

# The Bootstrap Variable Inductance: A New FACTS Control Element

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**Abstract**—A new FACTS controller, the Bootstrap Variable Inductance, can emulate variable positive and negative inductance. Unlike the conventional capacitive approach for reducing inductive reactance in ac power systems, it has the advantage that resonance is impossible. The Bootstrap Variable Inductance has a variety of FACTS applications: series compensation of lines, fault-current limiting, reactive-power control, and load power-factor improvement. We analyze its operation with particular attention to harmonics, and confirm the results through simulations.

## I. INTRODUCTION

Flexible ac transmission systems (FACTS) employ power electronic converters to control the active and reactive power flow within an ac transmission system, while maintaining an adequate stability margin [1]. A large number of FACTS controllers have been listed by an IEEE working group [2]. The action of many FACTS control elements is to introduce reactance in series with transmission lines, or in shunt at selected busbars.

The conventional approach in power systems is to add capacitive reactance in such a way as to compensate unwanted inductive reactance. This is satisfactory at 50 or 60Hz, but it introduces or exacerbates system resonances. A novel alternative is to use negative inductance, which can compensate inductive reactance without resonance. In this paper, we present an electronically emulated variable inductance — the Bootstrap Variable Inductance. The technique can be applied in power systems for several different purposes. It should result in improved transient and dynamic stability.

Let us first explore the concept of negative inductance. Inductance is defined by the branch equation  $di/dt = v/L$ : a positive voltage across a circuit branch is associated with an increasing current through it. The definition places no theoretical limitation on  $L$ , so it could be negative. As a shorthand, we use the term *reductance* as a synonym for negative inductance, and denote it by  $\Gamma$ . Thus  $\Gamma = -L$ , and we can define the new circuit element by

$$\frac{di}{dt} = \frac{-v}{\Gamma} \quad (1)$$

defining equation for reductance,  $\Gamma$

To reiterate, reductance is another name for negative inductance. It takes positive values. Table I compares the properties of inductance, reductance and capacitance. Note that reductance is *not* the same as capacitance: their frequency characteristics are quite different.

In an ac power system, inductive reactance often gives trouble at the synchronous frequency  $\omega_s$  (the ‘power frequency’). Inductive reactance is positive,  $X_L = \omega_s L$ . Capacitive reactance,  $X_C = -1/\omega_s C$ , has traditionally been used to compensate inductive reactance at  $\omega_s$ . However, total cancellation occurs at only a single frequency,  $\omega_0 = 1/\sqrt{LC}$ , the resonant frequency. For partial cancellation  $\omega_0 \neq \omega_s$ , and this can cause problems. If  $\omega_0 > \omega_s$ , the resonance can magnify any harmonics present. On the other hand if  $\omega_0 < \omega_s$ , the resonance could coincide with mechanical modes in the power system, e.g. shaft resonances, causing dangerous sub-synchronous resonance.

We contend that it is better to compensate inductance with reductance, because resonance is then impossible. Reductive reactance is negative, like capacitive reactance, but  $L$  and  $\Gamma$  in series have an impedance  $j\omega(L - \Gamma)$ . Adding reductance simply reduces the effective inductance.

Fig. 1 demonstrates that reductance and capacitance are equivalent at a single frequency, but differ when harmonics

TABLE I: INDUCTANCE, REDUCTANCE AND CAPACITANCE

|                           | Inductance<br>$L$ | Reductance<br>$\Gamma$ | Capacitance<br>$C$ |
|---------------------------|-------------------|------------------------|--------------------|
| Definition                | $v = L di/dt$     | $v = -\Gamma di/dt$    | $i = C dv/dt$      |
| Impedance $Z(j\omega)$    | $j\omega L$       | $-j\omega \Gamma$      | $-j/\omega C$      |
| Magnitude $ Z(j\omega) $  | $\propto \omega$  | $\propto \omega$       | $\propto 1/\omega$ |
| Phase $\angle Z(j\omega)$ | $90^\circ$        | $-90^\circ$            | $-90^\circ$        |
| Reactance at $\omega_s$   | $\omega_s L$      | $-\omega_s \Gamma$     | $-1/\omega_s C$    |

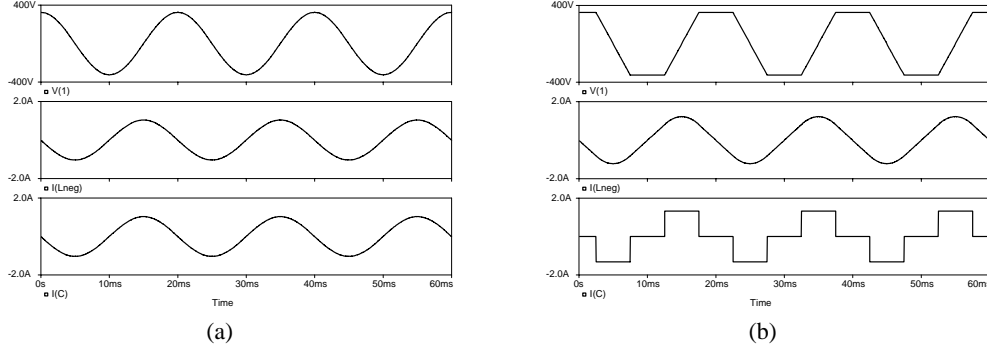


Fig. 1: Comparison of inductance and capacitance. (a) Sinusoidal voltage waveform, (b) trapezoidal voltage waveform. Top: impressed voltage; middle: current in inductance; bottom: current in capacitance. Although inductance and capacitance are indistinguishable at a single frequency, when harmonics are present their behavior is quite different.

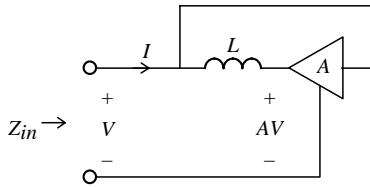


Fig. 2: The Bootstrap Variable Inductance (BVI).

are present. Suppose a voltage  $v(t) = \sum_{n=1}^{\infty} A_n \sin n\omega t$  is applied to each. The inductive current is proportional to the integral of the voltage waveform. (The initial condition  $i_L(0)$  can be chosen so that dc term vanishes.)

$$\begin{aligned} i_L(t) &= i_L(0) + \frac{-1}{\Gamma} \int_0^t v(\tau) d\tau \\ &= i_L(0) - \sum_{n=1}^{\infty} \frac{A_n}{n\omega\Gamma} (1 - \cos n\omega t) \end{aligned} \quad (2)$$

The high-order harmonics are diminished by a factor  $1/n$ , giving a smoother current waveform. In contrast, the capacitive current is proportional to the derivative of the voltage waveform:

$$i_C(t) = C \frac{dv}{dt} = \sum_{n=1}^{\infty} A_n n\omega C \cos n\omega t \quad (3)$$

This accentuates the high-order harmonics by a factor  $n$ , giving a highly distorted current waveform.

## II. REALIZATION OF VARIABLE INDUCTANCE / REDUCTANCE

Negative inductance was first proposed as a power-system component by Funato and Kawamura in the form of a ‘vari-

able active-passive reactance’ (VAPAR) [3]–[9]. As an alternative to the VAPAR, we recently proposed the Bootstrap Variable Inductance (BVI) [10]. The basic principle, shown in Fig. 2, is that of bootstrapping, a type of feedforward. The applied voltage  $V$  feeds an inductance  $L$  in series with a voltage amplifier of gain  $A$ . The effective input impedance is

$$Z_{in}(j\omega) = \frac{V}{I} = \frac{j\omega L}{1 - A} \quad (4)$$

When  $A < 1$ , the BVI appears as an inductance of magnitude greater than  $L$ : for  $A \in [0, 1)$ ,  $L_{in} \in [L, \infty)$ . When  $A = 1$ ,  $Z_{in} = \infty$  because  $I = 0$ . This principle is known in analog electronics as *bootstrapping*. When  $A > 1$ , negative impedance conversion takes place, and the BVI appears as a reductance ( $Z_{in} = -j\omega\Gamma$ ): for  $A \in [1, 2]$ ,  $\Gamma_{in} \in [L, \infty)$ . Thus by varying  $A$ , a wide range of inductance and reductance can be obtained. In general,

$$L_{in} = \frac{L}{1 - A} \quad (5)$$

where  $L_{in} \in (-\infty, \infty)$ .

## III. PRACTICAL CIRCUITS

### A. Single-Phase BVI

A proposed single-phase BVI circuit is shown in Fig. 3. Reference [10] contains further details of practical issues, and we summarize the main points here only briefly.

- For high efficiency, the amplifier must use a switched-mode output stage. Such amplifiers, using pulse-width modulation, have been proposed for FACTS [11].

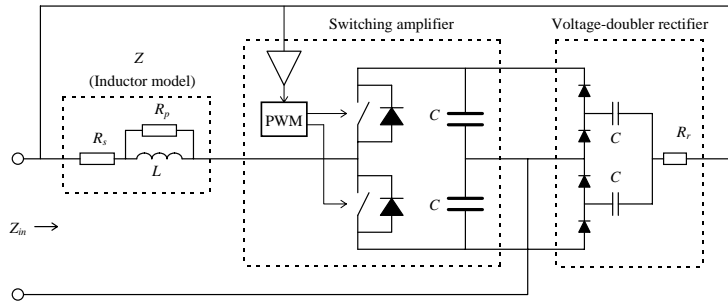


Fig. 3: A single-phase BVI, self-powered from the terminal voltage.

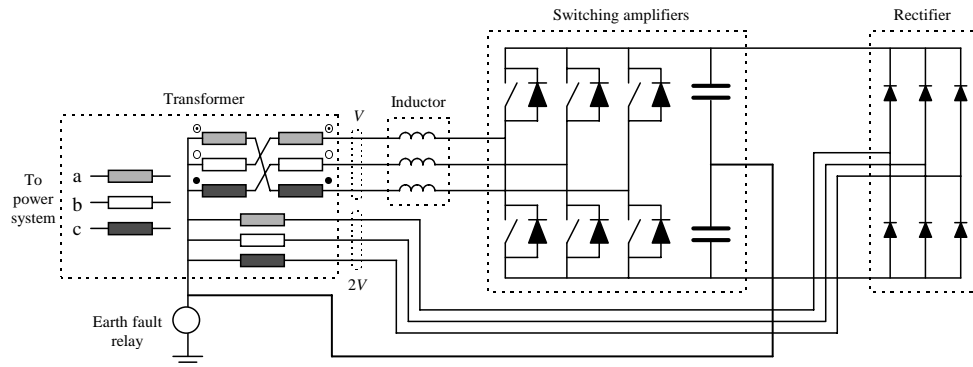


Fig. 4: Power circuit of a three-phase BVI. The step-down transformer has uncommitted primary windings and a zigzag-connected secondary. An additional low-power winding supplies the rectifier. Control circuits are not shown.

- The switching frequency must be several times the power frequency. To minimize the low-order harmonics introduced into the power system, the switching frequency should be high, but to minimize switching losses it should be low. A compromise is called for.
- At megawatt power levels, gate turn-off thyristors (GTOs) are the favored switching device. At lower powers, insulated-gate bipolar transistors (IGBTs) might be used.
- The voltage levels usually found in power systems (up to 1500kV) are incompatible with those of available GTOs (a few kilovolts). A step-down transformer can be added at the input to trade voltage for current.
- The power amplifier requires positive and negative dc supply rails at least  $A$  times the peak value of the terminal voltage  $V$ . If the circuit were ideal it would absorb no power, but in practice there are unavoidable parasitic losses. There are various schemes for supplying the appropriate amount of dc power. If the efficiency is high, the dc power requirement will be small.
- When  $A > 1$ , losses in the inductor  $L$  appear as unwanted negative resistance in  $Z_{in}$ . This effect can be overcome by making  $A$  suitably frequency-dependent.

### B. Three-Phase BVI

In power systems applications three-phase operation is the norm. Although it would be possible to use three separate

BVIs, there are advantages in combining them into a single circuit, as in Fig. 4. Again, only the main points will be given here.

- The three switching amplifiers are independent, each phase being treated individually. Only the dc circuit is shared. The midpoint of the two capacitors is connected as a return path to the ac source, so that unbalanced voltages can be tolerated. Zero-sequence components at the input should present no problems.
- The common point is grounded via a current-sensing relay. If an internal earth fault occurs, the circulating fault current flows through the relay. The BVI can be removed from the power system before there is a chance of further disruption.
- In series applications, the transformer primary windings are independently connected. In shunt applications, the primary is wye-connected to lower the voltage stress. The secondary is preferably zigzag connected. This provides a common point, and cancels the effect of triplen harmonics in each arm of the transformer core. The windings are all well utilized, and there are no problems with zero-sequence components caused by fault conditions in the power system.
- It may be possible to combine the transformer and the three inductors into a single integrated magnetic unit. In principle, each inductor could be formed by leakage inductance referred to the secondary.

#### IV. REACTIVE AND ACTIVE POWER

Besides controlling the reactive parameters of a power system, the BVI can also be made to supply active power from its dc energy storage elements (capacitors or batteries). This can be seen by considering Fig. 2. Suppose instead of a real gain  $A$  the amplifier provides a complex gain  $Ae^{j\alpha}$ , where  $\alpha$  is a phase shift deliberately introduced into the amplifier. The amplifier's output-voltage phasor is now  $A(\cos \alpha + j \sin \alpha)V$ , making the input impedance

$$Z_{in}(j\omega) = \frac{j\omega L(1 - A \cos \alpha)}{1 + A^2 - 2A \cos \alpha} - \frac{\omega L A \sin \alpha}{1 + A^2 - 2A \cos \alpha} \quad (6)$$

The first term in (6) is a variable inductive/reductive reactance, which reduces to that of (4) when  $\alpha = 0$ . The second term emulates a positive/negative resistance, which vanishes when  $\alpha = 0$ . The apparent power delivered to the power system is now

$$S = P + jQ = -\frac{A|V|^2}{\omega L} \sin \alpha + j \frac{|V|^2(1 - A \cos \alpha)}{\omega L} \quad (7)$$

If  $\alpha$  is positive (i.e. the amplifier introduces a phase lag), the BVI's energy storage elements supply real power to the power system. If  $\alpha$  is negative, the BVI absorbs real power from the power system, storing energy.

One outcome is that the BVI could be used for peak leveling: storing energy in batteries during off-peak times and delivering it at peak times. Another implication is that, like the VAPAR and other FACTS controllers, the BVI can be made self-powering without a rectifier:  $\alpha$  can be set so the BVI absorbs active power from the power system that exactly balances that dissipated in the parasitic losses. Then (except for a small ripple) the voltage on the dc capacitors remains constant. A control loop can set this to the desired value.

#### V. HARMONIC ANALYSIS OF PWM-BVI

When its terminal voltage is sinusoidal, the BVI would ideally draw a sinusoidal current. Hence we need to know the harmonic content of the switching amplifier's output voltage.

Ideally, the two switches are driven in anti-phase. (In practice overlap must be avoided.) The most straightforward way to produce a suitable switching function is by naturally-sampled pulse width modulation (PWM). The BVI's input voltage is the reference waveform, and is compared to a high-frequency carrier waveform. The comparator output provides the switching function.

##### A. Problem Definition and Assumptions

The present discussion is limited to naturally-sampled PWM with two special groups of carrier waveform:

Type 1: those that intersect with the reference waveform at exactly one point in each period (e.g. a ramp carrier and a

sinusoidal reference);

Type 2: those that intersect at exactly two points in each period (e.g. a triangular carrier and a sinusoidal reference).

We make the following assumptions:

- The reference waveform  $v_R(t)$  is periodic with period  $T_R = 2\pi/\omega_s$  (this depends on the power system).
- The carrier waveform  $v_C(t)$  is periodic with period  $T_C = 2\pi/\omega_C$  (a free choice).
- The carrier frequency is locked to the reference frequency, so that  $T_R = NT_C$ . This could be achieved by means of a phase-locked loop.
- The amplifier phase shift  $\alpha = 0$ .

The switching function is

$$f(t) = \begin{cases} 1 & \text{if } v_R(t) \geq v_C(t) \\ -1 & \text{if } v_R(t) < v_C(t) \end{cases} = \text{sgn}(v_R(t) - v_C(t)) \quad (8)$$

The objective is to find a Fourier series for  $f(t)$  in the form

$$f(t) = \sum_{n=-\infty}^{\infty} F_n e^{jn\omega_s t} \quad (9)$$

where  $\omega_s = 2\pi/T_R$  and  $F_n = (a_n - jb_n)/2$ . Here  $a_n$  and  $b_n$  are the cosinusoidal and sinusoidal Fourier-series coefficients respectively.

##### B. Lower Bound on $N$

We first find a conservative lower bound on the value of  $N$ . We show that, to ensure only one (Type-1) or two (Type-2) intersections within each  $T_C$ , in the worst case the carrier frequency must be at least four times the reference frequency.

*Type 1:* Let the carrier  $v_C(t)$  be a ramp that rises from  $-1$  at time  $t_0$  to  $+1$  at time  $t_1 = t_0 + T_C$ . Thus  $v_C(t) = 2(t - t_0)/T_C - 1$ . Let the reference  $v_R(t) = \sin \omega_s t$ , which also has a range of  $[-1, 1]$ , i.e. this is full modulation (the worst case). As in (8), we are interested in  $v_R(t) - v_C(t) = v_{RC}(t)$ , say. The reference-carrier intersections are given by the zero-crossings of  $v_{RC}(t)$ . It is clear that if  $v_{RC}(t)$  does not have a maximum or minimum within the interval, there can be only a single intersection. Now the derivative  $dv_{RC}/dt = \omega_s \cos \omega_s t - 2/T_C$ . Equating this to zero gives  $\cos \omega_s t = T_R/\pi T_C$ . But  $T_R/T_C = N$  and  $\cos \omega_s t \leq 1$ , so  $N \leq \pi$  if there is to be a possibility of more than one intersection. To avoid multiple intersections, therefore,  $N > \pi$ . But  $N$  is an integer, so  $N \geq 4$ . This conservative lower bound is easily met in practice, since  $N \geq 10$  in most cases (e.g. a 50Hz power system with a  $\geq 500$ Hz carrier).

*Type 2:* A triangular carrier waveform can be constructed from a rising ramp of duration  $kT_C$  and a falling ramp of duration  $(1 - k)T_C$ , where  $k < 1$ . (E.g.  $k = 1/2$  for a symmetrical triangle wave.) If each ramp intersects with  $v_R(t)$  at exactly

one point, the above analysis remains valid, provided  $T_C$  is replaced by  $kT_C$  for the rising ramp and  $(1-k)T_C$  for the falling ramp. For the rising ramp we find  $T_R/\pi T_C \leq k < 1$ , so  $T_R/T_C < \pi$  if multiple intersections are to occur, i.e. choosing  $N \geq 4$  will avoid them. For the falling ramp,  $T_R/\pi T_C \leq 1-k < 1$ , and we must have  $N \geq 4$  again. If  $k$  is known, the bound can be improved; e.g. if  $k = 1/2$ ,  $N \geq 2$ .

### C. Formula Derivation

By definition, the  $n$ th Fourier coefficients of  $f(t)$  are

$$a_n - jb_n = \frac{2}{T_R} \int_0^{T_R} f(t) \exp(-jn\omega_s t) dt \quad (10)$$

Suppose the intersections of the carrier and reference waveform occur at  $t_1, t_2, t_3, \dots, t_m$ . First consider a Type-1 carrier: one that intersects with the reference waveform at exactly one point in each period  $T_C$  (e.g. a ramp carrier and a sinusoidal reference). Substituting (8) into (10), integrating over intersection intervals and simplifying, we obtain:

$$a_n - jb_n = \frac{2}{-jn\pi} \sum_{k=1}^m e^{-jn\omega_s t_k} - e^{-jn\omega_s (k-1)T_C} \quad (11)$$

The dc component is  $a_0 = \frac{2(t_1 + t_2 + t_3 + \dots + t_m) - m^2 T_C}{T_R}$ .

Now consider a Type-2 carrier: one that intersects with the reference waveform at exactly two points in each period  $T_C$  (e.g. a triangular carrier and a sinusoidal reference). By a similar process, we obtain

$$a_n - jb_n = \frac{2}{-jn\pi} \sum_{k=1}^m e^{-jn\omega_s t_{2k-1}} - e^{-jn\omega_s t_{2k}} \quad (12)$$

and the dc component is

$$a_0 = \frac{2(t_1 - t_2 + t_3 - t_4 + \dots + t_{2m-1} - t_{2m}) + T_R}{T_R}$$

Equations (11) and (12) are the main results of this section. They give the complex Fourier coefficients of the amplifier's switching function.

### D. Assessment of the Results

Equations (11) and (12) have been evaluated numerically, and are assessed next.

The BVI's inductance/reductance is varied by changing the amplifier gain  $A$ . This can be done by varying the amplitude of the reference waveform fed to the PWM.

A program was written to evaluate (11) and (12) and hence find the spectrum of the switching function. To validate the results, the spectrum was also computed using Matlab's fast Fourier transform (FFT) for comparison. It should be noted that the FFT is slower and less accurate. This is because it

employs a Fourier transform approach, which does not assume any periodicity in the waveform, whereas our analysis makes use of the fact that the switching function is periodic with period  $T_R$ . Moreover, the FFT needs a windowing function, and the choice of this is a compromise between its time-domain and frequency-domain characteristics.

Fig. 5 shows the output of the switching amplifier (normalized by the input voltage). The numerical spectral analysis is compared with that from the FFT for two values of  $A$ . The results agree very well, confirming the validity of the analysis. The small amount of noise in the FFT results is due to the windowing. As the window time increases, we expect the noise to reduce (but computation to take longer).

Fig. 6(a) shows the variation of 'total' harmonic percentage (for  $n = 1$  to 250; i.e. 50Hz to 12.5kHz) in the amplifier's output versus the amplifier gain. The rms sum of the harmonic voltages is normalized with respect to the BVI's input voltage. (This is not THD by the conventional definition, but it makes more sense here. For example, at  $A = 0$  the amplifier's output is a square wave at  $\omega_C$ : the  $\omega_s$  component is zero, but there are components at  $\omega_C$  and its multiples, so the THD is infinite: this is not very informative.) The analytical and FFT curves of Fig. 6(a) agree well.

Fig. 6(b) shows how the 50Hz fundamental and three multiples of the switching frequency vary with the amplifier gain. Again, they are normalized by the BVI's input voltage. The switching harmonics are considerable, especially the 2kHz carrier. For example, at  $A = 0$  its magnitude is more than 2.5 times the input voltage, and feeds through to the input current of the BVI. Additional high-frequency filtering may be needed to remove the switching frequency and its harmonics.

## VI. APPLICATIONS IN AC POWER SYSTEMS

The BVI finds a number of different uses in ac power systems. Here we suggest four such applications.

### A. BVI as Series Compensator

Neglecting resistance and capacitance, the active power transmitted through an ac transmission line is given by

$$P = \frac{|V_S| |V_R|}{\omega_s L_{line}} \sin \delta \quad (13)$$

where  $V_S$  and  $V_R$  are the sending-end and receiving-end voltage phasors respectively,  $\delta$  is the phase angle between  $V_S$  and  $V_R$  (the 'power angle'),  $\omega_s$  is the synchronous frequency and  $L_{line}$  is the line inductance. In practice  $|V_S|$ ,  $|V_R|$ ,  $\omega_s$  and  $L_{line}$  are all fixed, and  $\delta$  is dictated by system configuration and stability requirements ( $\delta \approx 28^\circ$  typically).

The active power transmitted can be increased by placing a capacitor in series with the line (series compensation). For more flexibility the effective capacitance is varied in steps

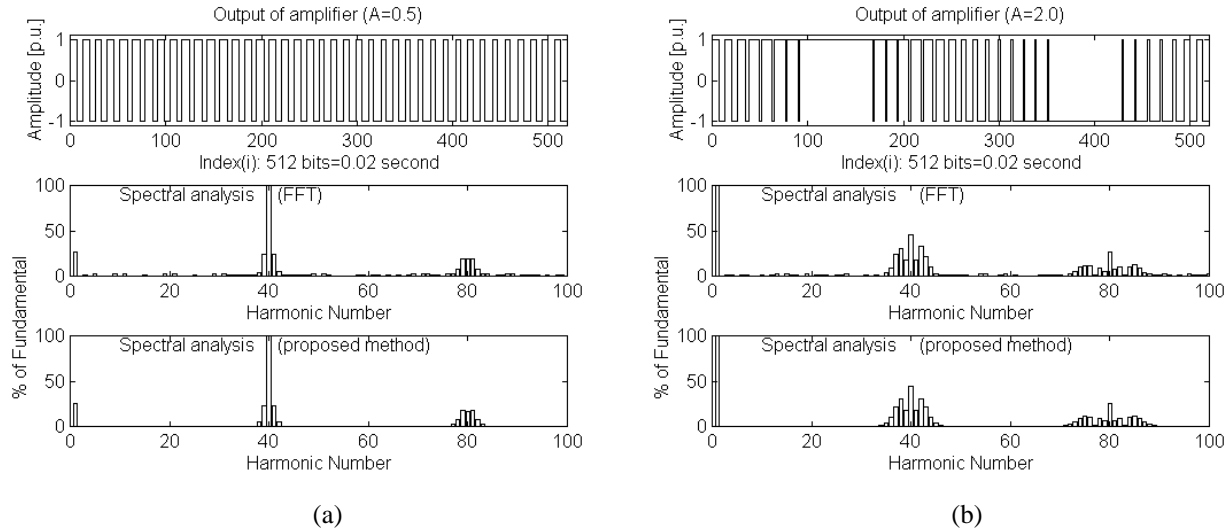


Fig. 5: Comparison of PWM spectra for (a)  $A = 0.5$  (inductance emulation), (b)  $A = 2.0$  (reductance emulation). PWM reference = 50Hz, carrier = 2kHz ramp.

Top: amplifier's output voltage waveform (per unit); middle: spectrum from Matlab FFT; bottom: spectrum from analysis.

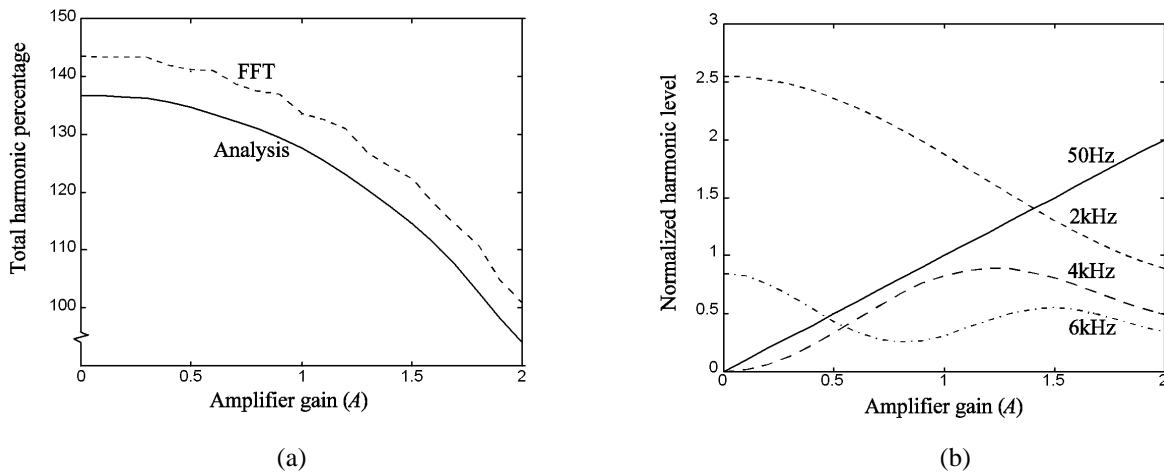


Fig. 6: (a) The total harmonic percentage of the amplifier's output voltage, normalized by the BVI's input voltage, by analysis and Matlab FFT; (b) the 50 Hz fundamental, 2kHz carrier frequency and two harmonics.

using a *thyristor switched series compensator* (TSSC), or continuously using a *thyristor controlled series compensator* (TCSC). Unfortunately, the capacitance resonates with inductances, a potential source of system instability. The resonance is typically around 15–30Hz. If this frequency coincides with mechanical modes, e.g. shaft resonances, the resulting sub-synchronous resonance could be dangerous.

It is important to realize that introducing a capacitance into a line not only affects that line, but also alters the eigenvalues of the whole power system, due to coupling. Thus its effect is global, not local.

To avoid these problems, we prefer to change the effective inductance of the line, compensating not only at  $\omega_s$  but over a

substantial bandwidth. Series reductance can be added by means of a BVI, as in Fig. 7(a). Because there is no resonance, we expect improved system stability. The BVI can also decrease the transmitted power, by operating in its inductive mode ( $A < 1$ ).

### B. BVI as Fault Current Limiter

A fault in a transmission line causes a large reactive power flow, because the line and other power-system components are essentially inductive. A series-connected BVI, as in Fig. 7(b), can be arranged to operate as a positive inductance under these conditions, limiting the fault current and reducing the rating of the associated circuit breakers. To avoid interfering with the operation of distance-protection relays, the

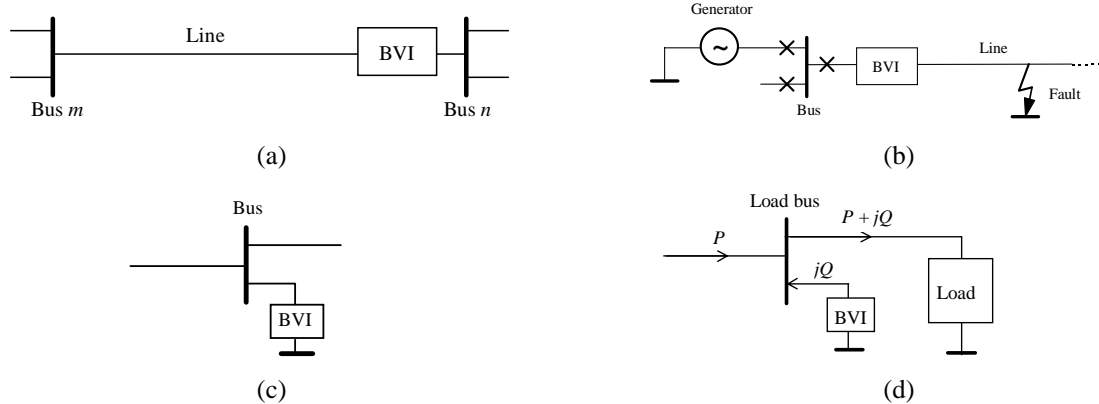


Fig. 7: Applications of the BVI: (a) series compensator; (b) fault current limiter; (c) shunt compensator; (d) load power factor corrector.

BVI should be coordinated with the protection system via a telecommunications link.

### C. BVI as Shunt Compensator

Reactive power flow in an ac power system must be controlled. This is usually managed by shunt compensation. Conventional methods include the *thyristor controlled reactor* (TCR), the *thyristor switched capacitor* (TSC) and the *static var compensator* (SVC), while a recent development is the *static synchronous compensator* (SSC). In all cases, the compensator draws a reactive (usually capacitive) current from the bus to regulate the bus voltage.

A BVI connected across an ac power system bus, as in Fig. 7(c), acts as a shunt compensator. Because high-frequency switching is used, the harmonic content is lower than that of conventional methods, and is more easily filtered. Moreover, unlike capacitive compensators, resonance is impossible, so there will be no amplification of existing harmonics.

### D. BVI as Load Power-Factor Corrector

The loads in a power system usually have a lagging power factor. Large consumers are penalized for low power factor, and it is incumbent upon them to fit power-factor correction equipment, usually shunt capacitors. An alternative is to employ a shunt BVI, as in Fig. 7(d). Again, reactance is advantageous in that it does not accentuate harmonic currents, and cannot resonate with an inductive load to produce large overvoltages if a disconnection occurs.

## VII. CONCLUSION

The Bootstrap Variable Inductance is a new FACTS element which can emulate variable inductance or reactance. Reactance is preferable to capacitance because it presents a negative reactance at the power frequency without introducing resonance. Single-phase and three-phase BVIs have been presented, self-powered from the terminal voltage. The BVI should find a variety of applications in power systems.

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