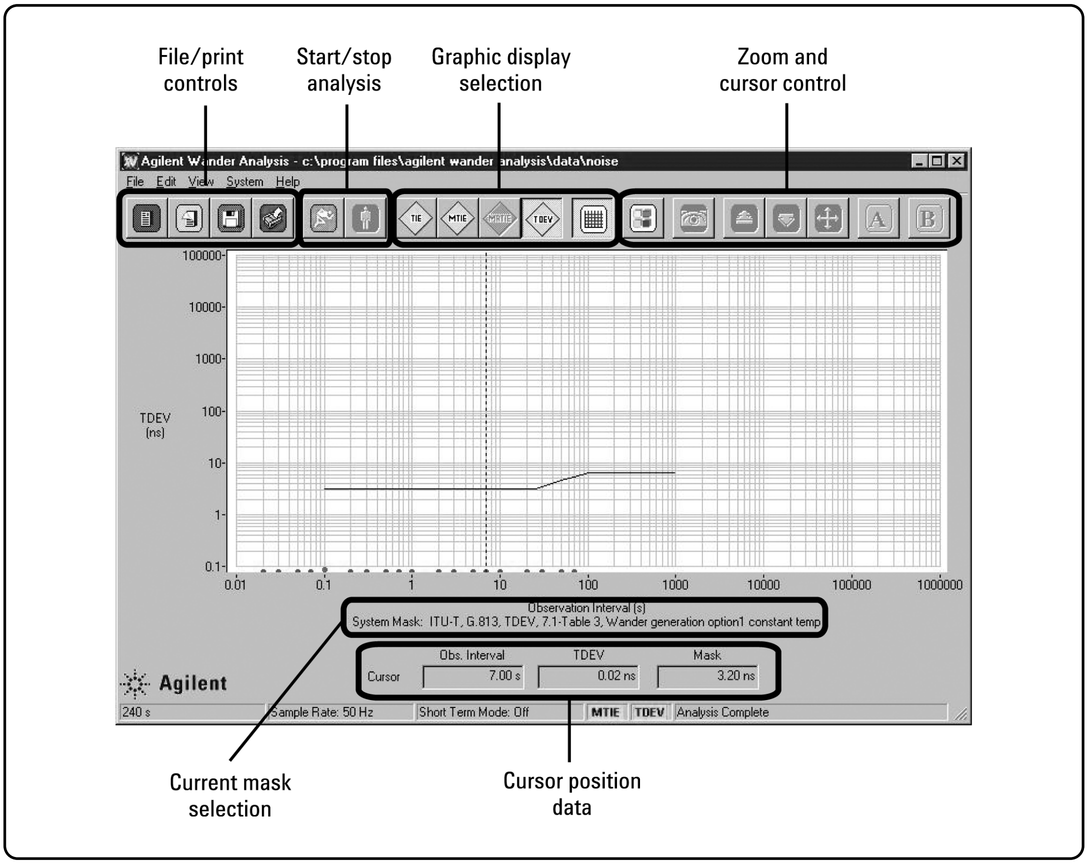


Figure 12.12 GUI description of Agilent E4547A Wander Analysis Package

Conclusions

Wander analysis requires the collection of a large amount of data over a relatively long period, followed by complex calculations to derive the various parameters such as TIE, MTIE and TDEV defined in the ITU-T standards. These parameters can help to analyze the performance of reference clocks and synchronization networks and pinpoint the possible sources of wander. By simplifying the collection and analysis of this data using a software-controlled communications performance analyzer, much useful information on network performance and possible long-term problems can be derived.

References

Standards

1. ITU-T Recommendation G.810 (08/96) “Definitions and terminology for synchronization networks”.

2. ITU-T Recommendation G.811 (09/97) “Timing characteristics of primary reference clocks”.

3. ITU-T Recommendation G.812 (06/98) “Timing requirements of slave clocks suitable as node clocks in synchronization networks”.

4. ITU-T Recommendation G.813 (08/96) “Timing characteristics of SDH equipment slave clocks (SEC)”.

5. ITU-T Recommendation G.823 (03/00) “The control of jitter and wander within digital networks which are based on the 2048 kb/s hierarchy”.

6. ANSI T1.101-1999 “Synchronization Interface Standard”.

General reading

1. “The Science of Timekeeping”, Agilent Application Note AN1289 (This is a very readable text with a good overview of time and frequency stability issues.)

2. “Properties of Signal Sources and Measurement Methods” by Howe, Allan and Barnes, Proceedings of the 35th Annual Symposium on Frequency Control, 1981.

3. “A Modified Allan Variance with Increased Oscillator Characterization Ability” by Allan and Barnes, Proceedings of the 35th Annual Symposium on Frequency Control, 1981.

4. “A Frequency-domain view of Time-domain Characterization of Clocks and Time and Frequency Distribution Systems” by Allan, Weiss and Jesperson, Proceedings of the 45th Annual Symposium on Frequency Control, 1991. (This is a very accessible treatment of the relative merits of the three stability measures.)

5. "Introduction to the Time Domain Characterization of Frequency Standards" by J. Jesperson, Proceedings of the 25th Annual Precise Time and Time Interval (PTTI) Meeting, Pasadena, CA, December 1991. (A largely non-mathematical overview, with background relevant to the band-pass filter concepts associated with TDEV.)

Acronym Guide

3R Reamplification, reshaping and retiming

ADEV Allan Deviation is a statistical measure of long-term clock stability, a forerunner of TDEV.

ADM Add Drop Multiplexer

AIS Alarm Indication Signal is the special signal transmitted downstream by a piece of network equipment when it loses the received signal or loses frame alignment.

ANSI American National Standards Institute is the North American forum for generating standards involving both equipment manufacturers and operators. The T1X1 committees produce the relevant proposals for transmission systems.

BER Bit Error Ratio is the ratio of errored bits in an observation period to the total number of bits received.

CSV Comma-separated Variable, a format for streaming data, for example from a measurement instrument to a computer.

dB Decibel, the logarithmic ratio of two voltages 20log10 (V2/V1)

DJ Deterministic Jitter is generated by systematic effects such as pattern dependency.

DSn Digital Signal where "n" describes the hierarchical level, typically DS0 at 64 kb/s, DS1(T1) at 1.5 Mb/s and DS3 (T3) at 45 Mb/s

DUT Device Under Test.

DWDM Dense Wave Division Multiplexing, providing multi-channel optical signal with wavelengths typically spaced at 50 or 100 GHz intervals.

ETSI European Telecommunications Standards Institute is the European forum for generating standards involving members from both equipment manufacturers and operators.

E/O Electrical to Optical convertor, usually a laser and its associated drive circuitry.

FEC Forward Error Correction, used to improve the error floor of an optical transport network, and enable longer optical spans without regeneration.

FM Frequency Modulation.

GbE Gigabit Ethernet, 1000E

GFP Generic Framing Procedure, which provides a simple encapsulation method for frame-based data traffic (Ethernet, IP/PPP, RPR, Fiber Channel, ESCON etc.) over the TDM transport path that could be SONET/SDH or the OTN.

GUI Graphic User Interface

High-Q High Q-factor denoting narrow relative bandwidth for a tuned-circuit used in clock recovery.

HPF High Pass Filter

IEEE Institute of Electrical and Electronic Engineers in North America through its industry associations produces standards, most notably for Ethernet in the 802.3 series.

ISI Inter-Symbol Interference refers to the characteristic of any bandwidth-limited system in which the memory of previous bits or symbols affects the value of the current symbol. ISI makes systems pattern-dependant.

ITU-T International Telecommunications Union, Telecommunications standards authority is the main body producing internationally agreed telecommunications standards, often with proposals from ANSI, ETSI, Telcordia etc.

J0, J1 etc. The Bessel function coefficients defining the magnitude of the carrier (J0) and sidebands (J(N)) of a frequency- or phase-modulated signal.

Low-Q Low Q-factor denoting wide relative bandwidth for a tuned-circuit used in clock recovery.

MDEV Modified Allan Deviation, a development of Allan Deviation (ADEV) for statistical analysis of clock stability, similar to TDEV.

MRTIE Maximum Relative Time Interval Error is the maximum peak-to-peak delay variation of an output timing signal with respect to a given input timing signal.

MTIE Maximum Time Interval Error is the peak-to-peak variation of TIE within a defined observation interval.

MTJ, MTIJ Maximum Tolerable Jitter, Maximum Tolerable Input Jitter is a measure of the maximum jitter at a particular jitter frequency that a piece of network equipment can tolerate at its input before generating bit errors.

NE Network Equipment.

NEM Network Equipment Manufacturer.

OAM Operations, Administration and Maintenance

OC-n Optical Carrier where "n" designates the hierarchical level similar to the electrical STS-n in the SONET standard. OC-1 (52 Mb/s), OC-3 (155 Mb/s), OC-12 (622 Mb/s), OC-48 (2488 Mb/s), OC-192 (9952 Mb/s).

OCh Optical Channel is the conversion of the framed Optical Transport Unit (OTU) electrical signal, typically forming one of the wavelengths in a DWDM signal.

ODCa/b/r/p ODUk Clock which generates the timing for signals produced by the OTN equipment types. Four variants are specified depending on client signals. Different jitter requirements are specified depending on the bit rate.

ODUk Optical channel Data Unit in the OTN (ITU-T G.709) which encapsulates the Optical channel Payload Unit (OPU) and adds the ODU overhead for such things as tandem connection monitoring and end-to-end path supervision. The integer "k" (1,2,3 ...) denotes the rate of 2.5, 10 or 40 Gb/s.

O/E, E/O Optical to Electrical and Electrical to Optical refer to the function of the optical receiver and transmitter in a network equipment or test instrument.

OPUk Optical channel Payload Unit in the OTN (ITU-T G.709) which encapsulates the client signal and adds an overhead describing the payload type and rate adaption. The integer "k" (1,2,3 ...) denotes the rate of 2.5, 10 or 40 Gb/s.

OmniBER 718 Agilent's family of Communications Performance Analyzers that offer a range of jitter measurement capability.

OTN Optical Transport Network is the ITU-T descriptor for the new generation of optical transmission and switching networks, and specifically refers to the new transport frame structure (digital wrapper) for optical networks described in ITU-T G.709

OTUk Optical Transport Unit in the OTN (ITU-T G.709) which encapsulates the Optical channel Data Unit (ODU) and adds an overhead transport supervisory functions and also the Forward Error Correction (FEC) overhead. The integer "k" (1,2,3 ...) denotes the gross bit rate of 2.66, 10.7 or 43 Gb/s

p-p peak-to-peak

PDF Probability Distribution Function applied to random processes such as noise or jitter generation.

PDH Plesiochronous Digital Hierarchy refers to legacy transport networks using asynchronous multiplexing (bit stuffing) such as DS1, DS3, E1, E3 etc.

PLL Phase Lock Loop used for clock recovery in regenerators and other line equipment. Also an essential part of a jitter measurement instrument.

PRBS Pseudo Random Binary Sequence used to simulate live traffic under test conditions. Most commonly used sequence is ITU-T 223 - 1. SONET/SDH scrambler uses a 27 - 1 PRBS.

PRC/PRS Primary Reference Clock/Source, usually an atomic clock which synchronizes all equipment directly or through slave clocks across the network

Q-factor Quality factor for a bandpass filter is the ratio of the center frequency to -3 dB bandwidth, higher Q means smaller relative bandwidth and more selectivity.

REI Remote Error Indication used by transmission equipment to relay back an error condition.

RJ Random Jitter, which has Gaussian noise-like statistics

RMS Root Mean Square applied to the average long-term value of a signal such as the level of jitter.

RS Reed-Solomon code, the forward error correcting code used in ITU-T G.709

SAW Surface Acoustic Wave filter used for clock recovery in regenerators.

SDH Synchronous Digital Hierarchy.

SEC SDH Equipment Slave Clock, synchronized to a primary or secondary reference clock. Usually an oven-stabilized crystal oscillator.

SMC SONET Minimum Clock similar to the SEC, above.

SMPS Switched-Mode Power Supply

SNR Signal-to-Noise Ratio

SONET Synchronous Optical NETwork, the North American version of SDH.

SSU/BITS	Synchronous Supply Unit/Building Integrated Timing Source is usually a Rubidium/quartz technology clock which may be synchronized to the primary reference clock (PRC).
STM-n	Synchronous Transmission Module where "n" is the hierarchical level. STM-1 (155 Mb/s), STM-4 (622 Mb/s), STM-16 (2488 Mb/s), STM-64 (9952 Mb/s). This is the descriptor for both electrical and optical signals in SDH
STS-n	Synchronous Transmission Signal where "n" is the hierarchical level. STS-1 (52 Mb/s), STS-3 (155 Mb/s), STS-12 (622 Mb/s), STS-48 (2488 Mb/s), STS-192 (9952 Mb/s). This is the descriptor for the electrical signals in SONET, equivalent to the OC-n descriptors for the optical signal.
TDEV	Time Deviation is a measure of the spectral content of wander and is a function of the observation interval.
Telcordia	North American standards organisation, formerly called Bellcore, responsible for the transmission standard GR-253
TIA	Telecommunications Industry Association in North America
TIE	Time Interval Error represents the cumulative time deviation of the clock signal under test, relative to the reference source.
TVAR	Time Variation used to qualify clock oscillator noise processes, TDEV is the square-root of TVAR.
Type A	Designation for low-Q clock recovery circuit in regenerators providing wide-band clock recovery.
Type B	Designation for high-Q clock recovery circuit in regenerators providing narrow-band clock recovery.
UI, UI p-p	Unit Interval and Unit Interval peak-to-peak is the measure of jitter, where 1 UI is equivalent to one bit period.
VC	Virtual Container which refers to the payload area in the SDH frame. May also mean Virtual Concatenation depending on the context.
VSR	Very Short Reach, applied to the high-speed interconnection between equipment in a facility, typically less than 300 meters.
WBFM	Wide Band Frequency Modulation which is the method described in this booklet for verifying the calibration of a jitter transmitter at high jitter amplitudes such as 800 UIp-p.

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