Design of Optical Receiver Modules for Digital Communications Analysis

These three bit-rate-specific optical plug-in modules are essential components of the HP 83480A Digital Communications Analyzer. They are for data rates of 155/622 Mbits/s, 2.488 Gbits/s, and 9.953 Gbits/s.

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The transmission rates of telecommunication systems based on fiber-optic standards such as the Synchronous Optical Network (SONET) and Synchronous Digital Hierarchy (SDH) are at multiples of 51.84 Mbits/s. A significant amount of optical telecommunications network equipment operates at three, twelve, or forty-eight times this fundamental rate. Currently, new equipment is being deployed that operates at approximately 10 Gbits/s, or one hundred ninety-two times the fundamental rate.

Differing measurement requirements at these specific bit rates, along with the modularity available from the hardware architecture of the HP 83480 communications analyzer, allowed the design team to tailor plug-in modules for each application. This modularity also benefits customers, who can configure an instrument that best meets their current needs, then later modify it easily as their needs change. To date, three optical plug-in modules have been released. They are:

- HP 83481A 155/622-Mbit/s optical-to-electrical module
- HP 83485A 2.488-Gbit/s optical-to-electrical module

Measurement Requirements
The measurement requirements inherent in the design and manufacturing of digital communications systems drove the specific design choices made for each plug-in module. The HP 83480 instrument family is used to characterize digital communications signals in the time domain. These signals are typically broadband and usually include a dc component. As an example, the inspection and analysis of eye diagrams are typical customer measurements.1 Eye diagrams are constructed from multiple overlays of successive bit patterns with a synchronized trigger. To display the eye diagram properly, the measurement system must have sufficient bandwidth to show the fast transitions.

A measurement system is often most easily characterized in terms of its frequency response, or the magnitude and phase of the transfer function. This can be related to the time-domain impulse or step response performance by an inverse Fourier transform. The optimum frequency response of a measurement system depends on the waveform measurement parameter of greatest interest (rise and fall times, overshoot, etc.). A rule of thumb for reasonable measurements of rise and fall times is that the ~3-dB bandwidth of the transfer function be at least three times and preferably five times the bit rate to be measured. Ideally, the transfer function should have a well-behaved roll-off and linear phase to prevent ringing or other measurement aberrations. The design target for the optical plug-in modules was to achieve a compromise between fast rise and fall times and excessive ringing. This was accomplished by controlling the amount of high-frequency peaking and striving for a Gaussian-like impulse response.

Very flat low-frequency performance is required for stable measurements of logic levels that extend over many bit periods, which occurs in both long-pattern-length pseudorandom binary sequences (PRBS) and live data transmissions. This is important, for instance, for accurate extinction ratio measurements. Extinction ratio is defined as the ratio of the signal power in the logic 1 state to the signal power in the logic 0 state. It is an important measurement of the distinction between logic states, the essential function of a digital communication system.

For a plug-in module to serve as a reference receiver, the frequency response must, at a minimum, comply with the low-pass Bessel-Thomson transfer function described in the SONET/SDH standards. In fact, mathematical simulations can demonstrate that even reference receiver frequency responses that technically meet the tight SONET/SDH standards can cause unacceptable time-domain artifacts that compromise the accuracy of extinction ratio measurements. Therefore, there is an advantage to a receiver frequency response that closely matches the ideal transfer function. It is critically important that the low-frequency transfer function, down to dc, be well-behaved. A low-frequency gain variation that either rises or droops will cause inaccurate extinction ratio measurements. For example, simulations have shown that a 0.2-dB low-frequency rise (which is within the SONET/SDH standard specifications) can cause an extinction ratio measurement of
10 to be in error by 10%\textsuperscript{2} in a manufacturing environment, such an error might cause the incorrect rejection of a good component.

Finally, the plug-in modules should have a dynamic range as large as possible. A high input power compression point extends the signal measurement range available to the user without adding an external optical attenuator. Noise considerations limit the low input signal range. Since these plug-in modules are often used for nonrepetitive waveform measurements (such as eye diagrams), waveform averaging often cannot be used to provide noise reduction.

**Plug-in Module Overview**

Each plug-in module contains an optical input channel with an optical-to-electrical (O/E) converter, at least one switchable SONET filter, and an electrical sampler with its associated pulse generation circuitry. In addition, each plug-in module has an electrical input channel and a trigger input to route the trigger signal to the mainframe. A generalized plug-in block diagram is shown in Fig. 1.

The O/E conversion starts with a photodiode that converts the incoming photons of light to a proportional electrical current. Because these receiver modules are intended to operate at the primary single-mode communication wavelengths of 1310 nm and 1550 nm, InP/InGaAs/InP p-i-n photodiodes are used. One of the major design choices is whether to add an electronic amplifier immediately after the O/E converter. This selection hinges mainly on a trade-off between the channel signal-to-noise ratio and frequency response. Amplified O/E converters can improve the sensitivity of the channel by

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**Fig. 1.** Generalized HP 83480 plug-in module block diagram.
reducing the effective noise contributions of the subsequent electronic circuit stages. However, amplified O/E converters present more challenges in meeting the stringent frequency response requirements for optical reference receivers, especially at the higher data rates.

The detected electrical signal can be filtered to comply with the communication standard for reference receivers. Each plug-in module comes with either one or two SONET/SDH filters appropriate for the 155-Mbit/s to 10-Gbit/s transmission rates. These filters have a fourth-order Bessel-Thomson low-pass transfer function with a characteristic frequency (-3-dB frequency) at 0.75 times the transmitted bit rate. Around this transfer function there is a narrow tolerance window that depends on the bit rate. One of the key contributions of the HP 83480 is that the entire instrument meets this filtered response, not just the optical receiver. In addition, other communication standard filters can be installed to meet special customer requirements. Of course, all the filters can be switched out of the detected signal path to allow the maximum available measurement bandwidth.

**Sampling Circuit Description.** Signals at these bit rates cannot be digitized directly in real time. There are currently no analog-to-digital converters sufficiently fast to meet these measurement bandwidth requirements. Instead, a sampling technique is used that allows the display of signals that are both repetitive and have a stable trigger. Many digital communications signals fit this description well enough to make the analyzer a very useful measurement tool. While many of the details of a sampling circuit are beyond the scope of this article, some insight into its function will provide an understanding of how the different plug-in modules have been optimized.

A sampler can be thought of as a very fast electrically controlled switch. A fast pulse is used to turn on the switch, which is connected to the analyzer input port. While the switch is on, a current related to the input signal flows into a capacitor. The amount of charge transferred during the sample interval is proportional to the signal present at that instant. If a stable trigger is available, we can eventually build up a representation of the input signal by scanning the time position of the sampling aperture relative to the trigger.

Some design complexity is required to build an electrical switch of sufficient speed. The sampling circuit consists of three main blocks: the sampler microcircuit, a step-recovery diode pulse generator which fires the sampler, and an amplifier and IF filter chain which reshapes the response of the sampler output into a bipolar pulse. The IF output thus generated is sent to the mainframe for further processing and display. The sampler and the step-recovery diode pulse microcircuits are leveraged from the previous-generation HP 54120 Series sampling oscilloscopes.

It is not easy to generate a fast symmetric electrical pulse directly. Instead, the step-recovery diode pulse microcircuit provides a single falling edge with ~70-ps transition time. This waveform is sent to the sampler where it propagates on a slotline transmission line structure. The slotline is terminated by the package wall, effectively forming a short circuit. When the step-recovery diode pulse edge reaches the wall, an inverted reflection is generated and propagates in the reverse direction. The total voltage across any point along the slotline is the sum of both the incident step-recovery diode pulse and the inverted reflected pulse. At an electrical distance from the package wall equivalent to one-half the pulse fall time, the voltage waveform is roughly triangular with 70-ps rising and falling edges. Sampling diodes are placed on the slotline structure at this point. These diodes form the actual switch element.

The sampling diodes turn on when the total voltage provided by the step-recovery diode pulse and its reflection exceed the effective reverse voltage on the diodes. At that time the signal from the O/E converter is connected (switched) to a holding capacitor (see Fig. 2), which collects charge. By controlling the reverse bias voltage across the sampling diodes, the sampling aperture and therefore the bandwidth can be adjusted. A DAC (digital-to-analog converter) within the plug-in module provides this function. Fourier transform theory shows that the sampling aperture time is inversely related to the effective bandwidth. For a 20-GHz bandwidth, the sampling aperture is approximately 20 ps.

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**Fig. 2. Generalized sampler circuitry.**
Why not make the sampling aperture as small as possible? As is usually the case, there is a trade-off involved. Here it is noise versus bandwidth. A larger sampling aperture means more charge can be collected per sample, resulting in a higher signal-to-noise ratio in the subsequent signal processing. It also means a reduction in measurement bandwidth. In reality, the trade-offs are not quite this simple. There are noise contributions from the sampling diodes that also increase at higher signal levels. In addition, larger signal levels also require more dynamic range in succeeding stages to avoid saturation effects. Therefore, selecting the appropriate sampler parameters is a key factor in the optimization of a plug-in module.

Once the charge has been collected, it must be amplified and shaped by filtering before being transferred to the mainframe for processing. The amplifier following the sampler is conceptually an operational amplifier with a resistor/capacitor feedback network. The feedback resistor needs to be as large as possible to minimize its noise contribution. Other system constraints and practical physical values limit the choice to a maximum of about 100 megohms. The input noise of the amplifier itself is minimized by using a discrete, low-noise, differential FET input followed by a precision, high-speed op amp. Significant design effort was expended on this circuit and the effective noise performance of the charge amplifier and IF circuitry is at least twice as good as the previous-generation products.

Because of the narrow sampling aperture, the sampled charge appears as a near-impulse to the much slower following amplifier. This amplifier stage eventually responds and produces an output pulse with an exponentially decaying tail that is proportional to the impulse amplitude. The subsequent filter sections in the plug-in module ac-couple the pulse, which prevents any drift in the dc level from altering the signal amplitude. These IF filter sections shape the exponentially decaying signal into a bipolar pulse, optimizing the signal-to-noise ratio before the analog-to-digital conversion is performed in the mainframe. In addition, the bipolar pulse provides a slowly changing peak amplitude so that the analog-to-digital conversion instant is less sensitive to timing variations in the mainframe.

By repetitively sampling the signal and using a synchronous trigger with a variable delay, a representation of the input signal can eventually be constructed. The sampler can only be triggered at a relatively low rate (40 kHz), so it takes multiple input signal cycles to complete the measurement process. If the IF signal is a faithful low-noise electrical representation of the input optical signal, the plug-in module hardware has done its job.

**Electrical and Optical Calibration.** The electrical sampling process is inherently nonlinear. To reproduce the electrical waveform accurately, a lookup table stored in the plug-in module provides the mapping from the input signal to the IF output for each sampler channel. For the best module performance, the user can perform an electrical calibration by selecting a softkey on the front of the mainframe. A built-in algorithm applies internal reference voltages to the sampler inputs, covering the allowable input range and automatically generating the calibration table.

The optical channel requires an additional calibration term to account for the optical-to-electrical conversion efficiency of the O/E converter. This conversion efficiency is dependent on the wavelength of the optical signal. Factory calibration constants are stored in the memory of each plug-in module for 1310-nm and 1550-nm wavelengths. The linearity of the O/E converters is excellent, so a single calibration constant is sufficient over the allowable optical input power range. Users can perform optical calibrations at other wavelengths by using an optical source with an accurately known power level. This also allows an operator to remove small insertion loss variations of the input connector for high-precision measurement applications.

A convenient feature of the plug-in module is the ability to monitor the average optical power accurately. The average current from a reverse-biased p-i-n photodiode is directly proportional to the average incident optical power. This proportionality constant is determined during the optical calibration process and is wavelength dependent. The average detected current is monitored by a variable-gain amplifier network, which provides a dynamic range of just over 30 dB.

Shown in Fig. 3 is a photograph of the three currently available optical plug-in modules. The side view of the plug-in module reveals the optical converter, samplers, step-recovery diode pulse generator, transfer switch, and filters.

**HP 83485A Plug-in Module**

The HP 83485A plug-in module is an integrated solution targeted for testing laser transmitters operating at rates up to 2.5 Gbits/s. For parametric measurements the optical channel offers a selection of either 12 or 20 GHz of well-behaved measurement bandwidth. As previously described, this bandwidth choice is made by changing the bias on the sampler diodes. The 12-GHz mode offers better sensitivity, having only about half as much channel noise as the 20-GHz mode. For SONET/SDH compliance testing at data rates of 622 Mbits/s or 2.5 Gbits/s (depending on which standard filter option is specified), the switchable low-pass filter shapes the overall frequency response of the channel. Filtered measurements are always made in the 12-GHz mode for best sensitivity, since the channel bandwidth is already limited by the narrower electrical filter bandwidth. In addition to the optical channel, a 20-GHz electrical channel and a 2.5-GHz trigger input are provided. Like the optical channel, the electrical channel can be switched to a 12-GHz mode for improved sensitivity.

Meeting the SONET/SDH specifications at 2.5 Gbits/s is a challenging task for optical transmitter manufacturers. These customers want a measurement system that contributes as little additional error as possible because measurement errors can reduce their manufacturing yields. Therefore, a significant effort in the design of the HP 83485A plug-in module was aimed at providing an accurate frequency response. The optical channel incorporates a 20-GHz unamplified O/E converter to maintain the highest integrity for eye diagram and mask compliance measurements. While many amplified optical receivers
Fig. 3. Three optical-to-electrical plug-in modules are available.

now on the market claim compliance with the SONET/SDH standards, some still suffer from unacceptable frequency response variations. In the HP 83485A plug-in module, these variations are significantly reduced.

**Optical Converter.** The optical-to-electrical converter in the HP 83485A is similar to that of the HP 83440 family of stand-alone optical receivers. The optical detection is performed by a custom 25-µm-diameter InP/In0.53Ga0.47As/InP p-i-n top-illuminated mesa photodiode, which absorbs the incoming infrared light and converts it to an electrical current. The device allows light with wavelengths from 1200 to 1600 nm to pass through the antireflection coating and the top p-type InP layer and be absorbed in the intrinsic InGaAs layer below. The absorption of photons creates electron-hole pairs in the active layer. These carriers are then swept out by an electrical field formed by an applied reverse bias. The carrier transit time across the InGaAs layer and the device capacitance determine the frequency response of the photodiode. A thicker intrinsic layer results in a lower device capacitance and a higher photodiode responsivity, which is the ratio of the detected photocurrent to the input optical power. This is true until the thickness is increased to a point where the quantum efficiency is unity or all the incident photons are absorbed. A thicker intrinsic layer also contributes to a longer transit time, so the thickness has to be carefully chosen to achieve the optimum photodiode bandwidth and responsivity.

The optical launch is required to have low optical reflections and maximum coupling to the photodiode. Optical reflections from a receiver can cause unpredictable measurement variations in some transmission systems. Any reduction in the power coupled to the photodiode produces a direct impairment in the signal-to-noise ratio of the measurement system. The optical launch in the HP 83485A O/E converter uses a single graded-index (GRIN) cylindrical lens to couple light into the small-area photodetector. The input fiber and the photodiode are tilted with respect to the optical axis to reduce reflections. The lens faces are antireflection-coated, resulting in optical return losses that are typically greater than 50 dB at each interface. The input optical fiber is locked in place using a specialized high-stability epoxy.

The optical converter package is hermetic. A mesa photodiode structure can be sensitive to moisture, resulting in increases in both dark current (reverse-bias photocurrent that flows without any light present) and noise. Water vapor can also cause a number of other semiconductor reliability problems. To provide hermeticity, the microcircuit package is sealed with glass and metal. The body and lid are made of gold-plated stainless steel. Glass-to-metal seals are soldered into the microcircuit body to provide electrical connections. A lens is soldered into the microcircuit lid to provide a hermetic optical port. This allows the input optical fiber to be attached with epoxy, since it is outside the hermetic wall. After the photodiode and other circuit elements are die-attached to the microcircuit floor and wire bonded, the lid is soldered to the body in a dry helium atmosphere.

The major modification made to the O/E converter used in the HP 83485A is the addition of a 50-ohm thin-film network at the photodiode output to provide a good electrical output match. This termination network is required to minimize reflections between components that can add frequency response ripple. Without the 50-ohm termination, the high output impedance of the photodiode would create a severe mismatch with the filter input, resulting in excessive ripple in the filtered response when used in the reference receiver mode. The two most important sets of reflection pairs are between the O/E converter and the filter input and between the O/E converter and the sampler. The 50-ohm termination network in the O/E converter provides at least a 26-dB output match from dc up to 5 GHz, where interactions with the filter are critical. The termination network, however, does introduce some peaking in the O/E converter response near 20 GHz. The frequency
response of the sampler is adjusted accordingly to generate the best overall channel frequency response. Shown in Fig. 4 are the measured unfiltered and filtered frequency responses of an HP 83485A-based digital communications analyzer.

Because of the low optical conversion gain (−20V/W) that results from the unamplified O/E converter, there is a potential for poor plug-in module sensitivity. To alleviate this concern, adjustments were made to the sampling parameters. This resulted in a typical optical noise level of only 8 μW rms for the 12-GHz bandwidth setting.

**HP 83485B Plug-in Module**

The HP 83485B is a companion module to the HP 83485A and provides an integrated solution for testing laser transmitters operating at 10 Gbits/s. The HP 83485B also consists of an optical channel, an electrical channel, and a trigger channel. Its optical channel includes a fourth-order Bessel-Thomson low-pass filter, which can be switched into the measurement path for making the compliance measurements.

The design of the HP 83485B is leveraged from the HP 83485A, allowing reuse of most of the internal printed circuit boards and mechanical structures. To achieve an increased channel bandwidth, a higher-speed, 14-μm-diameter photodiode is used in the O/E converter, along with higher-bandwidth samplers in both the optical and electrical channels. The combination of these new components provides over 30 GHz of bandwidth in the optical channel. For the electrical channel, the higher-bandwidth samplers are biased to provide 40 GHz of electrical bandwidth. The amplifier and IF filter circuits following the sampler and the optical average power monitor circuits are identical to those in the HP 83485A plug-in module. Calibration of both the electrical and optical channels is also the same for this plug-in module.

Currently there is no SONET/SDH industry standard for the frequency response performance of reference receivers used in testing 10-Gbits/s transmitters. Expectations are, however, that the frequency response will have a shape similar to the fourth-order Bessel-Thomson transfer function used for the lower bit rates, but with wider tolerances. The HP 83485B filtered performance specifications give the best possible fourth-order performance while maintaining good manufacturability.

The higher bandwidth required for the HP 83485B forced several design and performance trade-offs. The most important trade-offs concerned the O/E converter and the sampler. At these higher frequencies, parasitic components in the O/E converter are more difficult to control, which can result in a frequency response that includes some peaking. The photodiode is followed by a termination circuit to provide a 50-ohm output match. Without the termination circuit, the photodiode frequency response is very well-behaved. If the termination circuit were ideal, this frequency response would be preserved.

In practice, however, the bond wire required to connect the photodiode output introduces a parasitic inductance. This inductance resonates with the 80-fF photodiode capacitance, resulting in a peaked frequency response.

In the time domain, this type of resonance would manifest itself as overshoot and ringing in response to an optical impulse at the input. These measurement artifacts are undesirable for precision time measurements, so it was necessary to make adjustments to both the unfiltered and filtered frequency responses. In the unfiltered mode, the sampler frequency response is adjusted to minimize the effect. In the filtered mode, the roll-offs of both the filter and the sampler were customized to compensate for the high-frequency peaking.

At 10 Gbits/s, mismatch ripple between the O/E converter and the filter can be significant. At these higher frequencies, the output impedance of the photodiode is dominated by its capacitance. By adding the termination circuit, the output impedance of the O/E converter is nominally 50 ohms at lower frequencies, but eventually becomes capacitive at higher frequencies. The circuit was designed to minimize this effect, but it could not be completely eliminated. Other potential
mismatch ripple contributions are minimized by selecting the best possible components in the filtered path, such as switches and connectors.

The filtered frequency response also requires good control of the hardware filter parameters. At lower bit rates, a tolerance of ±0.1 dB can be achieved for the passband region. At 10 Gbits/s, however, the filter elements become quite small in value, making manufacturing consistency more difficult. Correlation between the frequency response measurements of isolated filters and measurements of the optical channel in filtered mode is also important. A few tenths of a dB of absolute error is possible from the network analyzer systems used to make the filter measurements. The tolerance on the filter response is specified as ±0.5 dB in the passband. When assembled into the complete instrument, the plug-in module meets the transfer function specification with a tolerance of ±1.25 dB to 7.5 GHz, the characteristic frequency. While in principle controlling the mismatch ripple is important to avoid distorted eye diagrams and measurement errors, computer simulations show that the level of mismatch ripple present in the HP 83485B filtered path does not significantly affect extinction ratio measurements. Shown in Fig. 5 are the measured unfiltered and filtered frequency responses of an HP 83485B-based digital communications analyzer.

Finally, the higher-bandwidth O/E converter and sampler have lower conversion gains, which results in a somewhat lower sensitivity for the HP 83485B plug-in module. Once again, the optical channel sampler bandwidth was adjusted in the filtered measurement mode to give the best possible sensitivity. This results in a 15-µW rms noise level for the 30-GHz bandwidth setting.

**HP 83481A Plug-in Module**

Compliance testing of 155-Mbit/s and 622-Mbit/s laser transmitters is the intended application for the HP 83481A plug-in module. Whereas laser transmitters for 2.5 Gbits/s and 10 Gbits/s are used primarily for long-haul transmission, transmitters at 155 Mb/s and 622 Mb/s are used for shorter-distance transmission and more varied applications. For some applications, the laser output power requirement is lower, which necessitates a measurement receiver with better sensitivity. When this is the case, a low-noise amplifier can be added to the O/E converter, before the sampler, to improve the noise figure of the entire receiver. The HP 83481A was designed with an amplified O/E converter for just such measurement needs.

In addition to an amplified O/E converter, the HP 83481A provides the customer with two switch-selectable Bessel-Thomson filters for compliance measurements at either 155 Mbits/s or 622 Mbits/s. Integration of these two filters within the receiver significantly improves measurement repeatability and reliability. The user can also select an unfiltered mode which bypasses the filters to provide the entire amplifier bandwidth for parametric measurements. The rest of the plug-in module, beginning at the sampler, is leveraged from the HP 83485A.

The amplifier approach chosen for the HP 83481A is a custom transimpedance design realized in a proprietary high-speed silicon bipolar IC process. This IC process is well-suited to delivering the level of performance needed to meet the demanding requirements of the SONET/SDH standards for reference receivers. An amplifier bandwidth of 3 GHz with a conversion gain of 500V/W was achieved. A low noise floor of approximately 1 µW rms results in an overall amplifier dynamic range of nearly 30 dB.

The design of the amplified O/E converter attempts to provide the user with a well-behaved pulse response for the required compliance measurements of laser transmitter waveforms. The amplifier is dc-coupled and the low-frequency gain is carefully controlled to give a stable pulse settling behavior. Pulse overshoot and ringing are damped to a suitable degree but not so much as to seriously degrade rise time. The low output impedance available with this IC technology meant that a series resistance was needed to match to 50 ohms. Consequently, some conversion gain was sacrificed to improve the output.
match. Fig. 6 shows the frequency response of the amplified converter used in the HP 83481A. Flatness to the 0.75-bit-rate

![Graph showing frequency response of amplified converter](image)

characteristic frequency of 466 MHz is better than 0.3 dB peak to peak. The small dip in gain from 500 MHz to 1 GHz is
caused by the bond pad capacitance of the transimpedance IC. Peaking, common to all transimpedance amplifiers, is below
1 dB and is well beyond the SONET/SDH specified frequency range. With this level of performance, the amplifier contributes
only about half the allowable deviation from the ideal reference receiver transfer function. Shown in Fig. 7 are the measured
unfiltered frequency response and the two filtered frequency responses of an HP 83481A-based digital communication
analyzer.

![Graph showing measured frequency responses](image)

Transimpedance Amplifier. The schematic of the complete amplifier is shown in Fig. 8. The transimpedance front end is
formed by transistors Q1 to Q3 with resistor Rf (1100 ohms) setting the overall transimpedance gain. A single-stage voltage
gain block (Q1, R1) was chosen for maximum phase margin and has a gain of 17. Transistor Q2 is a buffer and transistor Q3
provides an additional level shift to increase the maximum output voltage swing. Q2 is also part of the dc level temperature
compensation. Capacitors C1 and Cf introduce zeros into the preamplifier frequency response, further increasing the phase
margin. The driver amplifier was chosen for its simplicity and is implemented with a double emitter follower capable of
driving low-impedance loads. The overall transimpedance of this amplifier is 1000 ohms and its intrinsic bandwidth is
5.2 GHz, resulting in a gain-bandwidth product of 5.2 THz-ohm. The intrinsic bandwidth is the bandwidth with an ideal
zero-capacitance photodiode.

To stabilize the dc output level, the IC has passive temperature compensation. Active temperature compensation is capable
of achieving essentially zero dc drift but it usually affects the frequency response and ac stability of the amplifier. The
passive temperature compensation is obtained by causing the temperature drifts of the components in the signal path to
cancel. At the output of the transimpedance preamplifier there is a positive voltage drift equivalent to the drift of the
base-emitter junctions of Q1 and Q3 in series. This would ideally be canceled by the negative voltage drifts of base-emitter
junctions in the driver. However, because the current densities of these transistors are not identical there is residual
temperature drift of 76 µV/°C. This low value is suitable to minimize drift of the signal on the screen of the communications
analyzer and allows accurate extinction ratio computations.
To keep the amplifier response as flat as possible, a special constant-impedance ladder network was designed into the bias circuit of the IC. This circuit is made up of three sections of low-pass impedance transformers. The sections have progressively lower cutoff frequencies as the distance from the IC is increased. The advantage of this approach is that the unavoidable lead inductance is absorbed into the impedance transformer network. A separate network feeds the positive supply lines of the preamplifier and driver sections of the IC. This circuit provides a constant impedance of 7 ohms to the amplifier sections over the entire frequency range. Resonances are suppressed by the resistance of the bias network and a flat frequency response to dc is achieved.

The negative supply is less sensitive to low-frequency resonances but it has a different problem. The negative bias port feeds both the input and the output sections of the IC, and this is an undesired source of feedback. This coupling is made worse with increased bond wire inductance in the ground path. The effect manifests itself in increased frequency response peaking and some droop in the gain. The solution is to minimize the total ground bond length. All bias elements are integrated onto a thin-film substrate and a laser-shaped cutout is used to allow the IC to be mounted in close proximity to the bias circuit and flush with the top of the substrate. Multiple ground bonds of minimal length can be made to the substrate with this approach. The frequency peaking that results is typically less than 1 dB.

**Optical Launch and Package.** Besides the requirements that the optical launch have low optical reflections and maximum coupling to the photodiode, care must be exercised to ensure that no reflected rays can couple back into the input fiber.

The other design constraint is to minimize aberrations of the illuminated spot on the photodiode that falls predominantly within the 25-µm-diameter active region of the photodiode. This must be maintained over the temperature range of the instrument. To meet these goals an aspheric lens with a magnification of about 2 was chosen. This lens images light from the 9-µm fiber core diameter to a spot size of about 5 µm on the photodiode surface. The lens is coated with an antireflective layer and is mounted at a 20-degree angle to minimize reflections. This angle is enough to keep reflected rays from the photodiode surface from falling within the numerical aperture of the lens.

The lens and amplifier components (see Fig. 9) are mounted inside a hermetic package to protect the photodiode. A single-mode fiber is attached outside of the package. A sapphire window brazed into the lid of the package allows light to pass from the fiber to the lens. Since the window is tilted relative to the optical axis, an angle must be chosen that keeps the polarization dependent loss (PDL) at a reasonable level. An angle of 12 degrees was chosen for the window, which keeps PDL below 1.5%. This design eliminates the need to provide a hermetic seal around the fiber, thus simplifying the mechanical assembly process. The end of the fiber is polished at an angle to minimize reflection from this interface. Using this design a return loss in excess of 55 dB is readily obtained with negligible impact on coupled power.

The package design goal was to create a highly reliable hermetically sealed package to meet all functionality and manufacturability requirements. In selecting material for the package, Kovar (29% Ni, 17% Co + Fe) was chosen for its low thermal expansion, which closely matches that of the thin-film circuits and glass seals inside. Also, Kovar is an excellent material for laser welding and has good corrosion resistance. Fig. 10 shows a picture of the complete converter package.

A new ceramic brazing process called active metal brazing was introduced by the manufacturing engineers for attachment of the sapphire window to the Kovar lid. This new process eliminates the necessity of premetallization of the sapphire window edges by using a special active braze alloy of 63% Ag, 34.25% Cu, 1.75% Ti, 1% Sn. During brazing, the oxidization and diffusion of the active alloying element, Ti in this case, creates a chemical bond at the metal-ceramic interface. By means of
this relatively simple and robust new process, the cost saving realized in making this window/lid assembly is estimated at 70% compared with the conventional process.

The Kovar lid is hermetically attached to the body by laser welding using a Nd:YAG pulsed laser. The key to a sound weld joint is to keep a tight and consistent fit between the lid and the package body. Gold plating in the weld area has to be removed before welding to prevent cracking. The entire package assembly has withstood strife testing that involved temperature cycling from \(-50^\circ\text{C}\) to \(+85^\circ\text{C}\), random vibration at 8g rms, and mechanical shock to 1800g.

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