

# Total Compensation of Line-Side Switching Harmonics in Converter-Fed AC Locomotives

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**Abstract** — PWM controlled line-side converters of modern ac traction locomotives inject harmonic currents into the feeding overhead line. This causes problems of electromagnetic interference. Passive or active filters are usually provided for a partial reduction of the line harmonics. A novel and superior approach employs a switched electronic compensator. It generates an exact replica of the harmonic current, feeding it to the high-voltage transformer to produce a harmonic counter mmf. The transformer main flux is then forced to be sinusoidal, and so is the induced voltage in the primary. The line current, being low-pass filtered by the transformer leakage inductance, assumes a pure sinusoidal waveform. The compensator operates in the low switching frequency range of the main power converters. Its installed power is only 1% of the traction power. Operation, control and design considerations are described. Results from laboratory tests at full power level are presented.

## 1. INTRODUCTION

Controlled ac-drives have evolved as a preferred technology in railway traction. The concept offers the advantage of operation in an enlarged constant power range of the torque-speed characteristic. It reduces the size of the traction motor, minimizes maintenance requirements, and improves the power quality in the feeding overhead catenary [1], [2].

The electronic power conversion system of ac-drive locomotives consists of an arrangement of three-phase PWM inverters feeding the traction motors at variable voltage and variable frequency [3]. The associated dc-link circuits are supplied from the overhead line through a group of single-phase power converters, which are also operated in the PWM mode. The line-side converters generate current harmonics which are injected into the overhead line, spreading as traveling waves in either direction from the feeding point [4].

In the frequency range of the characteristic converter harmonics, the overhead supply reacts as a distributed electric wave guide, having various electric discontinuities such as feeding substations, moving traction vehicles, and line bifurcations. Partial reflections of the traveling waves at the points of discontinuity are the causes of multiple and time-varying electrical eigenresonances. Their excitation by the current harmonics of ac locomotives may stimulate electrical oscillations in the line. The resonating currents can be substantially higher in amplitude than the originating inverter harmonics injected by the locomotive [4]. This increases the intensity of electromagnetic interference (EMI) with track-side signaling lines and protection systems [5].

## 2. REDUCTION OF CONVERTER HARMONICS

Different methods have been employed in the past to reduce the harmonic distortion of the pantograph current of electric traction vehicles. The class 120 locomotives of the German railways (DB) and the traction equipment of the German InterCity Express (ICE) employ a tuned harmonic filter which is inserted between the pantograph and the high-voltage winding of the input transformer [3], Fig. 1. The filter consists of a series inductor, a capacitor and a damping resistor. Since the inductor carries the full load current and the capacitor must sustain the high line voltage of 15 kV, the total kVA rating of such harmonic filter is very high. This increases the cost of the traction equipment considerably. Even worse is the high weight of the harmonic filter, which is around 800 kg. It adds in an undesired way to the mechanical stresses in the locomotive body and increases the weight per axle, stressing the rails and their supporting structures. Other problems relate to electrical aspects. Transient overvoltages occur when the pantograph is lifted to connect the filter to the overhead line. The filter also absorbs those harmonics from the line which are generated by other traction vehicles. This increases the filter losses to an unpredictable degree.

Alternative solutions have been considered in an attempt to eliminate some of the disadvantages of a passive harmonic filter. The proposed design of the class 460 locomotive of the Swiss Federal Railways (SBB) makes provision for three-level inverters instead of two-level inverters on the line side [6]. The effective switching frequency is doubled by this approach, and the individual voltage steps at switching are reduced to one half. The penalty is that the number of power semiconductor devices doubles. Cost and complexity increase, and higher order harmonics still persist.

A different approach to the harmonic problem employs adaptive pulsewidth modulation for the feeding power converters [7]. The prevailing resonance conditions in the line are identified in a real-time process, which permits active damping of the line resonances by inverter control [8], [9].

Another alternative for the reduction of converter harmonics is the installation of active filters. This technique is well known in stationary power conversion systems. It is applied in the class 1014 locomotive of the Austrian Federal Railways (ÖBB). A fast-switching power converter is used for this purpose, forming part of a closed loop control system with the fundamental load current as the reference. The method transfers the harmonics of the traction power converters to a higher frequency range, which is determined by the switching rate of the compensating converter. Passive harmonic filters are still

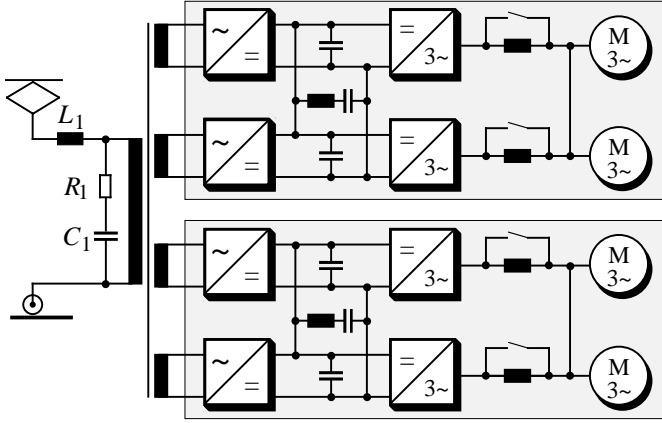


Fig. 1: The electric drive system of the class 120 locomotive

required. They consist of a group of tuned LC circuits, with each circuit being connected to a separate winding of the main transformer. These filters resonate in the higher frequency range given by the switching harmonics of the active filter. The filter components get therefore downsized in comparison to those of a passive filter [10], [11].

This paper describes a novel method which, in contrast to passive or active filters, eliminates the undesired converter harmonics completely. It operates on the principle of feedforward waveform compensation. A power electronic circuit is employed for this purpose, which is controlled at low switching frequency by the firing signals of the main power converter. The compensator is designed as an addition to the power conversion system of the existing class 120 locomotives and ICE high-speed trains.

### 3. TRACTION EQUIPMENT OF CLASS 120 LOCOMOTIVE

The configuration of the traction equipment of the class 120 locomotive is shown in Fig. 1. It comprises two separate and identical traction units which operate independently from each other for the purpose of redundancy. Each unit consists of a dc link circuit, two machine-side PWM inverters, and the respective traction motors. The dc link circuits are supplied from the single-phase ac catenary through the pantograph and the main step-down transformer. Two secondary windings are provided for each traction unit. PWM controlled rectification is performed by two single-phase PWM controlled power converters per traction unit. Pulsations of the dc link voltage at double line frequency, caused by the discontinuous power flow in the single-phase line, are attenuated by tuned series resonant circuits located in the respective dc links.

The four line-side converter bridges are effectively connected in series by virtue of the common step-down transformer. The equivalent voltage in Fig. 3 is a nine-level stepped ac waveform  $u_t$ , acting as the counter emf of the traction converters to the line voltage. A simplified equivalent circuit is shown in Fig. 2, where  $u_L$  is the line voltage and  $i_L$  is the line current;  $L_\sigma$  is the leakage inductance of the main trans-

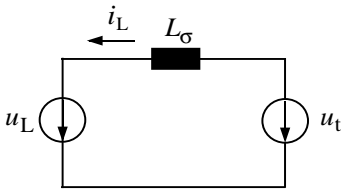


Fig. 2: Equivalent circuit

former, designed for a relatively high value. This inductance serves as a smoothing filter to attenuate the line-side harmonic currents.

Pulsewidth modulation for the line-side converters is performed using the ramp comparison method. Triangular carrier signals having eleven times the line frequency are synchronized with the line voltage. The carrier signals of the four line-side converter bridges are phase shifted by  $45^\circ$  with respect to each other. This eliminates the lower order converter harmonics, provided the load distribution on the four converter units is balanced. In the balanced case, an equivalent carrier of around 1.4 kHz [3] is created, although the switching frequency of the power semiconductors is only 183 Hz.

Load unbalances occur at heavy traction load, for instance when the locomotive breaks off from stand-still. The unbalance occurs because the respective rear axles develop a higher traction torque than the front axles, owing to the higher pressure existing between the rear wheels and the rail. The harmonic cancellation between the line-side converter groups is not complete at unbalanced load, and 700 Hz harmonic currents appear. The 700 Hz harmonics become predominant in the extreme case where one of the two traction units is shut down in emergency because of a system failure. The service is then continued with only two traction motors. The filter must be disengaged to avoid thermal overload, and the line harmonics become excessive [12].

### 4. COMPENSATION BY INVERSE DISTORTION CURRENT

#### 4.1 Basic principle

In the absence of a harmonic filter, the harmonic current  $i_h(t)$  in the line assumes a typical waveform as shown in Fig. 4 over a fundamental period. The novel concept of this paper eliminates the distortion of the line current by injecting the inverse of the harmonic current waveform into an auxiliary winding of the main transformer. The analysis leads to the equivalent circuit shown in Fig. 5. The transformer is modeled by three leakage inductances:  $L_{\sigma L}$  is associated to the leakage fluxes of the primary winding, and  $L_{\sigma m}$  accounts for the leakage fluxes of the main secondary windings which feed the traction con-

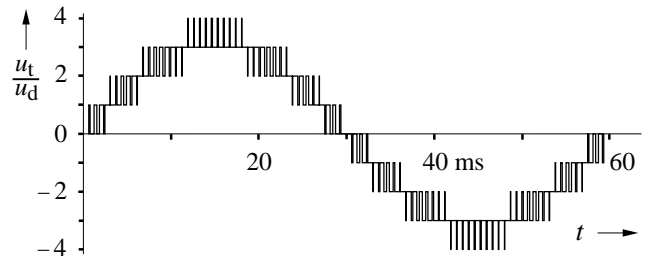


Fig. 3: The equivalent traction converter voltage  $u_t$

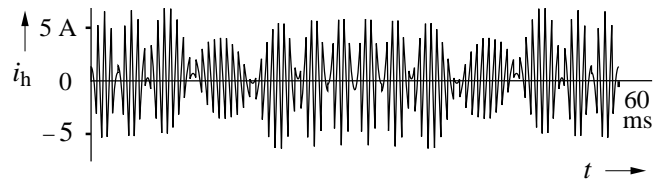


Fig. 4: Waveform of the harmonic current

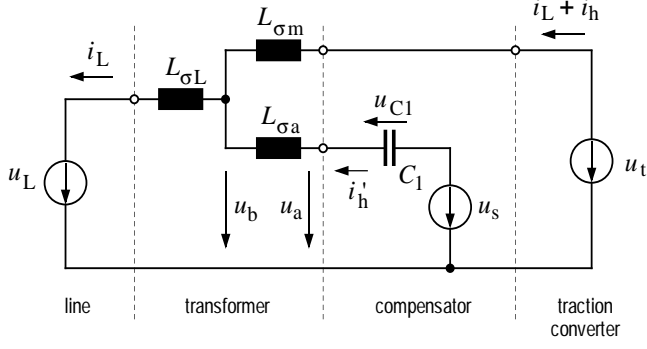


Fig. 5: Equivalent circuit, main transformer and compensator

verters. This inductance forms the major portion of the total transformer leakage so as to decouple the individual converter units from each other and to attenuate the circulating harmonics currents between them. Contrasting to that, the auxiliary winding of the compensator is closely coupled to the transformer main flux, and hence the third inductance  $L_{\sigma a}$  in Fig. 5 assumes a low value. The magnetizing current of the transformer is neglected in the equivalent circuit.

The voltage  $u_t$  represents the equivalent reaction on the transformer secondary of the four line-side converters. The primary winding is directly connected to the pantograph, saving additional power equipment on the high-voltage side. The transformer receives directly the line voltage  $u_L$ , which is sinusoidal in principle. To render the line current  $i_L$  is sinusoidal, the back emf  $u_b$  of the transformer must be sinusoidal, too.  $u_b$  is induced in the primary winding by the transformer main flux.

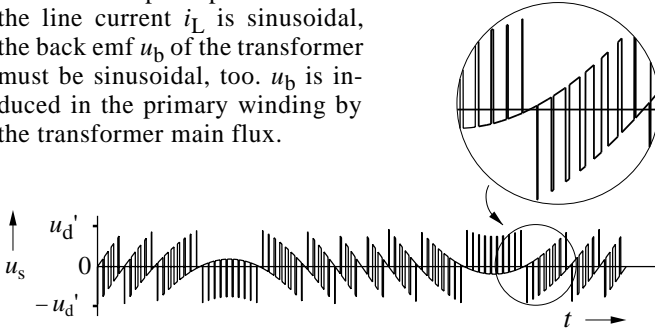


Fig. 6: The compensator voltage  $u_s$  (inset enlarged)

Since the equivalent voltage  $u_t$  of the traction converters contains high harmonic components, the equivalent current  $i_L + i_h$  through the traction converters gets distorted. It is characterized by the harmonic component

$$i_h(t) = \frac{1}{L_{\sigma m}} \int_0^t [u_t(t) - u_b(t)] dt + i_h(0). \quad (1)$$

The effect of this harmonic current on the primary side is annihilated by a compensator which forms part of the equivalent circuit Fig. 5. The compensator consists basically of a switched voltage source  $u_s(t)$  which injects a high-frequency current through a series capacitor  $C_1$  into the auxiliary transformer winding:

$$i'_h(t) = \frac{1}{L_{\sigma a}} \int_0^t [u_s(t) - u_{C1}(t) - u_b(t)] dt + i'_h(0) \quad (2)$$

For perfect harmonic cancellation  $i_h(t) = i'_h(t)$ , and hence from (1) and (2)

$$u_s(t) = -\frac{L_{\sigma a}}{L_{\sigma m}} [u_t(t) - u_{C1}(t)]. \quad (3)$$

The value of  $C_1$  is designed such that the capacitor voltage is almost not affected by the high-frequency ac current  $i'_h(t)$ . The resulting voltage  $u_{C1}$  is then sinusoidal and of fundamental frequency. On the other hand,  $u_s(t)$  in (3) is supposed not to contain a fundamental component. Hence

$$u_s(t) = -\frac{L_{\sigma a}}{L_{\sigma m}} u_{th}(t), \quad (4)$$

where  $u_{th}(t) = u_t(t) - u_{t1}(t)$  represents the harmonic content of the equivalent converter voltage  $u_t$ , and

$$u_{C1}(t) = -u_{t1}(t). \quad (5)$$

The required waveform of the harmonic compensation voltage  $u_s(t)$  is shown in Fig. 6. A comparison with Fig. 3 shows that this waveform can be made up from the same switching pattern that is used for the control of the traction converters.

For a more detailed design of the compensator circuit, the condition (5) must be satisfied.

#### 4.2 The compensator circuit

The harmonic compensation voltage  $u_s(t)$ , Fig. 6, can be decomposed into a pulsed rectangular waveform  $u_{B1}$ , shown in Fig. 7(a), and a piecewise sinusoidal voltage  $u_{B2}$ , Fig. 7(b). These voltage components are generated by two single-phase bridges B1 and B2 in series connection, Fig. 8. The gate signals  $g_1$  and  $g_2$  that control the respective bridges B1 and B2 are derived from the gate signals of the traction converters on the line side.

The bridge B1 operates from a constant dc source  $u'_d$  and generates the rectangular waveform  $u_{B1}$ . The bridge B2, seen

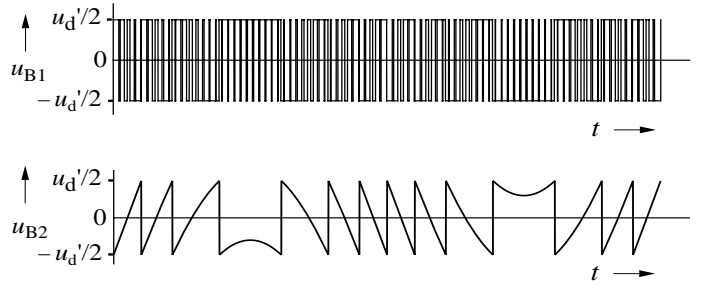


Fig. 7: Components of the compensator voltage  $u_s$ ; (a) rectangular waveform  $u_{B1}$ , (b) piecewise sinusoidal waveform  $u_{B2}$

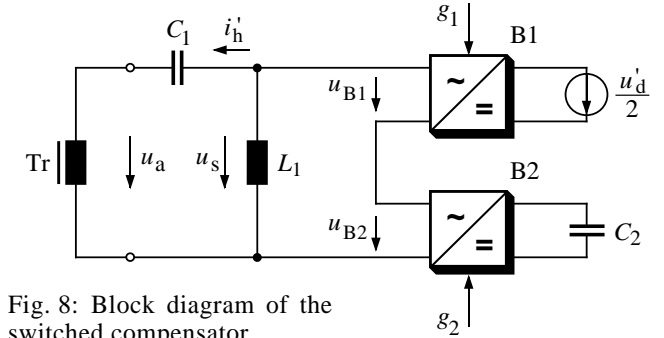


Fig. 8: Block diagram of the switched compensator

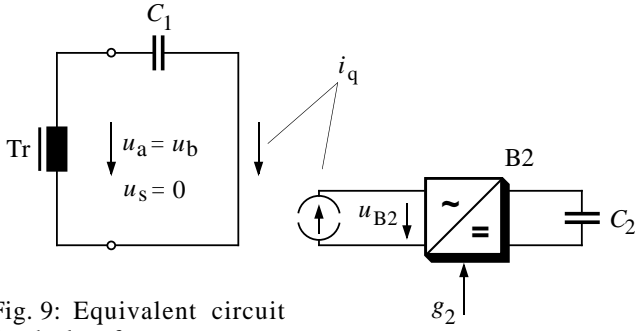


Fig. 9: Equivalent circuit for the low frequency range

in the lower portion of Fig. 8, operates in a different manner: it serves as a reversible electronic switch for the capacitor  $C_2$ . The capacitance of  $C_2$  is small, such that the capacitor voltage  $u_{C2}$  can assume the required waveform.

The two bridges generate the inverse harmonic current  $i_h'$ . This current flows through the capacitor  $C_1$  and is linked with the main flux of the transformer by the auxiliary winding.

The parallel inductance  $L_1$  is fairly large and hence the current  $i_{L1}$  is small. This makes  $L_1$  a low power component. It serves two purposes: i) it keeps the voltage across the two bridges B1 and B2 at a low value in the absence of the respective gate control signals; ii) during operation of the compensator, the current through  $L_1$  stabilizes the voltage across  $C_2$  within prescribed limits.

The operation of the compensator is described next, considering the fundamental and harmonic components separately.

#### 4.3 Performance at fundamental frequency

The fundamental frequency is 16 2/3 Hz. At this frequency, the leakage inductance  $L_{\sigma a}$  can be neglected against the reactance of the capacitor  $C_1$ .

The voltage  $u_s$ , which is forced across the inductor  $L_1$  by the two bridges B1 and B2, turns out to be modulated in pulse amplitude and pulse width; an inspection of the enlarged inset in Fig. 6 shows that the average voltage over a pulse period is always zero; hence  $\bar{u}_s = 0$  at low frequencies, which leads to the equivalent circuit Fig. 9. Interestingly, since an equivalent short-circuit path  $\bar{u}_s = 0$  is established by the two series

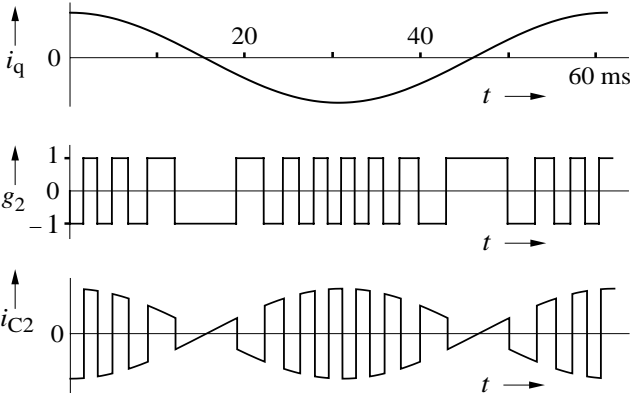


Fig. 10: Generation of a piecewise sinusoidal current waveform; (a) charging current of capacitor  $C_1$ , (b) gate control signal  $g_2$  of bridge B2, (c) charging current of capacitor  $C_2$

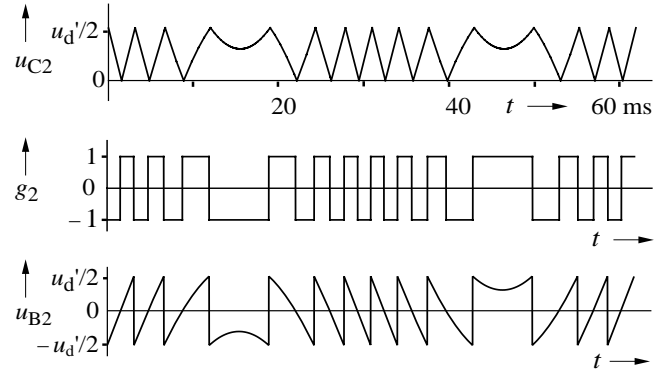


Fig. 11: Generation of a piecewise sinusoidal voltage waveform; (a) capacitor voltage  $u_{C2}$ , (b) gate control signal  $g_2$  of bridge B2, (c) ac side voltage  $u_{B2}$  of bridge B2

connected bridges B1 and B2, the short-circuit current  $i_q$  acts as the charging current for both capacitors  $C_1$  and  $C_2$ . However, independent control of the respective voltages  $u_{C1}$  and  $u_{C2}$  is possible through bridges B1 and B2.

The waveform of the charging current for  $C_1$  is shown in Fig. 10(a). The charging current of  $C_2$  has the same magnitude, but gets reversed in certain time instants as commanded by the gate control signal  $g_2$  of bridge B2. The signal  $g_2$  is shown in Fig. 10(b), and the resulting charging current  $i_{C2}$  is shown in Fig. 10(c).

Fig. 11(a) displays the voltage across  $C_2$ . This waveform builds up as the integral of the current  $i_{C2}$ , as shown in Fig. 10(c). The voltage  $u_{C2}(t)$  is reflected to the input of bridge B2 as the waveform  $u_{B2}(t)$ , Fig. 11(c), having its polarity changed as commanded by the gate control signal  $g_2$  shown in Fig. 10(b).

#### 4.4 Performance at higher frequencies

The equivalent circuit of the compensator Fig. 8 works out to be different in the frequency range of the traction converter harmonics. The lower order harmonics range around 88 times the fundamental frequency, which permits neglecting the impedance of  $C_1$  against the transformer leakage reactance  $L_{\sigma a}$ . This leads to the equivalent circuit Fig. 12, in which the harmonic compensation voltage  $u_s$  appears across the transformer auxiliary winding.

According to Fig. 6, the compensating voltage  $u_s$  exhibits a voltage step at every commutation of the traction converters. Each step in the compensating voltage is opposed in direction

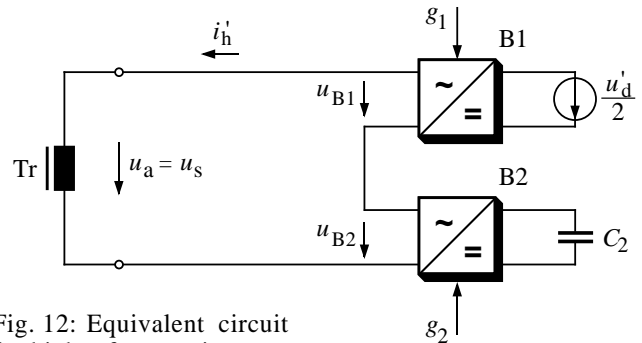


Fig. 12: Equivalent circuit for higher frequencies

to the respective step of the equivalent voltage  $u_t$  of the traction converters. Most frequently, positive and negative steps follow each other alternatingly. Such sequence can be handled by bridge B1 alone, while bridge B2 conserves its actual switching state.

A different situation occurs whenever the switched equivalent voltage  $u_t$  changes its alternating waveform to another pair of voltage levels. This process is accompanied by two subsequent voltage steps having the same polarity, as can be seen in the enlarged inset in Fig. 6. Of the two unipolar voltage steps, one is contributed by bridge B1, and the second by bridge B2.

## 5. CONTROL OF THE COMPENSATOR

### 5.1 Component losses

The inevitable losses in the power semiconductors and in the passive components lead to deviations from the idealized compensation waveform. It is especially the capacitor voltage  $u_{C2}$  which will gradually drift away from its desired average level since small errors of the charging current accumulate with time. For similar reasons, magnitude and phase angle errors are introduced to the capacitor voltage  $u_{C1}$ .

These deficiencies make it mandatory to operate the compensator with a superimposed control system.

### 5.2 Control of the capacitor voltage $u_{C2}$

According to the analysis in Section 4, the compensator bridge B1 is basically switched in synchronism, though at opposite polarity, with the traction converters. Unlike these, bridge B1 never assumes the freewheeling state of zero ac-side voltage. Hence its switching frequency results twice the sum of the switching frequencies of the traction converters, or  $88 \cdot 16 \frac{2}{3} \text{ Hz} \approx 1.47 \text{ kHz}$ .

Bridge B2 is switched at a much lower rate, primarily at every change between pairs of voltage levels that constitute the equivalent voltage  $u_t$ . The pertinent control signal  $g_2$  is shown in Fig. 11(b). Fig. 11(c) below shows that an undesired dc offset of the capacitor voltage  $u_{B2}$  can be compensated by displacing one of the switching instants of bridge B2 appropriately in time. In order to leave the waveform of the compensation voltage  $u_t$  unaffected by such displacement, bridge B1 must be simultaneously reversed.

The control of  $u_{C2}$  by pulse displacement of bridge B2 is explained more in detail with reference to Fig. 13. The time interval A marks the range of transition to a different switching level,  $3 u_d$  in this example. When entering range A at  $t = t_2$ ,

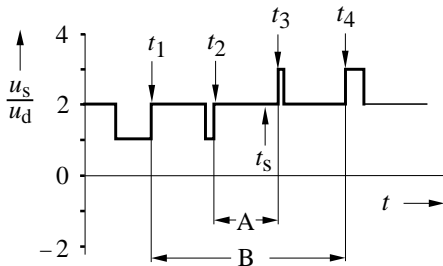


Fig. 13: Ranges A and B of possible pulse modifications for control of the capacitor voltages

$u_{B2}$  turns positive. The two bridges B1 and B2 can now be simultaneously reversed at any convenient time instant  $t_s$  within range A, so as to exert control on the capacitor voltage  $u_{C2}$  without affecting the waveform of  $u_s$ . After the event  $t_s$ ,  $u_{B2}(t > t_s)$  is nega-

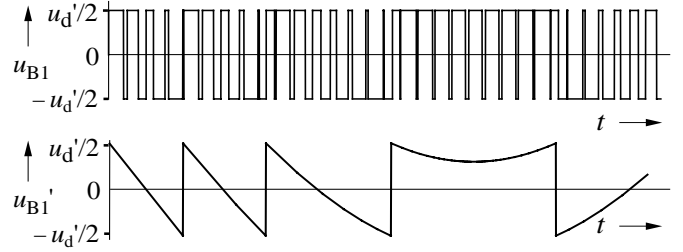


Fig. 14: The ac side voltage  $u_{B2}$  of bridge B2; (a) exact waveform, (b) equivalent piecewise continuous approximation

tive such that B2 can handle the second positive edge at  $t = t_3$ .

Higher ceiling values for the manipulation of  $u_{C2}$  can be achieved by displacing the simultaneous commutation of both bridges to a time instant within the larger interval B. A simultaneous commutation may then occur before, or after the two subsequent switching edges of equal polarity. When extending control to the larger interval B, the short notches which are located before the beginning and beyond the end interval A are executed by bridge B2. Both notches have very short durations such that their effect on the desired control action is negligible.

Although possible in principle, it was not found necessary to extend the simultaneous commutations beyond the limits of the larger time interval B.

### 5.3 Control of the capacitor voltage $u_{C1}$

The voltage of the capacitor  $C_1$  is principally determined by the fundamental component of  $u_a$ , which is the voltage across the auxiliary transformer winding. Minor deviations of phase angle and magnitude of  $u_{C1}$  are compensated by modifying the charging current  $i_q$  of capacitor  $C_1$ . To achieve this, an adequate fundamental component is added to the compensation voltage  $u_s$ . This waveform does not have a fundamental component at undisturbed operation, which was explained in Section 4.3.

The modification of  $u_s$  is achieved by pulse displacements in bridge B1. The pertinent pulsewidth modulated waveform  $u_{B1}(t)$  is shown in Fig. 7(a). The curve in Fig. 14(a) is a portion of this waveform, expanded in time such that the changing duty-cycles can be better observed. A good understanding of the nature of this waveform is conveyed by its approximation Fig. 14(b), which, as a piecewise continuous function, displays the time-varying average per pulse period of the switched waveform  $u_{B1}$ . If  $u_{B1}(t)$  was to be generated by pulsewidth modulation using a triangular carrier,  $u'_{B1}(t)$  would be the appropriate reference signal.

The most stringent condition for the generation of  $u_{B1}(t)$  is the accurate coincidence in time of its switching instants with those of the traction power converters. This coincidence accounts for the exact harmonic compensation and cannot be changed. Hence the piecewise continuous sections in the approximated waveform  $u'_{B1}(t)$  must remain unaffected. However, the transitions between two continuous sections of  $u'_{B1}$  may be displaced in time, provided a reaction on the compensating voltage  $u_s$  is offset by a simultaneous commutation of the complementary bridge, which in this case is B2.

It is obvious from the discussion of the characteristic waveforms in Fig. 10 and Fig. 11 that control of the charging

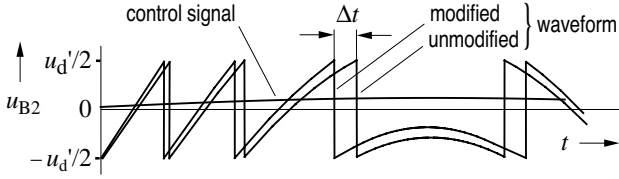


Fig. 15: Control of the the waveform  $u_{B2}$  by the current  $i_q$

current  $i_q$  of capacitor  $C_1$  must have an effect on the waveforms  $u_{C2}(t)$  and  $u_{B2}(t)$ . Exactly this is now achieved by the aforementioned pulse displacement in  $u'_{B1}(t)$  and the consequent simultaneous commutation of bridge B2. The effect of a control action on the waveform  $u_{B2}(t)$  is shown in Fig. 15.

#### 5.4 Control of the dc voltage $u_d'$

A mismatch between the waveform amplitudes of the compensator and the traction converter occurs when the dc link voltage of the traction converter changes subject to load conditions and line voltage fluctuations. Different commutation slopes in the high-power traction converter and the low-power compensator, as well as differences of gate signal delay times in the microsecond order are other sources of inaccuracy.

Such disturbances are compensated by adjusting the dc-side voltage of bridge B1, basically in proportion to the dc link voltage of the traction converters. This makes a near perfect cancelation of the harmonics possible.

### 6. CONTROL IMPLEMENTATION

The practical implementation of the compensator control is based on an event-oriented concept. This approach leads to very simple control algorithms, while providing the required robustness and high dynamic performance.

The control operates on the following rules:

- The basic switching pattern of bridge B1 is derived from the switching sequence of the traction converters.
- The magnitude  $u_d'$  of the dc-side voltage of bridge B1 is adjusted in proportion to the dc link voltage of the traction system.
- When the capacitor voltage  $u_{C2}$  reaches the level  $u_d'$ , bridge B2 is reversed. Bridge B1 is reversed at the same time, which leaves the compensating voltage unaffected.
- When the capacitor voltage  $u_{C2}$  reaches the zero level, bridge B2 is reversed while bridge B1 maintains its state. This ensures that  $u_{C2}$  remains always positive. The waveform  $u_{B2}$  is not affected by this commutation since the event occurs exactly at zero crossover.

### 7. DESIGN CONSIDERATIONS

The compensator was designed for operation in a class 120 ac locomotive. The nominal dc link voltage of the traction converter is 2 800 V. The leakage inductance  $L_{\sigma m}$  of the main secondary windings serves as the smoothing filter for the current harmonics of the traction converter. This inductance is fairly high, having a normalized value  $L_{\sigma m} = 0.28$ . The auxiliary leakage inductance  $L_{\sigma a}$  can be well chosen as  $0.2 L_{\sigma m}$ , which sets the step voltage amplitude of the compensator to 560 V. The nominal dc-side voltage of bridge B1 is then 280 V.

The harmonic amplitudes of the pantograph current reach

peak values of 6 A. Having set the ratio  $u_d/u_d' = 10$ , this current is matched by a compensating 60 A peak current in the auxiliary transformer winding. The power semiconductor of the two compensator bridges must carry an additional 50 A of fundamental frequency, which is the charging current of the capacitors  $C_1$  and  $C_2$ .

The lower order 700 Hz current harmonics that result from unbalanced operation of the two traction units cannot be absorbed by a conventional passive filter. In the extreme, this filter must be disconnected for overload protection when only one traction unit is active. The switched compensator can handle both situations just by proper gate control of its two bridges. The worst case is emergency operation with only one traction unit, which renders the 700 Hz harmonics predominant. Their peak value reaches 12 A in the pantograph. This translates into 120 A peak harmonic current of the compensator for complete harmonic elimination.

IGBTs are well-suited as power semiconductor switches for the compensator. They permit the accurate adjustment of the switching instants without introducing undesired switching delays by storage effects. 200-A modules are adequate.

The switching frequency of the power devices in bridge B1 is approximately 1.5 kHz, considering the additional commutations for pulse displacement control. The IGBTs of bridge B2 are operated around 200 Hz switching frequency.

The inherent fast switching of IGBTs requires low stray inductances in the commutation loops. Parallel-plate conductors are used for the wiring of the power components. Additional RC snubbers reduce the switching overvoltages. Owing to the low switching frequency, device switching losses are not a problem.

The total losses of the compensator range around 3% of the apparent harmonic power. The losses are restituted to the dc-side of bridge B1. A controlled dc-to-dc converter is provided for this purpose. It receives its input power of approximately 3 kW from one of the on-board auxiliary systems.

### 7. EXPERIMENTAL RESULTS

The compensator was built and laboratory tested at full power level. The 16-2/3-Hz supply was taken from a motor-

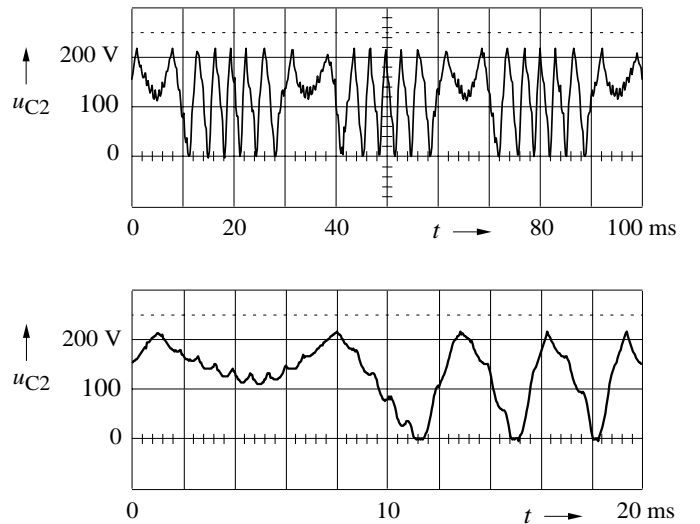


Fig. 16: (a) capacitor voltage  $u_{C2}$ , (b) as before, enlarged

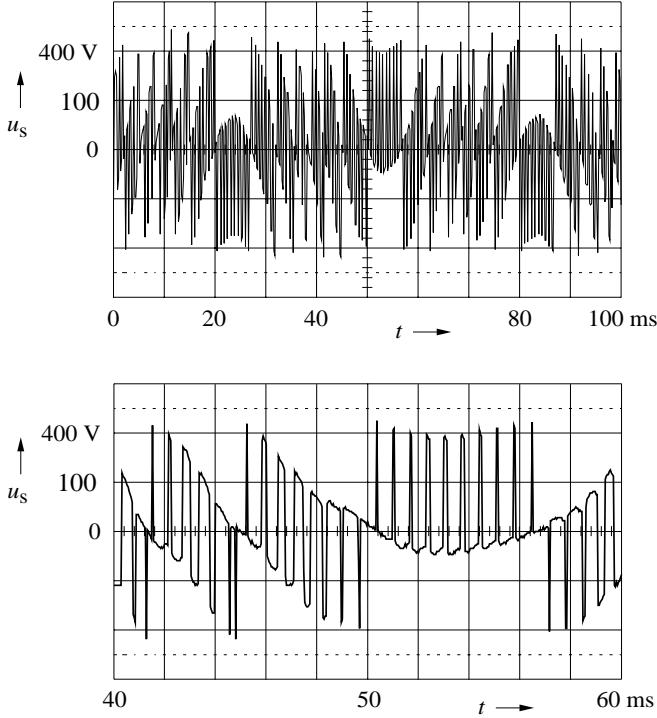


Fig. 17: (a) compensation voltage  $u_s$ , (b) as before, enlarged

generator set and matched to the impedance of an overhead line by a tuned LC circuit.

Fig. 16 shows the recorded capacitor voltage  $u_{C2}$ . The higher frequency components in this waveform are caused by the inverse harmonic current. The harmonic compensation voltage  $u_s$ , shown in Fig. 17, is a composition of  $u_{C2}$  and the higher frequency waveform  $u_{B1}$  generated by bridge B1.

The oscillogram Fig. 18(a) shows the waveform of the compensation current  $i_{h'}$ . It contains the inverse harmonic current and the fundamental charging current of the two ca-

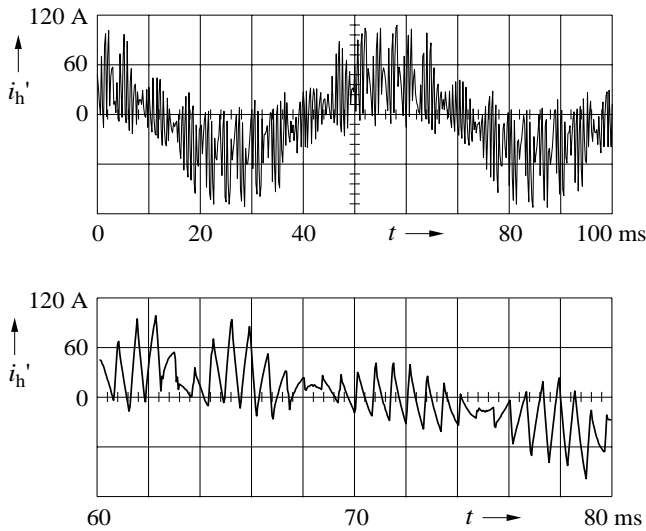


Fig. 18: Compensation current  $i_{h'}$ ; (a) waveform of a fundamental period, (b) as before, enlarged

pacitors  $C_1$  and  $C_2$ . The enlarged waveform Fig. 18(b) permits a better view of the inverse harmonic current. The current through the inductor  $L_1$  is oscillographed in Fig. 19.

A full scale locomotive power converter was not available for laboratory testing. The equivalent traction converter voltage  $u_t$  was therefore synthesized from the pulsewidth modulator signals. The superposition of the mmf waveforms in the main transformer was emulated at a low power level to obtain the waveform of the induced transformer voltage. This waveform is recorded in Fig. 20(a). Its residual ripple is not accurately reproduced in the graph due to aliasing effects. This is proved by observing the same waveform in an enlarged time scale. Fig. 20(b) shows that the voltage is in effect sinusoidal, having only very short pulses added to it. These result from delays in signal processing and in the switched power converter. The voltage-time area of the vestigial pulses is only  $10 \mu\text{Vs}$ .

Fig. 20(c) shows the induced transformer voltage without the compensator for comparison. The difference is also well documented by the measured harmonic spectra Fig. 21.

Note that the characteristics of Fig. 20 and Fig. 21 do not refer to the line current, but to the voltage induced by the main flux of the transformer. This voltage acts as a back voltage in an equivalent circuit as in Fig. 2. The line current is low-pass filtered by the primary leakage inductance  $L_{\sigma L}$ , making the effect of minor signal delays negligible.

Owing to the harmonic compensation, the main flux of the transformer is rendered sinusoidal and the harmonic transformer losses reduce considerably. The overall efficiency of the traction unit including the compensator losses is increased [11]. The installed compensator power amounts to about 1% of the traction power.

## 8. SUMMARY

A light-weight and low-power switched compensator is described which completely eliminates the harmonic current on the line-side of ac traction locomotives. The compensator consists of two single-phase IGBT bridges operated at a max-

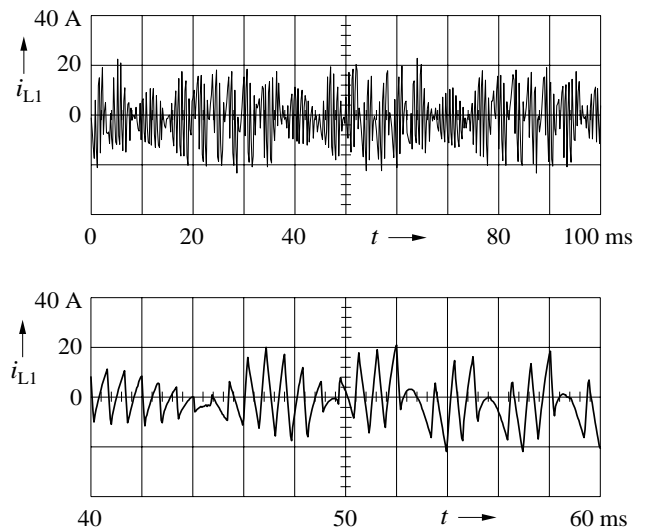


Fig. 19: Current through the inductor  $L_1$ ; (a) waveform of a fundamental period, (b) as before, enlarged

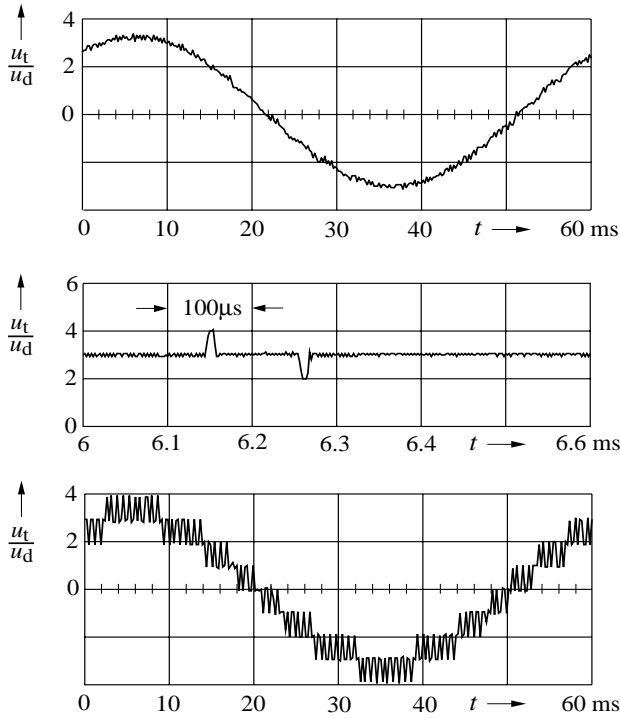


Fig. 20: Equivalent voltage  $u_t$  of the traction converters; (a) with harmonic compensation (20 kHz band-limited), (b) enlarged section of the waveform (a), (c) without compensator

imum switching frequency of 1.5 kHz. It generates a replica of the harmonic current of the traction power converters. This current produces a counter mmf in the main transformer through an auxiliary winding which forces the transformer flux sinusoidal. The induced voltage on the primary side is then also sinusoidal, and so is the line current.

The control of the compensator is event-oriented, which ensures robustness and high dynamic performance. Laboratory tests at full power level demonstrate its unique properties. It makes bulky and expensive high-voltage filter equipment obsolete and remedies other existing deficiencies such as turn-on transients, residual harmonic currents, and the inability to cope with lower order harmonics at unbalanced operation of the traction units. The novel principle of exact harmonic compensation constitutes the first static high-power converter having pure sinusoidal ac-side currents without employing reactive filter components.

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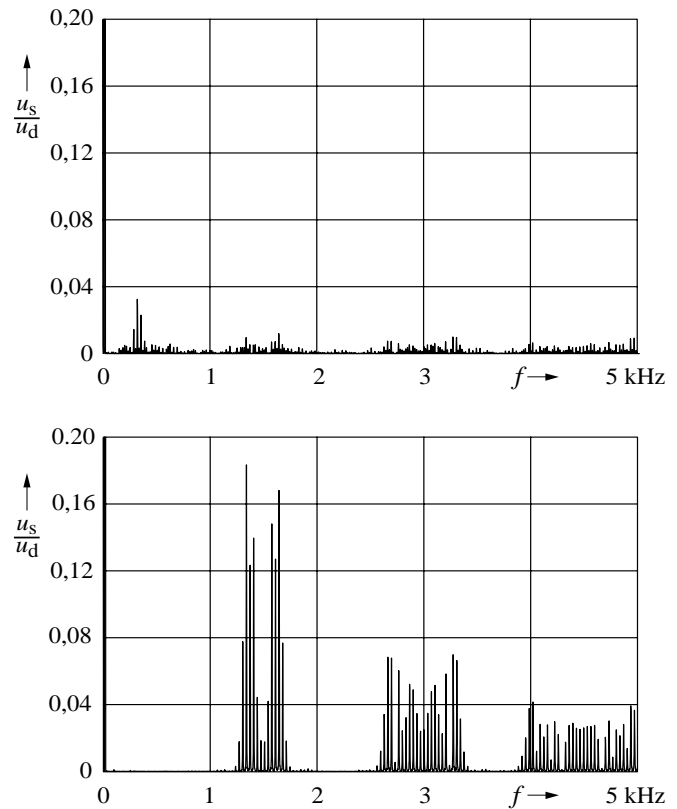


Fig. 21: Measured harmonic spectrum of the equivalent voltage  $u_t$ ; (a) with harmonic compensation, (b) without compensator

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