

Agilent HCPL-800J

Application Note 5074

Introduction

Low-frequency powerline communication modems (PLMs) have many applications in homes and businesses, including:

- Home appliance control
- Heating and ventilation control
- Lighting control
- Low-speed data communications networks
- Signs and information displays
- Fire and security alarm systems

Powerline communication modems transmit digital data on the live and neutral cables of the building power distribution network. Data is typically transmitted at carrier frequencies in the range from 50 kHz to 500 kHz using a variety of modulation techniques including amplitude keying, frequency-shift keying (FSK), binary phase shift keying (BPSK) or direct-sequence spread spectrum (DSSS).

Implementation of digital signal processing techniques in PLMs has helped to improve communication performance. However up to now there has been little or no improvement in the performance of the analog front-end (AFE) circuit that provides the analog signal processing needed to effectively couple the digitally-modulated carrier to and from the powder distribution wiring.

The AFE provides voltage and current signal amplification and galvanic isolation to protect the user and equipment from both the normal 120/230 VAC line voltage and from much higher transient voltages frequently appearing on the lines. The AFE is traditionally a complex circuit requiring the use of many discrete components including bulky isolation transformers.

Figure 1 shows an example of a conventional AFE circuit, which includes the key circuit functions:

- Transmitter filter
- Transmitter line driver
- Coupling transformer
- Coupling capacitor
- Receiver filter
- Receiver amplifier
- Surge protection

Such a conventional AFE can require as many as 60 individual components.

Agilent Technologies offers an alternative approach to analog front-ends in the form of the HCPL-800J DAA (data access arrangement) IC. The Agilent HCPL-800J provides an integrated, cost-effective and reliable solution for narrowband (up to approximately 10 kb/s) powerline modem AFEs that significantly simplifies power line communication system designs and helps improve performance.

The function of Agilent's HCPL-800J is to interface with the PLM transceiver/ENDEC (encoder-decoder) ICs available from several manufacturers with the

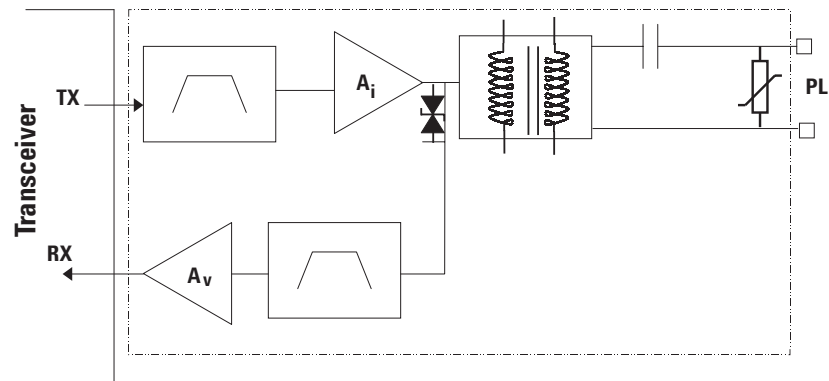


Figure 1. A conventional analog front-end (AFE) circuit using discrete components for a powerline communications modem.

AC power line that is used as the signal transmission medium. It is the industry's first integrated optical-based analog front-end solution.

The HCPL-800J incorporates two integrated circuits: a control IC and a line IC, optically isolated from each other. The modulated signal from the PLM transceiver is input to the control IC. This transmit signal is optically coupled to the line IC, then amplified and sent to the power line. In the other direction, a potentially weak and noisy signal from the power line is received by the line IC and optically coupled to the control IC, where it is amplified and output to the PLM transceiver. The HCPL-800J incorporates protection mechanisms including over-temperature, over-current and low supply voltage detection, and line condition monitoring. It operates from two +5 V power supplies and is supplied in a 16-pin SOIC (Small Outline IC) package. Figure 2 is a block diagram of the Agilent HCPL-800J, showing how it is applied in an AFE.

In considering the requirements of the AFE the power-line communication environment needs to be carefully considered.

Power-line Environment

Power-line Impedance

The effective impedance observed by the PLM transmitter is the cumulative impedance of all devices connected at power outlets plus the impedance of their respective interconnection cables. In the power-line communication frequency range from 50 kHz to 500 kHz the connection cable can normally be simplified as an inductive impedance.

With respect to designing an effective line driver for the AFE, it is the minimum power-line impedance that is the most important and that potentially causes the most problems.

Extremely low power line impedances are usually caused by one or both of the following:

1. A low impedance load connected in close proximity to the transmitter. This may be caused by an inductive load such as a motor or a resistive load such as a heater. However, it is more commonly caused by a large capacitive load such as the filter and supply decoupling capacitors used on the input of switching

power supplies, or noise suppression capacitors used on motors.

2. Resonance of the cable inductance with a capacitive load. In this instance the reactance of the power cable matches the reactance of the capacitive load. If this occurs the line impedance observed at the transmitter is equal to the resistive impedance of the power cable, which can be very low, especially for short separation distances.

The power-line impedance usually varies with time. One source of such variations is that loads, such as lamps or appliances that operate on an automatic cycle (refrigerators, air conditioners, furnace blowers) are connected and disconnected during the course of the day. In addition, the power line impedance can even change during each cycle of the alternating power line voltage. For example the rectifying circuits used on the front end of switching power supplies effectively disconnect/reconnect the power supply and its input capacitors each time the line voltage passes through the zero crossover point.

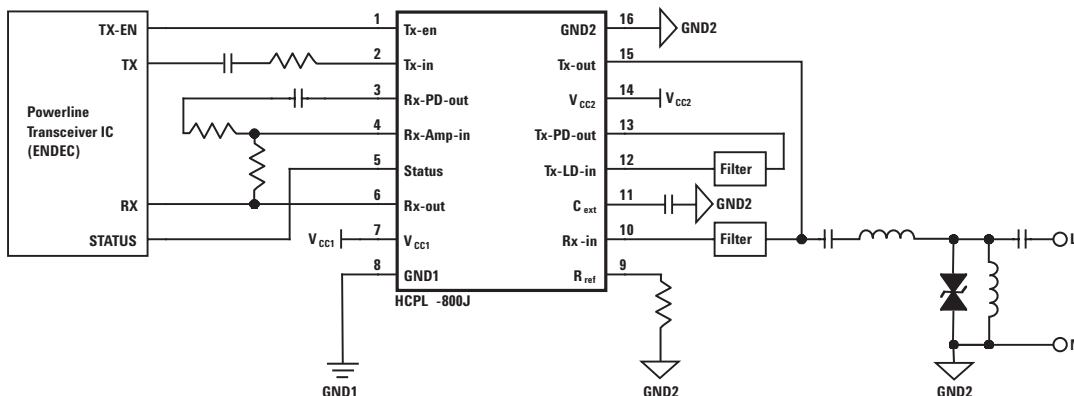


Figure 2. Block diagram of an AFE using Agilent's HCPL-800J data access arrangement IC.

Signal Attenuation

The finite coupling impedance of the transmitter stage together with the power-line impedance acts as a voltage divider. Subsequent attenuation takes place at each power outlet, due to the voltage-divider effect of the power-line cable inductance and the load connected to each power outlet. Total signal attenuation between the transmitter injection point and the receiver is therefore a function of the number and types of power-line loads and their interconnection lengths (impedances), as shown in Fig. 3.

Power Line Surges

High voltage transients and surges are a frequent occurrence on the power distribution network. The primary recurring source of surges is the switching of heavy loads such as large motors, but may be also result from less common events such as lightning strikes close to power lines.

Without sufficient protection, all electrical circuitry connected to the power distribution network is at risk from surge voltage-related damage. It is the low-voltage circuitry that is most vulnerable to high voltage transients. For most devices connected to the power utility, the only coupling mechanism the surge has to the sensitive low-voltage circuitry is via the power supply. The low-voltage circuitry is normally adequately protected from mains transients by the inherent low pass filter characteristics of the power supply circuit.

In the case of equipment incorporating a powerline modem, a high frequency path to the power

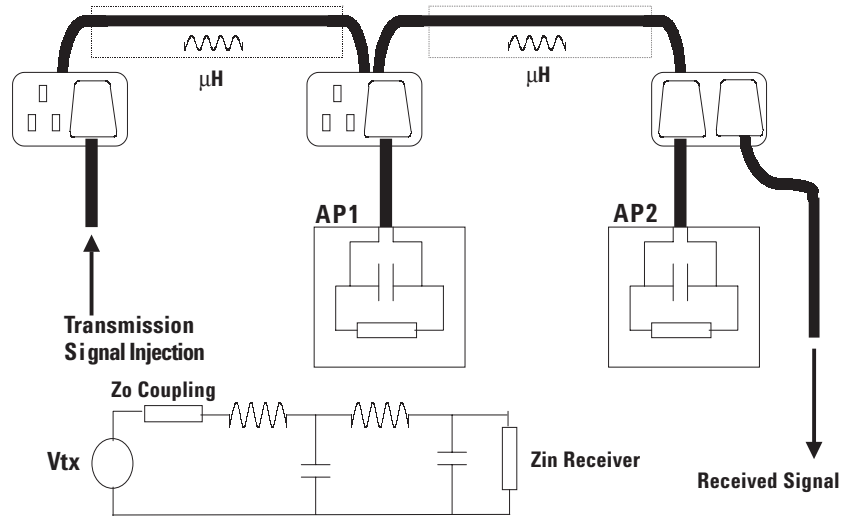


Figure 3. Total signal attenuation between the transmitter injection point and the receiver is a function of the number and types of power-line loads and their interconnection lengths (impedances).

line has to be opened up to allow data transfer. This signal coupling path is traditionally provided through an isolation transformer.

Unfortunately a transformer is not only able to transit and receive modulated signals, but is also capable of coupling surge transients into the low voltage control circuitry. These surges can potentially damage not only the powerline modem but also the circuitry of the associated appliance.

Power-line Noise

Power-line noise can be characterized into several categories having similar characteristics and influences on the powerline modem function.

Synchronous Impulse Noise

Impulse noise that is synchronous with the 50 or 60 Hz line frequency generates a broad range of frequencies for short periods of time. An example of a source of synchronous impulse

noise is a lamp dimmer in which the switching of a Triac results in periodic bursts of broad-spectrum noise.

Continuous Broad Spectrum Noise

A frequent source of continuous noise on the power line is the universal motor commonly used in electric drills, vacuum cleaners and kitchen appliances. Switching transients created by the motor's commutator and brushes cause the impulses. Although generally regarded as being continuous, the noise is in fact periodic with relatively short periods—related to the rotational speed of the motor—between impulses.

Nonsynchronous Impulse Noise

Nonsynchronous impulse noise creates broadband noise frequencies for short periods at irregular intervals. This form of noise is commonly caused by switching transients across electrical contacts such as mechanical switches and thermostats. Each time the circuit is

broken, the result is a single impulse or a group of impulses in quick succession. This type of noise appears in Fig. 4.

Continuous Narrowband Noise

Continuous narrowband noise occupies one or more discrete frequencies. Sources include television horizontal sweep oscillators, which for a PAL/SECAM system is 15,625 Hz, switch-mode power supplies and radio transmissions.

Many noise generators produce several categories of noise simultaneously. For example, triac-based lamp dimmers can generate continuous narrowband noise at harmonics of the power-line frequency in addition to the broadband noise.

In terms of the relative impact on the communication performance of PLMs, narrowband noise only causes problems if the noise is within or close to the transmitted frequency band. Impulse noise generates noise over a broad range of frequencies so it is always a concern.

Noise Coupling

When optimizing the design of the AFE it is important to appreciate the mechanisms by which noise is coupled from a source of interference. The two fundamental coupling mechanisms are differential-mode and common mode.

Differential-mode Noise:

Differential-mode noise is often referred to as conducted noise, since it originates from devices connected to the same power line. It appears as a differential voltage across the live and neutral terminals of the interference source.

The noise amplitude observed at any particular position on the power distribution network is therefore dependent on:

1. The noise source amplitude.
2. The noise source impedance, including cable impedance between the noise source and the measurement point
3. The power-line load impedance

Two important consequences are that low line impedances will result in lower noise levels, and that the greater the distance from the noise source the lower the noise level will be.

Common-mode Noise:

Unshielded cables can act as antennas, which can pick up RF signals from both unintentional emitters such as arcing electrical contacts and intentional emitters such as radio stations.

The power line is no exception. Furthermore since live and neutral conductors are run in close proximity to each other, RF noise gets equally coupled onto both live and neutral conductors creating a common-mode voltage with respect to the earth ground.

Common-mode noise can also come from the same sources as conducted differential noise when a portion of the noise voltage is propagated as a common-mode signal. An example of this is leakage of switching harmonics through the parasitic capacitance of a switch-mode power supply transformer.

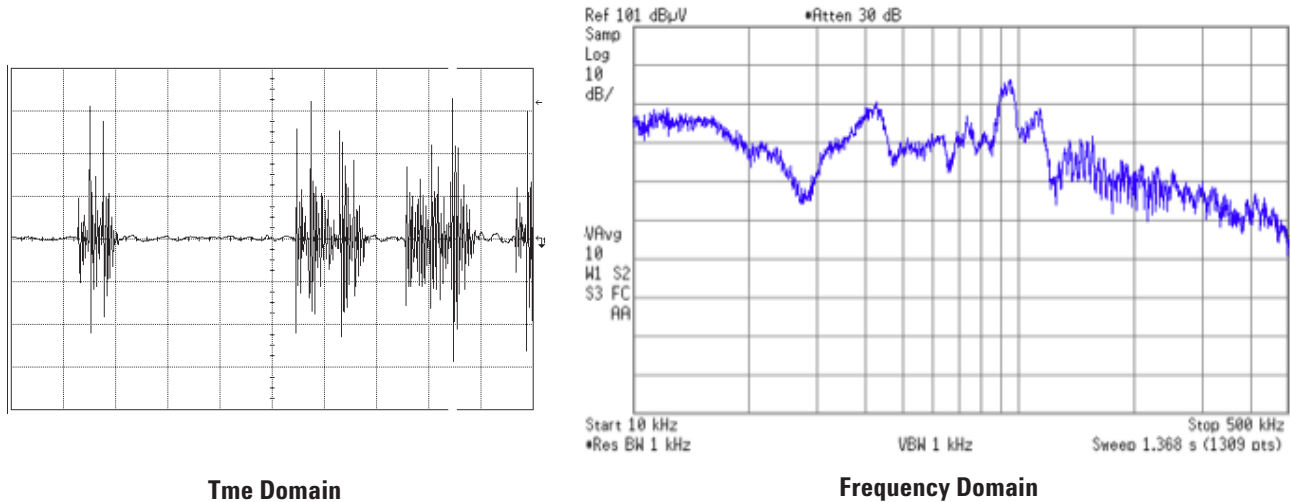


Figure 4. Typical nonsynchronous impulse power line noise in time and frequency domain.

Common-mode noise is not often distinguished from differential noise. One reason is that it is very difficult to accurately measure common-mode noise. Another is that the receiver circuit is not directly affected by common-mode noise, but by common-mode noise that is converted to differential noise by unavoidable asymmetry in the isolation and receiver circuitry.

It is important to differentiate common-mode from differential noise because common-mode noise is not attenuated by power-line load impedance nor distance as are both the transmitted PLM signal and differential noise. Consequently the impact of common-mode noise increases as the transmitted signal is attenuated.

Powerline Modem Conducted Interference Regulations

Because the unshielded, unbalanced nature of the power line means that it can potentially radiate radio frequencies, the use of the power line medium for transmitting data is often strictly controlled by national and international regulations. The regulations are intended to prevent interference both to other devices connected directly to the power line, and to RF services such as navigation systems and commercial TV and radio stations.

In addition to the interference concern, some international regulations go even further, treating the powerline modem communications frequencies as a resource, which should be managed to permit the coexistence of various devices and to ensure that the available bandwidth is most efficiently utilized.

North American Regulations
Powerline modem applications are governed by part 15 of the US Federal Communications Commission rules. In particular powerline modems are classified as unintentional radiators with limits placed on conducted emissions in the frequency bands from 450 kHz to 500 kHz and from 5 MHz to 30 MHz. The revised CISPR (International Special Committee on Radio Interference)/FCC limits for class A and class B devices are 79/73 dBµV and 56 dBµV/60 dBµV respectively.

European Regulations
In Europe, the operation of powerline modems is primarily governed by EN50065. This standard not only sets limits on the frequency bands which may be utilized but also sets limits and rules to facilitate the operation and coexistence of power-line modems in allocated frequency bands. The allocated frequency spectrum ranges from 3 kHz to 148.5 kHz as shown in Fig. 5.

The frequency band A and the frequency band from 3 kHz to 9 kHz are reserved for the use of utility companies and their partners. All other bands are available for general use with the exception that operation in band C requires the use of a media access protocol to prevent data collisions.

Not only does EN50065 specifically limit the peak amplitude of the modulated signal (upper curve in Fig. 6) but it also sets strict controls on spurious transmission harmonics within the overall power-line frequency band (lower curve in Fig. 6). For example, for a class 122 device transmitting on 130 kHz the transmitting amplitude is limited to an output amplitude of 116 dBµV and the resultant second and third harmonics have to be less than 62 dBµV and 58 dBµV, respectively.

IEC CISPR-22
Even if the region of the world in which the system is to be operated does not have any

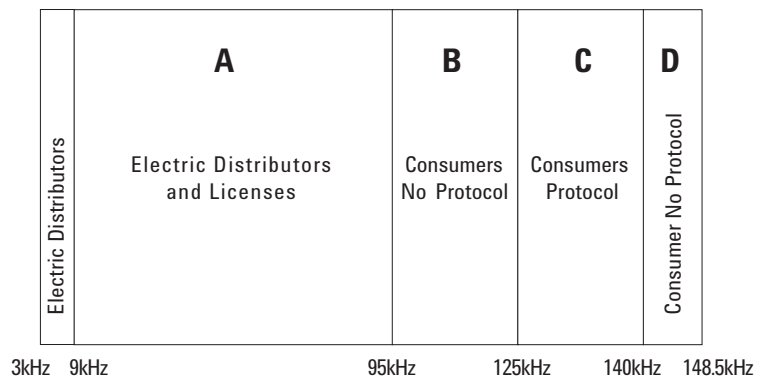


Figure 5. Frequency spectrum for powerline modems allocated under the EN50065 standard.

particular regulations governing the operation of powerline modems. To prevent interference to other users it is prudent to minimize unintentional and avoidable signal radiation, particularly within frequency bands used for TV and radio broadcasting stations and other over-the-air communications. An appropriate design criterion would be compliance with the conducted noise emission levels of an international standard such as IEC CISPR-22.

Measurement Method

To measure the conducted signal level and spurious harmonics a line impedance stabilization network (LISN) is required.

A LISN serves three main purposes:

1. It isolates the power line from the equipment under test and removes any noise on the power line which may incorrectly be attributed to the powerline modem under test.
2. It provides a stable line impedance, which is intended to simulate a typical power line environment.
3. It provides a coupling circuit from the power line to the spectrum analyzer or EMI analyzer.

Figure 7 shows the use of two LISNs for checking the compliance of an EN50065 application. A point to note here is that the use of two LISN networks effectively attenuates the measured signal by 6 dB, so the maximum transmitted signal and harmonics levels may be increased by 6 dB in relation to the curves shown in Fig. 6 .

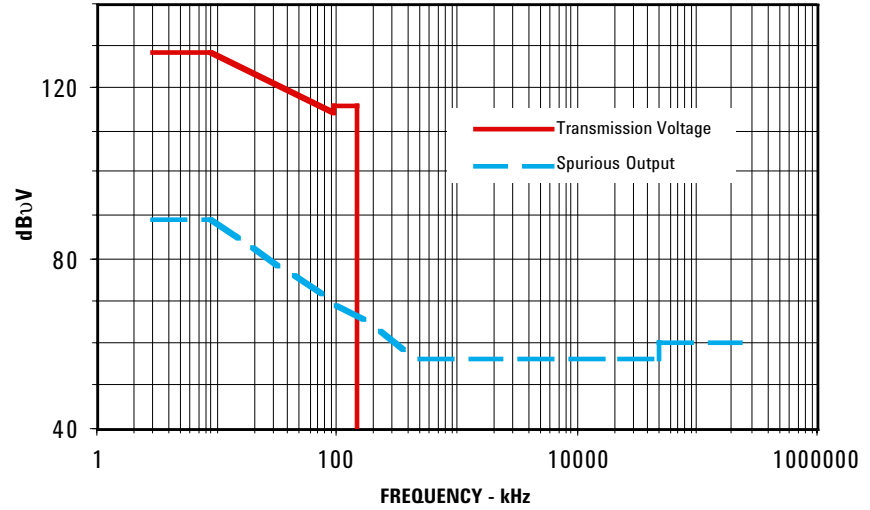


Figure 6. Transmission and spurious emission limits under EN50065.

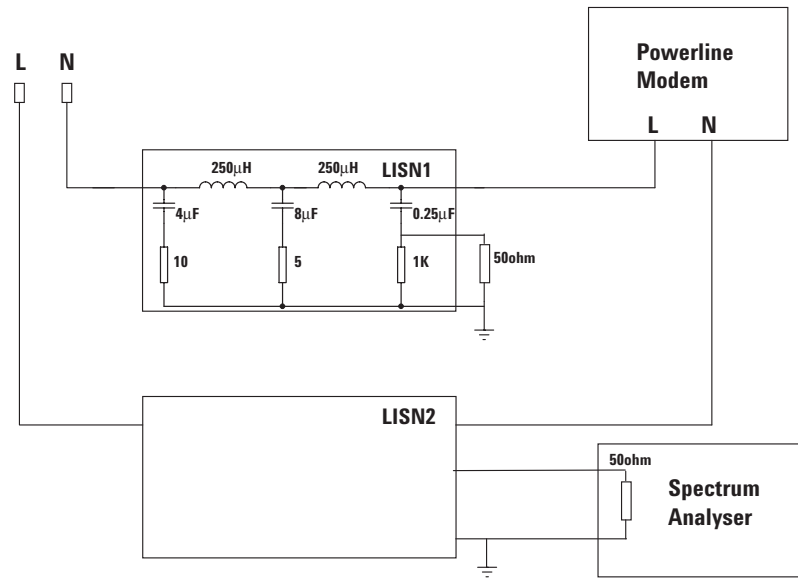


Figure 7. Using two line impedance stabilization networks (LISN) to check for compliance to EN50065.

Implementing the HCPL-800J

A block diagram of the HCPL-800J is shown in Fig. 8, with the pin-outs listed in Table 1.

Transmit (Tx) AGC Circuit

The transmit (Tx) AGC circuit ensures that optimal peak output signal is reached under a wide range of conditions, including variations in signal output from the modulator and variations in the HCPL-800J device's optical channel over temperature and time.

As soon as the Tx stage is switched to transmit mode the AGC starts to increase the output signal until it reaches a state of $3 V_{pk-pk}$. For this reason it is recommended to start transmitting either the data or preamble signal as quickly as possible after enabling the Tx stage. Ideally the modulated signal should be started within 100 μs of the Tx

Pin Descriptions

Pin No.	Symbol	Description
1	Tx-en	Transmit Enable Input
2	Tx-in	Transmit Input Signal
3	Rx-PD-out	Rx Photodetector Output
4	Rx-Amp-in	Receiver Output Amplifier Input
5	Status	Signal indicating Line Condition
6	Rx-out	Receiving Signal Output
7	V _{CC1}	+5 V Power Supply
8	GND1	V _{CC1} Power Supply Ground

Pin No.	Symbol	Description
16	GND2	V _{CC2} Power Supply Ground
15	Tx-out	Transmit Signal Output to Powerline
14	V _{CC2}	5 V Power Supply
13	Tx-PD-out	Tx Photodetector Output
12	Tx-LD-in	Tx Line Driver Input
11	C _{ext}	External Capacitor
10	Rx-in	Receiving Signal Input from Powerline
9	R _{ref}	Sets Line Driver biasing current, typically 24 k Ω

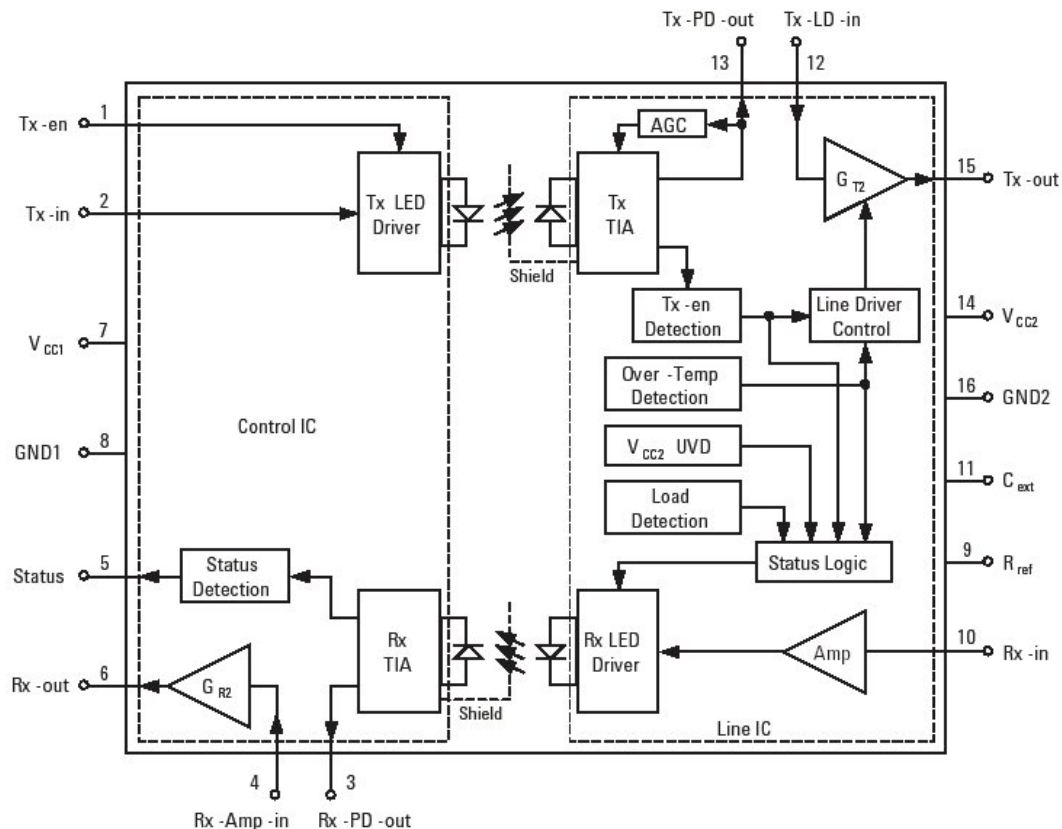


Figure 8. Block diagram of the Agilent HCPL-800J PLC powerline DAA IC.

pin being enabled. If the modulated signal is started much later than this there is the possibility of the output signal being distorted and clipped for a short period (<250 μ s) until the AGC is able to compensate. If it is not possible to start the modulation signal within 100 μ s of enabling the transmit function, the transmit enable (Tx-en) signal may be delayed either in software or with an external R-C delay circuit.

The optimum signal level on the Tx input is 250 μ A. An input signal significantly lower than this may be utilized with no effect on the output distortion level, but a signal level much below 180 μ A could cause the input signal to be outside the pull-in range of the AGC circuit, resulting in a drop in the output signal level.

Conversely, operating the input at input current levels greater than 250 μ A will result in progressive degradation of harmonic distortion performance. Although the input current is most simply set using an external capacitor and resistor combination, the basic input circuit may be modified to provide some input filtering.

Transmit Filter

A filter is normally required in the transmit path to remove spurious harmonics to ensure that the PLM meets pertinent international emissions regulations.

Spurious harmonics originate from two main sources: the internal clock frequencies used in the modulator, usually in the MHz range, and distortion harmonics resulting from the synthesis of the modulation waveform.

Before designing a suitable filter, it is necessary to consider in which region of the world the modem is to be operated. In Europe the EN50065 spurious emission limit curve specifically governs the maximum level of all harmonics for all frequencies in the range from 10 kHz to 148 kHz. For example if a modem transmits on a carrier frequency of 120 kHz, it is necessary to ensure that the second and third harmonics are at a level below 62 dB μ V and 58 dB μ V respectively.

In the US it's a little different because the regulations do not specifically limit all harmonics but rather only limit harmonic levels at frequencies greater than 500 kHz. This means that if the same 120 kHz modem is to be used in the US, only out of band harmonics greater than the fourth order are of concern.

The next point which needs to be considered, is the frequency synthesis method utilized by the PLM. The two most commonly used digital methods are square-wave modulation and a digital-to-analog converter (DAC) combined with a lookup table (Fig. 9):

With square-wave modulation the output signal contains very high odd-harmonic levels, which require considerable filtering to meet most international emissions regulations.

Generally speaking the harmonic levels of a DAC/lookup table synthesizer are significantly lower. The actual harmonic levels and frequencies depend on the number of effective bits and the sample intervals.

To remove the out of band harmonics a low-pass or bandpass filter is required.

The high frequency roll-off performance and hence the filter order is determined by the level of the spurious harmonics and the required attenuation. The higher the filter order the faster the high frequency roll off. The HCPL-800J includes a buffer output stage at pin 13, which is capable of driving a passive filter circuit. Figure 10 shows two example filter combinations. The output of the filter should be coupled to the line driver input (pin 12) with a coupling capacitor of less than 10 nF.

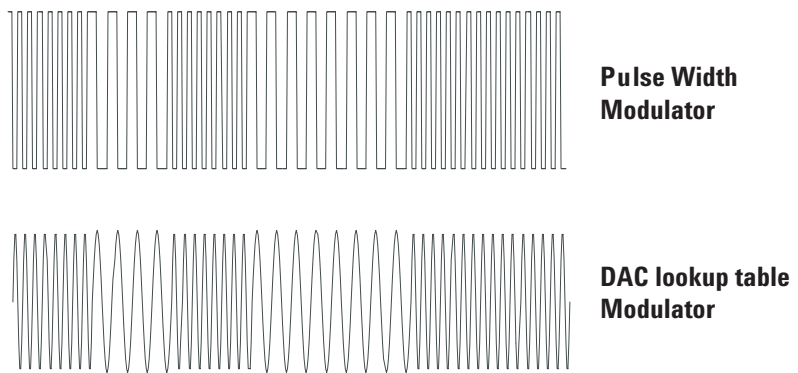


Figure 9. The two commonly used methods of frequency synthesis—square-wave modulation and a DAC with a lookup table—produce different waveforms.

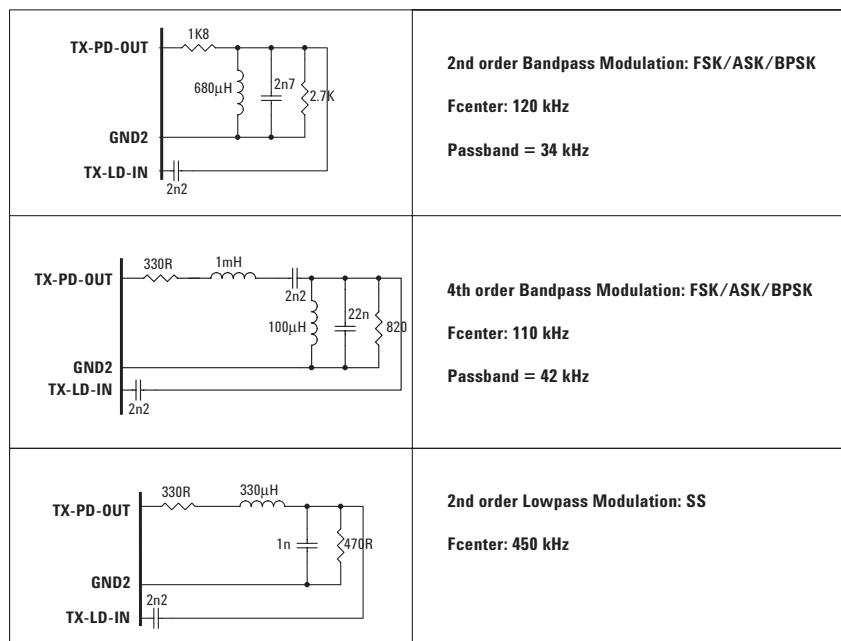


Figure 10. Transmission filters suitable for use with the Agilent HCPL-800J.

Line Driver

The advanced circuit design used in the HCPL-800J line driver stage ensures low distortion even for signal levels approaching the supply rail voltage. However the line driver has a fixed gain of 2, and the nominal output voltage at Tx-pd-out is 3.3V, so the output will saturate unless some attenuation is placed between Tx-pd-out and tx-in. The inherent attenuation of the filter circuit applied between Tx-pd-out and tx-in should normally suffice. To ensure low distortion operation it is recommended to keep the output signal less than $4 V_{pk-pk}$.

On the data sheet the test frequency used to rate the maximum distortion level is 132 kHz. Operating the line driver at higher frequencies will result in an increase in signal distortion due to the roll-off in the loop gain of the line driver with frequency.

This effect can be counteracted by increasing the biasing current in the line driver, which extends the loop gain roll-off frequency. This is achieved by reducing the value of the external resistor connected to pin 9. The trade off is that increasing the biasing current reduces the power efficiency of the line driver, since all biasing current is dissipated internally in the IC.

Load Detection

The unique load detection circuit in the HCPL-800J supplies an isolated feedback signal, which provides an instantaneous representation of the actual load conditions. This feature can be utilized in a wide range of configurations to enhance the modem performance.

Tx Output Current Limiting

Typically the peak output current is purposely limited by a minimum series coupling impedance. However this results in some signal attenuation during nominal load conditions.

A possible solution to avoiding this attenuation is to keep the coupling impedance as small as possible and limit the peak current by progressively reducing the output signal level. The load detection feature provides a convenient feedback signal, which may be used to create such a closed loop control system.

V_{cc2} Power Monitoring

The average power supply requirements for V_{cc2} are significantly lower than the peak power requirements. To reduce the size and cost of the power supply one solution is to use a large decoupling capacitor ahead of the V_{cc2} power supply regulator. This capacitor effectively provides a power reserve to supplement the limited capabilities of the power supply during peak load conditions.

The peak load on V_{cc2} occurs during transmission and is primarily dependent on the power line conditions. This means that the amount of time the reservoir capacitor can maintain a steady V_{cc2} supply is dependent on the length of transmissions and the power line load. The instantaneous load detection feedback signal allows the transceiver to calculate the power supply requirements at the start of transmission during the preamble phase, thus making it possible for the transmitter to determine an appropriate maximum packet length.

Adaptive Data Transmission

Another potential use of the load detection feature is to detect instances where there is a greater risk of poor signal reception at the receiver. For example, if the power line impedance at the transmitter is very low, the signal attenuation contribution at the transmitter stage can be very high and the likelihood that the SNR at the receiver is significantly reduced is increased. In such cases the data link performance could be compensated for by changing to a more robust transmission mode which, for instance, might simply involve a reduction in data rate.

Coupling Circuit

The coupling circuit (Fig. 11) basically acts as a bandpass filter for coupling signals on and off the power-line. In the transmit mode the coupling circuit must sufficiently block the 50/60 Hz power frequency from affecting the correct operation of the line driver output stage. More importantly, though, it must provide a low impedance signal path for the communication signal.

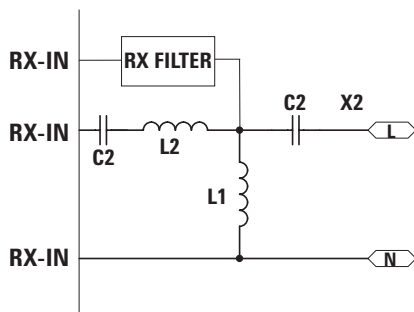


Figure 11. DAA-to-power line coupling circuit.

In receive mode good signal coupling is less critical, but the rejection of the 50/60 Hz harmonics and noise is important.

In practice a wide range of component values are capable of adequately rejecting 50/60 Hz harmonics. However there are some other factors, which help decide appropriate values and component types for the coupling circuit.

Choice of Capacitor C2 and Inductor L2
Capacitor C2 is continuously subjected to the full power line voltage, and to frequent power surges and transient overvoltage conditions. The type of capacitor normally used is a high voltage polyester metalized film type, with many equipment regulations stipulating that this capacitor should be X2 rated.

One of the primary considerations in selecting a value for C2 is to ensure low coupling impedance in the communication band. To minimize the series reactive impedance of the coupling capacitor, it is recommended that an inductor L2 be inserted in series with C2, forming a bandpass filter.

Inductor L2 should be high-Q (low resistance) with an appropriate peak current rating. In terms of coupling efficiency under a wide range of load conditions, all applications will benefit from a coupling circuit with as low a Q as possible. The lower the Q factor the larger C2 needs to be. There is, of course, an upper limit on the value of C2 due to the size and cost of the coupling capacitor.

Sometimes in legacy AFE applications the coupling circuit is purposely scaled to create a high-Q coupling circuit to assure compliance with emissions regulations by filtering out-of-band harmonics caused by nonlinearities in the line driver circuit. Coupling capacitors of 47 nF or lower are normally used. Unfortunately this type of filtering significantly compromises coupling efficiency.

The X2 capacitor type used for C2 is commonly available with $\pm 20\%$ tolerance. Such a large potential variation in capacitance can cause considerable variation of the coupling impedance in a real-world manufacturing environment. It is recommended that a coupling capacitor greater than 150 nF be used with the HCPL-800J, even if the original AFE circuit used a smaller value.

Inductor L1

When selecting the type of inductor for L1 the first point that needs to be considered is that it has 50/60 Hz current flowing through it. This means that a bobbin-type inductor is preferable to one wound on a ferrite toroidal core to prevent saturation. In terms of the parasitic resistance, it is preferable to select a low-cost inductor with a relative high series resistance in the 10 to 30 Ω range. The reason is that the L1/C2 combination has a tendency to ring when hit with impulse noise from the power line. The parasitic series resistance helps to damp out such ringing.

The actual value of the inductor does not significantly affect the ability of the coupling circuit to filter out 50/60 Hz harmonics. In practice the series resistance of the inductor limits the efficiency of the 50/60 Hz filtering. Furthermore very high rejection of 50/60 Hz is in fact not required since the receiver filter stage supplements the filter operation.

The inductance value does significantly affect other parameters, in particular the input impedance. The ideal situation would be to keep the input impedance as high as possible to minimize the loading effect when there are a number of receivers in use. However a trade off is the amount of noise picked up at the receiver. A suggested compromise is to design for an input impedance of around 100 to 200Ω at the communication frequency.

Transmission Mode Power Line Termination

During transmission mode the power line is effectively terminated with an impedance that is equal to the series impedance of the coupling circuit and the output impedance of the line driver. This low impedance path is not only beneficial to reducing signal coupling losses, it also helps to attenuate noise in the transmission frequency band, increasing the SNR at the signal injection point.

However a side effect is that it may, in theory, also attenuate other power-line signals occupying adjacent communication bands. In most practical systems the series resistance and the Q factor of the L2/C2 coupling cir-

cuit limits the minimum impedance observed at other communication bands. However in Europe this requirement is more specific in that EN50065-7 specifies that the minimum impedance at other PLM frequency bands should be greater than 5Ω.

HCPL-800J Surge Protection

To protect the HCPL-800J from transients on the power line some external protection components are required.

The minimum level of protection required is the use of a transient voltage suppression avalanche diode, such as Transil or Transorb, connected across the filter inductor L1. Connecting Schottky clamp diodes at tx-out (pin 15) will provide additional protection, if required.

To further enhance surge protection, particularly for very large voltage spikes, a MOV— (metal oxide varistor) should be connected across the power line terminals. For safety reasons it is important that the correctly rated fuse is used together with the MOV—appropriate fuse ratings may be found in manufacturer's data sheets.

Power Supply Requirements

The power supply requirements of a typical powerline modem are not insignificant, often adding considerably to the overall cost and complexity of the complete application. One of the biggest areas of power consumption is the transmitter circuit. The high efficiency of the line driver used in the HCPL-800J typically reduces the power consumption to just a third of that normally required by conventional designs.

When designing a suitable power supply, the influence it might have on the performance of the modem needs to be considered. Switched power supplies are particularly problematic.

Switching power supplies typically operate in frequency ranges from 30 kHz up to several hundred kHz. So, if care is not taken, the power supply switching frequency or its harmonics can result in disturbances within the communication band. Including a filter can, of course, reduce the switching harmonics of the power supply, but the irony of this is that if care is not taken the filter could also attenuate the modem signals, degrading communication performance. Linear power supplies, in contrast, are benign and, if at all possible, they would be the preferred choice.

V_{cc2} Power Supply Requirements

The inherent galvanic isolation of the AFE circuit requires that the V_{cc2} supply voltage rail be referenced to the line and neutral, as opposed to the V_{cc1} supply, which needs to be galvanically isolated from the L/N.

The V_{cc2} supply can be obtained from a winding on the primary side of the application's power transformer. An alternative to sourcing the V_{cc2} supply from the application power supply is to implement a self-contained supply, which is totally decoupled from the application power supply.

A couple of advantages of decoupling the V_{cc2} power supply are:

1. The powerline modem subjects the application power supply to large transient load demands when switching from receive mode to transmit mode. This places great demands on the power supply regulation.
2. If a separate V_{cc2} supply is provided for the powerline modem, this reduces the power consumption from the appliance's main power supply to just 20 mA at 5 V. This is particularly beneficial in instances where the powerline modem is to be added to an existing appliance with limited power supply capabilities.

The simplest, lowest-cost method of obtaining a standalone V_{cc2} supply is to use a line-voltage-rated capacitor to step down the line voltage, as shown in Fig. 12.

Inductor $L5$ is required to prevent attenuation of the communication signals. Coupling capacitor $C10$ is required between $Gnd2$ and the neutral connection to provide a continuous low-impedance return path for the modulated transmit and receive signals. The protection diode $TV2$ is added to limit the peak current through the diode bridge during a surge or transient event. This diode may be omitted depending on the maximum expected surge voltage and the peak current ratings of the bridge.

The current capability of the power supply is effectively set by the value of capacitor $C2$. For example with a 230 V, 50 Hz supply a 2.2 μF capacitor would be capable of supplying up to 150 mA of continuous current which is equivalent to a continuous peak-peak output current level of 0.65 A_{pk-pk} in transmit mode.

Because of the master/slave half-duplex communication protocol, each power line modem only transmits for short periods. The consequence is that the average supply current is much lower than the peak current requirements.

One way of meeting the peak and average supply requirements using a smaller high voltage capacitor is to increase the size of the storage capacitor $C12$ and increase the voltage rating of the zener clamp diode $ZD1$.

Another possible variation is to use the same high voltage capacitor used to couple the transmit/receive signal onto the power line to also provide the power source for V_{cc2} as shown in Fig. 13. To accomplish this, the value of the coupling capacitor normally needs to be increased, however the end result is a very cost-effective power supply requiring the addition of just

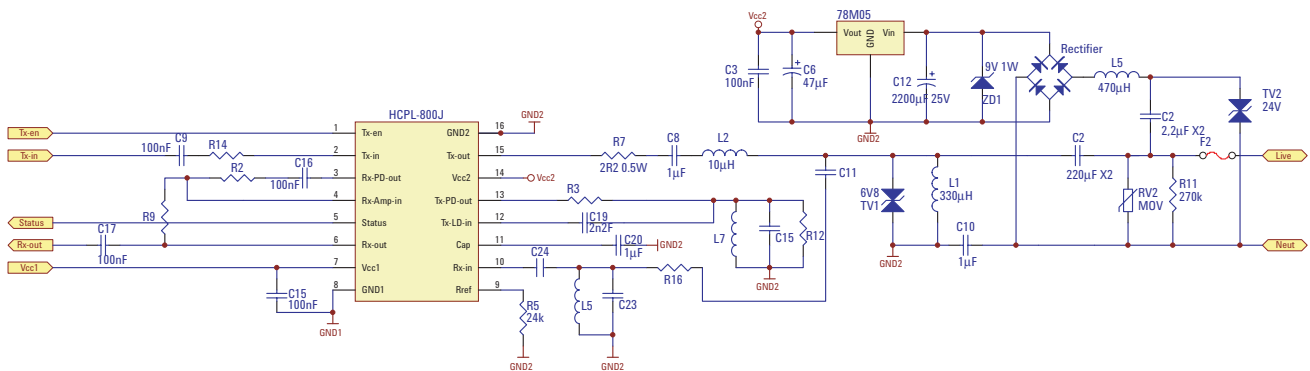


Figure 12. Obtaining V_{cc2} using a capacitor to step down the line voltage.

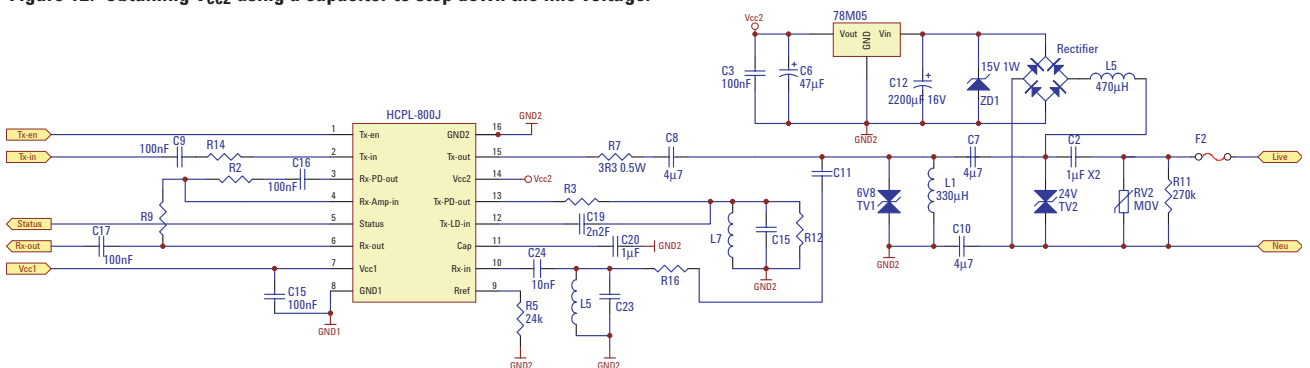


Figure 13. The high voltage capacitor used to couple the Tx/Rx signal onto the power line may also provide the power source for V_{cc2} .

seven components (ZD1, bridge rectifier, avalanche diodes TV1 and TV2, capacitors C10 and C7, 5V regulator).

Receiver Filter

To utilize as much of the receiver's input dynamic range as possible, the received signal from the power line should be filtered to remove as much out-of-band noise as possible.

The transmission characteristics of the power line causes the noise amplitude to increase with decreasing frequency (Fig. 4), so particular emphasis should be placed on the filter performance at lower frequencies. High-amplitude high-frequency noise can also appear on the power line, but such signals are normally caused by RF transmissions, and the very high common-mode noise rejection of the HCPL-800J effectively blocks such signals.

Typically a bandpass filter arrangement is used. Ideally the filter passband should be as steep as possible, however if the filter is made too steep, component variations and temperature drift could result in significant attenuation of the received signal.

The impulse response of the filter also needs to be considered. The frequent occurrence of impulses on the power line could cause the filter to ring excessively every time an impulse is received. During this ringing period, the receiver is blocked. Ideally any ringing of the filter should not extend much beyond the duration of the impulse.

Figure 14 shows two typical filters, a second-order Butterworth filter for a 120 kHz FSK application and a fourth-order 100-450 kHz Butterworth filter for a SS application.

Receiver Gain and Dynamic Range

For demodulators using analog-to-digital converters and digital signal processing, the receiver's input sensitivity is primarily determined by the effective resolution of the ADC.

To extend the dynamic range, signal gain is typically applied in the analog signal path. The HCPL-800J includes a fixed 20 dB gain in the receiver for this purpose. This can further be increased through the use of the uncommitted op-amp located between pins 4 and 6. The maximum theoretical gain that can be usefully applied is only limited by the noise floor of the receiver front end.

The input-referred noise floor of the HCPL-800J is around 25 nV/√Hz. With a receiver bandpass of 30 kHz, this is equivalent to a noise floor (at pin 3) of approximately 32 dBμV. The peak output signal at pin 3 is approximately 1.3 V (113 dBμV). So in this case the maximum theoretical

dynamic range of a system utilizing the HCPL-800J is around 100 dB. This calculation assumes that the fixed 20 dB gain on the front of the HCPL-800J receiver circuit is used.

The receiver front-end gain is fixed with an internal op-amp feedback arrangement as shown in Fig. 15, so the receiver gain may be reduced by increasing the series impedance to pin 10. However reducing the fixed gain will not significantly reduce the noise floor measured at pin 3.

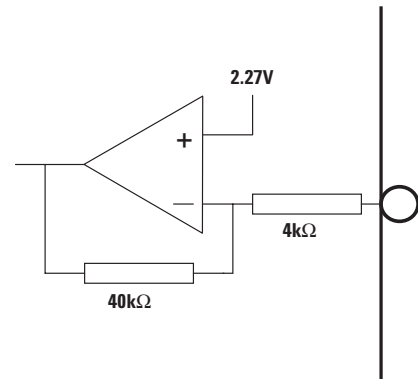


Figure 15. Receiver front end amplifier.

	<p>Modulation: SS</p> <p>Passband = 100–450 kHz</p>
	<p>Modulation: FSK/ASK/BPSK</p> <p>Fcenter: 105 kHz</p> <p>Passband = 55 kHz</p>

Figure 14. Two typical receiver filters: (a) Second-order Butterworth filter for a 120 kHz FSK application, and (b) Fourth-order 100-450 kHz Butterworth filter for a spread-spectrum application.

So if the receiver front-end gain is reduced to 10 dB the dynamic range would be effectively reduced to 90 dB. For this reason it is recommended that the fixed 20 dB gain in the HCPL-800J receiver front end is fully utilized if maximum dynamic range is required, and the overall system gain reduced using other means, if required.

P.C.B Layout Recommendations

To maintain optimum signal isolation and insulation capabilities, maximum distance should be maintained between input and output sides of the circuits, and no ground plane or traces should pass directly below the HCPL-800J. To minimize injected noise level in the receiver path, connections to the high impedance receiver inputs (pins 3 and 10) should be kept as short as possible (Fig. 16).

To minimize line driver distortion, the V_{cc2} decoupling capacitors should be positioned as close as possible to the supply input pin and interconnections should be as kept as short and wide as possible.

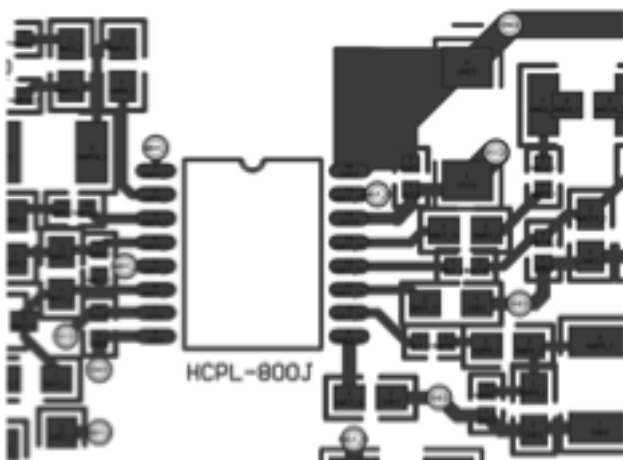


Figure 16. PC board layout recommendations.

The external R_{bias} input pin is very high impedance so it is particularly susceptible to external noise pickup, which can distort or add noise to the line driver output signal. The external resistor connected to this pin should be positioned as close as possible to the input pin, and large amplitude signals should not be routed in close proximity to this pin or its interconnects.

The position and orientation of inductors used in the external filters can affect the performance of filters and result in noise pickup. In particular mutual-inductance effects need to be considered. Ideally inductances should not be placed in close proximity to each other: if this cannot be avoided they should be placed at 90-degree orientation to each other. Other scenarios to avoid are magnetic pickup from sources such as switch-mode power supply transformers.

Output IC Power Dissipation

The output IC is the most power-intensive in the HCPL-800J and, in terms of meeting the maximum junction temperature limit of $+125^{\circ}\text{C}$, it is this area which requires special attention.

In the very unpredictable power line environment the worst-case load condition has to be assumed to be a virtual short circuit. In this case the series resistance of the coupling impedance limits the peak output current, so the worst-case load condition is equal to the series coupling resistance. The power dissipation in the output IC is simply the input power minus the power dissipation in the series resistance of the coupling circuit. The power dissipation in the total series resistance is simply calculated as $I_{\text{rms}} \times 2R$. With FSK, BPSK and ASK communication systems the output waveform is sinusoidal, so it is relatively straightforward to calculate the RMS output current.

Example

An FSK modem transmits an output signal of $3.5 V_{\text{pk-pk}}$. The coupling circuit has a total series resistance of 3.5Ω . The output signal is $3.5 V_{\text{pk-pk}}$ and the maximum ambient temperature in the modem is $+70^{\circ}\text{C}$.

The series resistance effectively limits the peak output current to $1 A_{\text{pk-pk}}$ or $354 \text{ mA}_{\text{rms}}$.

The power dissipation in the total external series resistance is $0.354 \times 2 \times 3.5 = 0.44 \text{ W}$.

V_{cc2} input power = $0.2 \times 5 = 1 \text{ W}$

Output IC power dissipation = $1 - 0.44 = 0.56 \text{ W}$.

The thermal gradient between ambient and the junction of the output IC = $0.56 \times 85 = +47.6^\circ\text{C}$ ^[1]

So operating the line driver continuously in this configuration at an ambient temperature of $+70^\circ\text{C}$ would result in an output IC temperature of $+117^\circ\text{C}$ which is below the maximum allowable temperature of $+125^\circ\text{C}$.

Note 1: When calculating the output IC junction temperature strictly speaking the influence of the thermal heating of the power dissipation in the input IC should also be considered. However the power dissipation in the input IC in transmit mode is just 100 mW ($I_{cc1} = 20\text{ mA}$). So disregarding its influence will result in less than 4°C error, which can generally be regarded as insignificant, unless the part is being operated very close to its limit.

Frequently Asked Questions

Question: *The line driver circuit used in the modem demonstration kit we are evaluating generates a peak output signal level of $6 V_{pk-pk}$ and has a peak output current capability of $1 A_{pk-pk}$. Although the HCPL-800J is capable of matching the peak output current capabilities its peak output voltage level is limited to $4 V_{pk-pk}$. If I use the HCPL-800J am I likely to see any noticeable degradation in communication performance with high amounts of noise, long distances or heavy signal attenuation?*

The simple answer to this is no.

Most powerline modems have input dynamic ranges of at least 60 dB, so heavy signal attenuation alone will not affect communication performance. In

reality the primary limitation on powerline modem performance is not the absolute level of the signal observed at the receiver but rather the SNR at the receiver. This means that an increase in transmission signal level is only beneficial if it improves the SNR at the receiver.

The actual signal power coupled onto the power line is dependent on the coupling impedance and the power line load impedance.

The coupling impedance is effectively set by the peak output current capabilities of the line driver. To limit the peak output current under worst-case power line conditions, the 6 V line driver will require a minimum output impedance of 6Ω whereas a 4 V driver limited to a peak output current of 1A will require a output impedance of 4Ω .

If we assume a typical power line impedance of 10Ω , the actual signal coupled onto the power line by the 6V and the 4V driver would be 3.75V and 2.85V, respectively. As mentioned previously, it is the SNR that is more important than the absolute signal level.

The noise level measured at the transmission injection point will depend not just on the noise source amplitude but also the interaction of the power-line load impedance, the noise source impedance and the modem coupling impedance.

Since in this case the coupling impedance is lower than the power line impedance, and in most cases the noise source impedance, the noise at the transmission injection point will be attenuated by the presence of

the modem operating in a transmission mode. Furthermore since the coupling impedance required for the 6V driver is higher than that for the 4V driver, the noise level measured with the 4V driver will be lower. So, in many cases there is little to no improvement in the effective SNR at the transmitter injection point.

Of course the actual position of the noise source relative to the receiver will affect the SNR measured at the receiver. But even in the worse case scenario, with a low impedance noise source positioned right next to the receiver, the improvement in SNR would be just 2.4 dB.

Furthermore if we consider a condition where the power-line impedance is very low, e.g. 1Ω , since the power line load impedance is significantly lower than the coupling impedance, the presence of the modem operating in transmission mode has little effect on the noise level measured at the transmission injection point. However the coupled signals on the power line have been significantly attenuated to just 0.85V and 0.8V respectively. The 6V line driver in this case provides a SNR advantage of just 0.5 dB.

It can clearly be seen that the main limitation on improving the SNR with a large amplitude transmission signal is the peak output current of the line driver and the associated coupling impedance. An increase in output swing at the line driver output is only really useful if it comes with a similar increase in output current.

Question: *The current AFE I am using uses a transformer together with a differential receiver stage. The manufacturer of our modem transceiver recommends that I make use of the differential input stage, to reduce the influence of common-mode noise on the powerline modem. The HCPL-800J does not use a differential receiver front end. Am I likely to see a reduction in rejection of common-mode noise?*

The manufacturer of your transceiver IC is right. The use of a differential input stage does help to reduce the effect of common-mode noise. Transformers are very poor signal isolators even at relatively low frequencies. At 130 kHz a typical PLM transformer will only provide around 35-45 dB of common-mode noise rejection when used in a single-ended receiver operation. The use of a differential input stage can, in theory, completely compensate for the poor isolation-mode rejection of the transformer. However in practice the improvement is limited by the matching of components used in the differential amplifier and the longitudinal

balance of the transformer. This effectively limits the improvement in isolation mode rejection to around 35 dB.

So in the best case the combined isolation mode rejection of the transformer and differential receiver circuit is going to be around 80 dB.

The HCPL-800J uses optical isolation, which provides much higher signal isolation than a transformer. Even without a differential input stage the HCPL-800J provides an isolation mode rejection of 80 dB at 130 kHz.

In practice the HCPL-800J with a single-ended receiver stage matches the performance of a transformer-based AFE with a differential receiver stage.

Question: *Is it possible to use the HCPL-800J in a duplex communication mode, operating the receiver and transmitter channels simultaneously?*

No. The transmitter and receiver paths in the HCPL-800J cannot be operated simultaneously.

Question: *Is the HCPL-800J compatible with 3.3V transceiver ICs?*

Yes, the HCPL-800J is compatible with 3.3V transceiver ICs. The input signals are fully compatible with 3.3V analog or digital input levels. The output signals may require additional

clamping or divider circuits to prevent electrical overstress of the inputs of the transceiver.

Question: *What effect will light output degradation in the LEDs used in the HCPL-800J have over the operational lifetime of my product?*

In addition to the same wear-out/failure mechanisms experienced by all semiconductor devices, LEDs are affected by an additional aging process, which causes their light output to diminish over time. The type of LED used in the HCPL-800J has a relatively small amount of light degradation with time and has been successfully used in many of Agilent's optoisolators for over a decade.

The transmit path used in the HCPL-800J also has an AGC circuit which compensates for any reduction in light output.

The receiver path, however, does not have any compensation circuit. As a general rule Agilent recommends a lifetime design margin of 40 percent on LED light output degradation for all its optoisolator products. A 40 percent degradation in light output in the receiver channel corresponds to a reduction in receiver signal of around 4.5 dB. This amount of change is in most cases not significant to the correct operation of the receiver, so even if the LED light output was to reduce by as much as 40 percent the part would still in fact be functional.

www.agilent.com/semiconductors

For product information and a complete list of distributors, please go to our web site.

For technical assistance call:

Americas/Canada: +1 (800) 235-0312 or (916) 788-6763

Europe: +49 (0) 6441 92460

China: 10800 650 0017

Hong Kong: (65) 6756 2394

India, Australia, New Zealand: (65) 6755 1939

Japan: (+81 3) 3335-8152(Domestic/International), or 0120-61-1280(Domestic Only)

Korea: (65) 6755 1989

Singapore, Malaysia, Vietnam, Thailand, Philippines, Indonesia: (65) 6755 2044

Taiwan: (65) 6755 1843

Data subject to change.

Copyright © 2004 Agilent Technologies, Inc.

July 25, 2004

5989-1242EN



Agilent Technologies