A NEW INPUT-VOLTAGE FEEDFORWARD HARMONIC-INJECTION 
TECHNIQUE WITH NONLINEAR GAIN CONTROL FOR 
SINGLE-SWITCH, THREE-PHASE, DCM BOOST RECTIFIERS 

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Abstract — A new input-voltage feedforward harmonic-injection 
technique for a single-switch, three-phase, discontinuous-
conducton-mode boost rectifier is introduced. With this 
technique, rectification with a low total harmonic distortion 
that meets the IEC555-2 requirements can be achieved. In addition, 
the rectifier shows an excellent transient performance which 
dramatically reduces the rectifier’s output voltage overshoots 
during line-voltage step-up transients. Moreover, by the 
addition of a nonlinear gain-control circuit, the dc gain of the 
DCM boost rectifier at light load is adaptively reduced so that 
the stability of the rectifier at light load is achieved. The 
performance of the proposed injection technique was verified on 
a 6-kW prototype rectifier. 

I. Introduction 
Three-phase, single-switch, discontinuous-conduction-mode 
(DCM), pulse-width-modulated (PWM) boost rectifiers are 
commonly used for three-phase, high-power-factor (HPF) 
applications since their input-current waveform automatically 
follows the input-voltage waveform. In addition, they can achieve 
extremely high efficiencies because the reverse-recovery-related 
losses of the boost diode are eliminated [1], [2]. However, if a DCM 
PWM boost rectifier is implemented with the conventional constant-
frequency low-bandwidth output-voltage feedback control, which 
keeps the duty cycle of the switch constant during a rectified 
line period, the rectifier input current exhibits a relatively large 5th-order 
harmonic. As a result, at power levels above 5 kW, the 5th-order 
harmonic imposes severe design, performance, and cost trade-offs 
in order to meet the maximum permissible harmonic-current levels 
defined by the IEC555-2 document [3]. Recently, robust harmonic-
injection methods for three-phase, DCM boost rectifiers have been 
introduced [4], [5]. These methods reduce the 5th-order harmonic of 
the input current so that the power level at which the input-current-
harmonic content still meets the IEC555-2 standard is extended. 

Since the bandwidth of the output-voltage feedback control loop 
of the boost rectifier used in HPF applications is very low to achieve 
a low THD, the transient response of the control to the line and load 
changes is very slow causing high transient deviations of the output 
voltage with respect to the steady-state value. Due to an output-
voltage overshoot during a step-up transition, power-stage 
semiconductor components with a higher voltage rating are usually 
required to maintain the necessary design margin between the 
maximum voltage stress of the components and their voltage rating. 
Inevitably, higher voltage-rated semiconductor components are 
more expensive and usually more lossy than their counterparts with 
lower voltage ratings. To reduce the transient output-voltage 
overshoot caused by the input-voltage change, it is necessary to 
make the duty cycle of the controller to respond to the input-voltage 
changes instantaneously. This can be accomplished by the 
feedforward control technique described in [6], [7]. In this 
technique, instead of a fixed-slope saw-tooth ramp, a ramp whose 
slope is proportional to the input voltage is used at input of the 
PWM modulator to achieve an instantaneous response of the 
controller to the line-voltage changes. 

Generally, the dc gain of the DCM boost rectifier is inversely 
proportional to the duty cycle of the rectifier. As a result, the output-
voltage-feedback control loop of the rectifier may become unstable 
at light loads because of the increased dc gain [8]. To achieve the 
loop stability at light loads or even at no load, the effect of the 
increased dc gain of the power stage on the control loop gain must 
be compensated. This loop gain compensation can be achieved by 
employing a nonlinear gain control circuit [8]. 

In this paper, a new harmonic-injection technique with the 
feedforward control is introduced. By this technique a low THD of 
the rectifier current can be achieved with an excellent transient 
performance which dramatically reduces the rectifier’s output 
voltage overshoots during line-voltage step-up transients. Moreover, 
by the addition of a simple nonlinear gain control circuit, the dc 
gain of the DCM boost rectifier at light load is automatically 
compensated and the light-load stability of the rectifier is achieved. 

Fig. 1. Conventional, single-switch, three-phase, DCM boost rectifier with 
a harmonic injection technique.
The proposed injection technique was verified on a 6-kW prototype boost rectifier. The measured full-load efficiency of the experimental converter at nominal line of 380 V_(L-L, rms) was about 97%. In addition, the rectifier meets the IEC555-2 harmonic limits in the entire line-voltage range from 304 V_(L-L, rms) to 456 V_(L-L, rms).

II. Brief Review of Previous Harmonic Injection Method for Single-Switch Three-Phase, DC/DC Boost Rectifiers

To meet the IEC555-2 specifications at power levels above 5 kW, the three-phase, constant-frequency, constant-duty-cycle DC/DC boost rectifier needs to be designed either with a higher voltage-conversion ratio \( M = \sqrt{3} V_p / V_{in} \) (i.e., higher output voltage \( V_{out} \) compared to the peak input phase-neutral voltage \( V_{in} \)) or with a control which employs a harmonic-injection technique.

Generally, for a given line voltage, a larger \( M \) requires a boost switch with a higher voltage rating because of an increased voltage stress. On the other hand, the harmonic-injection approach does not increase the voltage stress of the boost switch, and requires only a few additional components for its implementation. Figure 1 shows the block diagram of the robust, simple injection technique introduced in [5]. In this technique a voltage signal which is proportional to the inverter component of the rectified, three-phase, line-to-neutral input voltages is injected into the output-voltage feedback loop. The injected signal varies the duty cycle of the rectifier within a line cycle in order to reduce the 5th-order harmonic and improve the THD of the rectifier input current.

Various circuit implementations of this injection technique were described in [5]. The simplest implementation of the harmonic injection circuit and its key waveforms are shown in Fig. 2. In the implementation in Fig. 2, the three-phase line voltage is first rectified by three-phase bridge rectifier BR, and then attenuated by the resistive voltage divider \( R_A \cdot R_B \). The scaled-down line voltage developed across \( R_A \), \( V_{in} \), is then inverted by difference amplifier OP1 before it is processed through the high-pass filter \( C_5 \cdot R_5 \) to remove the dc component of \( V_{in} \) and generate injection signal \( V_{inj} \). Finally, \( V_{inj} \) is injected in the circuit in Fig. 1 at point A through summing resistor \( R_1 \).

The ratio between the peak-to-peak value of injected signal \( V_{inj}(p-p) \) shown in Fig. 2(b) and feedback control signal \( V_{fb} \) (error-amplifier output voltage) shown in Fig. 1 defines the modulation index \( m \) as

\[
m = \frac{V_{inj}(p-p)}{V_{fb}(1 - \cos \frac{\pi}{2})}.
\]  

At any given voltage-conversion ratio \( M \), optimal modulation index \( m \) which produces the minimum THD should be determined. Figure 3 shows the calculated values of optimal modulation index \( m \) for the minimum THD (solid line) as a function of \( M \). To maximize the input power of the rectifier at which the IEC555-2 specifications are met, modulation index \( m \) should be determined so that the ratio of the 7th-order harmonic and the 5th-order harmonic is equal to the ratio of the corresponding IEC555-2 limits. It should be noted that the effects of the higher-order harmonics are not significant in comparison with the 5th and 7th-order harmonics. Moreover, the higher frequency harmonics can be easily attenuated by an EMI input filter. Figure 3 also shows the calculated values of optimal modulation index \( m \) for the maximum input power (dashed line) at which IEC555-2 limits are met as a function of \( M \).

To implement an harmonic-injection scheme with a variable modulation index, it is necessary to add a variable-gain amplifier in the harmonic-injection circuit in Fig. 2(a). The implementation of the harmonic-injection technique with a variable modulation index is described in the next section.

Generally, to achieve a low THD, the bandwidth of the output-voltage-feedback control loop of the boost rectifier used in HFP applications is very low. Specifically, it is much smaller than the line frequency. As a result, the transient response of the control to the line and load changes is very slow causing high transient deviations of the output voltage with respect to the steady-state value. To further explain the effect of a low loop bandwidth on the performance of the converter, Fig. 4(a) shows the block diagram of the conventional output-voltage-feedback control. The controller in Fig. 4(a) consists of an error amplifier (EA), PWM modulator,
controller respond to the input-voltage changes instantaneously. This can be accomplished by the feedforward control technique whose block diagram is shown in Fig. 5(a) [1], [2]. In this technique, instead of a fixed-slope sawtooth ramp, a ramp whose slope is proportional to the input voltage is used at input of the PWM modulator. As can be seen in Fig. 5(a), the ramp is generated by integrating a voltage proportional to the input voltage. The input voltage is first sensed and attenuated by voltage divider $R_1 - R_2$ and then inverted before it is brought to the input of the integrator. The integrator is reset at the beginning of each switching cycle by an external fixed-frequency clock signal.

Since in the feedforward control in Fig. 5(a) the duty cycle of the switch is determined by comparing the output voltage of the error amplifier with the ramp whose slope depends on the input-voltage, any change in the input voltage causes immediate (within one switching cycle) change of the duty cycle even if the bandwidth of the voltage loop is very low (i.e., $V_{EA}$ is constant for a short period after the change). As shown in Fig. 5(b), after the input voltage is increased at $t = T_O$, the ramp slope increases causing an immediate decreases of the duty cycle in order to maintain the output voltage constant. Because of the instantaneous change in the duty cycle, the overshoot of the output voltage caused by the input-voltage step is reduced. It should be noted that in Fig. 5(b), the output voltage of constant-frequency sawtooth ramp ($V_{RAMP}$), reference voltage ($V_{REF}$), and a voltage divider ($R_3 - R_4$). In Fig. 4(a), the divider is used to scale down the sensed output voltage $V_O$ so that it can be compared to reference voltage $V_{REF}$ at the input of the error amplifier. The voltage at the output of the error amplifier, which is proportional to the error (difference) between the scaled output voltage and reference voltage, is then compared to the sawtooth ramp voltage at the input of the modulator to generate a signal with a desirable duty cycle to drive the switch. Due to the negative feedback in the voltage loop ($T_V$), the error-amplifier output voltage changes in such a manner that the duty cycle of the converter is modulated so that the output voltage is maintained constant. Compensation impedances $Z_1$ and $Z_2$ of the error amplifier in Fig. 4(a) are used to provide a proper gain, bandwidth, and frequency compensation of the loop so that the loop is stable for all operating conditions.

Figure 4(b) shows the transient responses of the key waveforms of a low-bandwidth output-voltage-feedback control in Fig. 4(a) for a positive line-voltage transient. As can be seen from Fig. 4(b), at $t = T_O$, voltage $V_{IN}$ experiences a positive step change. However, because of the slow control loop, control voltage $V_{EA}$ starts changing slowly some time after $t = T_O$. Since immediately after the voltage change the duty cycle stays unchanged for some time, the output voltage experiences a high transient overshoot.

To reduce the transient output-voltage overshoot caused by the input-voltage change, it is necessary to make the duty-cycle of the

Fig. 4. Conventional output-voltage-feedback control scheme: (a) Block diagram; (b) key waveforms during a step-up input-voltage transient.

Fig. 5. Conventional input-voltage-feedforward control scheme: (a) Block diagram; (b) key waveforms during a step-up input-voltage transient.
the error amplifier, $V_{E,A}$, does not change immediately after the input-voltage change because of the assumed low bandwidth of the voltage loop.

III. New Feedforward Harmonic Injection Method for A Single-Switch Three-Phase, DCM Boost Rectifier

By combining the feedforward control with the harmonic-injection technique, the performance of the three-phase, single-switch, DCM PWM converter can be optimized so that it meets the IEC655-2 requirements with an excellent transient response to the line-voltage changes. Figure 6 shows the block diagram of the implementation of the proposed feedforward control with the harmonic injection.

In the implementation in Fig. 6, the harmonic-injection and the feedforward paths use a common line-voltage and scaling (attenuating) circuit. In the feedforward path, the scaled line voltage, which contains both a dc and a relatively small 360-Hz ac component, is integrated to generate a ramp with a line-voltage-dependent slope, $V_{PF}^RAMP$. This ramp is then added to a constant-slope, constant-frequency external ramp at the inverting input (point B in Fig. 6) of the PWM modulator.

Because the sensed voltage and, therefore, the feedforward ramp contains an ac component proportional to the rectified-line voltage, the feedforward control inherently possesses the harmonic-injection property which helps in reducing the 5th-order harmonic of the line current. However, since the optimization of the amplitude of the ac injection signal for the harmonic reduction, and the optimization of the feedforward ramp slope for the reduction of the line-voltage transients require different integrator gains, it is necessary to separate the feedforward and the harmonic-injection paths, as shown in Fig. 6, so that both paths can be optimized independently.

As can be seen from Fig. 6, in the harmonic-injection path, the sensed, scaled line voltage is first amplified by the variable-gain amplifier (VGA) and then the amplified signal is passed through the high-pass filter to remove its dc component. In the implementation in Fig. 6, the ac component generated at the output of the high-pass filter is integrated to generate ramp $V_{RAMP}_{PF}$ with the slope proportional to the injection signal. Finally, $V_{RAMP}_{PF}$ is summed up with external ramp $V_{EXT}^RAMP$ and feedforward ramp $V_{PF}^RAMP$ at the inverting input of the PWM modulator (point B in Fig. 6).

The circuit diagram of the line-voltage sensing and scaling circuit is given in Fig. 7(a), whereas its output signal $V_S = -V_{g}R_{g}R_{C}$ waveform is shown in Fig. 7(b). It should be noted that in the sensing and scaling circuit in Fig. 7(a) the control ground is isolated from the ground of three-phase input voltages by $R_A = 4\, \Omega$ resistors. As shown in Fig. 7(b), sensed input voltage $V_S$ possesses the information about the peak input voltage and the ac component of the rectified three-phase input voltage which is the desired injection component. Since the proposed injection-signal generator does not contain a bandpass filter, the injection-signal which contains 6th and higher-order harmonics does not suffer from any significant delay. As a result, the phase of injection signal $V_S$ is

![Fig. 6. Block diagram of input-voltage-feedforward control with harmonic injection for a single-switch three-phase DCM boost rectifier.](image_url)
naturally well synchronized with the input currents and line-to-neutral voltages. Moreover, this phase synchronization does not drift with time and it is not very sensitive to component tolerances.

Figure 7(a) also shows the implementation of the feedforward integrator, whereas Fig. 7(b) shows the key waveforms of the modulator in the presence of the feedforward modulation. In the integrator in Fig. 7(a), the integrator capacitor $C_{i1}$ is charged by current $I_{s} = V_{S}/R_{i1}$ during a switching period, and discharged by the clock pulse at the end of the switching period. Due to a short discharge time, the waveform at the output of the integrator is sawtooth ramp $V_{kAMP}$.

As shown in Fig. 7(b), as sensed voltage $V_{S}$ changes, the ramp slope also changes causing a modulation of the boost-switch duty cycle. It should be noted that in Fig. 7(b) sensed voltage $V_{S}$ changes because of the ac component in the rectified line voltage. However, it should be noticed that the purpose of the feedforward path is to improve the transient response of the circuit to line-voltage changes, and not to serve as the harmonic injection path. In fact, with the feedforward integrator in Fig. 7(a), the modulation index of the injected signal cannot be optimized because different integrator gains are required for the optimal feedforward ramp and the optimal modulation index of the injection signal. To optimize modulation index $m$ as a function of $M$, the ac component of sensed input voltage $V_{S}$ should be properly amplified and added to the slope of the feedforward PWM compensation ramp as shown in Fig. 6.

Figure 8(a) shows the schematic diagram of the variable-gain amplifier (VGA) block in Fig. 6, which is used to generate an approximate optimal modulation index of the injected signal. The VGA in Fig. 8(a) is implemented with three Zener diodes with different breakdown voltages, which make the gain of the amplifier to increase as sensed voltage $V_{S}$ increases, as shown in Fig. 8(b).

The VGA circuit in Fig. 8(a) has four distinct regions of operation. When input voltage $V_{S}$ is lower than $8.2 \, V$, which approximately corresponds to the zero line voltage of $304 \, V_{(L-Lrms)}$, output voltage $V_{SAMP}$ of the VGA is nearly zero. When sensed voltage $V_{S}$ is higher than $8.2 \, V$ but lower than $10 \, V$, which corresponds to the nominal line voltage of $380 \, V_{(L-Lrms)}$, the voltage gain of the circuit is approximately $0.09$ (which is the ratio of the $5.1 \, k\Omega$ and $56 \, k\Omega$ resistor in Fig. 8(b)). Similarly, when the sensed voltage is higher than $10 \, V$ but lower than $12 \, V$ which corresponds to high line voltage of $456 \, V_{(L-Lrms)}$, the voltage gain of the VGA circuit is approximately $0.57$ (which is the ratio of the $5.1 \, k\Omega$ resistor and the parallel resistance of the $10 \, k\Omega$ and $56 \, k\Omega$ resistor). Finally, when sensed voltage $V_{S}$ is higher than $12 \, V$, the gain of the VGA is $1.25$. As shown in Fig. 8(b), since the transition of a Zener diode into the avalanche region is not abrupt but gradual, the output-voltage vs. input-voltage curve of the VGA is not piece-wise linear, but is represented by the smooth dashed curve in Fig. 8(b).

Figure 8(a) also shows a schematic diagram of the high-pass filter block in Fig. 6. The high-pass filter consists of blocking capacitor $C_{f}$ and filter resistor $R_{f}$. In the circuit in Fig. 8(a), the dc component of rectified voltage $V_{SAMP}$ is eliminated by the blocking capacitor $C_{f}$. Since the impedance of the blocking capacitor $C_{f}$ at the line frequency is much smaller than $R_{f}$, the voltage across $R_{f}$ is nearly identical to the ac component of $V_{SAMP}$. As a result, the scaled ac component of the rectified three-phase line-to-line input-voltages which contains the $6th$ and higher-order harmonics can pass through the filter without a phase shift.

Finally, Fig. 9 shows the implementation of the nonlinear-slope external-ramp generator. As can be seen from Fig. 9, the external

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**Fig. 8.** Variable-gain amplifier (VGA), the high-pass filter, and the harmonic signal integrator: (a) Schematic diagram; (b) output voltage versus input voltage curve of VGA.
ramp $V_{\text{EXT RAMP}}$ is the exponential waveform which has a nonlinear dv/dt over a switching period. As a result, at a light load when the duty cycle of the DCM boost rectifier is very small, the dv/dt of external ramp $V_{\text{EXT RAMP}}$ is much greater than at the full load. Since the dc gain of the rectifier is inversely proportional to the slope of the ramp signal, the high dc gain of the DCM boost rectifier at light load is automatically compensated.

Figure 10 shows the calculated dc gain of the rectifier without and with the nonlinear gain control as a function of the output power. As can be seen from Fig. 10, with the nonlinear gain control the dc gain is very much reduced at lower power levels.

**IV. Experimental Results**

To verify the performance of the proposed input-voltage feedforward control technique with harmonic injection, a three-phase, 6-kW, 45 kHz, DCM boost rectifier for 304 $V_{\text{rms}}$ - 456 $V_{\text{rms}}$ line-to-line input voltage range and $V_{O} = 750 V_{\text{PC}}$ was built. The prototype boost rectifier was tested without and with the proposed feedforward harmonic injection technique. Figure 11 shows the oscillograms of the input voltage and current waveform of the experimental circuit with and without harmonic-injection control at full power. The measured input-current harmonics of the experimental rectifier with and without harmonic-injection control at full power and at different input voltages are summarized in Figs. 12 and 13, respectively. As can be seen from Fig. 13, the magnitudes of the 5th-order harmonic as well as the higher harmonics are well below the IEC555-2 limit in the entire input voltage range. The minimum THD of 7.5% occurs at the low line, whereas the maximum THD of 14.7% occurs at the high line.

Figure 14 shows the oscillogram of the output voltage response during line-voltage transient test. For the transient-response test, the three-phase input voltage was stepped from 304 $V_{\text{rms}}$ to 456 $V_{\text{rms}}$ within 0.2 ms, and it was kept at 456 $V_{\text{rms}}$ for 400 ms before it was stepped back to 304 $V_{\text{rms}}$ within 0.2 ms at 6 kW and 60 W output power levels. Figure 14 shows the measured waveforms of output voltage $V_{O}$ and input line-to-neutral voltage $V_{\text{IN}}$ of the DCM boost rectifier implementation with and without the feedforward control scheme.

As can be seen from Fig. 14, the proposed input-voltage feedforward harmonic-injection technique significantly reduces the output voltage overshoot. For the rectifier without feedforward control the maximum output-voltage overshoot is approximately 150V, whereas the corresponding overshoot is below 50 V for the implementation with the feedforward control.

Figure 15 shows the key waveforms of the controller. It should be noted that external ramp $V_{\text{EXT RAMP}}$ is an exponential waveform which has a nonlinear dv/dt over a switching period.

As shown in Fig. 16, the dv/dt of ramp signal $V_{\text{RAMP}}$ is about 0.75 V/µs when the pulse width of the gate signal is 0.38 ms for 60 W output power. Since the dv/dt of $V_{\text{RAMP}}$ is about 0.1 V/µs when the pulse width is approximately 6 ms for 6 kW output power, the dc gain of the DCM boost rectifier at 60 W is only 13.3% of the dc gain at full load. Therefore, the high dc gain of the DCM boost

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**Fig. 12.** Measured full-power input-current harmonics of the experimental DCM boost rectifier without harmonic-injection control at different input voltages.

**Fig. 13.** Measured full-power input-current harmonics of the experimental DCM boost rectifier with harmonic-injection control at different input voltages.

**Fig. 14.** Input line-to-neutral voltage $V_{\text{IN}}$ (200 V/div) and output voltage $V_{\text{OUT}}$ (50 V/div) measurements of the experimental rectifier (a) without and (b) with the input-voltage feedforward harmonic-injection circuit during the input-voltage transition (304 $V_{\text{rms}}$ - 456 $V_{\text{rms}}$) at $V_{0} = 750 V$ and $P_{0} = 6 kW$. Time base: 10 ms/div.

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**Fig. 11.** Input voltage $V_{\text{IN}}$ (100 V/div) and input current $I_{\text{IN}}$ (10 A/div) waveforms of the experimental rectifier (a) without and (b) with the input-voltage feedforward harmonic-injection circuit at $V_{\text{IN(LLL)}} = 380 V$, $V_{\text{OUT}} = 750 V$, $P_{\text{OUT}} = 6 kW$. Time base: 2 ms/div.
Fig. 15. Measured key waveforms of the input-voltage feedforward controller. Time base: 5 μs/div.

Fig. 16. Detailed view of key waveforms in Fig. 15. Time base: 1 μs/div.

The rectifier at light load is automatically compensated by the nonlinear gain-control circuit.

Figure 17 shows the measured transient waveforms of output voltage $V_o$ and input line-to-neutral voltage $V_{in}$ of the rectifier with and without the nonlinear gain-control circuit at $P_{in}=60$ W. Since the dc gain of the DCM boost rectifier at light load is extremely high, the output-voltage exhibits an oscillatory response because of a small phase margin as shown in Fig. 17(a). By the addition of the nonlinear gain-control circuit shown in Fig. 9, the dc gain of the control loop at light load is adaptively reduced so that the output voltage response is improved due to an increased phase margin as shown in Fig. 17(b).

The rectifier exhibits the maximum efficiency of 98.1% at the maximum input voltage of 456 Vrms. The minimum efficiency of 96.1% occurs at the low-line voltage of 304 Vrms.

V. Summary

A new harmonic-injection technique with the feedforward control has been introduced. With this technique a low THD of the rectifier current can be achieved with an excellent transient performance which reduces the rectifier’s output voltage overshoots during step-up line-voltage transients. The proposed injection technique was verified on a 6-kW prototype boost rectifier. The rectifier meets the IEEE55-2 harmonic limits in the entire line-voltage range with the maximum THD of 14.7% that occurs at the high line. The full load overshoot of the output voltage for the line step from 304 Vrms to 456 Vrms was below 50 V.

References


Fig. 17 Input line-to-neutral voltage $V_{in}$ (200 V/div) and output voltage $V_{out}$ (5 V/div) measurements of the experimental rectifier (a) without and (b) with the non-linear gain control circuit during the input-voltage transition (304 Vrms → 456 Vrms → 304 Vrms) at $V_{in}=750$ V and $P_{in}=60$ W. Time base: 100 ms/div.