IN THE UNITED STATES DISTRICT COURT
FOR THE NORTHERN DISTRICT OF ILLINOIS EASTERN DIVISION

OLE K. NILSSEN and GEO FOUNDATION, LTD.	)	03C 2962
Plaintiffs,	)	Civil Action No. ________
v.	)	
	)	JURY TRIAL DEMANDED
OSRAM SYLVANIA, INC. and OSRAM SYLVANIA PRODUCTS, INC.	)	
Defendants.	)	

COMPLAINT

Plaintiffs, Ole. K. Nilssen ("Nilssen") and Geo Foundation, Ltd. ("Geo Foundation"), by their undersigned attorneys, complain of Defendants, Osram Sylvania, Inc. and Osram Sylvania Products, Inc., and allege as follows:

1. This Complaint comprises a single count for patent infringement.

A. Jurisdiction and Venue

2. Jurisdiction arises under 28 U.S.C. § 1338(a).

3. Venue is proper in this Court pursuant to 28 U.S.C. § 1391.

B. The Parties

4. Plaintiff, Ole K. Nilssen, is a domiciliary, and therefore a citizen, of Florida.

5. Plaintiff, Geo Foundation, is a not-for-profit corporation incorporated in the Cayman Islands, British West Indies.

1-1

6. Defendant, Osram Sylvania, Inc., is a Delaware corporation with headquarters in Danvers, Massachusetts.

7. Defendant, Osram Sylvania Products, Inc., is a wholly owned subsidiary of Osram Sylvania, Inc. and a Delaware corporation with headquarters in Danvers, Massachusetts.

8. Defendants have been, and are currently, in the business of making and selling electronic ballasts and compact fluorescents. Defendants sell electronic ballasts and compact fluorescents throughout the United States, including locations within the Northern District of Illinois.

9. Defendants' selling and offering for sale electronic ballasts and compact fluorescents within the Northern District of Illinois demonstrate continual and systematic contacts by the Defendants within the Northern District of Illinois.

10. In addition, Defendants' selling and offering for sale electronic ballasts and compact fluorescents within the Northern District of Illinois establish minimum contacts as such contacts were made for purposes of availing the Defendants of the privilege of doing business within the Northern District of Illinois.

11. Defendants' selling and offering for sale electronic ballasts and compact fluorescents within the Northern District of Illinois give rise to and are related to Plaintiffs' cause of action for patent infringement.

12. Exercising jurisdiction over Defendants in the Northern District of Illinois is consistent with traditional notions of fair play and substantial justice.

C. Background

13. Nilssen is in the business of identifying, formulating plans for, developing knowhow and technology for, and implementing (via license agreements) promising new business opportunities in the field of electronics, including electronic ballasts.

14. Nilssen is the inventor and owner of United States Patents Nos. 4,882,663, 4,935,669, 5,049,787, 5,404,083, 5,489,823, 5,550,439, 5,710,488, 5,710,489, 5,757,140, 6,121,733, 6,172,464, 6,211,619, 6,211,625, 6,472,827, 6,479,074 and 6,495,969 ("the patents-in-suit").

15. Geo Foundation has been an exclusive licensee of the patents-in-suit since January 1, 2000, with an exclusive right to license others.

16. Geo Foundation has authorized Nilssen to negotiate patent license agreements on its behalf as a prospective licensor of the patents-in-suit, and Nilssen has negotiated on behalf of Geo Foundation, Ltd.

17. The electronic ballasts and compact fluorescents that Defendants manufacture and sell infringe the patents-in-suit.

18. On information and belief, Defendants have had knowledge of the patents-in-suit since sometime after their issuance and have knowingly and without justification infringed upon these patents.

19. Plaintiffs have the right to bring suit with respect to each of the patents-in-suit.

20. Defendants have in the past and continue to make, use and sell electronic ballasts and compact fluorescents embodying the inventions claimed in each of the patents-in-suit and will continue to do so unless enjoined by this Court.

21. On information and belief, Defendants have in the past and continue to willfully infringe each of the patents-in-suit.

22. In the United States, purchasers of electronic ballasts made and sold by Defendants have used in the past and continue to use the electronic ballasts in combination with other components, including power sources and fluorescent lamps, thereby infringing each of the patents-in-suit.

23. On information and belief, each electronic ballast made and sold by Defendants was designed to be used in connection with a power source and one or more fluorescent lamps.

24. On information and belief, Defendants knew of each of the patents-in-suit at all relevant times before selling electronic ballasts to said purchasers.

25. Defendants have in the past and continue to manufacture, use, offer for sale, and sell electronic ballasts that constitute a material component of each patent-in-suit and which have no substantial use other than as an infringement of the patents-in-suit.

26. On information and belief, Defendants knew and intended that purchasers of Defendants' electronic ballasts use the electronic ballasts in combination with other components, including power sources and fluorescent lamps, so as to infringe each of the patents-in-suit.

27. On information and belief, Defendants have actively induced purchasers of Defendants' electronic ballasts to use the electronic ballasts in combination with other components, including power sources and fluorescent lamps, so as to infringe each of the patents-in-suit.

WHEREFORE, Plaintiffs pray that judgment be entered against Defendants:

(a) awarding damages and prejudgment interest to plaintiff under 35 U.S.C. §284;

(b) enjoining Defendants from making, using or selling electronic ballasts and compact fluorescents embodying the patented invention;

(c) enjoining Defendants from contributorily infringing and inducing the infringement of the patented inventions;

(d) increasing Plaintiffs' actual damages under 35 U.S.C. §284;

(e) awarding Plaintiffs reasonable attorney fees under 35 U.S.C. §285; and

(f) awarding such other relief as the Court deems proper.

Dated: May 2, 2003

Respectfully Submitted,

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EXHIBIT A

United States Patent [19]

Nilssen

[11] **Patent Number: 4,882,663**

[45] **Date of Patent: Nov. 21, 1989**

[54] MOSFET FLYBACK CONVERTER

[76] Inventor: Ole K. Nilssen, Caesar Dr., Rte. 5, Barrington, Ill. 60010

[21] Appl. No.: 338,513

[22] Filed: Apr. 10, 1989

Related U.S. Application Data

[63] Continuation of Ser. No. 812,721, Dec. 23, 1985, abandoned.

[51] Int. Cl.[4] H02M 3/338

[52] U.S. Cl. .... 363/19; 219/10.55 B; 323/222; 331/112; 363/131

[58] Field of Search 363/18, 19, 131; 331/111, 112; 323/222

[56] **References Cited**

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Primary Examiner—William H. Beha, Jr.

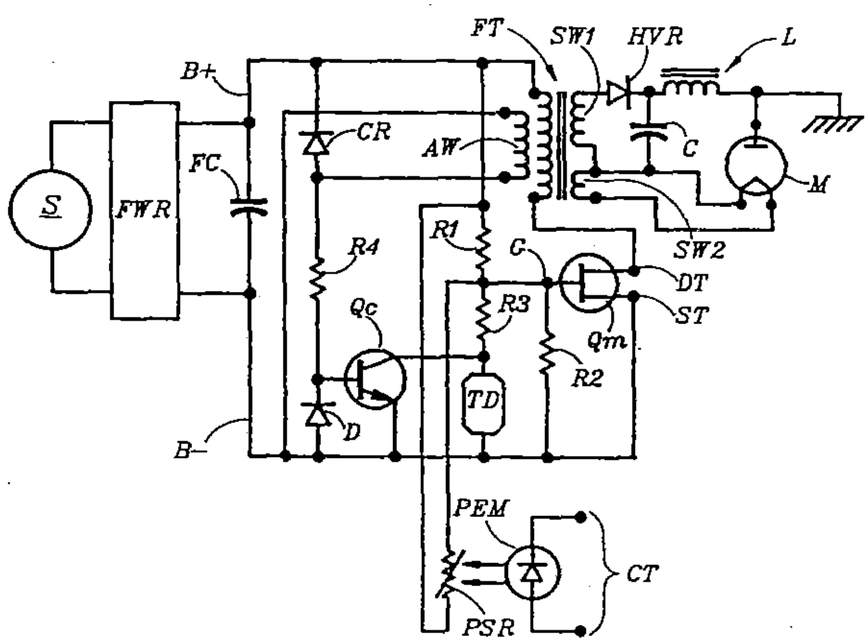
[57] ABSTRACT

In a flyback converter, a 15 Amp N-channel power MOSFET is driven to saturation by a gate voltage derived by connecting the gate to the relatively high-magnitude (150 volt) B+ voltage by way of a resistor of relatively high resistance (about 20 kilo-Ohm). Current flowing through this resistor causes the gate capacitance (3600 pico-Farad) to charge at a rate of about 2 Volt per micro-second. Since the forward transconductance of the MOSFET is 6 mhos or more, a situation has been established where the MOSFET is effectively fully switched ON as long as the MOSFET's drain current does not rise at a rate higher than about 12 Amp per micro-second.

Eventually, the magnitude of the gate voltage reaches a predetermined maximum level (about 20 Volt), at which point a threshold device, which is connected between gate and source, breaks down and rapidly discharges the base capacitance, thereby rapidly switching the MOSFET into a non-conducting state.

An ordinary bi-polar control transistor is also connected between gate and source, and this control transistor is made conductive by a small current from a secondary winding on the flyback inductor. This control transistor makes the threshold device de-latch and also keeps the gate shorted to the source for as long as the flyback inductor is in the process of discharging its energy.

12 Claims, 2 Drawing Sheets



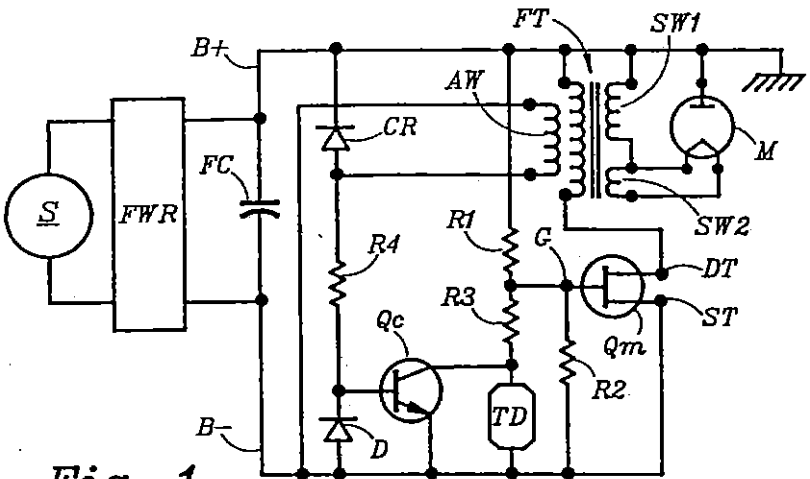


Fig. 1

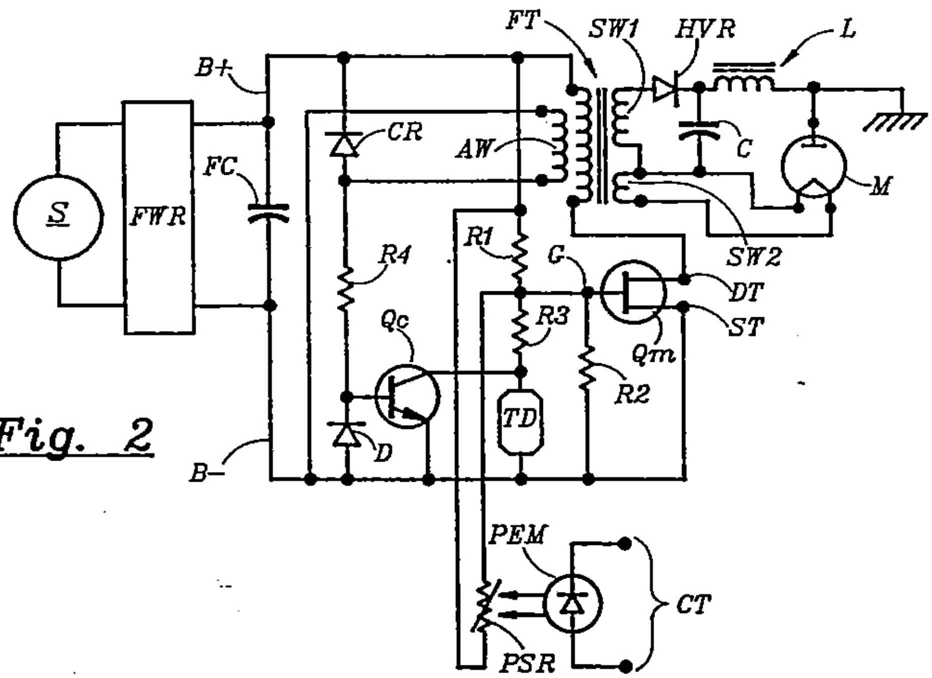


Fig. 2

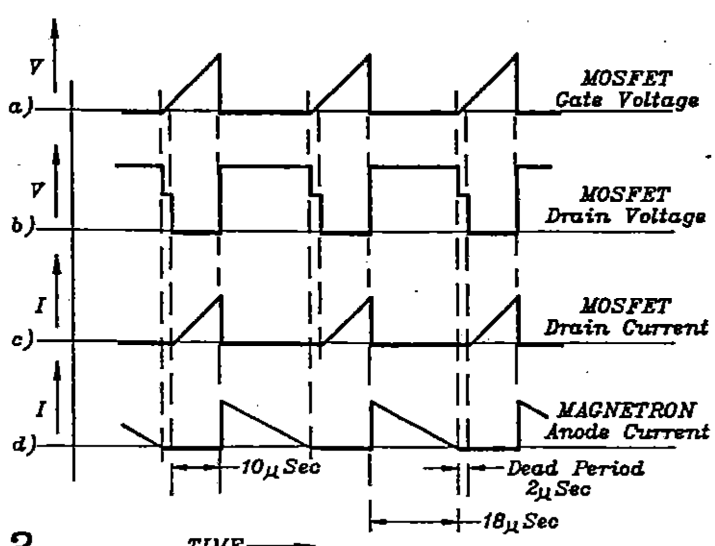


Fig. 3

MOSFET FLYBACK CONVERTER

This is a continuation of Ser. No. 812,721 filed Dec. 23, 1985, now abandoned.

BACKGROUND OF THE INVENTION

1. Field of Invention

The present invention relates to flyback converter-type power supplies, particularly of a type using MOSFET devices and having means to control the output.

2. Prior Art

Flyback converter-type power supplies using MOSFET devices are well known. However, compared with the actual drive requirements of a MOSFET in a flyback converter, present drive circuits are relatively complex and/or costly.

SUMMARY OF THE INVENTION

Objects of the Invention

An object of the present invention is that of providing a particularly simple and cost-effective drive and control circuit for a MOSFET flyback converter.

These as well as other important objects and advantages of the present invention will become apparent from the following description.

BRIEF DESCRIPTION

In a flyback converter, a 15 Amp N-channel power MOSFET (an MTM15N40 from Motorola) is driven to saturation by a gate voltage derived by connecting the gate to the relatively high-magnitude (150 Volt) B+ voltage by way of a resistor of relatively high resistance (about 20 kilo-Ohm). Current flowing through this resistor causes the gate capacitance (3600 pico-Farad) to charge at a rate of about 2 Volt per micro-second. Since the forward transconductance of this particular MOSFET is 6 mhos or more, a situation is thereby established by which the MOSFET exists in a substantially fully conductive state as long as the MOSFET's drain current does not rise at a rate higher than about 12 Ampere per micro-second.

In this connection, it should be noted that—with a B+ voltage of 150 Volt—the flyback inductor may have an inductance as low as about 15 micro-Henry without giving rise to a situation wherein the drain current rises so fast as to cause the MOSFET to go out of saturation.

Eventually, the magnitude of the gate voltage reaches a predetermined maximum level (20 Volt), at which point a threshold device, such as a low-voltage so-called Sidac, connected between gate and source, breaks down and rapidly discharges the base capacitance, thereby rapidly switching the MOSFET into a non-conducting state.

An ordinary bi-polar control transistor is also connected between gate and source, and this control transistor is made periodically conductive by a small current from a secondary winding on the flyback inductor. This control transistor makes the threshold device de-latch and also keeps the gate shorted to the source for as long as the flyback inductor is in the process of discharging its energy. Thus, since the MOSFET can not enter its conductive state until after the flyback inductor is fully discharged, effective protection against short circuiting is attained.

Adjusting the magnitude of the resistor connected between B+ and the gate causes a corresponding adjustment of the length of the period during which the MOSFET exists in a conductive state; which means that the amount of power provided at the output of the flyback converter power supply is correspondingly adjusted: the shorter the length of the MOSFET ON-time, the less power transferred.

In a typical application, the flyback converter is used for powering a magnetron in a microwave oven, and operates at a frequency adjustable around 30 kHz.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 provides a schematic circuit diagram of a basic version of the preferred embodiment of the invention.

FIG. 2 provides a schematic circuit diagram of a modified version of the preferred embodiment.

FIG. 3 shows voltage and current waveforms associated with both embodiments.

PROBLEM SITUATION UNDERLYING INVENTION

The present invention is not aimed at solving any expressly enunciated problems associated with flyback converter-type power supplies. Rather, it is based on a combination of perceptions and recognitions related to how it may be possible to improve and simplify such power supplies to a significant degree, especially in connection with the use of MOSFET devices. Some of these perceptions and recognitions are identified as follows.

1. In a MOSFET flyback-type power supply, the normal modus operandi is that of the MOSFET being used as an ON/OFF switch in such manner as periodically to connect and disconnect an inductor (the flyback inductor) with/from a source of DC voltage. When initially connected, current through the inductor, and therefore through the MOSFET, is zero; but thereafter, as long as the MOSFET conducts in an effective manner, the current rises in a substantially linear manner at a rate determined by the magnitude of the applied DC voltage divided by the inductance of the flyback inductor.

When the current through the MOSFET is low, a relatively low-magnitude gate voltage is sufficient to keep the MOSFET saturated. However, as current through the MOSFET increases, the gate voltage must exhibit a correspondingly increasing magnitude.

Thus, it is seen that it is not necessary to drive the MOSFET with a squarewave voltage—which is what is normally done in such situations. Rather, it becomes possible to drive the MOSFET to saturation by way of a relatively slowly rising voltage—such as may be attained by a simple R-C integrating arrangement, where the C could be the gate input capacitance of the MOSFET.

As a result, a fully adequate gate drive may be attained by the very simple means of connecting a resistor between the gate and the B+ terminal of the DC voltage supply.

2. When the gate voltage in a MOSFET exhibits a more-or-less linearly rising magnitude, it becomes particularly simple to provide for a control means operative to abruptly bring the magnitude of the gate voltage back down to near-zero (i.e., to discharge the base input capacitance), and thereby to efficiently and rapidly switch the MOSFET into a non-conductive state.

This control means can simply be a bistable threshold device, such as a Sidac-type or unijunction-type device, so selected or designed as to precipitously change from a relatively high resistance value to a relatively low resistance value as soon as the magnitude of the voltage across it reaches a predetermined level.

3. When the MOSFET switches into a non-conductive state, the voltage across the flyback inductor reverses its polarity; which implies that an auxiliary winding on this inductor can provide a reverse voltage for as long as the inductor is in the process of discharging its energy. This reverse voltage can then be used to cause the bistable threshold device to de-latch as well as to maintain the gate voltage at a near-zero magnitude until the inductor has finished discharging its energy.

DESCRIPTION OF THE PREFERRED EMBODIMENTS

Details of Construction

FIG. 1 shows an AC voltage source S, which in reality is an ordinary 120 Volt/60 Hz electric utility power line.

Connected to S is a full-wave rectifier FWR that rectifies the AC voltage from S and provides the rectified current to a filter capacitor FC, wherefrom is provided a substantially constant-magnitude DC voltage between a positive power bus B+ and a negative power bus B−.

Connected between the B+ bus and drain terminal DT of a MOSFET Qm is the primary winding of a flyback transformer FT. Source terminal ST of MOSFET Qm is connected with the B− bus.

Transformer FT has a first secondary winding SW1 and a second secondary winding SW2. One terminal of winding SW1 is connected with the anode of a magnetron M; which anode is connected to ground. The other terminal of winding SW1 is connected with one of the terminals of winding SW2. The two terminals of winding SW2 are connected with the two terminals of the thermionic cathode of magnetron M.

Flyback transformer FT has an auxiliary winding AW, one terminal of which is connected with the B− bus, the other terminal of which is connected with the anode of a clamping rectifier CR. The cathode of rectifier CR is connected with the B+ bus.

A first resistor R1 is connected between the B+ bus and gate G of MOSFET Qm and a second resistor R2 is connected between gate G and the B− bus. A third resistor R3 is connected in series with a threshold device TD to form a series-combination; which series-combination is connected between gate G and the B− bus.

A control transistor Qc is connected with its collector to gate G and with its emitter to the B− bus. A diode D is connected with its cathode to the base of transistor Qc and with its anode to the B− bus. A fourth resistor R4 is connected between the base of transistor Qc and the anode of clamping rectifier CR.

The circuit of FIG. 2 is identical to that of FIG. 1 except in two respects.

First, a photo-sensitive resistor PSR is connected in parallel with resistor R1; and this photo-sensitive resistor is placed near to and in photo-responsive relationship with a photo-emitting means PEM, which is connected with and actuated from a pair of control terminals CT.

Second, between the terminals of winding SW1 and magnetron M are interposed: (i) a high voltage rectifier HVR having its anode connected with one of the terminals of winding SW1, (ii) a capacitor C connected between the cathode of rectifier HVR and the other terminal of winding SW1, and (iii) an inductor L connected between the cathode of rectifier HVR and the anode of magnetron M.

Explanation of Waveforms

FIG. 3*a* shows the waveform of the voltage at gate G of MOSFET Qm as observed with reference to the B− bus.

FIG. 3*b* shows the corresponding voltage at drain terminal DT of the MOSFET.

FIG. 3*c* shows the corresponding current flowing through the primary winding of transformer FT and into drain terminal DT of the MOSFET.

FIG. 3*d* shows the corresponding current flowing out from first secondary winding SW1 and into the anode of magnetron M.

Description of Operation

The operation of the power supply arrangement of FIG. 1 may be explained as follows.

In FIG. 1, after having been connected with the power line, a substantially constant-magnitude DC voltage exists between the B+ bus and the B− bus. As a result of this DC voltage, a unidirectional current flows through resistor R1 (about 20 kOhm) and into the capacitance (about 3600 pF) of gate G, thereby causing the voltage on the gate to rise in a substantially linear manner. (The value of R2 is about 1.0 megOhm and has negligible effect on the circuit's operation as herein relevant.)

As the magnitude of the gate voltage rises—assuming a starting point of near-zero voltage—the MOSFET (Ex: Motorola type MTM15N40) will soon become conductive (after 2 micro-seconds or so and at about 3.5–4.0 Volt), and current will start to flow through the primary winding of flyback transformer FT.

The magnitude of the gate voltage will keep on rising until it reaches a magnitude (just under 20 Volt) at which threshold devide TD will break down and become conductive, thereby effectively placing resistor R3 (about 12 Ohm) in shunt between the gate and source terminal ST; which, in turn causes the gate capacitance to rapidly discharge, thereby bringing the magnitude of the gate voltage back down to near-zero within a timespan of 100 nano-seconds or so.

As long as the MOSFET existed in a saturated conductive state, substantially the full B+ voltage was applied across the primary winding of the flyback transformer; which transformer has a substantial built-in inductance; which, in turn, is to say that the transformer must have an air-gap. In other words, during this MOSFET's ON-period, the magnitue of the voltage on the drain terminal (DT) is near-zero when referenced to the source terminal (ST) or the B− bus.

As long as the B+ voltage is present across this primary winding, energy becomes stored in the inductance of the transformer; and at the end of the MOSFET conduction period—i.e., at the point just before the MOSFET ceases to conduct—the energy stored is equal to the amount of energy needed by the magnetron for each cycle of the inverter. Thus, at an inversion frequency of 33 kHz and a magnetron power requirement of 800 Watt, the energy required per inversion cycle (or per flyback cycle) is about 24 milli-Joule;

which is to say that the energy that must be stored in the inductance of the flyback transformer at the end of the period of the MOSFET being conductive, must be about 24 milli-Joule. (With a B+ voltage of 150 Volt, this implies that the magnitude of the current flowing through the MOSFET just prior to turn-off is on the order of about 16 Ampere, assuming an effective MOSFET ON-period of about 10 micro-seconds per cycle.)

With inductive energy stored in the flyback transformer, and with the gate voltage reduced to near-zero—thereby switching OFF the MOSFET—the voltage on the drain terminal (DT) rises to the point of becoming limited by whatever might be loading the transformer. Under normal operating conditions, this loading would be due to the magnetron.

The voltage transformation ratio of the flyback transformer is so arranged that the magnitude of the reverse-voltage resulting across the primary winding of the flyback transformer (during the period when the stored-up energy discharges itself into the magnetron) is about 85 Volt. As a necessary consequence, it takes about 18 micro-seconds for the inductive energy in the flyback transformer to discharge itself into the magnetron.

According to above considerations, just after the point is reached at which the threshold device (TD) breaks down and causes the magnitude of the gate voltage to drop to near-zero (thereby switching the MOSFET off), the magnitude of the voltage on the drain terminal (DT) increases from near-zero to about 85 Volt higher than the B+ voltage; which, due to the chosen primary-to-auxiliary turns-ratio, makes the magnitude of the voltage across auxiliary winding AW about 60 Volt.

With about 60 Volt present at the point to which resistor R4 is connected with the anode of rectifier CR, current starts flowing into the base of transistor Qc, which therefore will become conductive, thereby preventing the voltage on the gate of the MOSFET from rising as long as the inductive energy in the flyback transformer is being discharged (i.e., for as long as the 60 Volt is present).

As soon as the inductive energy has been completely discharged, the magnitude of the voltage feeding resistor R4 falls to near-zero; and transistor Qc now ceases to conduct. At this point, the MOSFET gate voltage starts rising again (at a rate of about 2 Volt per micro-second); and, about 2 micro-seconds later, the gate voltage will have reached a magnitude (4 Volt large enough to cause the MOSFET once more to start conducting; from which point the cycle repeats.

With transformer winding polarities as indicated, the magnetron will conduct during the period when the flyback transformer discharges its energy. However, being in effect an electronic diode, the magnetron does not conduct (between its cathode and anode) during the period when the flyback transformer is being charged up. During that time, it only draws the relatively modest level of power associated with heating the cathode.

During the short period before the magnetron is operable to represent an effective load to the flyback transformer—i.e., while the thermionic cathode is in the process of becoming incandescent—most of the energy stored in the flyback transformer will be discharged back into filter capacitor FC by way of clamping rectifier CR. During this mode of operation, the voltage present across the auxiliary winding must by necessity be equal to the B+ voltage (i.e., about 150 Volt).

The operation of the circuit arrangement of FIG. 2 is in most respects identical to that of FIG. 1. However, in addition to the obvious differences associated with the filtering of the magnetron current and the control provided by the photo-sensitive resistor (PSR) and the photo-emitting means (PEM), a few changes in timing and turns-ratios have been made.

In FIG. 2, the MOSFET ON-time has been increased to about 13 micro-seconds; and the MOSFET OFF-time has been decreased to about 15 micro-seconds (which still leaves a dead period of about 2 micro-seconds). Also, as a necessary corollary, the voltage present across the auxiliary winding during the discharge of the inductive energy from the flyback transformer has been increased from 85 Volt to about 130 Volt.

As additional consequences of these different values of MOSFET ON-time and OFF-time, the peak MOSFET drain current is reduced from 16 Ampere to about 13 Ampere, and the maximum voltage presented to the magnetron is now limited to being only a little higher than its normal operating voltage.

The operation of the control arrangement consisting of the photo-sensitive resistor and the photo-emitting means is explained as follows.

With no light provided by the photo-emitting means, the resistance of the photo-sensitive resistor is very high in comparison with that of R1; which means that the control arrangement has no effect under this condition, and that the magnetron now receives its maximum flow of power. However, as light is provided to PSR (as emitted from PEM—which, in turn, results from current provided to control terminals CT) its resistance decreases, thereby giving rise to a shortening of the time it takes for the capacitance of the MOSFET gate to charge to a given voltage level.

Thus, with light provided to the photo-sensitive resistor, the MOSFET ON-time is shortened; which implies that the power provided to the magnetron will be reduced.

In this connection, it should be noted that—while flyback conversion frequency will increase essentially as a linear function of shortened ON-time—the energy stored and transferred to the magnetron per cycle will decrease as a square function of the shortening of the ON-time; which explains why the net power provided to the load will decrease substantially in linear relationship with the decreased ON-time.

Additional Comments

(a) The waveforms of FIG. 3 are principally relevant in connection with the circuit arrangement of FIG. 1. However, except for the waveform of FIG. 3*d* and for the somewhat different proportioning of MOSFET ON-times versus OFF-times, they are also applicable to the circuit arrangement of FIG. 2.

(b) There are simple ways by which the 2 micro-second dead period (see FIG. 3) may be substantially eliminated. For instance, by the use of a Zener diode connected in series between the emitter of transistor Qc and the B− bus, and by having the one terminal of the PUT connected with this emitter rather than with the B− bus, it is readily possible to assure that the MOSFET gate capacitance never gets discharged further than necessary to assure complete MOSFET turn-off. With most presently common MOSFETS, this would imply the use of a Zener diode with a Zener voltage of about 3 Volt.

(c) Due to the filtering by capacitor C and inductor L in the circuit arrangement of FIG. 2, the anode current provided to the magnetron in that arrangement becomes continuous and substantially constant in magnitude.

(d) With but a change in the number of turns on secondary winding SW1, the circuit of FIG. 2 would be directly applicable as a conventional DC power supply or a battery charger; in which case, of course, secondary winding SW2 could be removed.

(e) Instead of using clamping rectifier CR, as combined with the DC supply voltage, as a voltage clamping means to limit the maximum magnitude of the voltage developing across the primary winding of the flyback transformer during its discharge of inductive energy, a Zener diode may be used. However, unless a scheme of inverter disablement is used, the power rating of this Zener diode would have to be quite large.

(f) Threshold device TD can be any one of a variety of devices, such as: (i) a high-frequency SCR combined with a Zener diode; (ii) a so-called Sidac; (iii) a unijunction transistor; (iv) a programmable unijunction transistor (PUT); (v) a Silicon Controlled Switch; (vi) a custom-made IC threshold means; etc. The particular choice in a given application would depend on the particular characteristics desired in terms of speed, threshold voltage, forward conductivity and voltage drop, etc.

(g) It is believed that the present invention and its several attendant advantages and features will be understood from the preceeding description. However, without departing from the spirit of the invention, changes may be made in its form and in the construction and interrelationships of its component parts, the form herein presented merely representing the presently preferred embodiment.

I claim:

1. Converter means adapted to convert a DC voltage to an alternating voltage, comprising:

transformer means having a primary winding;

switch means having a pair of switched terminals and a pair of control terminals, the switch means being operative to permit relatively unimpeded flow of current between the switched terminals, but only as long as: (i) a control voltage is applied between its control terminals, and (ii) the magnitude of this control voltage exceeds a minimum level, this minimum level being approximately proportional to the magnitude of the current flowing between the switched terminals;

connect means operative to connect the primary winding and the switched terminals in circuit with the DC voltage in such manner that, as long as the switch means is operative to permit relatively unimpeded flow of current between its switched terminals, the DC voltage is effectively applied across this primary winding; and

control means connected with the control terminals and operative to provide a control voltage having a periodically varying magnitude that, whenever current is flowing between the switched terminals, is: (i) substantially proportional to the magnitude of this current, and (ii) in excess of said minimum level.

2. The converter means of claim 1 wherein periodically the magnitude of the control signal is abruptly reduced below the the minimum level for a brief period of time, thereby to prevent current from flowing freely between the switched terminals.

3. The converter means of claim 1 wherein the current flowing between the switched terminals and the control voltage are both substantially characterized by consisting of periodically-occurring triangularly shaped pulses.

4. The converter means of claim 1 wherein a load means is connected with the transformer means, this load means being operative to absorb energy from the transformer means, but substantially only during periods when no current flows between the switched terminals.

5. The converter means of claim 1 wherein the transformer means comprises inductive energy-storing means.

6. A DC-to-DC converter adapted to be powered from a DC voltage and to provide DC power to a load, comprising:

inductor means;

switch means connected in series with the inductor means to form a series-combination, the series-combination being connected across the DC voltage, the switch means having control means operative on receipt of a control voltage at a pair of control terminals to render the switch means: (i) substantially non-conductive as long as the magnitude of the control voltage does not exceed a threshold level; and (ii) substantially fully conductive as long as the magnitude of the control voltage exceeds the threshold level by an amount that is substantially proportional to the magnitude of the current flowing through the switch means;

control source operative to provide a control voltage to the control terminals, this control voltage alternating periodically in magnitude between a first state of being below the threshold level to a second state of being above the threshold level, the second state being characterized by the magnitude of the control voltage being above the threshold level by an amount substantially proportional to the magnitude of the current flowing through the switch means; and

load circuit means operative to connect the load in circuit with the inductor means in such manner as to cause the load to receive DC power substantially only during periods when the switch means is substantially non-conductive.

7. The DC-to-DC converter means of claim 6 wherein the control signal is further characterized by having a waveshape comprising a first portion of substantially linearly rising magnitude followed by a relatively rapid drop in magnitude to a second portion of substantially non-varying magnitude.

8. The DC-to-DC converter of claim 6 wherein the control source comprises a threshold device operative, whenever the magnitude of the voltage between the control terminals reaches a certain magnitude, to cause a shunt of relatively low impedance to be placed across the control terminals.

9. The DC-to-DC converter of claim 6 wherein the magnitude of the control voltage is prevented from exceeding the threshold level as long as there is any substantive amount of energy stored in the inductor means.

10. A flyback converter adapted to be powered from a DC voltage and comprising:

a semiconductor switching device having a pair of switched terminals and a pair of control terminals, electric current being: (i) permitted to flow substantially freely between the switched terminals as long as the control terminals are provided with a control voltage of minimum magnitude that exceeds a threshold level by an amount that is substantially proportional to the magnitude of any current flowing between the switched terminals; and (ii) substantially prevented from flowing between the switched terminals as long as the control terminals are provided with a control voltage of magnitude less than the threshold level;

an inductor means connected between the DC voltage and the switched terminals in such manner that, as long as current is permitted to flow freely between the switched terminals, substantially the full magnitude of the DC voltage is applied across the inductor means, thereby to cause current to flow through the inductor means and therefore between the switched terminals; and

control means operative to provide a control voltage to the control terminals, this control voltage alternating periodically between: (i) having a magnitude less than the threshold level, thereby substantially preventing current from flowing between the switched terminals; and (ii) having, substantially for as long as current is flowing between the switched terminals, a magnitude that exceeds the threshold level by an amount that is substantially proportional to the magnitude of the current flowing between the switched terminals.

11. The flyback converter of claim 10 wherein:

(i) the control means is operative to provide the control voltage to the control terminals by way of a current-limiting means; and

(ii) a bistable threshold means is connected across the control terminals, this threshold means being operative to change from a first state constituting a relatively high resistance to a second state constituting a relatively low resistance as soon as the magnitude of any voltage present across it reaches

a predetermined level, thereby to cause the magnitude of the voltage present across the control terminals to precipitously decrease as soon as the magnitude of the control voltage reaches the predetermined level.

12. A flyback converter adapted to be powered from a DC voltage and to provide power to a load, comprising:

a semiconductor switching device having a pair of switched terminals and a pair of control terminals receptive of a control voltage, electric current being: (i) permitted to flow between the switched terminals as long as the magnitude of the control voltage exceeds a threshold level; and (ii) prevented from flowing between the switched terminals as long as the magnitude of the control voltage is below the threshold level;

inductor means connected between the DC voltage and the switched terminals in such manner that, as long as current is permitted to flow between the switched terminals, inductive energy becomes stored in the inductor means;

load means connected with the inductor means and operative to absorb inductive energy therefrom, but only during periods when current is prevented from flowing between the switched terminals; and

control means operative to provide a control voltage to the control terminals, this control voltage alternating periodically between: (i) having a magnitude in excess of the threshold level, thereby causing inductive energy to become stored in the inductor means; and (ii) having a magnitude less than the threshold level, thereby preventing current from flowing between the switched terminals and therefore causing any inductive energy stored in the inductor means to be absorbed by the load means; the control means being operative to prevent the magnitude of the control voltage from exceeding the threshold level as long as inductive energy is being absorbed by the load means.

* * * * *

EXHIBIT B

United States Patent [19]

Nilssen

[11] **Patent Number:** **4,935,669**

[45] **Date of Patent:** **Jun. 19, 1990**

[54] **TWO-MODE ELECTRONIC BALLAST**

[76] Inventor: Ole K. Nilssen, Caesar Dr., Barrington, Ill. 60010

[21] Appl. No.: **145,925**

[22] Filed: **Jan. 20, 1988**

[51] **Int. Cl.⁵** **H05B 41/29**

[52] **U.S. Cl.** .... **315/105**; 315/205; 315/209 R; 315/278; 315/DIG. 5; 315/DIG. 7

[58] **Field of Search** .... 315/DIG. 5, 209 R, 291, 315/DIG. 7, 205, 278, 105

[56] **References Cited**

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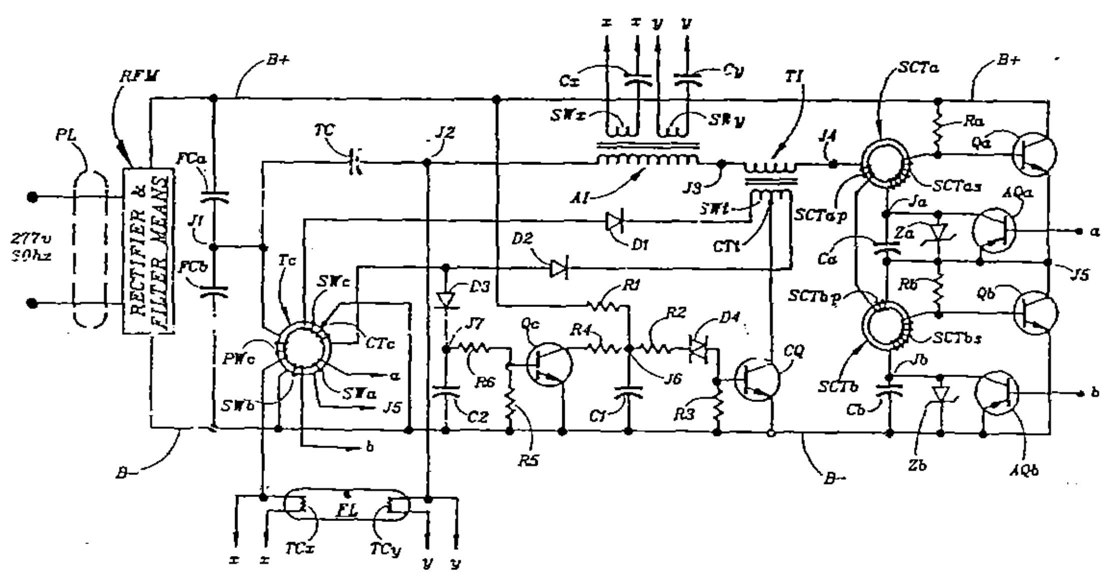
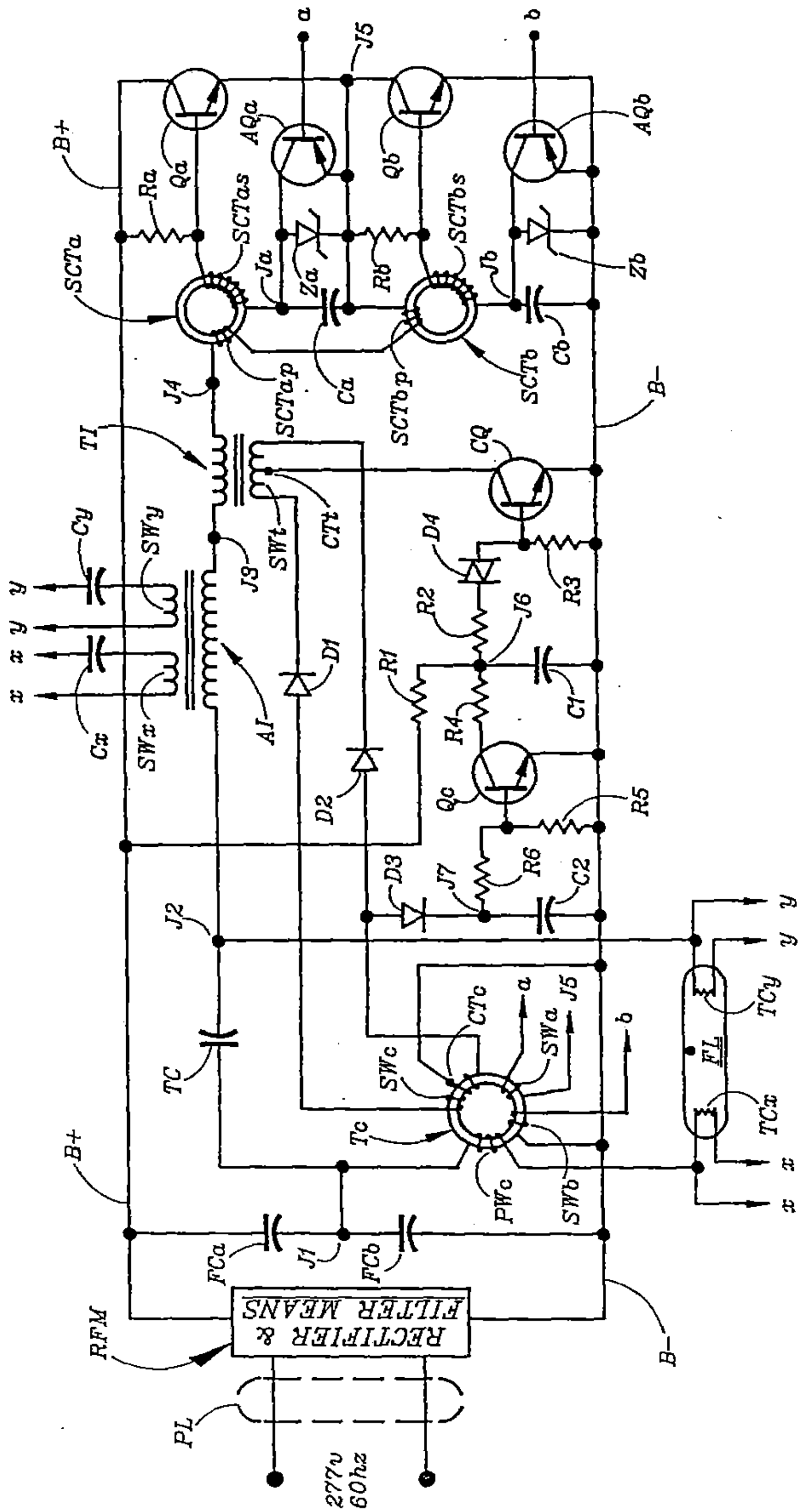
Primary Examiner—Eugene R. LaRoche
Assistant Examiner—Michael B. Shingleton

[57] **ABSTRACT**

A self-oscillating inverter-type fluorescent lamp ballast has two modes of operation: (a) a first mode in which the inversion frequency is about 70 kHz and is resonant with a first tuned circuit by which power is supplied to the cathodes of the fluorescent lamp; and (b) a second mode in which the inversion frequency is about 30 kHz and is resonant with a second tuned circuit by which main lamp power is supplied. When the ballast is initially powered-up, it starts operation in its first mode, thereby providing cathode heating power without yet providing main lamp power. About one second later, after the cathodes have reached full incandescence, the inverter automatically changes into its second mode, thereby providing main lamp power while at the same time removing cathode heating power. If for some reason the lamp were not to ignite within about 10 milliseconds, the inverter reverts back into its first mode; thereafter cycling between its two modes until the lamp does ignite.

16 Claims, 1 Drawing Sheet





RECTIFIER & FILTER MEANS
RFM
PL
277v 60hz
B+
B-
FCa
FCb
J1
J2
J3
J4
J5
J6
J7
TC
Tc
PWc
SWa
SWb
SWc
CTc
FL
TCx
TCy
SCTa
SCTb
SCTas
SCTap
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Qa
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AQb
Ra
Rb
Za
Zb
Ja
Jb
Ca
Cb
Ct
T1
AI
SWt
SWx
SWy
Cx
Cy
CQ
Qc
D1
D2
D3
D4
R1
R2
R3
R4
R5
R6
C1
C2
a
b
x
y

TWO-MODE ELECTRONIC BALLAST

BACKGROUND OF THE INVENTION

1. Field of the Invention

The present invention relates to electronic ballasts for gas discharge lamps, particularly to ballasts wherein the load is powered by way of a series-excited parallel-loaded resonant L-C circuit.

2. Description of Prior Art

There are two predominant types of electronic ballasts for gas discharge lamps: (a) a first type may be referred-to as the parallel-resonant type and involves the use of a current-excited (i.e., parallel-excited) parallel-loaded resonant L-C circuit; and (b) a secnd type that may be referred-to as the series-resonant type and involves the use of a voltage-excited (i.e., series-excited) parallel-loaded resonant L-C circuit.

An example of the parallel-resonant type of electronic ballasts is described in U.S. Pat. No. 4,277,726 to Burke. An example of the series-resonant type of electronic ballasts is described in U.S. Pat. No. 4,538,095 to Nilssen.

Of these two types of electronic ballasts, the parallel-resonant type is conducive to yielding a stable easy-to-control self-oscillating inverter-type ballast; whereas the series-resonant type, although potentially simpler and more efficient, is harder to control in that it has a natural tendency to self-destruct in case the lamp load be removed.

To mitigate this tendency to self-destruct under no-load conditions, various protection circuits have been developed, such as for instance described in U.S. Pat. No. 4,638,562 to Nilssen.

GENERAL PURPOSE OF PRESENT INVENTION

The general purpose of the present invention is that of providing a method for cost-effectively controlling the operation of a series-resonant electronic inverter-type ballast for fluorescent lamps.

SUMMARY OF THE INVENTION

1. Objects of the Invention

An object of the present invention is the provision of a cost-effective control arrangement for attaining proper operation of an electronic ballast wherein the lamp load is powered by way of a series-excited predominantly parallel-loaded resonant L-C circuit.

This as well as other objects, features and advantages of the present invention will become apparent from the following description and claims.

2. Brief Description

A self-oscillating inverter-type fluorescent lamp ballast has two modes of operation: (a) a first mode in which the inversion frequency is about 70 kHz and is resonant with a first tuned L-C circuit by which power is supplied to the cathodes of the fluorescent lamp; and (b) a second mode in which the inversion frequency is about 30 kHz and is resonant with a second tuned L-C circuit by which main lamp power is supplied.

When the ballast is initially powered-up, it starts operation in its first mode, thereby providing cathode heating power without yet providing main lamp power. About one second later, after the cathodes have reached full incandescence, the inverter automatically changes into its second mode, thereby providing main lamp power while at the same time removing cathode heating power. If for some reason the lamp were not to ignite within about 10 milli-seconds, the inverter reverts back into its first mode; thereafter cycling (with a period of about one second) between its two modes until the lamp does ignite.

Thus, the first tuned L-C circuit is resonant at 70 kHz; and, due to inherent frequency-selectivity characteristics, this first tuned circuit provides cathode heating power only when being excited at or near 70 kHz. Likewise, the second tuned L-C circuit provides main lamp starting voltage and operating power only when being excited at or near 30 kHz.

BRIEF DESCRIPTION OF THE DRAWING

The drawing diagrammatically illustrates the circuit arrangement of the invention in its preferred embodiment.

DESCRIPTION OF THE PREFERRED EMBODIMENT

1. Details of Construction

The drawing schematically illustrates the preferred embodiment of the invention in the form of a half-bridge inverter-type two-mode electronic ballast for a fluorescent lamp.

In the drawing, 277 Volt/60 Hz power line voltage from an ordinary electric utility power line PL is provided to the AC power input terminals of a rectifier and filter means RFM, the DC output from which is applied between a B+ bus and a B− bus.

A filter capacitor FCa is connected between the B+ bus and a junction J1; a filter capacitor FCb is connected between junction J1 and the B− bus. A tank capacitor TC is connected between junction J1 and a junction J2. An auxiliary inductor AI is connected between junction J2 and a junction J3; and a main tank inductor TI is connected between junction J3 and a junction J4.

Junction J4 is connected with a junction J5 by way of series-connected primary windings SCTap and SCTbp of saturable current transformers SCTa and SCTb, respectively.

A first main inverter transistor Qa is connected with its collector to the B+ bus and with its emitter to junction J5; a second main inverter transistor Qb is connected with its collector to junction J5 and with its emitter to the B− bus.

Secondary winding SCTas of saturable current transformer SCTa is connected between the base of transistor Qa and a junction Ja. A capacitor Ca is connected between junctions Ja and J5. A Zener diode Za is connected with its anode to junction Ja and with its cathode to junction J5. An auxiliary transistor AQa is connected with its collector to junction Ja and with its emitter to junction J5. A resistor Ra is connected between the B+ bus and the base of transistor Qa. The base of auxiliary transistor AQa is designated a.

Secondary winding SCTbs of saturable current transformer SCTb is connected between the base of transistor Qb and a junction Jb. A capacitor Cb is connected between junction Jb and the B− bus. A Zener diode Zb is connected with its anode to junction Jb and with its cathode to the B− bus. An auxiliary transistor AQb is connected with its collector to junction Jb and with its emitter to the B− bus. A resistor Rb is connected between junction J5 and the base of transistor Qb. The base of auxiliary transistor AQb is designated b.

Tank inductor TI has a secondary winding SWt, which has a center tap CTt connected with the collector of a control transistor CQ. The emitter of control transistor CQ is connected with the B− bus.

The terminals of secondary winding SWt are connected with the cathodes of two diodes D1 and D2; whose anodes are connected with the terminals of a secondary winding SWc of a control transformer Tc; which secondary winding has a center tap CTc connected with the B− bus.

A fluorescent lamp FL has two thermionic cathodes TCx and TCy; which has power input terminals x—x and y—y, respectively. One of the power input terminals of cathode TCx is connected with junction J1 by way of primary winding PWc of control transformer Tc. One of the power input terminals of cathode TCy is connected with junction J2.

Power input terminals x—x and y—y of cathodes TCx and TCy are connected with power output terminals x—x and y—y of secondary windings SWx and SWy of auxiliary inductor AI, all respectively; which secondary windings have series-connected capacitors Cx and Cy, also respectively.

A resistor R1 is connected between the B+ bus and a junction J6; and a capacitor C1 is connected between junction J6 and the B− bus. A resistor R2 and a Diac D4 are connected in series between junction J6 and the base of control transistor CQ. A resistor R3 is connected between the base of transistor CQ and the B− bus.

A resistor R4 is connected between junction J6 and the collector of a transistor Qc, whose emitter is connected with the B− bus. A resistor R5 is connected between the base of transistor Qc and the B− bus. A resistor R6 is connected between the base of transistor Qc and a junction J7. A capacitor C2 is connected between junction J7 and the B− bus. A diode D3 is connected with its anode to the anode of diode D2 and with its cathode to junction J7.

Control transformer Tc also has two secondary windings SWa and SWb. The terminals of secondary winding SWa are connected between base a of transistor AQa and junction J5; the terminals of secondary winding SWb are connected between the B− bus and base b of transistor AQb.

2. Details of Operation

The operation of the circuit arrangement schematically illustrated by the drawing may be explained as follows.

In the arrangement of the drawing, ordinary 277 Volt/60 Hz power line voltage is provided from the power line (PL) and is rectified and filtered by conventional rectifier and filter means RFM such as to provide a DC voltage between the B+ and the B− buses, with the B+ bus carrying the positive polarity.

The half-bridge inverter, which principally consists of capacitors FCa and FCb, transistors Qa and Qb, and saturable current feedback transformers SCTa and SCTb, is self-oscillating and functions in a substantially ordinary manner, such as for instance described in conjunction with FIG. 8 of U.S. Pat. No. Re. 31,758 to Nilssen.

The output of the half-bridge inverter is provided to and between junctions J1 and J4; between which junctions are connected in series: tank capacitor TC, tank inductor TI, and auxiliary inductor AI.

Auxiliary inductor AI is tuned to about 70 kHz by way of capacitors Cx and Cy; which two capacitors are connected with the secondary windings of the auxiliary inductor as well as with the loads connected to the output of these secondary windings. Thus, only when lamp cathodes TCx and TCy are indeed connected with the two secondary windings is the auxiliary inductor tuned to about 70 kHz. As a result, at about 70 kHz, the auxiliary inductor appears like a parallel-resonant circuit as viewed from between junctions J2 and J3.

Tank inductor TI is tuned to series-resonate with tank capacitor TC at about 30 kHz; which is to say that the total impedance between junctions J1 and J4 appears substantially like a series-resonant circuit at about 30 kHz.

At 30 kHz, the impedance of auxiliary inductor AI is inductive and relatively small, and is at that frequency simply considered as a small part of tank inductor TI.

At 70 kHz, the impedance of tank capacitor TC is capacitive and relatively small, whereas the impedance of tank inductor TI is inductive and relatively high.

Thus, when the inverter oscillates at 70 kHz, its output voltage is applied by way of high-impedance tank inductor TI to the parallel-resonant circuit represented by auxiliary inductor AI; which parallel-resonant circuit then operates to power the two thermionic cathodes of fluorescent lamp FL. During this mode, the power provided to these two cathodes is about two watts, and the magnitude of the current then drawn from the inverter output is quite small. As a result, the magnitude of the 70 kHz voltage resulting across tank capacitor TC is very small.

On the other hand, when the inverter oscillates at 30 kHz, essentially no power is provided to the thermionic cathodes. However, at that frequency, the resonant series-tuned L-C circuit then loading the inverter's output causes a 30 kHz voltage of very large magnitude to develop across the tank capacitor. The magnitude of this 30 kHz voltage is so large as to cause the fluorescent lamp to ignite; whereafter the magnitude of the 30 kHz voltage across the tank capacitor will be determined by the current-voltage characteristices of the fluorescent lamp. In reality, at the 30 kHz series-resonance, the output provided from the output terminals to the fluorescent lamp (i.e., from junctions J1 and J2) will essentially be a 30 kHz constant-magnitude current.

The frequency of inverter oscillation is determined by the saturation characteristics of saturable current transformers SCTa and SCTb in conjunction with the magnitude of the voltage presented to their secondary windings SCTas and SCTbs.

The magnitude of the voltage presented to secondary windings SCTas and SCTbs will be determined by the base-emitter voltage of transistors Qa and Qb in combination with the magnitude of the voltage present at junctions Ja and Jb as referenced to the emitters of transistors Qa and Qb, respectively.

With no control signals provided to the bases a and b of auxiliary transistors AQa and AQb, the magnitude of the voltage at junctions Ja and Jb will be determined by the Zener voltages of Zener diodes Za and Zb; which Zener voltages are chosen to be about 4.0 Volt each. However, when sufficient control current is provided to each of bases a and b, transistors AQa and AQb become conductive and therefore operative to shunt Zener diodes Za and Zb, thereby to cause the magnitudes of the voltages at junctions Ja and Jb to become very low (about 1.0 Volt each).

Thus, absent control currents at bases a and b, the inverter will oscillate at about 70 kHz; whereas, with control currents, the inverter will oscillate at about 30 kHz.

When the inverter is initially powered-up, no lamp current is flowing through the primary winding of control transformer Tc and control transistor CQ is non-conductive; which means that no control currents are provided to bases a and b of transistors AQa and AQb. Thus, when initially powered-up, the inverter will initiate oscillations at a frequency of about 70 kHz.

However, after about one second, capacitor C1 will have reached a voltage high enough to cause Diac D4 to break down; which, in turn, causes capacitor C1 to discharge into the base of control transistor CQ, thereby causing this transistor to become conductive.

Control transistor CQ will remain conductive for a period of about 10 milli-seconds; and, during this period, current from secondary winding SWt of tank inductor TI will flow through secondary winding SWc of control transformer Tc; thereby—via secondary windings SWa and SWb on control transformer Tc—providing control currents to bases a and b of transistors AQa and AQb; thereby causing capacitors Ca and Cb to discharge to a voltage level of about 1 Volt; thereby, in turn, to cause the inverter's oscillating frequency to become about 30 kHz.

With the inverter frequency at 30 kHz, the magnitude of the voltage provided between the lamp's cathodes becomes large enough to cause lamp ignition within the 10 milli-second period; which, in turn, gives rise to the flow of lamp current; which lamp current flows through primary winding PWc of control transformer Tc, thereby continuing to provide control currents to bases a and b of transistors AQa and AQb; thereby continuing to maintain the inverter's oscillation frequency at 30 kHz.

On the other hand, if the fluorescent lamp were to fail to ignite within the 10 milli-second time-window during which the control transistor CQ be conductive, control currents to bases a and b would not be sustained; thereby causing the inverter to revert to its 70 kHz oscillating frequency.

In short, with a properly operational fluorescent lamp connected, the ballast arrangement of the drawing operates as follows.

(1) Upon initial connection to the power line, the inverter starts oscillating at a 70 kHz frequency; which, via a 70 kHz resonating circuit associated with auxiliary inductor AI, therefore causes cathode heating power to be provided to the thermionic cathodes of the fluorescent lamp.

(2) After about one second, at which time the cathodes are fully thermionic, control transistor CQ suddenly becomes conductive and thereafter remains conductive for a period of about 10 milli-seconds. With transistor CQ conductive, control current is provided to auxiliary transistors AQa/AQb; which then become conductive, thereby to cause a reduction in the magnitudes of the voltages across capacitors Ca/Cb; which, in turn, causes the frequency of inverter oscillation to reduce to 30 kHz and to remain at 30 kHz for at least 10 milli-seconds.

(3) With the inverter oscillating at 30 kHz, series-resonance occurs between tank capacitor TC and tank inductor TI (including the net inductance of auxiliary inductor AI and its associated circuitry); which series-resonance, due to so-called Q-multiplication effects, results in a high-magnitude 30 kHz voltage developing across the tank capacitor.

(4) The high-magnitude 30 kHz voltage developing across the tank capacitor is applied across the fluorescent lamp and, because its cathodes are already thermionic, causes it to ignite immediately. The resulting lamp current will then, via control transformer Tc, continue to provide control current to auxiliary transistors AQa/AQb; thereby, even after the initial 10 millisecond period, ensuring that the inverter's frequency of oscillation remains at 30 kHz.

(5) When the inverter is operating at 30 kHz, essentially no power is being delivered to the cathodes of the fluorescent lamp, thereby providing for improved energy efficiency as compared with the situation where cathode power be supplied on a continuous basis.

(6) If the lamp were to be removed or if lamp current otherwise were to fail to flow, control current would cease to be provided to auxiliary transistors AQa/AQb, thereby causing the inverter's oscillation frequency to revert to 70 kHz. Thus, as long as no lamp current is flowing, the inverter will alternate between two modes: a first mode of oscillating at 70 kHz and a second mode of oscillating at 30 kHz, spending about one second (1000 milli-seconds) at 70 kHz for each 10 milliseconds at 30 kHz.

3. Additional Comments

(a) To protect against possible self-destruction of the inverter circuit (which might occur if the circuit were to operate for a period of time without being connected with a properly functioning lamp load), it may be advantageous to connect a voltage-limiting means, such as a Varistor, in parallel with the tank capacitor.

(b) For further details relative to the biasing arrangement used in connection with main inverter transistors Qa/Qb, reference is made to FIG. 3 of U.S. Pat. No. 4,307,353 to Nilssen.

(c) By providing for additional levels of adjustment for the magnitude of the bias voltage (i.e., the voltage across capacitors Ca/Cb), corresponding adjustment of the magnitude of lamp current may be attained, thereby to provide for lamp dimming.

(d) It is believed that the present invention and its several attendant advantages and features will be understood from the preceeding description. However, without departing from the spirit of the invention, changes may be made in its form and in the construction and interrelationships of its component parts, the form herein presented merely representing the preferred embodiment.

I claim:

1. An arrangement comprising:

self-oscillating inverter means connected with a source of DC voltage and operative to provide an inverter voltage at an inverter output, the inverter voltage having a frequency, the inverter means being self-oscillating by way of positive feedback means and having control means operative in response to a control input to control the self-oscillation frequency;

gas discharge lamp means having: (i) main lamp terminals operative to receive main lamp operating power, and (ii) thermionic cathode means having cathode terminals operative to receive cathode heating power;

impedance means connected in circuit between the inverter output, the main lamp terminals, and the cathode terminals, the impedance means being operative to supply from the inverter output main lamp operating power to the main lamp terminals and cathode heating power to the cathode terminals, the amount of main lamp operating power and the amount of cathode heating power supplied both being dependent on the frequency of the inverter voltage; and

control means operative to provide the control input in such manner as to control the frequency of the inverter voltage, thereby to control the amount of main lamp operating power as well as the amount of cathode heating power.

2. The arrangement of claim 1 wherein: (i) the amount of cathode heating power supplied to the cathode terminals is a first function of the frequency of the inverter voltage, and (ii) the amount of main lamp operating power is a second function of the frequency of the inverter voltage, the second function being substantively different from the first function.

3. The arrangement of claim 2 wherein the amount of cathode heating power supplied decreases as the amount of main lamp operating power increases.

4. The arrangement of claim 1 wherein the impedance means comprises a first and a second tuned circuit means, the first tuned circuit means being operative to determine the amount of cathode heating power being supplied to the cathode terminals, the second tuned circuit means being operative to determine the amount of main lamp operating power being supplied to the main lamp terminals.

5. The arrangement of claim 1 combined with current sensing means connected in circuit with the main lamp terminals as well as with the control means, the current sensing means being operative to sense lamp current flowing between the main lamp terminals and, in response to this lamp current, to provide at least part of the control input.

6. The arrangement of claim 1 wherein the impedance means comprises a series-tuned L-C circuit connected across the inverter output.

7. An arrangement comprising:

inverter means connected with a source of DC voltage and operative to provide an inverter voltage at an inverter output, the inverter voltage having a frequency, the inverter means having control means operative in response to a control input to control this frequency;

gas discharge lamp means having: (i) main lamp terminals operative to receive main lamp operating power, and (ii) thermionic cathode means having cathode terminals operative to receive cathode heating power;

impedance means connected in circuit between the inverter output, the main lamp terminals, and the cathode terminals, the impedance means being operative to supply from the inverter output main lamp operating power to the main lamp terminals and cathode heating power to the cathode terminals, the amount of main lamp operating power and the amount of cathode heating power supplied both being dependent on the frequency of the inverter voltage; and

control means operative to provide the control input in such manner as to control the frequency of the inverter voltage between a first frequency and a second frequency;

such that the arrangement is operative to provide: (i) a substantive amount of cathode heating power but only a negligible amount of main lamp operating power at the first frequency, and (ii) a negligible amount of cathode heating power but a substantive amount of lamp operating power at the second frequency.

8. An arrangement comprising:

inverter means connected with a source of DC voltage and operative to provide an inverter voltage at an inverter output, the inverter voltage having a frequency, the inverter means having control means operative in response to a control input to control this frequency;

gas discharge lamp means having: (i) main lamp terminals operative to receive main lamp operating power, and (ii) thermionic cathode means having cathode terminals operative to receive cathode heating power;

impedance means connected in circuit between the inverter output, the main lamp terminals, and the cathode terminals, the impedance means being operative to supply from the inverter output main lamp operating power to the main lamp terminals and cathode heating power to the cathode terminals, the magnitude of the main lamp operating voltage and the magnitude of the cathode heating voltage both being dependent on the frequency of the inverter voltage; and

control means operative to provide the control input in such manner as to control the frequency of the inverter voltage between a first frequency and a second frequency;

such that the arrangement is operative to provide: (i) at the first frequency, a substantive magnitude of cathode heating voltage but only a negligible magnitude of main lamp operating voltage, and (ii) at the second frequency, a substantive magnitude of main lamp operating voltage.

9. In a power supply means connected with and operative to power a gas discharge lamp, the lamp having a pair of main lamp terminals and a cathode, the cathode having a pair of cathode terminals, an improvement comprising:

(1) control input means operative in response to a control input to cause the power supply means to function in either of two modes:

(a) a first mode wherein: (i) a first cathode voltage is provided to the cathode terminals, and (ii) a first lamp voltage is provided to the main lamp terminals, the magnitude of the first lamp voltage being insufficient to cause lamp ignition; and

(b) a second mode wherein: (i) a second cathode voltage is provided to the cathodes, the magnitude of the second cathode voltage being substantially lower than that of the first cathode voltage, and (ii) a second lamp voltage is provided to the main lamp terminals, the magnitude of the second lamp voltage being sufficient to cause lamp ignition; and

(2) control output means connected with the control input means and operative to provide the control input, thereby to cause the power supply means to exist in the first mode for a period of time before causing it to change to the second mode.

10. The improvement of claim 9 wherein the magnitude of the second lamp voltage is substantially larger than that of the first lamp voltage.

11. The improvement of claim 9 wherein the first lamp voltage has a first frequency and the second lamp voltage has a second frequency, the first frequency being different from the second frequency.

12. An arrangement comprising:

power supply means operative to power a gas discharge lamp, the gas discharge lamp having a pair of main lamp terminals and a thermionic cathode, the thermionic cathode having a pair of cathode terminals, the power supply means having: (i) a first and a second pair of output terminals operative, respectively, to connect with the main lamp terminals and the cathode terminals and to provide thereto, respectively, a main lamp voltage and a cathode voltage, and (ii) control input means receptive of a control input and operative in response thereto to control the magnitude of the main lamp voltage; and

control output means connected with the control input means and operative to provide the control input such as to cause the power supply means to exist in either of two modes:

(1) a first mode in which: (i) the cathode voltage is of a first magnitude sufficient to cause the thermionic cathode to become incandescent, and (ii) the main lamp voltage is of a magnitude sufficient to cause the gas discharge lamp to ignite; and

(2) a second mode in which: (i) the cathode voltage is of a second magnitude, the second magnitude being substantially lower than the first magnitude, and (ii) the main lamp voltage is of a magnitude sufficient to cause the gas discharge lamp to ignite.

13. The arrangement of claim 12 wherein: (i) at any given time, the frequency of the main lamp voltage is the same as that of the cathode voltage, and (ii) the frequency of the main lamp voltage during the first mode is different from the frequency of the main lamp voltage during the second mode.

14. The arrangement of claim 12 wherein the control input means is operative to control the magnitude of the main lamp voltage by way of controlling its frequency.

15. The arrangement of claim 12 wherein the power supply comprises tuned circuit means.

16. An arrangement comprising:

inverter operative to provide an inverter output voltage at an inverter output means; the inverter having switching transistor means including transistor

drive input means; the inverter being self-oscillating by way of positive feedback means connected in circuit between the inverter output means and the transistor drive input means; the inverter having frequency control input means operative, on receipt of a control signal, to control the frequency of the inverter output voltage;

gas discharge lamp having: (i) main lamp terminals receptive of a main lamp voltage, and (ii) a thermionic cathode having cathode terminals receptive of a cathode heating voltage;

frequency-discriminating circuit means connected between: (i) the inverter output means, (ii) the lamp input terminals, and (iii) the cathode terminals; the frequency-discriminating circuit means being operative to cause a cathode voltage to be applied at the cathode terminals and a main lamp voltage to be applied at the main lamp terminals; the magnitude of the cathode voltage being a first function of the frequency; the magnitude of the main lamp voltage being a second function of frequency; the second function being substantially different from the first function; and

control means connected with the frequency control input means and operative to provide said control signal, thereby to cause the frequency of the inverter output voltage to vary such as: (i) initially to cause the cathode voltage to be of a first magnitude and the main lamp voltage to be of a second magnitude, such as to provide a sufficient amount of cathode heating power to the termionic cathode to reach incandescence without causing the gas-discharge lamp to ignite, and (ii) subsequently, after the termionic cathode has reached incandescence, to cause the first magnitude to decrease by a substantial degree while at the same time to cause the second magnitude to increase substantially, thereby to cause the gas discharge lamp to ignite and operate while at the same time causing the amount of cathode heating power provided to the thermionic cathode to decrease to a substantial degree.

* * * * *

UNITED STATES PATENT AND TRADEMARK OFFICE

CERTIFICATE OF CORRECTION

PATENT NO. : 4,935,669

DATED : June 19, 1990

INVENTOR(S) : Ole K. Nilssen

It is certified that error appears in the above-indentified patent and that said Letters Patent is hereby corrected as shown below:

Claim 12, Column 9, line 22, "lamp voltage is of a magnitude sufficient to cause", should read --lamp voltage is of a magnitude insufficient to cause--.

Signed and Sealed this

Third Day of May, 1994

Attest:

Attesting Officer

BRUCE LEHMAN

Commissioner of Patents and Trademarks

EXHIBIT C

United States Patent [19]

Nilssen

[11] Patent Number: 5,049,787

[45] Date of Patent: Sep. 17, 1991

[54] CONTROLLED ELECTRONIC BALLAST

[76] Inventor: Ole K. Nilssen, Caesar Dr., Barrington, Ill. 60010

[21] Appl. No.: 312,217

[22] Filed: Feb. 21, 1989

Related U.S. Application Data

[63] Continuation-in-part of Ser. No. 80,865, Aug. 3, 1987, Pat. No. 4,819,146, and a continuation-in-part of Ser. No. 730,596, May 6, 1985, abandoned, which is a continuation-in-part of Ser. No. 640,240, Aug. 13, 1984, abandoned, which is a continuation of Ser. No. 412,771, Aug. 30, 1982, abandoned, said Ser. No. 80,865, is a continuation-in-part of Ser. No. 917,788, Oct. 10, 1986, Pat. No. 4,727,470.

[51] Int. Cl.[5] H05B 37/02; H05B 39/04; H05B 37/02; G05F 1/00

[52] U.S. Cl. .... 315/209 R; 315/224; 315/291; 315/307; 315/DIG. 5; 315/DIG. 7

[58] Field of Search .... 315/DIG. 7, DIG. 5, 315/291, 224, 307, 209 R

[56] References Cited

U.S. PATENT DOCUMENTS

4,079,292 3/1978 Kaneda 315/DIG. 7

4,245,177 1/1981 Schmitz 315/DIG. 7

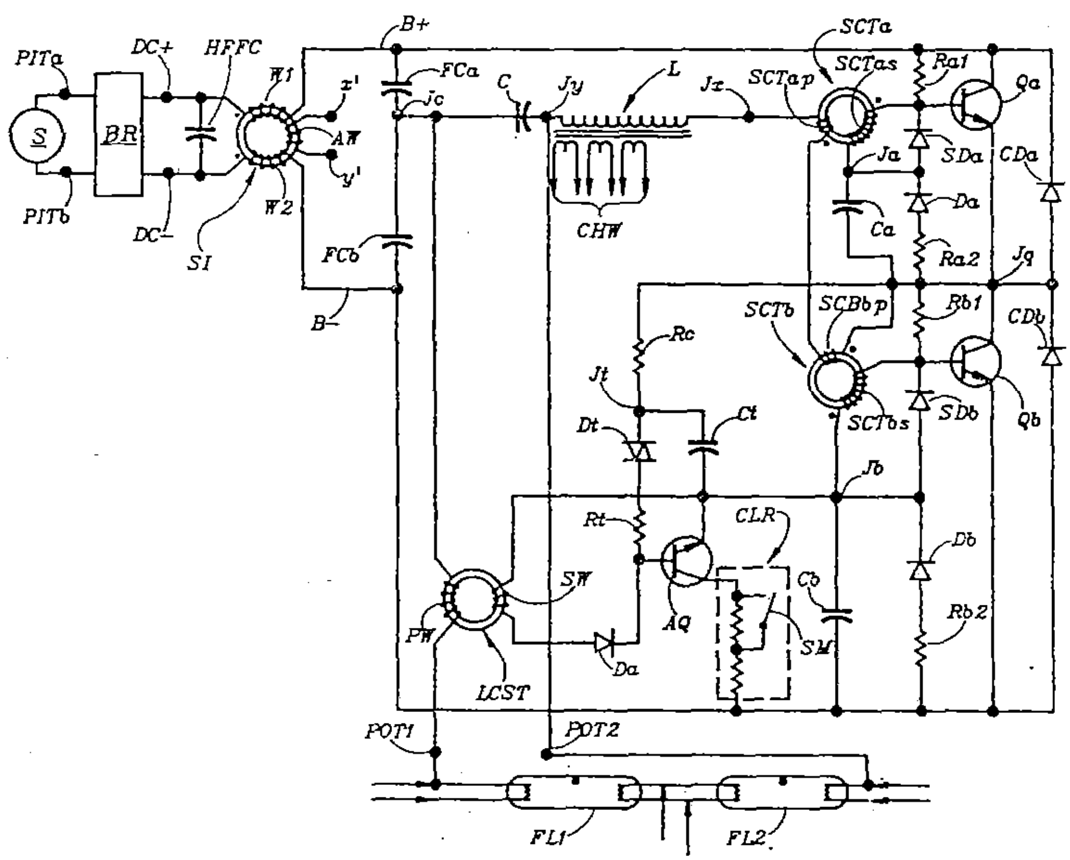
Primary Examiner—Eugene R. LaRoche

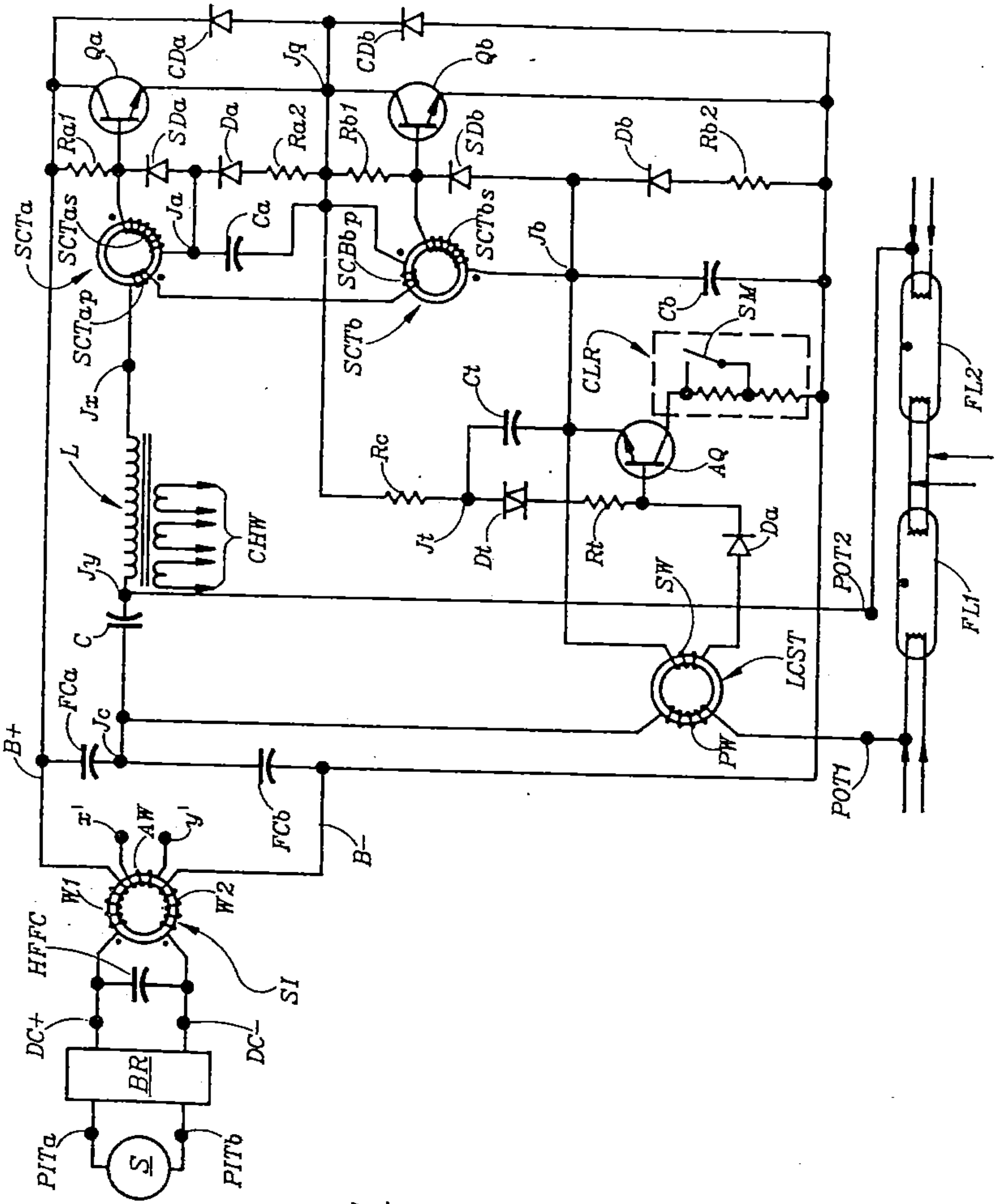
Assistant Examiner—Michael Shingleton

[57] ABSTRACT

A self-oscillating half-bridge inverter is powered from a power-line-operated DC voltage source. The inverter is loaded by way of a series-tuned high-Q LC circuit connected across its output. A pair of fluorescent lamps is series-connected across the tank-capacitor of the LC circuit. The inverter has two bipolar transistors, each driven by an associated saturable current transformer that provides for a transistor ON-time dependent upon the magnitude of an associated bias voltage. One of the transistors has a control arrangement connected in circuit with its associated saturable transformer and operative to control the magnitude of its associated bias voltage. As the magnitude of this bias voltage is controlled, the magnitude of the voltage across the tank-capacitor, as well as of the current available therefrom, is correspondingly controlled. The magnitude of the bias voltage is automatically controlled such that: (a) with the lamps not-yet-ignited, the magnitude of the voltage across the tank-capacitor is maintained at a level somewhat higher than normal lamp operating voltage, except that for 10 milli-seconds once each second the magnitude is increased to a level high enough to cause lamp ignition; (b) after the lamps have ignited, the magnitude of the lamp current is limited to an adjustably preset level; and (c) the magnitude of any ground-fault current is limited to a level considered safe from shock hazard.

24 Claims, 2 Drawing Sheets




Qa
CDa
Jq
CDb
Qb
Ra1
SDa
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Ra2
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SCBbp
SCTbs
Jb
SCTap
SCTb
Cb
SM
CLR
Ct
FL2
Jx
L
Rc
AQ
CHW
Jt
Dt
Rt
Da
SW
POT2
Jy
FL1
C
LCST
FCa
Jc
B+
PW
POT1
x'
AW
y'
FCb
B−
W1
W2
HFFC
SI
DC+
DC−
BR
PITa
S
PITb

Fig. 1

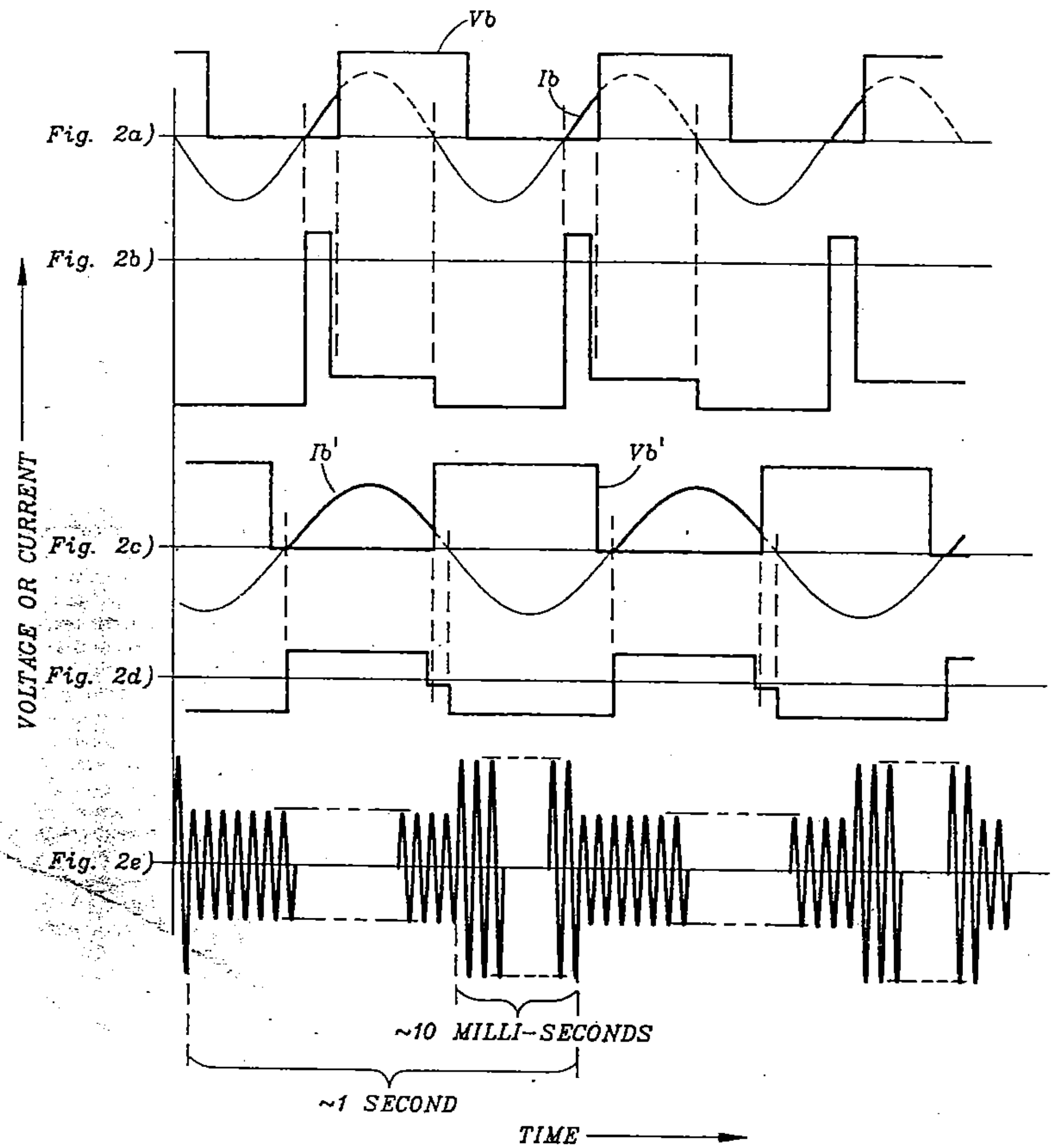
Vb
Ib
Fig. 2a)
Fig. 2b)
Ib'
Vb'
Fig. 2c)
Fig. 2d)
Fig. 2e)
VOLTAGE OR CURRENT
~10 MILLI-SECONDS
~1 SECOND
TIME

Fig. 2

CONTROLLED ELECTRONIC BALLAST

RELATED APPLICATIONS

Instant application is a continuation-in-part of Ser. No. 07/080,865 filed Aug. 3, 1987, now U.S. Pat. No. 4,819,146; which Ser. No. 07/080,865 is a continuation-in-part of Ser. No. 06/917,788 filed Oct. 10, 1986, now U.S. Pat. No. 4,727,470.

Instant application is also a continuation-in-part of Ser. No. 06/730,596 filed May 6, 1985, now abandoned; which is a continuation-in-part of Ser. No. 06/640,240 filed Aug. 13, 1984, now abandoned; which is a continuation-in-part of Ser. No. 06/412,771 filed Aug. 30, 1982, now abandoned.

BACKGROUND OF THE INVENTION

1. Field of the Invention

The present invention relates to ground-fault-protected series-resonance-loaded inverters, particularly as used for controllably powering gas discharge lamps.

2. Description of Prior Art

Ballasts with built-in ground-fault-protection means have been previously described, such as in U.S. Pat. No 4,563,719 to Nilssen.

Ballasts using a series-resonance-loaded output have also been previously described, such as in U.S. Pat. No. 4,370,600 to Zansky.

In an inverter where a gas discharge lamp load is parallel-connected across the tank capacitor of a high-Q LC circuit that is resonantly series-excited by a high-frequency voltage output of the inverter, it is necessary to provide some means to protect against the high currents and voltages resulting due to so-called Q-multiplication whenever the lamp load is removed or otherwise fails to constitute a proper load for the LC circuit.

In U.S. Pat. No. 4,370,600 to Zansky, circuit protection is provided by way of providing to the LC circuit an alternative load in the form of a voltage-clamping means; which voltage-clamping means acts to load the LC circuit during any period when the lamp does not constitute a proper load therefor.

The voltage-clamping is accomplished by rectifying the Q-multiplied voltage output of the LC circuit and by applying the resulting DC output to the inverter's DC power source.

However, during any period when voltage-clamping does occur, a relatively large amount of power circulates within the electronic ballast means: from the inverter's output, through the LC circuit, and back into the inverter's DC power source by way of the voltage-clamping means.

SUMMARY OF THE INVENTION

Objects of the Invention

One object of the present invention is that of providing for a cost-effective ballasting means for powering gas discharge lamps.

Another object is that of providing for control means in a series-resonance-loaded inverter ballast.

These as well as other objects, features and advantages of the present invention will become apparent from the following description and claims.

Brief Description

A self-oscillating half-bridge inverter is powered from a power-line-operated DC voltage source. The

inverter is loaded by way of a series-tuned high-Q LC circuit connected across its output. A pair of fluorescent lamps is series-connected across the tank-capacitor of the LC circuit. The inverter has two bipolar transistors, each driven by an associated saturable current transformer that provides for a transistor ON-time dependent upon the magnitude of an associated bias voltage.

One of the transistors has a control arrangement connected in circuit with its associated saturable transformer and operative to control the magnitude of its associated bias voltage. As the magnitude of this bias voltage is controlled, the magnitude of the voltage across the tank-capacitor, as well as of the current available therefrom, is correspondingly controlled.

The magnitude of the bias voltage is automatically controlled such that: (a) with the lamps not-yet-ignited, the magnitude of the voltage across the tank-capacitor is maintained at a level somewhat higher than normal lamp operating voltage, except that for 10 milli-seconds once each second the magnitude is increased to a level high enough to cause lamp ignition; (b) after the lamps have ignited, the magnitude of the lamp current is limited to an adjustably preset level; and (c) the magnitude of any ground-fault current is limited to a level considered safe from shock hazard.

As a consequence of not having to provide, on a continuous basis, an output voltage of magnitude large enough to cause lamp ignition, the cost and weight of the tank-inductor and the tank-capacitor of the LC circuit may be reduced by a considerable factor.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 provides a basic electrical circuit diagram of the preferred embodiment of the invention.

FIG. 2 illustrates waveshapes of various high frequency voltages and currents present within the circuit during different modes of operation.

DESCRIPTION OF THE PREFERRED EMBODIMENT

Details of Construction

In FIG. 1, a source S of ordinary 120 Volt/60 Hz power line voltage is applied to power input terminals PITa and PITb; which terminals, in turn, are connected with a bridge rectifier BR. The DC output from bridge rectifier BR is applied to a DC+ terminal and a DC− terminal, with the DC+ terminal being of positive polarity. A high-frequency filter capacitor HFFC is connected between the DC+ terminal and the DC− terminal. A first winding W1 of an EMI suppression inductor SI is connected between the DC+ terminal and a B+ bus; and a second winding W2 of EMI suppression inductor SI is connected between the DC− terminal and a B− bus.

An auxiliary winding AW is wound on EMI suppression inductor SI; which auxiliary winding has output terminals x' and y'.

A filter capacitor FCa is connected between the B+ bus and a junction Jc; a filter capacitor FCb is connected between junction Jc and the B− bus. A switching transistor Qa is connected with its collector to the B+ bus and with its emitter to a junction Jg; a switching transistor Qb is connected with its collector to junction Jg and with its emitter to the B− bus. A commutating diode CDa is connected between the B+ bus and junction Jg, with its cathode connected with the B+

bus; a commutating diode CDb is similarly connected between junction Jg and the B− bus.

A saturable current transformer SCTa has a secondary winding SCTas connected between the base of transistor Qa and a junction Ja; a saturable current transformer SCTb has a secondary winding SCTbs connected between the base of transistor Qb and a junction Jb. Saturable current transformers SCTa and SCTb, respectively, have primary windings SCTap and SCTbp; which primary windings are series-connected between junction Jg and a junction Jx.

A resistor Ra1 is connected between the collector and the base of transistor Qa; a resistor Rb1 is connected between the collector and the base of transistor Qb. A capacitor Ca is connected between junction Ja and the emitter of transistor Qa; a capacitor Cb is connected between junction Jb and emitter of transistor Qb. A diode Da is connected with its cathode to junction Ja and, by way of a leakage resistor Ra2, with its anode to the emitter of transistor Qa; a diode Db is connected with its cathode to junction Jb and, by way of a leakage resistor Rb2, with its anode to the emitter of transistor Qb. A shunt diode SDa is connected between the base of transistor Qa and junction Ja, with its anode connected with junction Ja; a shunt diode SDb is similarly connected between the base of transistor Qb and junction Jb.

An auxiliary transistor AQ is connected with its emitter to junction Jb and with its collector to the B− bus by way of a current-limiting resistor CLR; which current-limiting resistor is switch-adjustable by way of a switch means SM.

A tank-inductor L is connected between junction Jx and a junction Jy; and a tank-capacitor C is connected between junctions Jy and Jc. A power output terminal POT1 is connected with junction Jc by way of a primary winding PW of a lamp current sensing transformer LCST; another power output terminal POT2 is connected directly with junction Jy. First and second fluorescent lamps FL1 and FL2 are series-connected between power output terminals POT1 and POT2. Tank-inductor L has three cathode heater windings CHW; which are connected with the cathodes of fluorescent lamps FL1 and FL2.

A capacitor Ct is connected between junction Jb and a junction Jt. A Diac Dt is series-connected with a resistor Rt to form a series-combination, which series-combination is connected between junction Jt and the base of transistor AQ. A resistor Rc is connected between junctions Jg and Jt.

A secondary winding SW of transformer LCST is connected between the emitter of auxiliary transistor AQ and the anode of a diode Da, whose cathode is connected with the base of transistor AQ.

Details of Operation

The basic operation of the half-bridge inverter of FIG. 1 is conventional and is explained in conjunction with FIG. 3 of U.S. Pat. No. 4,307,353 to Nilssen.

For a given magnitude of the DC supply voltage, due to the effect of the high-Q LC circuit, the magnitude of the current provided to the fluorescent lamp load (or to any other load presented to the output) is a sensitive function of the frequency and the waveshape of the inverter's output voltage; which output voltage is a substantially squarewave voltage of controllable frequency and with peak-to-peak magnitude about equal to that of the instantaneous magnitude of the DC voltage present between the B− bus and the B+ bus.

The frequency of the inverter's squarewave output voltage is a sensitive function of the natural resonance frequency of the high-Q LC circuit as well as of the duration of the forward conduction period (i.e., the ON-period) of the two inverter switching transistors; which duration, in turn, is a sensitive function of the saturation characteristics of saturable current transformers SCTa and SCTb as combined with the magnitude of the bias voltages present on capacitors Ca and Cb. That is, the duration of the forward conduction period (the ON-time) of each switching transistor is determined by the volt-second product sustainable by its associated saturable current transformer as well as by the magnitude of the negative bias on capacitors Ca and Cb: the higher the volt-second product available before saturation, the longer the ON-time; the higher the negative bias on the Ca/Cb capacitors, the shorter the ON-time.

In the circuit arrangement of FIG. 1, the magnitude of the negative voltage on capacitors Ca and Cb is determined by the magnitude of the current provided to the bases of transistors Qa and Qb, less any current drained away through resistors R2a and Rb2/CLR, all respectively. (Of course, a small amount of current is also drained away from bias capacitors Ca and Cb by resistors Ra1 and Rb1, respectively. However, this amount of charge leakage is in most situations negligible. Resistors Ra1 and Rb1 are principally used for getting the inverter to initiate oscillation.)

The magnitude of the base current provided to each transistor is directly proportional to the magnitude of the current flowing through the primary windings of saturable current transformers SCTa and SCTb. Thus, assuming transistor AQ to be conducting, for given values of resistors Ra2 and Rb2/CLR: the higher the magnitude of the inverter's output current, the higher the magnitude of the negative voltage on capacitors Ca and Cb.

Thus, for given values of Ra2 and Rb2/CLR, the circuit of FIG. 1 provides for a high degree of automatic regulation of the magnitude of the inverter's output current.

By selecting a suitable resistance value for resistor Ra2, and assuming transistor AQ to be conducting, the magnitude of the inverter's output current may be adjusted by adjusting the resistance value of CLR by way of switch means SM: a relatively low resistance value leads to an inverter output current of relatively high magnitude; a relatively high resistance value leads to an inverter output current of relatively low magnitude.

The higher the magnitude of the negative voltage on each bias capacitor, the higher the magnitude of the voltage that has to be provided from the secondary winding of each saturable current transformer; which, in turn, leads to a correspondingly shorter period before saturation is reached. Thus, as the magnitude of the negative bias on each bias capacitor is increased, the duration of each transistor's forward conduction period (ON-time) is decreased; which, in turn, leads to a reduction in the magnitude of the inverter's output current in comparison with what it otherwise would have been.

Whereas the base current provided to each transistor has to flow from its associated bias capacitor, the reverse or reset current provided from each of the saturable current transformer's secondary windings does not flow from the bias capacitor, but rather flows in a separate path through the reverse shunt diode (SDa or SDb) shunting the secondary winding of each saturable current transformer.

More particularly, the circuit and control arrangement of FIG. 1 operates as follows.

As power is applied at power input terminals PIT1/PIT2, the inverter starts to oscillate at a frequency near the natural self-resonance frequency of the LC circuit. The resulting inverter output current results in a positive feedback current provided to each base; and this feedback current, in turn, causes a negative bias to build up on each of bias capacitors Ca/Cb. As the magnitude of the negative bias voltage increases, the inverter's oscillation frequency increases as well. As a result, the magnitude of the inverter output current will stabilize at a some level determined by the effective resistance values of resistors Ra2 and Rb2/CLR.

With the fluorescent lamps non-connected or otherwise non-functional, transistor AQ is effectively non-conducting; and under this condition the magnitude of the high-frequency (30 kHz or so) inverter current stabilizes at a level determined by the resistance values of resistors Ra2 and Rb2.

With the fluorescent lamps connected and fully operating (i.e., fully loading the LC circuit), transistor AQ is conducting by virtue of the current provided to its base by way of lamp current sensing transformer LCST. Under this condition the magnitude of the high-frequency current stabilizes at a level effectively determined by the resistance values of resistors Ra2 and CLR—the resistance value of Rb2 being much higher than that of CLR.

More particularly, the resistance value of resistor Ra2 is selected such that the ON-time of transistor Qa corresponds to nearly a 50% duty-cycle; the resistance value of Rb2 is then selected such as to result in an inverter output current of magnitude such as to provide for a high-frequency voltage across the tank-capacitor that is approximately equal in magnitude to that of the voltage across the fluorescent lamps under normal full-power lamp operation; and the resistance value of resistor CLR is then selected such as to provide for the proper magnitude of the lamp operating current.

Under the condition of normal lamp loading of the LC circuit, the RMS magnitude of the voltage provided to the lamp cathodes is so selected as to provide for proper cathode heating. Thus, as the circuit is initially powered, even though the lamps have not yet ignited, the cathodes are provided with a heating voltage of RMS magnitude appropriate for cathode heating; yet, the magnitude of the high-frequency voltage then provided across the lamps is too low to cause lamp ignition. However, by action of the trigger arrangement consisting of elements Rc, Ct, Dt and Rt, after about one second (and once each second thereafter), a current pulse of about 10 milli-second duration will be provided to the base of transistor AQ, thereby making this transistor conduct for a period of about 10 milli-seconds. During this 10 milli-second period, the magnitude of the high-frequency voltage across the tank-capacitor will increase substantially, and the lamps will ignite.

During the 10 milli-second period, the magnitude of the cathode heating voltages also increases substantially, thereby aiding in lamp ignition. However, on an integrated RMS basis, this brief period of increased cathode heating voltage is of little consequence.

As soon as the lamps ignite, lamp current will start to flow; and, by way of transformer LCST, transistor AQ will now be conducting on a nearly continuous basis. This implies that the 10 milli-second pulses that will continue to be provided every second or so will be of little consequence.

If the lamps were to be disconnected, however, the repeatedly provided 10 milli-second pulses will assure that affirmative lamp ignition will occur as soon as fully functional lamps are indeed connected.

As long as transistor AQ is conductive, the fluorescent lamps will be powered in a normal manner; and the magnitude of the lamp current flowing will depend on the particular setting of adjustable resistor CLR. With switch means SM closed, the magnitude of the lamp current will be relatively large; with switch means Sm open, the magnitude of the lamp current will be relatively low.

In case of a ground-fault condition—which typically might occur if fluorescent lamp FL2 were to be connected at its one end with power output terminal POT2 while a ground-connected person held onto the lamp's terminals at its other end—high-frequency current would flow out from the ballast at terminal POT2, through the lamp and the person to ground, and then eventually back to the ballast by way of one or both of the power input terminals (PIT1/PIT2. However, since necessarily this condition would require that lamp current stop flowing through lamp current sensing transformer LCST, transistor AQ would cease to conduct and the magnitude of any current then flowing from output terminal POT2 would be limited to a level that is substantially lower than that of full-power lamp current. In particular, the magnitude of the resulting ground-fault current will only be on the order of 30 milli-Ampere; which is a level that—at a frequency of 30 kHz or so—is considered by authoritative entities, such as Underwriters Laboratories, Inc. of Northbrook, Ill., as being substantially non-hazardous.

FIG. 2 depicts various voltage and current waveforms associated with the circuit of FIG. 1.

For a situation with no loading presented to the high-Q LC circuit—that is, with the lamps disconnected, or before the lamps have ignitied—FIG. 2*a* shows the collector-to-emitter voltage Vb of transistor Qb and the corresponding inverter output current Ib. The part of Ib actually flowing through transistor Qb in the forward direction is shown in heavy solid line, the part of Ib flowing through commutating diode CDa is showed in light dashed line, and the part of Ib flowing through either Qa or CDb is shown in light solid line.

FIG. 2*b* shows the base-emitter voltage of transistor Qb as it corresponds to the waveforms of FIG. 2*a*.

For a situation where the LC circuit is substantially fully loaded by the two fluorescent lamps. FIG. 2*c* shows the collector-to-emitter voltage Vb' of transistor Qb and the corresponding inverter output current Ib'. The part of Ib' actually flowing through transistor Qb in the forward direction is shown in heavy solid line, the part of Ib' flowing through commutating diode CDa is showed in light dashed line, and the part of Ib' flowing through either Qa or CDb is shown in light solid line.

FIG. 2*d* shows the base-emitter voltage of transistor Qb as it corresponds to the waveforms of FIG. 2*c*.

FIG. 2*e* shows the waveshape of the high-frequency voltage present across the tank-capacitor under the condition of an unloaded LC circuit: a continuous substantially sinusoidal voltage of a relatively low magnitude, interrupted once each second with a 10 milli-second long burst of relatively high-magnitude substantially sinusoidal voltage.

Additional Comments

(a) Detailed information relative to a fluorescent lamp ballast wherein the fluorescent lamp is powered by way of a series-excited parallel-loaded L-C resonant circuit is provided in U.S. Pat. No. 4,554,487 to Nilssen.

(b) The instantaneous peak-to-peak magnitude of the squarewave output voltage provided by the half-bridge inverter between junctions Jg and Jc is substantially equal to the instantaneous magnitude of the DC supply voltage.

(c) Saturable current transformers SCTa and SCTb require only a miniscule amount of voltage across their primary windings. Hence, the magnitude of the voltage-drop between junctions Jg & Jx is substantially negligible, and the inverter's full output voltage is therefore effectively provided across the LC circuit, which consists of tank-capacitor C and tank-inductor L.

(d) In FIG. 2, the inverter frequency associated with the waveforms of FIGS. *2a* and *2b* is substantially higher than that associated with FIGS. *2c* and *2d*.

Also, current Ib is nearly 180 degrees out of phase with the fundamental frequency component of voltage Vb, while current Ib' is almost in phase with voltage Vb'.

(e) In the situation associated with the waveform of FIG. *2b*, the magnitude of the voltage "seen" by the secondary winding of saturable current transformer SCTb is about five times as high as that "seen" by the same secondary winding in the situation associated with FIG. *2d*.

Correspondingly, the duration of the transistor ON-time in the situation associated with FIG. *2d* is about five times longer than the transistor ON-time in the situation associated with the waveform of FIG. *2b*.

(f) As may be noticed in FIG. *2a*, transistor Qb ceases to conduct in its forward direction while a substantial amount of current is still flowing from the inverter's output. After transistor Qb has ceased to conduct, the inverter's output current will continue to flow until the energy in the tank inductor has dissipated itself. However, the output current will continue its flow through commutating diode CDa, thereby discharging its energy into the DC power supply.

(g) Forward conduction of a transistor is defined as current flowing, with the aid of forward base drive current, directly between the collector and the emitter; which, in case of transistor Qb for, instance, means that forward current is defined as positive current flowing from its collector to its emitter while drive current is being provided to its base.

A transistor's ON-time is defined as the period during which it conducts current in the forward direction.

(h) In FIG. 2 it is noted that the fundamental frequency of the waveforms depicted in FIGS. *2c* and *2d* is lower by a certain factor as compared with the frequency associated with the waveforms of FIGS. *2a* and *2b*; yet the indicated duration of transistor ON-time associated with the waveforms of FIGS. *2a* and *2b* is shorter by a much larger factor as compared with the indicated duration of transistor ON-time associated with the waveforms of FIGS. *2c* and *2d*.

In fact, when the transistor ON-time is shortened by a given proportion, the fundamental frequency of the inverter's output voltage increases by a much smaller proportion. In instant case, with each transistor's ON-time shortened by a factor of about five, the inverter frequency increased only by a factor of about 1.3: from about 30 kHz to about 40 kHz.

(i) The time constant associated with each bias capacitor and its associated leakage resistance means is normally longer than a complete cycle of the high frequency inverter output voltage, and it is typically on the order of several such complete cycles. For instance, for a situation where the power line input voltage is 120 Volt/60 Hz, the frequency of the inverter output voltage/current is on the order of 30 kHz, and the total inverter power output falls in the range between 10 and 100 Watt, the values of bias capacitors Ca and Cb might reasonably be in the range from one to ten micro-Farad, the value for leakage resistor Ra2 might reasonably be in the range between 10 and 100 Ohm, and adjustable resistor CLR might reasonably be adjustable over a range between 2 and 100 Ohm.

Thus, in general, the magnitude of the bias voltage on the bias capacitors is responsive to the average magnitude of the inverter's output current—normally as averaged over at least a full cycle of this output current.

(j) With power input terminals PIT1 and PIT2 connected with an ordinary electric utility power line, all electrical parts of the circuit arrangement of FIG. 1 are effectively connected with earth ground by way of those power input terminals.

(k) Although not expressly shown in FIG. 1, it is emphasized that fluorescent lamps FL1 and FL2 are connected with terminals POT1 and POT2 by way of lamp socket means of a type that has electrical terminal means that are non-accessible to a person's fingers or the like. However, the terminal means of the fluorescent lamps are of a type that does permit a person to touch them directly, provided they are not inserted into their sockets.

Thus, in a situation where the circuit arrangement of FIG. 1 is used as a fluorescent lamp ballasting means, and under otherwise ordinary circumstances, the only way that a person is able to get exposed to a ground-fault current from the circuit arrangement of FIG. 1 is by holding on to the terminals at one end of a fluorescent lamp while inserting the other end into a lamp socket connected with the POT2 or the "hot" terminal.

(l) In the circuit arrangement of FIG. 1 there are two distinctly different kinds of current-magnitude-limiting provided. One is the ordinary kind associated with the natural characteristics of a series-excited parallel-loaded resonant LC circuit; another is due to the action of the control circuit associated with auxiliary transistor AQ.

The former is the principal means for limiting the lamp current; the latter is the principal means for limiting the output current in the absence of proper circuit loading.

(m) Due to basic factors related to magnetic hysteresis and leakage iductance, lamp current sensing transformer LCST requires the lamp current to have a certain minimum magnitude before producing an output signal of magnitude adequate to cause auxiliary transistor AQ to become conductive. Hence, there is a minimum threshold level automatically built into the control means used for sensing lamp current. This threshold level may readily be changed, for instance by changing the number of turns on primary winding PW, or by placing a shunting impedance across secondary winding SW.

(n) In the control circuit related to and including auxiliary transistor AQ, there are to main control effects: (i) one associated with the fact that the magnitude of the bias voltage on capacitor Cb tends to vary around an average level as a function of the average absolute magnitude of the inverter's output current, and (ii) one associated with the fact that the average level around which the bias voltage varies may itself be varied, such as by varying the magnitude of the base current provided to auxiliary transistor AQ.

(o) Without any substantial loading on the LC circuit, its Q-factor is quite high, and—absent proper control measures—the magnitude of the voltage developing across the tank-capacitor tends to become destructively high.

A large-magnitude voltage across the tank-capacitor must by necessity be associated with an inverter output current of correspondingly large magnitude. However, a large-magnitude inverter output current will cause a correspondingly high-magnitude negative bias on both bias capacitors, but—absent lamp current of magnitude sufficient to place transistor AQ into conduction—particularly on bias capacitor Cb: the two bias capacitors have to provide a positive current of magnitude proportional to the average absolute magnitude of the inverter's output current; which means that the magnitude of the negative bias voltage will have to increase with increasing magnitude of the inverter output current.

It is this negative feedback feature, which relates to negative feedback of the rectified average magnitude of the inverter's output current, that provides for stable controllable operation of a self-oscillating inverter whose output is connected across an unloaded high-Q series-connected LC circuit.

(p) In fact, the circuit arrangement of FIG. 1 may be defined as an inverter that is loaded by way of a high-Q tuned LC circuit and arranged to self-oscillate by way of positive feedback derived from the inverter's instantaneous output current (and/or voltage) while at the same time arranged to provide for controllable-magnitude output current (and/or voltage) by way of negative feedback derived from the average absolute magnitude of the inverter's output current (and/or voltage).

(q) So as to fully reset the saturable cores each cycle, diodes SDa and SDb should each have a relatively high-magnitude forward voltage drop, such as might be obtained by using two ordinary diodes in series. However, instead of using special diodes with high-magnitude forward voltage drops, it is acceptable to use ordinary diodes with added series-resistors, thereby effectively to increase their forward voltage drops.

(r) The magnitude of the relatively high-magnitude high-frequency voltage of FIG. 2*e* may be established by various means, such as by properly sizing the resistance value of resistor Rt.

(s) Some of the values associated with operating the ballast with the kind of waveform indicated by FIG. 2*e* are as follows: (i) substantially relaxed specifications for the tank-inductor; (ii) similarly relaxed specifications for the tank-capacitor; (iii) reduced glow current prior to lamp ignition, thereby providing for increased lamp life; (iv) much improved lamp starting; (v) substantially reduced idling power; and (vi) more cost-effective compliance with U.L. specifications related to ground-fault current.

(t) The RMS magnitude of the cathode heating voltage, which voltage is provided to each of the lamps'-thermionic cathodes by way of cathode heating windings CHW, is such as to provide for proper cathode heating during the period before the lamps ignite, as well as on a continuous basis thereafter.

During the brief pulses provided by way of elements Rc, Ct, Dt and Rt, the RMS magnitude of the cathode heating voltage is increased to about twice normal value. However, since the duration of each of these pulses is so very brief (about 10 milli-seconds) compared with the duration of each of the periods between such pulses (about 1000 milli-seconds), the net effect on the temperature of the cathodes is negligible. However, with respect to lamp ignition, the effect is substantial and beneficial. The briefly elevated RMS magnitude of the cathode voltage gives rise to ionization of the lamp gas along the cathodes' surfaces, thereby greatly facilitating the ignition of the main gas columns of the lamps.

(u) While the RMS magnitude of the high-frequency output voltage, as provided at output terminals POT1-/POT2, may indeed be determined by appropriate choice of resistor Rt, the resulting magnitude is highly dependent on the gain of transistor AQ.

To eliminate such dependence, an additional auxiliary transistor may be placed in parallel with transistor AQ; which additional transistor would have its emitter connected with junction Jb and its collector connected with the B— bus by way of a separate resistor. Then, resistor Rt would be connected with the base of this additional transistor instead of to the base of transistor AQ. The resistance value of the separate resistor would then be chosen such as to provide for the desired RMS magnitude of the high-frequency output voltage; whereas the value of resistor Rt would now be chosen such as to provide adequate base current to bring transistor AQ into a fully conductive state.

(v) It is believed that the present invention and its several attendant advantages and features will be understood from the preceeding description. However, without departing from the spirit of the invention, changes may be made in its form and in the construction and interrelationships of its component parts, the form herein presented merely representing the presently preferred embodiment.

I claim:

1. An arrangement comprising:

a source providing a manifestly current-limited AC voltage at an AC output; the frequency of the AC voltage being substantially higher than the frequency of the power line voltage on an ordinary electric utility power line; the source having a control input; the AC voltage having an RMS magnitude; the source having control means operative, on receipt of an action at the control input, to affect said RMS magnitude;

connect means operative to permit connection of a gas discharge lamp means with said AC output; and

control means connected with the control input; the control means being operative, provided said gas discharge lamp means is not connected with the AC output, to provide said action in such manner as to cause the RMS magnitude to be modulated at a modulation frequency lower than the frequency of the power line voltage.

2. The arrangement of claim 1 wherein the frequency of the AC voltage is substantially higher than that of the power line voltage.

3. The arrangement of claim 1 wherein: (i) the RMS magnitude is modulated between a relatively low RMS magnitude and a relatively high RMS magnitude; and

(ii) the relatively high RMS magnitude is approximately twice as large as the relatively low RMS magnitude.

4. The arrangement of claim 1 wherein the modulation frequency is higher than one cycle per minute but lower than ten cycles per second.

5. The arrangement of claim 4 wherein the modulation frequency is approximately equal to about one cycle per second.

6. The arrangement of claim 1 wherein, a brief period after said gas discharge lamp is indeed connected with the AC output, the RMS magnitude ceases to be modulated.

7. The arrangement of claim 6 wherein said brief period has a duration of about one second.

8. The arrangement of claim 1 in actual combination with said gas discharge lamp.

9. The arrangement of claim 1 wherein the control means is additionally operative to permit control of the maximum magnitude of any current flowing from the AC output.

10. The arrangement of claim 9 wherein the control means comprises adjustment means operative to permit manual adjustment of said maximum magnitude.

11. An arrangement comprising:

a source of electric power; and

a ballast means connected with the source of electric power and operative to provide an AC voltage at an AC output; the AC voltage being of frequency subsequently higher than the frequency of the power line voltage on an ordinary electric utility power line: the ballast means being operative to power a gas discharge lamp connected with the AC output: the AC voltage being characterized by: (i) whenever the gas discharge lamp is indeed connected with the AC output and powered therefrom, being of a substantially constant RMS magnitude; and (ii) whenever the gas discharge lamp is not so connected, periodically varying, at a frequency lower than that of the power line voltage, between a minimum relatively low RMS magnitude and a maximum relatively high RMS magnitude.

12. The arrangement of claim 11 wherein said minimum relatively low RMS magnitude is approximately equal to said substantially constant RMS magnitude.

13. The arrangement of claim 11 wherein said maximum relatively high RMS magnitude is more than about 50% larger than said substantially constant RMS magnitude.

14. The arrangement of claim 11 wherein, whenever the discharge lamp is not connected with the AC output and powered therefrom, the AC voltage is amplitude-modulated at a frequency substantially lower than 60 Hz.

15. An arrangement comprising:

a source of electric power;

a gas discharge lamp; and

a ballast means connected with the source of electric power and operative to provide an AC voltage at an AC output; the fundamental frequency of the AC voltage being substantially higher than that of the power line voltage on an ordinary electric utility power line; the gas discharge lamp being connected with the AC output and operative to be properly powered by the AC voltage provided thereat; the AC voltage being characterized by: (i) before the gas discharge lamp has ignited, having an RMS magnitude periodically alternating, at a frequency lower than the fundamental frequency of the power line voltage, between a relatively low minimum RMS magnitude and a relatively high maximum RMS magnitude; and (ii) after the gas discharge lamp has ignited, being of a substantially constant RMS magnitude.

16. The arrangement of claim 15 wherein said relatively high maximum RMS magnitude is larger that said relatively low minimum magnitude RMS magnitude by at least one third.

17. An arrangement comprising:

a source of electric power;

a gas discharge lamp; and

a ballast means connected with the source of electric power and operative to provide an AC voltage at an AC output; the fundamental frequency of the AC voltage being substantially higher than that of the power line voltage on an ordinary electric utility power line; the gas discharge lamp being connected with the AC output and operative to be properly powered by the AC voltage provided thereat; the AC voltage being characterized by: (i) before the gas discharge lamp has ignited, periodically varying at a frequency lower than about 60 Hz between a relatively low RMS magnitude and a relatively high RMS magnitude; and (ii) after the gas discharge lamp has ignited, being of a substantially constant RMS magnitude.

18. The arrangement of claim 17 wherein said relatively low RMS magnitude is about equal to said substantially constant RMS magnitude.

19. The arrangement of claim 17 wherein said first period of time is substantially longer than said second period of time.

20. The arrangement of claim 17 wherein said relatively low RMS magnitude is insufficiently high to cause lamp ignition.

21. An arrangement comprising:

a source providing an ordinary power line voltage at a pair of source terminals; and

frequency-converting ballast means connected with the source terminals and operative to provide an AC voltage at an AC output; the fundamental frequency of the AC voltage being substantially higher than that of the power line voltage: a gas discharge lamp being disconnectably connected with the AC output; the AC voltage being amplitude-modulated at a low frequency whenever the gas discharge lamp means is disconnected otherwise fails to draw a substantial current from the AC output; the low frequency being substantially lower than the fundamental frequency of the power line voltage.

22. The arrangement of claim 21 wherein, whenever the gas discharge lamp is not connected with the AC output: (i) the RMS magnitude of the AC voltage varies periodically between a minimum level and a maximum level; and (ii) the maximum level being larger than the minimum level by about one third or more.

23. The arrangement of claim 21 wherein the low frequency is higher than about 0.01 Hz but lower than about 10 Hz.

24. An arrangement comprising:

source terminal means across which is provided an ordinary power line voltage; and

ballast means connected with the source terminal means and operative to provide an AC voltage at an AC output; the AC output being operative to connect with and to be properly loaded by a lamp load: the frequency of the AC voltage being substantially higher than that of the power line voltage; the AC voltage being characterized by: (i) whenever the AC output is indeed being properly loaded with a lamp load, being of a substantially

constant RMS magnitude; and (ii) whenever the AC output is not being properly loaded, being amplitude-modulated at a frequency lower than that of the power line voltage.

* * * * *

EXHIBIT D

US00:)83A

United States Patent [19]

Nilssen

[11] Patent Number: 5,404,083

[45] Date of Patent: Apr. 4, 1995

[54] ENERGY-EFFICIENT COST-EFFECTIVE ELECTRONIC BALLAST

[76] Inventor: Ole K. Nilssen, Caesar Dr., Barrington, Ill. 60010

[21] Appl. No.: 908,924

[22] Filed: Jul. 2, 1992

Related U.S. Application Data

[63] Continuation of Ser. No. 614,037, Nov. 19, 1990, abandoned, which is a continuation of Ser. No. 96,461, Sep. 15, 1987, abandoned, which is a continuation-in-part of Ser. No. 730,596, May 6, 1985, Pat. No. 4,855,860, which is a continuation-in-part of Ser. No. 640,240, Aug. 13, 1984, Pat. No. 4,563,719, which is a continuation of Ser. No. 412,771, Aug. 30, 1982, abandoned.

[51] Int. Cl.[6] H05B 41/36

[52] U.S. Cl. 315/244; 315/106; 315/224; 315/DIG. 5; 315/DIG. 7

[58] Field of Search .... 315/244, 243, 315/245, 106, 107, 224, DIG. 4, 315/DIG. 5, DIG. 7

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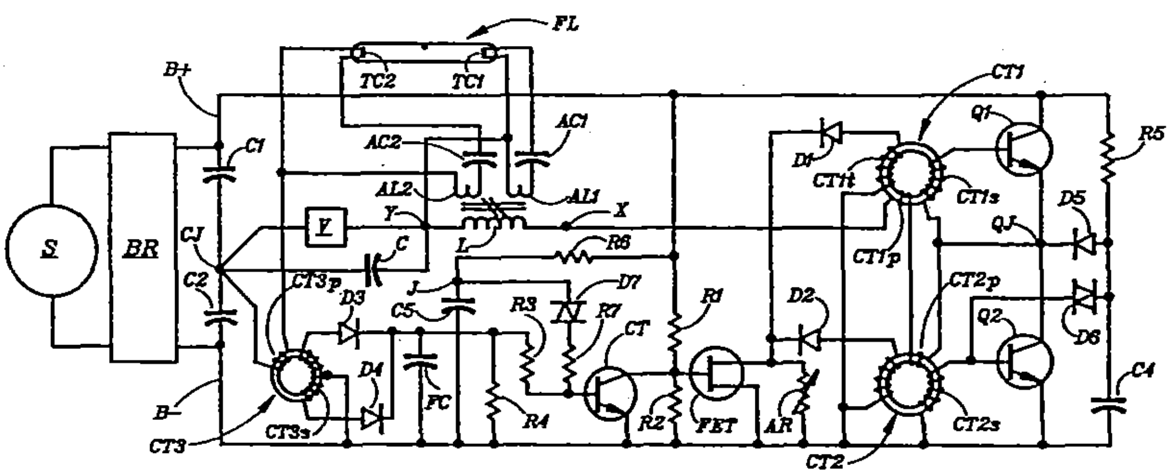
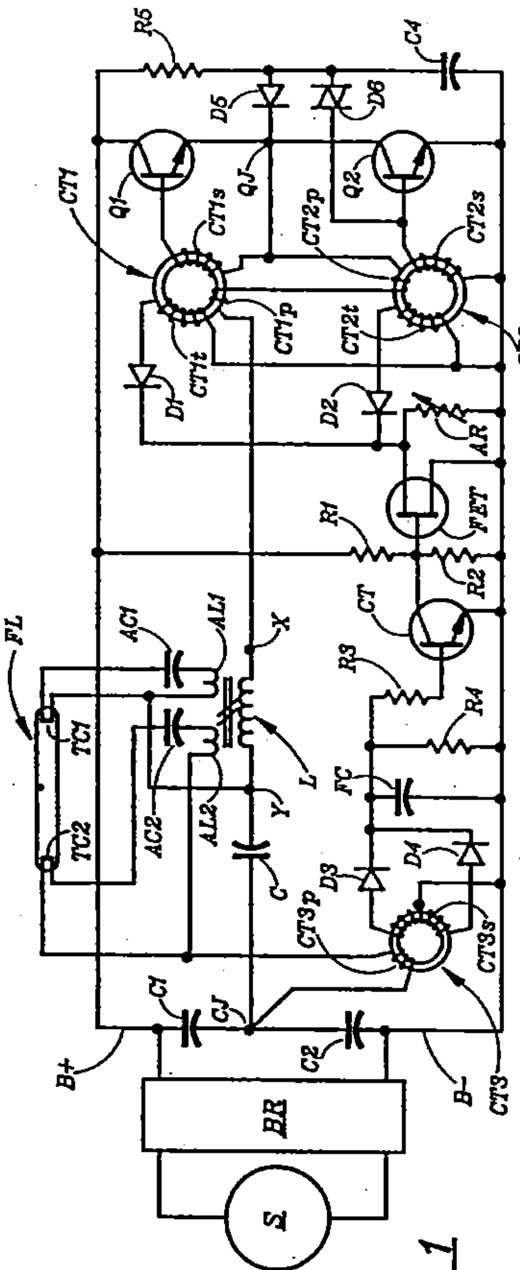
Primary Examiner—Robert J. Pascal
Assistant Examiner—Michael Shingleton

[57] ABSTRACT

An inverter-type electronic fluorescent lamp ballast normally powers a fluorescent lamp by way of a series-excited parallel-loaded resonant L-C circuit. During the lamp starting phase, as well as whenever the lamp is inoperative or not connected, inverter frequency is automatically increased substantially beyond resonance, thereby preventing circuit self-destruction which would otherwise probably result whenever an inverter is used for series-exciting an unloaded resonant L-C circuit.

12 Claims, 2 Drawing Sheets

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C4
D5
D6
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Q2
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D1
D2
AR
FET
R1
R2
CT
R3
R4
FL
AC1
AL1
X
TC1
TC2
L
AC2
AL2
Y
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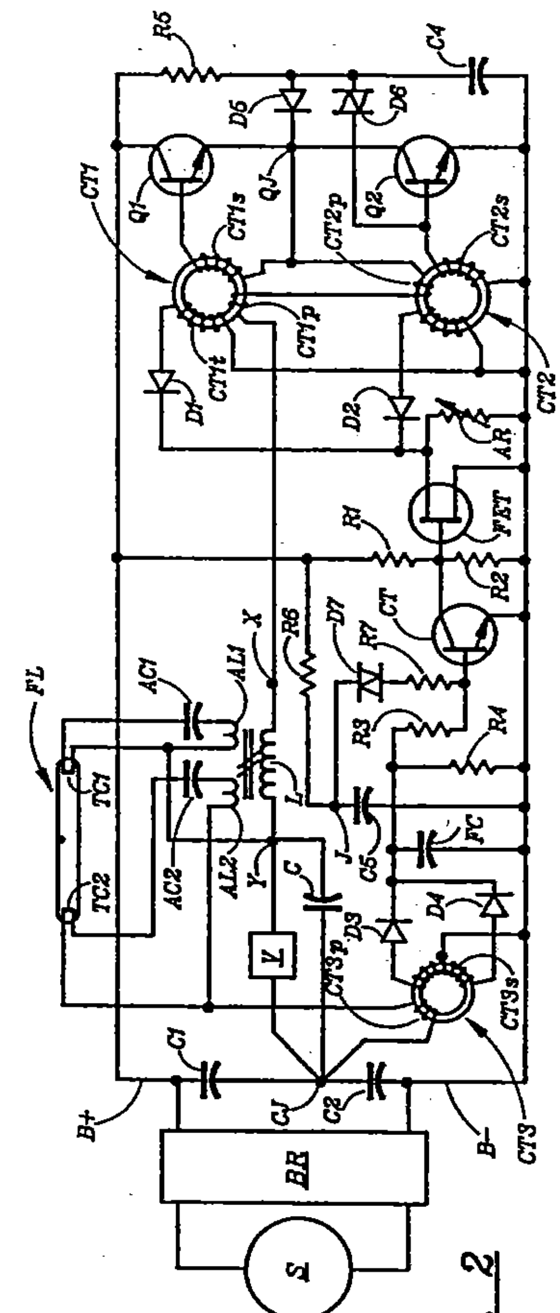
S

Fig. 1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D7
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R4
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J
C5
FC
AC2
AL2
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S

Fig. 2

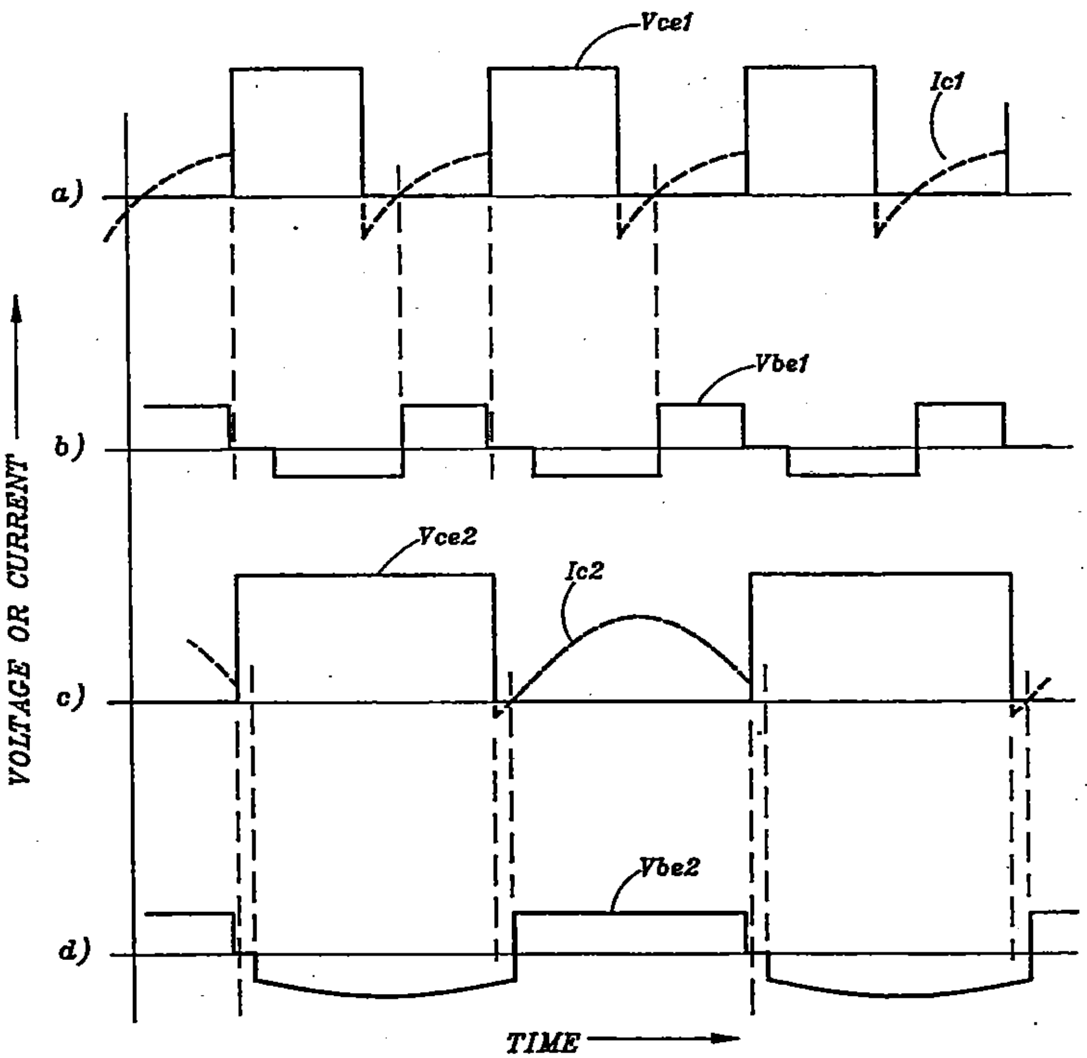

Fig. 3

ENERGY-EFFICIENT COST-EFFECTIVE ELECTRONIC BALLAST

Related Applications

This application is a Continuation of Ser. No. 07/614,037 filed Nov. 19, 1990, now abandoned; which is a Continuation of Ser. No. 07/096,461 filed Sep. 15, 1987, now abandoned; which is a Continuation-in-Part of Ser. No. 06/730,596 filed May 6, 1985, now U.S. Pat. No. 4,855,860; which is a Continuation-in-Part of Ser. No. 06/640,240 filed Aug. 13, 1984, now U.S. Pat. No. 4,563,719; which is a Continuation of Ser. No. 06/412,771, filed Aug. 30, 1982, now abandoned.

BACKGROUND OF THE INVENTION

1. Field of Invention

The present invention relates to inverter-type electronic ballasts for gas discharge lamps, particularly of the type wherein a gas discharge lamp is connected with the inverter's output by way of a series-excited parallel-loaded resonant L-C circuit.

2. Description of Prior Art

Inverter-type electronic ballasts for gas discharge lamps of the type wherein the inverter output is connected with the gas discharge lamp by way of a series-excited parallel-loaded resonant L-C circuit are fundamentally cost-effective and energy-efficient. Such ballasts are described in prior art, such as in U.S. Pat. Nos. 4,461,980, 4,581,562 and 4,663,571 to Nilssen.

However, a very basic problem associated with such series-resonance-loaded inverter-type ballasts is that of the likelyhood of self-destruction in the event that the lamp is removed or otherwise fails to constitute a proper load on the series-resonant L-C circuit.

The prior art has dealt with that problem in various ways; and the issue now is basically that of finding a still more cost-effective way of so doing.

SUMMARY OF THE INVENTION

Objects of the Invention

An object of the present invention is that of providing an energy-efficient cost-effective inverter-type electronic ballast for gas discharge lamps.

More specifically, an object is that of providing an energy-efficient cost-effective inverter-type ballast of a type wherein the inverter is powering a gas discharge lamp by way of a series-excited parallel-loaded resonant L-C circuit.

This as well as other objects, features and advantages of the present invention will become apparent from the following description and claims.

BRIEF DESCRIPTION

In its preferred embodiment, subject invention constitutes a series-excited parallel-loaded fluorescent lamp ballast comprising the following key component parts:

a source of DC voltage, which DC voltage is derived by rectification of the AC voltage from a regular 60 Hz power line;

an inverter connected with the source of DC voltage and operative to provide across an output a high-frequency square-wave voltage, the inverter having control input means operative in response to a control signal to control the frequency of the squarewave voltage between a minimum frequency and a maximum frequency;

an L-C circuit series-connected across the output, the L-C circuit having: i) a main tank-capacitor, ii) a main tank-inductor, and iii) a natural resonance frequency equal to the fundamental frequency of the squarewave voltage at its minimum frequency;

a pair of auxiliary tank-inductors, each magnetically coupled to the main tank-inductor and connected by way of an auxiliary capacitor to a pair of cathode power output terminals, each auxiliary tank-inductor being series-resonant with its auxiliary tank-capacitor at the fundamental frequency of the squarewave voltage at its maximum frequency;

a fluorescent lamp having a pair of main lamp power input terminals and two pairs of cathode power input terminals;

connect means operative to connect: i) the main lamp power input terminals across the tank-capacitor, and ii) each pair of cathode power input terminals with one of the pairs of cathode power output terminals; and

control means: i) responsive to lamp current flowing through the fluorescent lamp, ii) connected with the control input means, and iii) operative to provide the control signal in such manner as to increase the frequency of the squarewave voltage in response to the flow of lamp current;

whereby:

a) the inverter is protected from self-destruction by making the frequency of the squarewave voltage substantially higher than the L-C circuit's natural resonance frequency whenever the L-C circuit is inadequately loaded, as signified by absence of lamp current; and

b) the amount of cathode heating power is reduced as the magnitude of lamp current is increased, thereby improving overall operating efficiency.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 schematically illustrates the preferred embodiment of the invention.

FIG. 2 illustrates a modified version of the preferred embodiment.

FIG. 3 shows various voltage and current waveforms associated with the preferred embodiment.

DESCRIPTION OF THE PREFERRED EMBODIMENT

Details of Construction

In FIG. 1, a source S of 120 Volt/60 Hz voltage is applied to a full-wave bridge rectifier BR, the unidirectional voltage output of which is applied directly between a B+ bus and a B+ bus, with the positive voltage being connected to the B+ bus.

Between the B+ bus and the B− bus are connected a series-combination of two transistors Q1 and Q2 as well as a series-combination of two energy-storing capacitors C1 and C2.

A secondary winding CT1s of positive feedback current transformer CT1 is connected directly between the base and the emitter of transistor Q1; a secondary winding CT2s of positive feedback current transformer CT2 is connected directly between the base and the emitter of transistor Q2.

The collector of transistor Q1 is connected directly with the B+ bus; the emitter of transistor Q2 is connected directly with the B− bus; and the emitter of transistor Q1 is connected directly with the collector of transistor Q2, thereby forming junction QJ.

One terminal of capacitor C1 is connected directly with the B+ bus, while the other terminal of capacitor C1 is connected with a junction CJ. One terminal of capacitor C2 is connected directly with the B− bus, while the other terminal of capacitor C2 is connected directly with junction CJ.

An inductor L and a capacitor C are connected in series with one another and with primary windings CT1p and CT2p of current transformers CT1 and CT2.

The series-connected primary windings CT1p and CT2p are connected directly between junction QJ and a point X. Inductor L is connected with one of its terminals to point X and with the other of its terminals to another point Y; and capacitor C is connected between point Y and junction CJ.

A first auxiliary inductor AL1 is coupled loosely with tank-inductor L and is connected in series with a first auxiliary capacitor AC1 and a first thermionic cathode TC1 of a fluorescent lamp FL; a second auxiliary inductor AL2 is also coupled loosely with tank-inductor L and is connected in series with a second auxiliary capacitor AC2 and a second thermionic cathode TC2 of fluorescent lamp FL.

One of the terminals of thermionic cathode TC2 is connected by way of a primary winding CT3p of a current transformer CT3 to junction CJ; one of the terminals of thermionic cathode TC1 is connected with point Y. A secondary winding CT3s has two terminals and a center-tap; which center-tap is connected with the B− bus.

Current transformer CT1 has a tertiary winding CT1t connected between the B− bus and the anode of a diode D1; the cathode of diode D1 is connected with the cathode of a diode D2, whose anode is connected with one of the terminals of a tertiary winding CT2t of current transformer CT2. The other terminal of tertiary winding CT2t is connected with the B− bus.

A field effect transistor FET is connected with its drain terminal to the cathodes of diodes D1 and D2 and with its source terminal to the B− bus. An adjustable resistor AR is connected between the drain and source terminals. The gate terminal of transistor FET is connected with the B+ bus by way of a resistor R1 and with the B− bus by way of a resistor R2.

A control transistor CT is connected with its collector to the gate of transistor FET and with its emitter to the B− bus. Its base is connected by way of a resistor R3 to the cathode of a diode D3, whose anode is connected with one of the terminals of secondary winding CT3s of current transformer CT3. A diode D4 is connected with its anode to the other terminal of secondary winding CT3s and with its cathode to the cathode of diode D3.

A filter capacitor FC and a resistor R4 are both connected between the cathodes of diodes D3/D4 and the B− bus.

A resistor R5 is connected between the B+ bus and the anode of a diode D5, whose cathode is connected with junction QJ. A Diac D6 is connected between the anode of diode D5 and the base of transistor Q2. A capacitor C4 is connected between the anode of diode D5 and the B− bus.

FIG. 2 is identical to FIG. 1 except for having: a Varistor V connected between junction CJ and point Y; a resistor R6 connected between the B+ bus and a junction J; a capacitor C5 connected between junction J and the B− bus; and a Diac D7 and a resistor R7 series-connected between junction J and the base of control transistor CT.

DETAILS OF OPERATION

In FIG. 1, source S represents an ordinary electric utility power line, the voltage from which is applied directly to the bridge rectifier identified as BR. This bridge rectifier is of conventional construction and provides for the rectified line voltage to be applied to the inverter circuit by way of the B+ bus and the B− bus.

The two energy-storing capacitors C1 and C2 are connected directly across the output of the bridge rectifier BR and serve to filter the rectified line voltage, thereby providing for the voltage between the B+ bus and the B− bus to be substantially constant in magnitude. Junction CJ between the two capacitors serves to provide a power supply center tap.

The inverter circuit of FIG. 1, which represents a so-called half-bridge inverter, operates in a manner that is analogous with circuits previously described in published literature, as for instance in U.S. Pat. No. 4,184,128 entitled High Efficiency Push-Pull Inverters.

Inverter oscillation is initiated by one or a few trigger pulses applied to the base of transistor Q2 by way of the combination of resistor R5, capacitor C4 and Diac D6. Once the inverter starts operating, the provision of trigger pulses ceases because diode D5 then prevents capacitor C4 from reaching a voltage high enough to cause Diac D6 to break down.

The output of the half-bridge inverter is a substantially squarewave AC voltage provided between point X and junction CJ. By controlling the degree by which the saturable feedback current transformers CT1/CT2 are re-set after each time they have been operative to supply base current to their respective transistors Q1/Q2, the frequency of this squarewave AC voltage can be controlled between about 30 kHz and 60 kHz.

The degree to which the saturable feedback current transformers are re-set is determined by the magnitude of the voltage presented to the tertiary windings CT1t/CT2t during the re-set period. By controlling the magnitude of this voltage, the degree of re-set is controlled correspondingly: the lower the magnitude of the voltage present across the tertiary windings during the re-set period, the lower the degree of re-set of the saturable magnetic cores of feedback transformers CT1/CT2.

And, the lower the degree of re-set, the shorter will be the duration of the periods where the feedback transformers provide drive current to the bases of transistors Q1/Q2, and the higher will be the frequency of the squarewave AC voltage.

FIG. 3 illustrates the situation.

FIG. 3*a* depicts the collector-emitter voltage Vce1 of transistor Q2 during a first situation where the magnitude of the voltage across the tertiary windings of saturable feedback transformers CT1/CT2 is prevented from exceeding a relatively low level—as indicated in FIG. 3*b*, which depicts the corresponding base-emitter voltage Vbe1.

FIG. 3*c* depicts the collector-emitter voltage Vce2 of transistor Q2 during a second situation where the magnitude of the voltage presented to the tertiary windings of saturable feedback transformers CT1/CT2 is permitted to reach a relatively high level—as indicated in

FIG. 3*d*, which depicts the corresponding base-emitter voltage Vbe2.

The frequency of inverter operation prevailing during the first situation is about twice that prevailing during the second situation (60 kHz or so versus 30 kHz or so).

Saturable feedback transformers CT1 and CT2 are both current transformers; which means that the magnitude of the voltage developing across a secondary or tertiary winding is a function of the magnitude of the associated primary current as multiplied by the turns-ratio and affected by the impedance characteristics of the load presented to this secondary or tertiary winding.

In particular, when transistor FET is fully conductive (i.e, acting like a short circuit)—which is the state it does indeed assume as long as no current flows through the fluorescent lamp (FL)—each of tertiary windings CT1t/CT2t is loaded with a forward-conducting diode during the re-set periods, while each of secondary windings CT1s/CT2s is loaded with a forward-conducting base-emitter junction during the drive periods. In other words, both the tertiary and the secondary windings are then loaded with a single forward-conducting diode junction.

However, the tertiary windings have about three times as many turns as do the secondary windings; which implies that the forward voltage drops presented by diodes D1/D2 to the tertiary windings have substantially less effect (per unit time) in terms of re-setting the magnetic cores of transformers CT1/CT2 than do the forward voltage drops presented to the secondary windings by the base-emitter junctions of transistors Q1/Q2 have in terms of setting the magnetic cores.

As a consequence of positive feedback, each transistor receive base current until its associated saturable feedback transformer reaches saturation; and the length of time it takes for this saturation to occur is proportional to the degree by which the magnetic core of the saturable feedback transformer has been reset.

FIGS. 3*a* and 3*c* also indicate the collector currents Ic1 and Ic2 flowing through transistor Q2 in correlation with collector-emitter voltages Vce1 and Vce2 and base-emitter voltages Vbe1 and Vbe2, all respectively.

When transistor FET is conducting, the situation of FIGS. 3*a* and 3*b* prevails; when transistor FET is non-conducting, the situation of FIGS. 3*c* and 3*d* prevails.

The conditions prevailing when transistor FET is nonconducting can be adjusted by adjustable resistor AR; which means that the lower inverter frequency can be adjusted by adjusting adjustable resistor AR.

The loosely coupled auxiliary inductors AL1 and AL2 are each tuned to series-resonate with auxiliary capacitors AC1 and AC2, respectively, at the higher inverter frequency; which means that, when the inverter frequency changes to the lower frequency, the amount of power provided to the cathodes will diminish significantly. The degree of diminishment can be chosen by way of choosing the loaded (operating) Q of the series-resonant circuits consisting of AL1/AC1 and AL2/AC2.

In the arrangement of FIG. 1, in the initial mode of the ballast, when the inverter oscillates at its higher frequency, the magnitude of the voltage present across tank-capacitor C is so arranged as to be just adequate to cause lamp current to start flowing after the cathodes have become thermionic. Then, as soon as some lamp current is flowing, current will be provided to the base of control transistor CT; which will then act to cause

transistor FET to change to its non-conductive state, thereby causing the inverter to reduce its frequency to the lower frequency, which will then increase lamp current to its proper operational level.

If the lamp is non-connected, or if the lamp otherwise fails to conduct current, the ballast will remain in its initial mode of oscillating at the higher frequency.

In the arrangement of FIG. 2, the initial higher-frequency inverter mode is such as to provide proper cathode heating, but inadequately high voltage across the tank-capacitor to cause any significant amount of the current to flow through the lamp. Instead, to get the lamp ignited, after the initial mode has existed for about one second, a pulse is provided to the base of control transistor CT; which pulse is arranged to last for about 5 milli-seconds, thereby causing transistor FET to become non-conductive for a period of about 5 milli-seconds; which means that the inverter will oscillate at its lower frequency for that length of time.

After the cathodes have been pre-heated for about one second (or 1000 milli-seconds), the lamps are ready to ignite; and they then do indeed ignite within the 5 milli-second period during which the inverter oscillates at its lower frequency—this being so for the reason that the high-Q L-C circuit (which consists of tank-inductor L and tank-capacitor C) is resonant at or near this lower frequency; which means, due to so-called Q-multiplication, that the magnitude of the voltage developing across the tank-capacitor will increase until limited by whatever load is present thereacross.

After the lamps ignite (i.e., as soon as lamp current starts flowing), by means of lamp current sensing transformer CT3, control current will be provided to the base of control transistor CT, which will then assure that the inverter will remain in its position of oscillating at the lower frequency as long as lamp current is indeed flowing.

However, if the lamps were to fail to conduct current—perhaps because they were to become inoperative or removed—the inverter will revert to its initial mode of oscillating at its higher frequency; whereafter each 1000 milli-seconds it will for a period of 5 milli-seconds change mode to oscillate at the lower frequency.

If lamp current were to fail to flow, the magnitude of the voltage developing across the tank-capacitor will be limited by the Varistor, the (non-linear) characteristics of which are so chosen as to clamp the voltage magnitude to just the proper level to provide for proper lamp starting. Then, after the lamps have ignited, the magnitude of the voltage across the tank-capacitor will decrease to a lower level due to the loading provided by the lamps; which lower level is substantially lower than the level at which the Varistor provides for voltage clamping. Thus, after the lamps have ignited, current will cease to flow through the Varistor.

Since, in a series-excited parallel-loaded resonant high-Q L-C circuit, the power provided to the load is approximately proportional to the magnitude of the voltage developing across the load, the power provided to the Varistor when it is operative to effect voltage clamping is higher than that provided to the lamps during normal operation; and it is higher by a degree corresponding to the degree by which the lamps' starting voltage is higher than the lamps' operating voltage.

With two series-connected rapid-start lamps, the ratio between starting voltage and operating voltage is about 1.5.

Since the power provided to the lamps during normal operation is about 60 Watt, the power dissipated in the Varistor during any periods when it is constituting the load on the resonant L-C circuit will be about 90 Watt. However, even under the worst of circumstances, the Varistor can only be subject to this 90 Watt load for only about 5 milli-seconds once each 1000 milli-seconds; which means that the average dissipation of the Varistor can not exceed 0.5 Watt. Of course, these worst of circumstances would only occur if the lamp load were to be disconnected (or if it were to fail to ignite) for an extended period of time; in which case the output voltage provided from the ballast would alternate about once each second between a relatively low-magnitude minimum level and a relatively high-magnitude maximum level: the minimum level corresponding to a relatively high frequency, the maximum level corresponding to a relatively low frequency.

Additional Comments

a) The setting of adjustable resistor AR will determine the amount of power provided to the lamps during their normal operation; which implies that adjustable resistor AR may be used as a dimming means: the higher the resistance value of AR, the higher the power level provided to the lamps.
b) Transistor FET is a field effect transistor. However, a bi-polar transistor could just as well have been used.
c) It is possible by varying the amount of initial bias on the gate of transistor FET to control the effective initial impedance of this transistor, thereby effectively permitting a gradual or continuous feedback arrangement rather than the abrupt ON/-OFF feedback arrangement actually described.
d) It is believed that the present invention and its several attendant advantages and features will be understood from the preceeding description. However, without departing from the spirit of the invention, changes may be made in its form and in the construction and interrelationships of its component parts, the form herein presented merely representing the presently preferred embodiment.

I claim:

1. A combination comprising:

a source providing an AC voltage at an AC output, the source: (i) having control means responsive to a control input, (ii) when receiving the control input, being operative to cause the AC voltage to be of a substantially constant frequency, and (iii) when not receiving the control input, being operative to cause the frequency of the AC voltage to alternate between a higher frequency and a lower frequency;

gas discharge lamp means connected with the AC output and sometimes operative to be ignited and to draw a lamp current therefrom; and

sensing means connected in circuit with the gas discharge lamp means and operative: (i) to sense the presence of the lamp current whenever it is flowing, and (ii) in response to the presence of the lamp current, to provide the control input;

whereby: (i) whenever the gas discharge lamp means fails to draw the lamp current, the frequency of the AC voltage alternates between the higher and the lower frequency, and (ii) whenever the gas discharge lamp means does indeed draw the lamp current, the AC voltage is of the substantially constant frequency.

2. The combination of claim 1 wherein the substantially constant frequency equals the lower frequency.

3. The combination of claim 1 wherein, during a period when the frequency of the AC voltage does alternate between the higher and the lower frequency, most of the time during that period the AC voltage would be of the higher frequency.

4. The combination of claim 1 wherein: (i) the gas discharge lamp means comprises thermionic cathode means operative, after having been provided with cathode heating power for a brief period, to condition the gas discharge lamp to ignite from the AC voltage provided to it, and (ii) the source comprises cathode heating power supply means operative, whenever the control input is not being provided, to provide cathode heating power to the thermionic cathode means at a certain rate.

5. The combination of claim 4 wherein, whenever the control input is being provided, any cathode heating power provided by the cathode heating power supply means is substantially lower than said certain rate.

6. In a ballasting means for a gas discharge lamp, the ballasting means providing a first AC voltage at a first output and a second AC voltage at a second output, both outputs being connectable with the gas discharge lamp, both AC voltages being of the same frequency, the gas discharge lamp having a thermionic cathode connected with the first output and a pair of main lamp terminals connected with the second output, the improvement comprising:

control means connected within the ballasting means and operative to cause the ballasting means to operate in either of two modes: (i) a first mode wherein the magnitude of the first AC voltage is relatively high, thereby providing a relatively high rate of cathode heating power to the thermionic cathode, the first mode prevailing as long as substantially no power is being drawn from the second output, and (ii) a second mode where the magnitude of the first AC voltage is relatively low, thereby providing a relatively low rate of cathode heating power to the thermionic cathode, the second mode prevailing as long as a substantial amount of power is being drawn from the second output.

7. In a ballasting means for a gas discharge lamp, the ballasting means providing a first AC voltage at a first output and a second AC voltage at a second output, both outputs being connectable with the gas discharge lamp, both AC voltages being of the same frequency, the gas discharge lamp having a thermionic cathode connected with the first output and a pair of main lamp terminals connected with the second output, the improvement comprising:

control means connected within the ballasting means and operative to cause the ballasting means to operate in either of two modes: (i) a first mode wherein the magnitude as well as the frequency of the first AC voltage is relatively high, thereby providing a relatively high rate of cathode heating power to the thermionic cathode, and (ii) a second mode where the magnitude as well as the frequency of the first AC voltage is relatively low, thereby providing a relatively low rate of cathode heating power to the thermionic cathode, the second mode prevailing as long as a substantial amount of power is being drawn from the second output.

8. The improvement of claim 7 wherein: (i) the first mode represents a lamp conditioning mode, and (ii) the second mode represents a lamp operating mode.

9. A combination comprising:

a source providing an AC power line voltage at a pair of power line terminals; the AC power line voltage having a first frequency;

gas discharge lamp means having a pair of lamp terminals; and

ballast means having a pair of input terminals connected with the power line terminals and a pair of output terminals connected with the lamp terminals; the ballast means being operative to provide an alternating output voltage at the output terminals; the alternating output voltage having a second frequency; the second frequency being substantially constant whenever the gas discharge lamp draws more than a certain amount of power from the output terminals; the second frequency exhibiting periodic variations whenever the gas discharge lamp draws substantially less than said certain amount of power from the output terminals.

10. The combination of claim 9 wherein: (i) the second frequency is substantially higher than the first frequency; and (ii) the periodic variations have a period defined as the period of the periodic variations, which period of the periodic variations is substantially longer than the period of the AC power line voltage.

11. A combination comprising:

a source providing an AC power line voltage at a pair of power line terminals; the AC power line voltage having a first frequency and a first period;

gas discharge lamp having a pair of lamp terminals; and

a ballasting circuit having a pair of input terminals connected with the power line terminals and a pair of output terminals connected with the lamp terminals; the ballastig circuit being operative to provide an alternating output voltage at the output terminals; the alternating output voltage having a second frequency substantially higher than the first frequency as well as a second period substantially shorter than the first period; the second frequency being substantially constant whenever the gas discharge lamp draws more than a certain amount of power from the output terminals; the second frequency exhibiting periodic variations whenever the gas discharge lamp draws substantially less than said certain amount of power from the output terminals; the periodic variations being characterized by having a third frequency and a third period; the third frequency being substantially lower than the first frequency; the third period being substantially longer than the first period.

12. The arrangement of claim 11 wherein the ballasting circuit is also characterized by having frequency-discriminating circuitry connected in circuit with its output terminals.

* * * * *

EXHIBIT E

US005489823A

United States Patent [19]

Nilssen

[11] **Patent Number:** **5,489,823**

[45] **Date of Patent:** *** Feb. 6, 1996**

[54] **ELECTRONIC BALLAST FOR GAS DISCHARGE LAMP**

[76] Inventor: **Ole K. Nilssen**, 408 Caesar Dr., Barrington, Ill. 60010

[*] Notice: The portion of the term of this patent subsequent to Nov. 17, 2009, has been disclaimed.

[21] Appl. No.: **997,138**

[22] Filed: **Dec. 22, 1992**

Related U.S. Application Data

[63] Continuation of Ser. No. 751,587, Aug. 22, 1991, abandoned, which is a continuation of Ser. No. 546,267, Jun. 29, 1990, abandoned, which is a continuation-in-part of Ser. No. 787,692, Oct. 15, 1985, abandoned, which is a continuation of Ser. No. 644,155, Aug. 27, 1984, abandoned, which is a continuation of Ser. No. 555,426, Nov. 23, 1983, abandoned, which is a continuation-in-part of Ser. No. 330,599, Dec. 14, 1981, Pat. No. 4,441,087, which is a continuation of Ser. No. 973,741, Dec. 28, 1978, abandoned, which is a continuation-in-part of Ser. No. 890,586, Mar. 20, 1978, Pat. No. 4,184,128, said Ser. No. 751,587is a continuation-in-part of Ser. No. 717,860, Jun. 19, 1991, which is a continuation of Ser. No. 636,246, Dec. 31, 1990, abandoned, which is a continuation of Ser. No. 787,692, Oct. 15, 1985, said Ser. No. 555,426, is a continuation of Ser. No. 178,107, Aug. 14, 1980, abandoned, which is a continuation-in-part of Ser. No. 23,849, Mar. 26, 1979, Pat. No. 4,279,011.

[51] **Int. Cl.**^{6} **H05B 37/00**

[52] **U.S. Cl.** .... **315/227 R**; 315/244; 315/247; 315/209 R

[58] **Field of Search** .... 315/226, 227, 315/244, 247, 171, 172, 174, 209 R, 219, 223, 239, DIG. 7; 363/65, 67, 68

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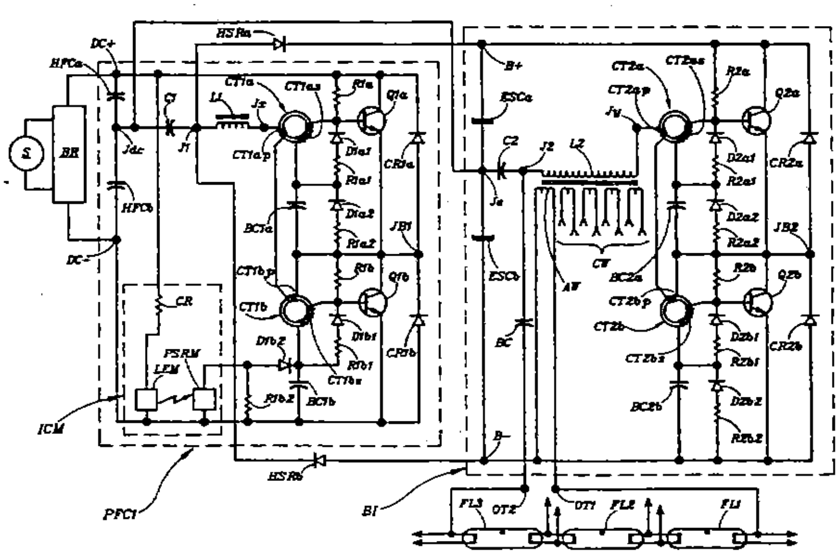
Primary Examiner—Frank Gonzalez
Assistant Examiner—Reginald A. Ratliff

[57] **ABSTRACT**

A first half-bridge inverter is powered from an unfiltered full-wave-rectified ordinary 60 Hz electric utility power line voltage. This first inverter provides at a first inverter output, across which is series-connected a first tuned L-C circuit, a first squarewave voltage of fundamental frequency between about 30 and 33 kHz; which first squarewave voltage is magnitude-modulated at 120 Hz. The first tuned L-C circuit, which is series-resonant at about 30 kHz, is parallel-loaded by a full-wave high-frequency rectifier whose DC output is applied to substantially constant-magnitude DC voltage existing across a pair of energy-storing capacitors. At a constant 30 kHz inverter frequency, the waveshape of the current drawn from the power line is substantially that of a squarewave in phase with the power line voltage, thereby giving rise to a power factor of about 90%. However, by frequency-modulating the first inverter at 120 Hz, the waveshape of the line current is made to be substantially that of a sinewave in phase with the power line voltage, thereby giving rise to a power factor close to 100% and a total harmonic distortion of negligible magnitude. A second half-bridge inverter is powered from the substantially constant-magnitude DC voltage and provides at a second inverter output, across which is series-connected a second tuned L-C circuit, a second squarewave voltage of fundamental frequency between about 30 and 33 kHz. The second tuned L-C circuit, which is series-resonant at about 30 kHz, is parallel-loaded by three series-connected fluorescent lamps. The magnitude of the current supplied to these three lamps is adjustable be adjusting the frequency of the second inverter between about 30 and 33 kHz.

14 Claims, 3 Drawing Sheets



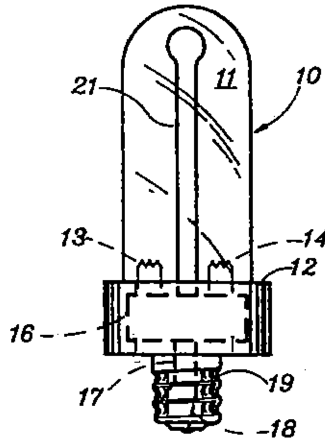


Fig. 1

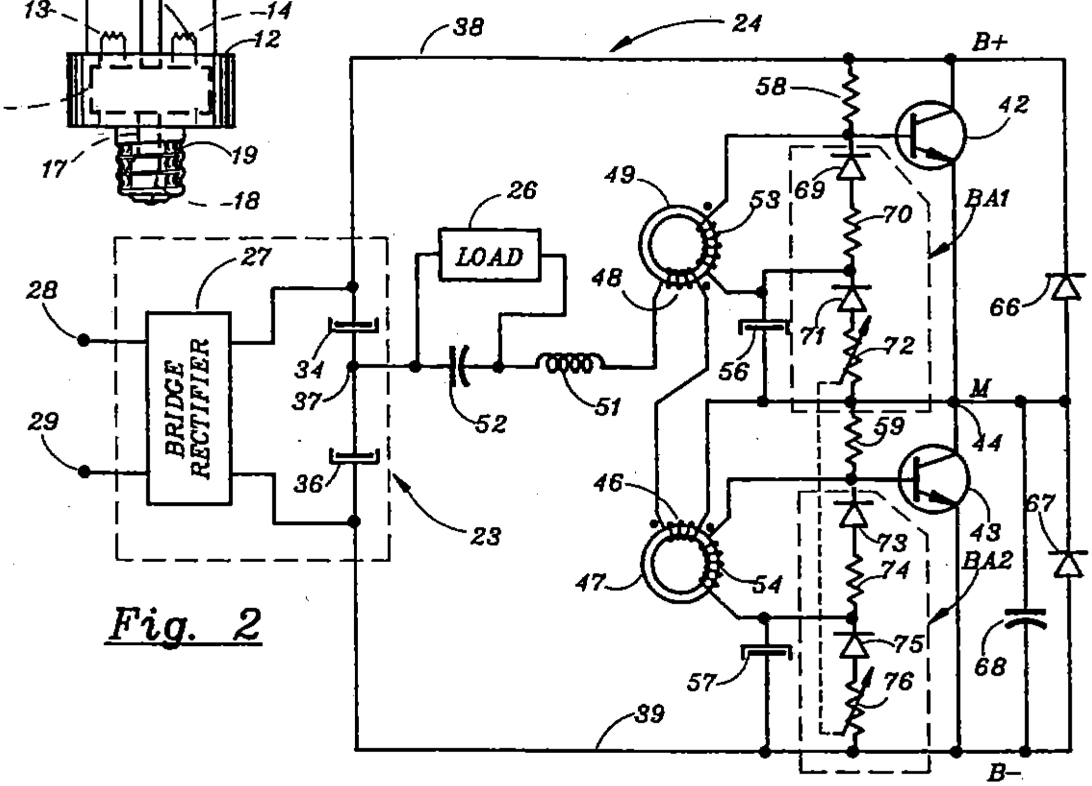


Fig. 2

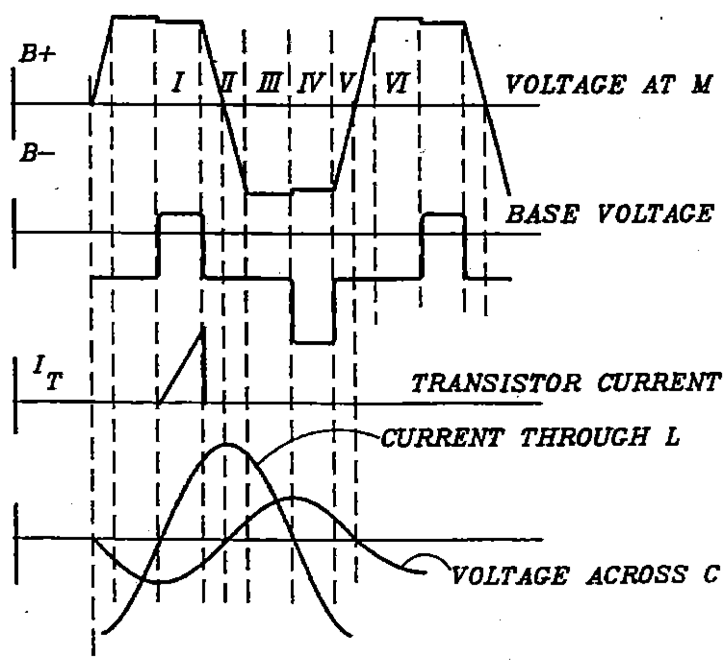
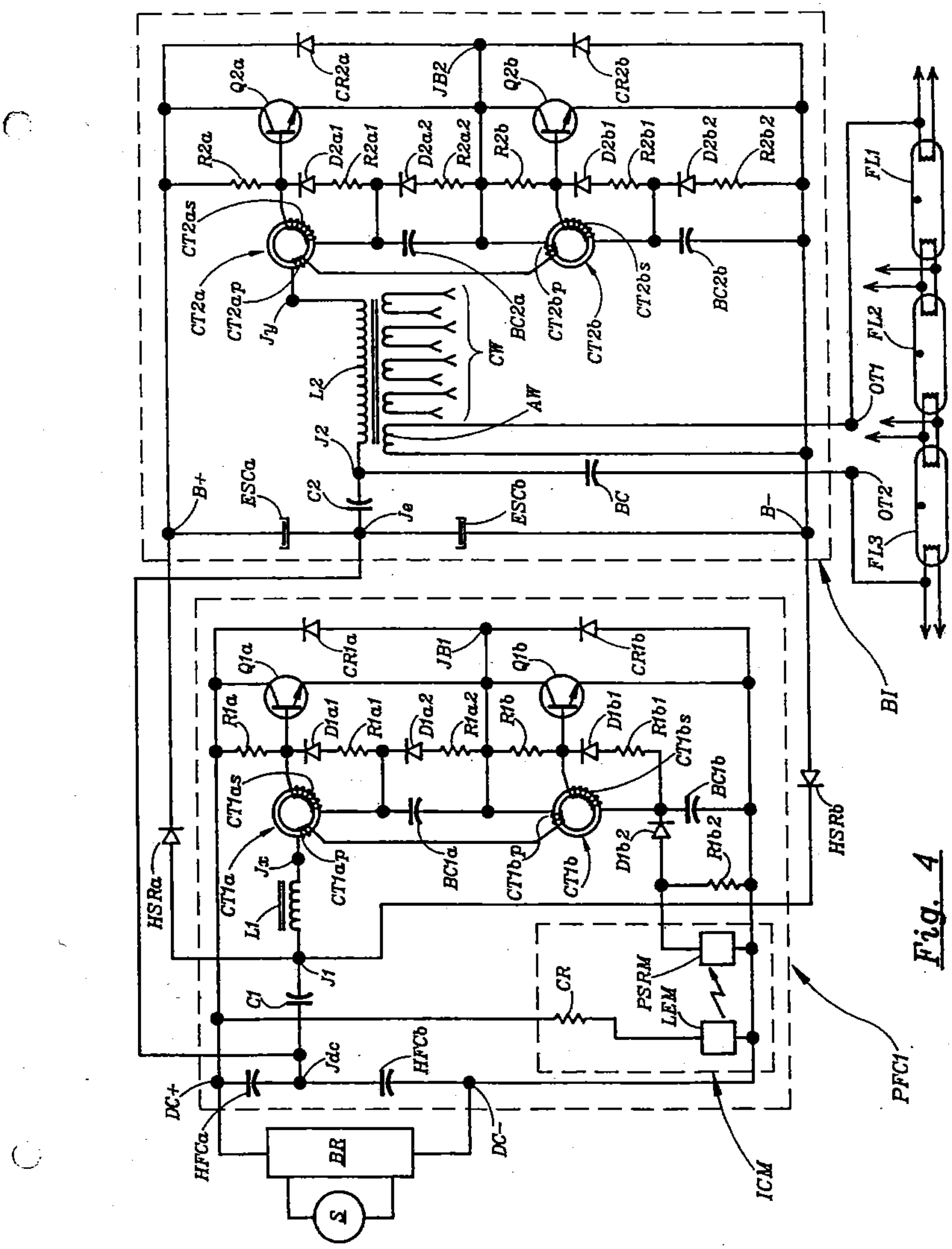


Fig. 3A

Fig. 3B

Fig. 3C

Fig. 3D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CT2bs
BC2b
FL2
L2
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AW
OT1
J2
B+
ESCa
C2
Je
ESCb
BC
B-
OT2
FL3
BI
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JB1
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CR1b
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R1a1
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R1a2
R1b
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R1b1
CT1bs
CT1as
BC1b
HSRb
HSRa
CT1a
Jx
CT1ap
BC1a
CT1bp
CT1b
D1b2
R1b2
L1
J1
C1
CR
PSRM
LEM
Jdc
HFCb
DC+
HFCa
BR
S
DC-
ICM
PFC1

Fig. 4

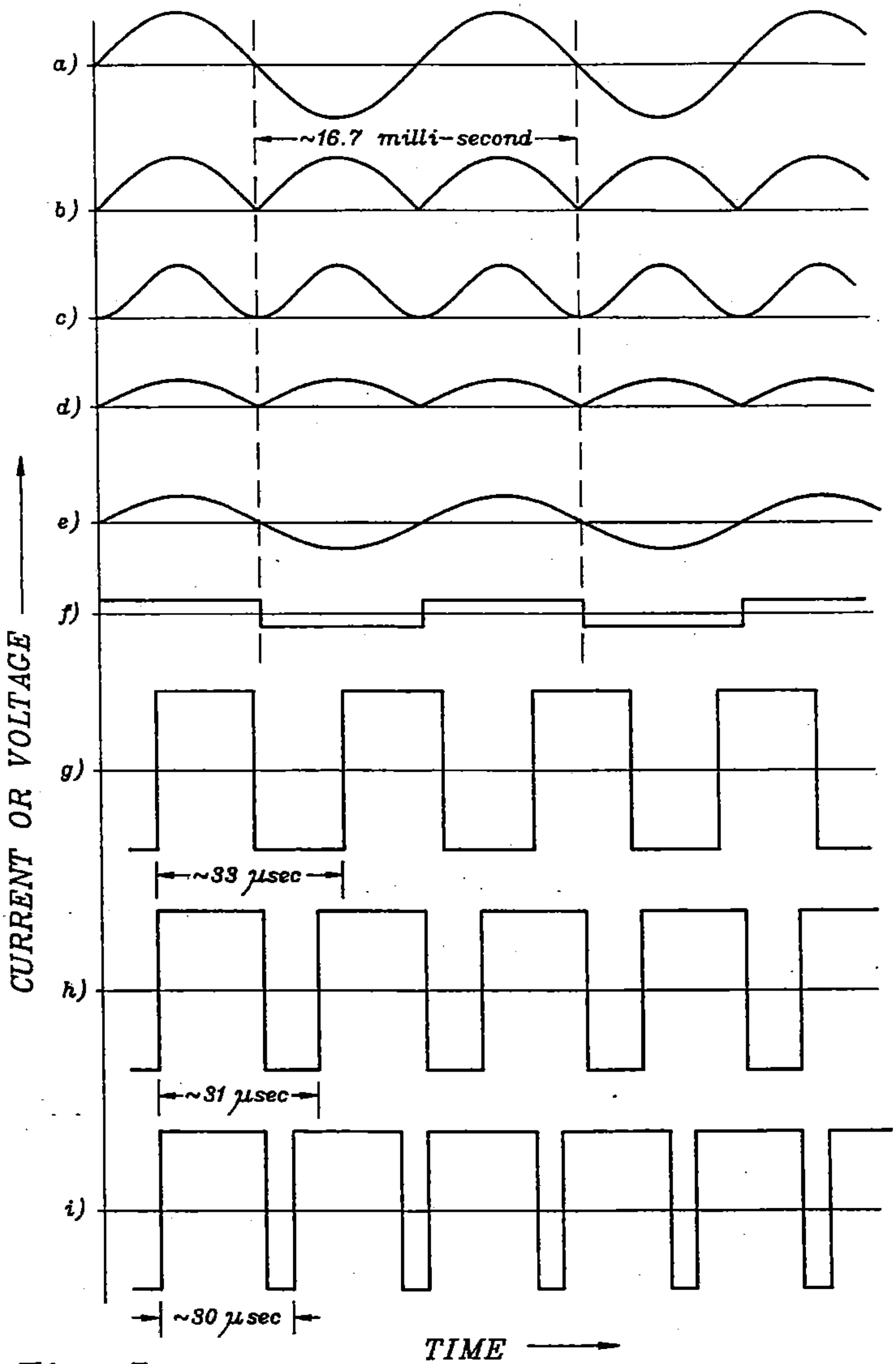
a)
~16.7 milli-second
b)
c)
d)
e)
f)
g)
~33 μsec
h)
~31 μsec
i)
~30 μsec
CURRENT OR VOLTAGE
TIME

Fig. 5

ELECTRONIC BALLAST FOR GAS DISCHARGE LAMP

RELATED APPLICATIONS

The present application is a continuation of Ser. No. 07/751,587 filed Aug. 22, 1991, now abandoned; which is a continuation of Ser. No. 07/546,267 filed Jun. 29, 1990, now abandoned; which is a continuation-in-part of Ser. No. 06/787,692 filed Oct. 15, 1985, now abandoned; which is a continuation of Ser. No. 06/644,155 filed Aug. 27, 1984, now abandoned; which is a continuation of Ser. No. 06/555, 426 filed Nov. 23, 1983, now abandoned; which was a continuation of Ser. No. 06/178,107 filed Aug. 14, 1980, now abandoned; which application Ser. No. 07/751,587 is also a continuation-in-part of Ser. No. 07/717,860 filed Jun. 19, 1991; which is a continuation of Ser. No. 07/636,246 filed Dec. 31, 1990, now abandoned; which is a continuation of Ser. No. 06/787,692 filed Oct. 15, 1985, now abandoned; which is a continuation of Ser. No. 06/644,155 filed Aug. 27, 1984, now abandoned; which is a continuation of Ser. No. 06/555,426 filed Nov. 23, 1983, now abandoned; which is a continuation of Ser. No. 06/178,107 filed Aug. 14, 1980, now abandoned; which application Ser. No. 06/555,426 is also a continuation-in-part of Ser. No. 06/330,599 filed Dec. 14, 1981, now U.S. Pat. No. 4,441,087; which is a continuation of Ser. No. 973,741 filed Dec. 28, 1978, now abandoned; which is a continuation-in-part of Ser. No. 890,586 filed Mar. 20, 1978, now U.S. Pat. No. 4,184,128; which application Ser. No. 06/178,107 is also a continuation-in-part of Ser. No. 23,849 filed Mar. 26, 1979, now U.S. Pat. No. 4,279,011.

BACKGROUND OF THE INVENTION

Field of Invention

Instant invention relates to power-factor-corrected inverter-type fluorescent lamp ballasting means operable to be powered from an ordinary electric utility power line.

SUMMARY OF THE INVENTION

1. Objects of the Invention

An object of the present invention is that of providing a reliable cost-effective fluorescent lamp ballasting means.

This as well as other objects, features and advantages of the present invention will become apparent from the following description and claims.

2. Brief Description

In its preferred embodiment, instant invention comprises a first half-bridge inverter that is powered from an unfiltered full-wave-rectified ordinary 60 Hz electric utility power line voltage. This first inverter provides at a first inverter output, across which is series-connected a first tuned L-C circuit, a first squarewave voltage of fundamental frequency between about 30 and 33 kHz; which first squarewave voltage is magnitude-modulated at 120 Hz.

The first tuned L-C circuit, which is series-resonant at about 30 kHz, is parallel-loaded by a full-wave high-frequency rectifier whose DC output is applied to substantially constant-magnitude DC voltage existing across a pair of energy-storing capacitors.

At a constant 30 kHz inverter frequency, the waveshape of the current drawn from the power line is substantially that of a squarewave in phase with the power line voltage, thereby giving rise to a power factor of about 90%. However, by frequency-modulating the first inverter at 120 Hz,

the waveshape of the line current is made to be substantially that of a sinewave in phase with the power line voltage, thereby giving rise to a power factor close to 100% and a total harmonic distortion of negligible magnitude.

A second half-bridge inverter is powered from the substantially constant-magnitude DC voltage and provides at a second inverter output, across which is series-connected a second tuned L-C circuit, a second squarewave voltage of fundamental frequency between about 30 and 33 kHz.

The second tuned L-C circuit, which is also series-resonant at about 30 kHz, is parallel-loaded by three series-connected fluorescent lamps. The magnitude of the current supplied to these three lamps is adjustable be adjusting the frequency of the second inverter between about 30 and 33 kHz.

BRIEF DESCRIPTION OF THE DRAWING

FIG. 1 schematically illustrates a compact, screw-in, self-ballasted fluorescent lamp assembly; which lamp may advantageously comprise the type of ballasting means represented by the present invention.

FIG. 2 is a schematic diagram of a half-bridge inverter and lasting circuit of the basic type used in the preferred embodiment of the present invention.

FIGS. 3A–3D illustrate the waveforms of various voltages and currents associated with the ballasting circuit of FIG. 2.

FIG. 4 illustrates the preferred embodiment of the present invention.

FIG. 5(*a*–*i*) illustrates various voltage and current waveforms associated with the operation of the preferred embodiment of the present invention.

DESCRIPTION OF THE INVENTION

Details of Construction of Screw-in Fluorescent Lamp

FIG. 1 illustrates a screw-in gas discharge lamp unit **10** comprising a folded fluorescent lamp **11** secured to an integral base **12**. The lamp comprises two cathodes **13**, **14** which are supplied with the requisite high operating voltage from a frequency-converting power supply and ballasting circuit **16**; which, because of its compact size, conveniently fits within base **12**.

Circuit **16** is connected by leads **17**, **18** to a screw-type plug **19** adapted for screw-in insertion into a standard Edison-type incandescent lamp socket at which ordinary 120 Volt/60 Hz power line voltage is available.

Details of Construction of the FIG. 2 Circuit

In FIG. 2, a power supply **23** is connected with the 120 Volt/60 Hz power line voltage and provides a center-tapped DC output voltage for supplying a high-efficiency half-bridge inverter circuit **24**. The inverter circuit is operable to provide a high-frequency (20–30 kHz) high-magnitude current-limited voltage to a load **26**, which actually represents fluorescent lamp **11** of FIG. 1.

Power supply **23** comprises bridge rectifier **27** which connects with 120 Volt/60 Hz power line terminals **28**, **29** and provides full-wave rectified power line voltage to two series-connected filter capacitors **34**, **36**; which filter capacitors are: i) connected together at a center-tap **37**, and ii) connected between a positive B+ bus **38** and a negative B− bus **39**.

Inverter circuit **24** is a half-bridge inverter comprising transistors **42**, **43** connected in series between the B+ bus and the B− bus. The collector of transistor **42** is connected to the B+ bus 38, the emitter of transistor 42 and the collector of transistor 43 are connected to a midpoint line 44 ("M"), and the emitter of transistor 43 is connected to the B− bus 39.

Midpoint line 44 is connected to center-tap 37 through a primary winding 46 of a toroidal saturable core transformer 47, a primary winding 48 on an identical transformer 49, a tank inductor 51 (L) and a series-connected tank capacitor 52 (C). Inductor 51 and capacitor 52 are energized upon alternate transistor conduction in manner to be described later. Load 26 is connected in parallel with capacitor 52.

Drive current to the base terminals of transistors 42 and 43 is provided by secondary windings 53, 54 of transformers 49, 47, respectively. Winding 53 is also connected to midpoint line 44 through a bias capacitor 56, while winding 54 is connected to the B− bus 39 through an identical bias capacitor 57. The base terminals of transistors 42 and 43 are also connected to lines 38 and 44 through bias resistors 58 and 59, respectively. Shunt diodes 66 and 67 are connected across the collector-emitter terminals of transistors 42 and 43, respectively. A capacitor 68 is connected across the collector-emitter terminals of transistor 43 to restrain the rate of voltage rise across those terminals.

A first optional biasing arrangement BA1 comprises a diode 69 connected with its cathode to the base of transistor 42 and with its anode to the cathode of a diode 71 by way of a resistor 70; the anode of diode 71 is connected with the emitter of transistor 42 by way of a resistor 72; the cathode of diode 71 is connected with the un-dotted side of secondary winding 53 of transformer 49. A second optional biasing arrangement BA2 comprises a diode 73 connected with its cathode to the base of transistor 43 and with its anode to the cathode of a diode 75 by way of a resistor 74; the anode of diode 75 is connected with the emitter of transistor 43 by way of a resistor 76; the cathode of diode 76 is connectred with the un-dotted side of secondary winding 54 of transformer 47.

Details of Operation of the FIG. 2 Circuit

The operation of the circuit of FIG. 2 can best be understood with additional reference to FIG. 3, which illustrates significant portions of the waveforms of the voltage at midpoint M (FIG. 3A), the base-emitter voltage on transistor 42 (FIG. 3B), the current through transistor 42 (FIG. 3C), and the capacitor 52 voltage and the inductor 51 current (FIG. 3D).

Starting at a point where transistor 42 first starts to conduct, current flows from the B+ bus 38 through windings 46 and 48 and inductor 51 to charge capacitor 52 and returns to the B+ bus through capacitor 34 (refer to the time period designated I in FIG. 3). When the saturable transformer 49 saturates at the end of period I, drive current to the base of transistor 42 will terminate, causing voltage on the base of the transistor to drop to the negative voltage stored on bias capacitor 56 in a manner to be described, causing this transistor to become non-conductive. As shown in FIG. 3c, current-flow in transistor 43 terminates at the end of period I.

However, since the current flowing through inductor 51 cannot change instantaneously, this current will now continue to flow from the B− bus 39 through capacitor 68, eventually causing the voltage at midpoint line 44 to drop to the voltage level on the B− bus (period II in FIG. 3). Thus, capacitor 68 restrains the rate of voltage change across the collector and emitter terminals of transistor 42.

The current through inductor 51 reaches its maximum value when the voltage at midpoint line 44 is zero. During period III, the current will continue to flow through inductor 51 but will be supplied from the B− bus through shunt diode 67. It will be appreciated that during the latter half of period II and all of period III, positive current is being drawn from a negative voltage; which, in reality, means that energy is being returned to the power supply through a path of relatively low impedance.

When the inductor current reaches zero at the start of period IV, the current through the primary winding 46 of the saturable inductor 47 will cause a current to flow out of its secondary winding 54 to cause transistor 43 to become conductive, thereby causeing a reversal in the direction of current through inductor 51 and capacitor 52. When transformer 47 saturates at the end of period IV, the drive current to the base of transistor 43 terminates and the current through inductor 51 will be supplied through capacitor 68, causing the voltage at midpoint 44 to rise (Period V). When the voltage at the midpoint line M reaches the voltage on the B+ bus, the current will then flow through shunt diode 66 (period VI). The cycle is then repeated.

As seen in FIG. 3, saturable transformers 47, 49 provide transistor drive current only after the current through inductor 51 has diminished to zero. Further, the the transistor drive current is terminated before the current through inductor 51 has reached its maximum amplitude. This coordination of base drive current and inductor current is achieved because of the series-connection between the inductor 51 and the primary windings 46, 48 of saturable transformers 47, 49, respectively.

The series-connected combination of inductor 51 and capacitor 52 is energized upon the alternate conduction of transistors 42 and 43. With a large value of capacitance of capacitor 52, very little voltage will be developed across its terminals. As the value of this capacitance is decreased, however, the voltage across this capacitor will increae. As the value of capacitor 52 is reduced to achieve resonance with inductor 51, the voltage on the capacitor will rise and become infinite in a loss-free circuit operating under ideal conditions.

It has been found desirable to regulate the transistor inversion frequency, determined mainly by the saturation time of saturable transformers 47, 49, to be equal to or higher than the natural resonance frequency of the inductor and capacitor combination in order to provide a high voltage output to external load 26.

Due to so-called Q-multiplication, a high-magnitude voltage develops across capacitor 52 as the transistor inversion frequency approaches the natural resonance frequency of the series-combination of inductor 51 and capacitor 52.

When inverter circuit 24 is used in the self-ballasted fluorescent lamp of FIG. 1, it has been found that the inversion frequency may be about equal to the natural resonance frequency of the series L-C tank circuit consisting of inductor 51 and capacitor 52. However, if the capacitance value of capacitor 52 is reduced below the point of resonance, unacceptably high transistor currents will result and transistor burn-out will occur.

The sizing of capacitor 52 is determined by the particular application of inverter circuit 24; but, as long as the combined load presented to the output of inverter transistors 42, 43, has an effective inductance value sufficient to provide adequate energy storage for self-sustained transistor inverter action, the current-feedback provided by saturable transformers 47, 49 will effect alternate transistor conduction without the need for additional voltage-feedback.

Because the voltages across transistors 42, 43 are relatively low (due to the absolute voltage-clamping effect of capacitors 34, 36), the half-bridge inverter 24 is very reliable. The absence of switching transients minimizes the possibility of transistor burn-out.

Inverter circuit **24** comprises means for supplying reverse bias to the conducting transistor upon saturation of its associated saturable transformer. For this purpose, capacitors **56** and **57** are charged to negative voltages as a result of reset currents flowing into secondary windings **53**, **54** from the bases of transistors **42**, **43**, respectively. This reverse current rapidly turns off a conducting transistor to increase its switching speed and to achieve high inverter switching efficiency.

When a transistor base-emitter junction is reversely biased, it exhibits the characteristics of a Zener diode, having a reverse breakdown voltage on the order of 8 to 14 Volt for transistors typically used in high-voltage inverters.

Since load **56** comprises a fluorescent lamp, the maximum magnitude of the voltage across capacitor **52** will be limited by the lamp's ignition and operating characteristics, thereby effectively preventing voltages across inductor **51** and capacitor **52** from ever reaching destructive levels.

The above-presented explanation of the operation of the FIG. **2** inverter circuit was based on the two biasing arrangements (BA1 and BA2) being non-connected.

With these biasing arrangements actually connected as indicated, the inverter's operation will become independent of the exact magnitudes of the transistors' base-emitter Zenering voltages. Instead, the magnitude of the negative bias voltage established on each of capacitors **56** and **57** can now be chosen by choice of resistance value of resistor **72** and/or resistor **76**: the lower the resistance value, the lower the magnitude of the associated negative bias voltage; and, in turn, the longer the transistors' ON-time, the lower the inverter's self-oscillating frequency, and the higher the magnitudes of the inverter's output current and power.

By providing for means whereby the resistance values of resistors **72** and **76** can be manually adjusted (in tandem and/or individually), the power provided to the fluorescent lamp may be correspondingly adjusted: the lower the resistance values, the more power provided to the lamp.

Moreover, due to the negative feedback effect inherently provided by resistors **72** and **74**, the inverter may be made to operate safely even with the fluorescent lamp being non-connected.

This negative feedback effect is due to the fact that, as the magnitude of the current flowing through the L-C circuit increases, the magnitudes of the drive currents provided to the transistors' bases increase, and the magnitudes of the currents drawn out of capacitors **56** and **57** increase correspondingly; which, in turn, increases the magnitudes of the negative bias voltages present on these capacitors to the point where the magnitudes of the currents flowing through resistors **72** and **76** equal those of the increased base currents. However, the increased negative bias voltages will inherently shorten the transistors' ON-times; which, in turn, will increase the inverter frequency, thereby reducing the inverter's output current; etc. In other words, the indicated biasing arrangements provide for an automatic self-limiting of the magnitude of the inverter's output current.

Additional Comments re the FIG. **2** Circuit

a) With commonly available components, inverter circuit **24** can be made to operate efficiently at any frequency between a few kHz to perhaps as high as 50 kHz. However, for various well-known reasons (i.e., eliminating audible noise, minimizing physical size, and maximizing efficiency), the frequency actually chosen for the lamp unit of FIG. **1** was in the range of 20 to 30 kHz.

b) The fluorescent lighting unit of FIG. **1** could be made in such manner as to permit fluorescent lamp **11** to be disconnectable from its base **12** and ballasting means **16**. However, if powered with normal line voltage without its lamp load connected, frequency-converting power supply and ballasting circuit **16** is apt to self-destruct.

To avoid such self-destruction, arrangements can readily be made whereby the very act of removing the load automatically establishes a situation that prevents the possible destruction of the power supply and ballasting means. For instance, with the tank capacitor (**52**) being permanently connected with the lamp load (**11**)—thereby automatically being removed whenever the lamp is removed—the inverter circuit is protected from self-destruction.

c) At frequencies above a few kHz, the load represented by a fluorescent lamp—once it is ignited—is substantially resistive. Thus, with the voltage across lamp **11** being of a substantially sinusoidal waveform (as indicated in FIG. 3*d*), the current through the lamp will also be substantially sinusoidal in waveshape.

d) In the fluorescent lamp unit of FIG. **1**, fluorescent lamp **11** is connected with power supply and ballasting circuit **16** in the exact same manner as is load **26** connected with the circuit of FIG. **2**. That is, it is connected in parallel with the tank capacitor (**52**) of the L-C series-resonant circuit. As is conventional in instant-start fluorescent lamps—such as lamp **11** of FIG. **1**—the two terminals from each cathode are shorted together, thereby to constitute a situation where each cathode effectively is represented by only a single terminal. However, it is not necessary that the two terminals from each cathode be shorted together; in which case—for instant-start operation— connection from a lamp's power supply and ballasting means need only be made with one of the terminals of each cathode.

e) It is noted that the transistor's ON-time is shorter than half the period of the inverter's high frequency squarewave voltage output; which voltage output is illustrated by FIG. 3A.

The fact that each of the transistors' ON-times is shorter than half the period of the inverter's high frequency output voltage (or output current) is important: it inherently provides for a situation where the two transistors are manifestly prevented from conducting at the same time, thereby providing protection against circuit failure due to excess-magnitude transistor currents.

f) By adjusting the resistance values of resistors **72** and/or **76**, the ON-times of the associated transistors are adjusted accordingly. For instance, by increasing the resistance value of resistor **76**, the ON-time associated with transistor **43** is shortened; and, as a result, the magnitude of the current prvided to the load **26** is reduced.

Details of Construction of the Preferred Embodiment

In FIG. 4, a source S represents an ordinary electric utility power line providing 120 Volt/60 Hz power line voltage to the input of a bridge rectifier BR, whose DC output is applied between a DC− bus and a DC+ bus.

A high-frequency filter capacitor HFCa is connected between a junction Jdc and the DC+ bus; a high-frequency filter capacitor HFCb is connected between the DC− bus and junction Jdc.

A tank capacitor C1 is connected between junction Jdc and a junction J1; a tank inductor L1 is connected between junction J1 and a junction Jx.

A transistor Q1a is connected with its collector to the DC+ bus and with its emitter to a junction bus JB1; a transistor Q1b is connected with its collector to junction bus JB1 and with its emitter to the DC− bus. A commutating rectifier CR1a is connected with its cathode to the DC+ bus and with its anode to junction bus JB1; a commutating rectifier CR1b is connected with its cathode to junction bus JB1 and with its anode to the DC− bus.

Primary windings CT1ap and CT1bp of saturable current transformers CT1a and CT1b, respectively, are series-connected between junction Jx and junction bus JB1.

Secondary winding CT1as of transformer CT1a is connected between the base of transistor Q1a and the cathode of a diode D1a2, whose anode is connected with junction bus JB1 via a resistor R1a2. A diode D1a1 is connected with its cathode to the base of transistor Q1a and with its anode to the cathode of diode D1a2 via a resistor R1a1. A resistor R1a is connected between the DC+ bus and the base of transistor Q1a. A bias capacitor BC1a is connected between the cathode of diode D1a2 and junction bus JB1.

Secondary winding CT1bs of transformer CT1b is connected between the base of transistor Q1b and the cathode of a diode D1b2, whose anode is connected with the DC− bus via a resistor R1b2. A diode D1b1 is connected with its cathode to the base of transistor Q1b and with its anode to the cathode of diode D1b2 via a resistor R1b1. A resistor R1b is connected between junction bus JB1 and the base of transistor Q1b. A bias capacitor BC1b is connected between the cathode of diode D1b2 and the DC-bus.

A high-speed rectifier HSRa is connected with its anode to junction J1 and with its cathode to a B+ bus; a high-speed rectifier HSRb is connected with its cathode to junction J1 and with its anode to a B− bus.

An energy-storing capacitor ESCa is connected between a junction Je and the B+ bus, junction Je being connected with junction Jdc; an energy-storing capacitor ESCb is connected between junction Je and the B− bus.

A tank capacitor C2 is connected between junction Je and a junction J2; a tank inductor is connected between junction J2 and a junction Jy.

A transistor Q2a is connected with its collector to the B+ bus and with its emitter to a junction bus JB2; a transistor Q2b is connected with its collector to junction bus JB2 and with its emitter to the B− bus. A commutating rectifier CR2a is connected with its cathode to the B+ bus and with its anode to junction bus JB2; a commutating rectifier CR2b is connected with its cathode to junction bus JB2 and with its anode to the B-bus.

Primary windings CT2ap and CT2bp of saturable current transformers CT2a and CT2b, respectively, are series-connected between junction Jy and junction bus JB2.

Secondary winding CT2as of transformer CT2a is connected between the base of transistor Q2a and the cathode of a diode D2a2, whose anode is connected with junction bus JB2 via a resistor R2a2. A diode D2a1 is connected with its cathode to the base of transistor Q2a and with its anode to the cathode of diode D2a2 via a resistor R2a1. A resistor R2a is connected between the B+ bus and the base of transistor Q2a. A bias capacitor BC2a is connected between the cathode of diode D2a2 and junction bus JB2.

Secondary winding CT2bs of transformer CT2b is connected between the base of transistor Q2b and the cathode of a diode D2b2, whose anode is connected with the B-bus via a resistor R2b2. A diode D2b1 is connected with its cathode to the base of transistor Q2b and with its anode to the cathode of diode D2b2 via a resistor R2b1. A resistor R2b is connected between junction bus JB2 and the base of transistor Q2b. A bias capacitor BC2b is connected between the cathode of diode D2b2 and the B-bus.

An auxiliary winding AW is wound as a loosely coupled secondary winding on tank inductor L2 and connected between the B− bus and an output terminal OT1. A DC blocking capacitor BC is connected between junction J2 and an output terminal OT2.

Also wound on tank inductor L2 are four cathode windings CW; which four cathode windings are connected with corresponding pairs of cathode terminals of three series-connected fluorescent lamps FL1, FL2 and FL3; which three fluorescent lamps are series-connected across output terminals OT1 and OT2.

An inverter control means ICM is connected between the DC-bus and the DC+ bus, as well as with the anode of diode D1b2; which inverter control means consists of: (i) a photo-sensitive resistive means PSRM connected between the anode of diode D1b2 and the DC− bus; and (ii) a light-emitting means LEM, such as a light-emitting diode (or LED), connected in series with a control resistor CR between the DC− bus and the DC+ bus. Light-emitting means LEM is so positioned and arranged that its light output impinges on a light-receptive part of photo-sensitive resistive means PSRM.

The half-bridge inverter with Q1a and Q1b as its switching transistors is identified as power-factor-correcting inverter PFCI; and the half-bridge inverter with Q2a and Q2b as its switching transistors is identified as ballast inverter BI.

Details of Operation of the Preferred Embodiment

The operation the preferred embodiment of instant invention may best be understood when read with reference to FIG. 5; which illustrates various current and voltage waveforms associated with the operation of the circuit arrangement of FIG. 4.

With reference to the waveforms of FIG. 5 and the circuit arrangement of FIG. 4, as long as the magnitude of the DC voltage existing between the B− bus and the B+ bus remains substantially constant, waveform: (a) represents that of the 120 Volt/60 Hz power line voltage supplied from source S; (b) represents the corresponding DC voltage present between the DC− bus and the DC+ bus; (c) represents the net current provided via high-speed rectifiers HSRa and HSRb to energy-storing capacitors ESCa and ESCb; (d) represents the current drawn from the DC output of bridge rectifier BR; and (e) represents the waveform of the current drawn from source S. Waveform (f) represents the current that would be drawn from source S in case the light from light-emitting means LEM were to be kept at a constant intensity, such as would occur if a filter capacitor were to be connected thereacross.

The two half-bridge inverters (PFCI and BI) both operate in the same basic manner as does the half-bridge inverter of FIG. 2.

In case of the PFCI inverter, the load constitutes the substantially constant-magnitude DC voltage present across energy-storing capacitors ESCa and ESCb; in case of the BI inverter, the principal load constitutes the three series-connected fluorescent lamps FL1, FL2 and FL3.

In the overall operation of the circuit of FIG. 4, the PFCI inverter continually charges the two energy-storing capacitors (ESCa/ESCb), while the BI inverter continually discharges these same energy-storing capacitors. Thus, the magnitude of the DC voltage across these energy-storing capacitors will stabilize at a point where the average rate of capacitor-charging equals the average rate of capacitor-discharging.

While the average rate of capacitor-charging is essentially fixed for a given magnitude of the voltage supplied by the source (S), the average rate of capacitor-discharging increases with increasing magnitude of the DC voltage present across the energy-storing capacitors (i.e., between the B− bus and the B+ bus).

Absent any control by inverter control means ICM, a basic characteristic of the series-driven parallel-loaded PFCI inverter powering a constant-voltage-magnitude load, such as indeed represented by capacitors ESCa/ESCb, is that the instantaneous magnitude of the current provided to this load will be roughly proportional to the instantaneous magnitude of the DC voltage powering the inverter. Thus, absent control, the magnitude of the charging current supplied to capacitors ESCa/ESCb will vary in proportion with the magnitude of the DC voltage present between the DC– bus and the DC+ bus.

However, another basic characteristic of the PFCI inverter is that, with a constant-magnitude DC voltage present between the DC– bus and the DC+ bus, the magnitude of the charging current supplied to capacitors ESCa/ESCb is roughly proportional to the conductance of photo-sensitive resistive means PSRM; which, in turn, is roughly proportional to the amount of light emitted from light-emitting means LEM; which, in turn, is roughly proportional to the magnitude of the current flowing through light-emitting means LEM; which, in turn, is roughly proportional to the magnitude of the DC voltage present between the DC– bus and the DC+ bus.

Thus, with the amount of light emitted by light-emitting means LEM being roughly proportional to the instantaneous magnitude of the DC voltage present between the DC– bus and the DC+ bus, the instantaneous magnitude of the charging current supplied to capacitor ESCa/ESCb will be roughly proportional to the square of the instantaneous magnitude of that DC voltage.

As an overall consequence, the instantaneous magnitude of the current drawn by the PFCI inverter will be roughly proportional to the instantaneous magnitude of the DC voltage present betwen the DC– bus and the DC+ bus; which is to say that the instantaneous magnitude of the current drawn from the power line (i.e., source S) will be proportional to the instantaneous magnitude of the voltage provided from the power line; which, in turn, provides for the power drawn from the power line by the circuit arrangement of FIG. 4 to exhibit a power factor close to unity (i.e., 100%). Moreover, the waveform of the current drawn from the power line will exhibit an exceptionally low degree of harmonic distortion.

In effecting its control action, inverter control means ICM causes both symmetry-modulation and frequency-modulation of the inverter output voltage-provided between junction Jdc and junction Jx. As illustrated by FIG. 5(*g*), this inverter output voltage is a nearly symmetrical squarewave of a certain frequency (about 30 kHz) whenever the instantaneous magnitude of the DC voltage between the DC– bus and the DC+ bus is at its maximum (about 170 Volt). However, with the instantaneous magnitude of this DC voltage being about half its maximum, the inverter output voltage—as illustrated by FIG. 5(*h*)—is a clearly assymmetrical squarewave. Moreover, its frequency is now higher (about 32 kHz). As indicated by FIG. 5(*i*), at a still lower magnitude of the DC voltage between the DC– bus and the DC+ bus, the inverter output voltage is still more assymmetrical and of still higher frequency.

Thus, the PFCI inverter effects its control action by a combination of symmetry-modulation and frequency-modulation.

Increasing frequency—other things being equal—provides for reduced-magnitude charging current to energy-storing capacitors ESCa/ESCb. This is so for the reason that the natural series-resonance-frequency of the L2-C2 tuned circuit is below the inverter's actual frequency

Increasing asymmetry—other things being equal—also provides for reduced-magnitude charging current to energy-storing capacitors ESCa/ESCb. This is so for the reason that the magnitude of the fundamental frequency component of the inverter's output voltage decreases roughly in proportion to the degree of asymmetry.

As the degree of conductance of photo-sentitive resistive means PSRM increases, the magnitude of the negative bias voltage present across bias capacitor BC1b decreases; which results in a longer ON-time for transistor Q1b.

As for the BI inverter, the operation is substantially as described in connection with the circuit of FIG. 2, except for the particular feature associated with auxiliary winding AW.

The phasing of the AW winding is such that the fundamental frequency component of the high-frequency voltage provided at output terminal OT1 is substantially out-of-phase with the high-frequency voltage provided at output terminal OT2. That way, the magnitude of the net voltage provided across the three series-connected lamps is larger than it would be if the lamps had been connected directly across tank capacitor C2.

In case of ordinary F40 or F34 T-12 Rapid-Start fluorescent lamps, each lamp requires an operating voltage of RMS magnitude equal to about 100 Volt; which implies a total RMS magnitude of about 300 Volt across the three series-connected lamps.

To reduce potentially dangerous electric shock effects, as might result from capacitive coupling directly from the glass envelope of the fluorescent tube to the hand of a person installing and/or removing the fluorescent lamps from their sockets, it is important that the magnitude of the lamp's arc voltage be not much higher than about 200 Volt RMS (at about 30 kHz) with respect to ground. If it were to exceed this 200 Volt RMS magnitude, a person might receive a potentially dangerous electric shock effect merely from grabbing the fluorescent tube at a point where the potential of the ionized gas inside the glass envelope were higher than 200 Volt RMS with respect to ground.

In the particular arrangement of FIG. 4, the maximum magnitude of the 30 kHz potential of the ionized gas within the fluorescent glass tube is indeed maintained below approximately 200 Volt RMS with respect to ground, in spite of the fact that the three lamps require an operating voltage of 300 Volt RMS magnitude. This result has been achieved by way of the AW winding, whose output voltage—during lamp operation—is about 100 Volt RMS with respect to the B– bus; which, with respect to the 30 kHz inverter voltage, is indeed at ground potential.

On the other hand, the magnitude of the voltage contributed by the AW winding should be as low as reasonably possible for the reason that its waveform is of such nature as to cause degradation of the lamp current crest factor; which degradation is due to the fact that—contrary to the case with the voltage across the tank capacitor (C2)—the waveshape of the voltage present across the tank inductor (L2) includes the full magnitude of the squarewave voltage provided at the inverter's output (i.e., the voltage provided between junctions Je and Jy).

With the DC voltage provided between the B– bus and the B+ bus being of substantially constant magnitude, the lamp current crest factor would be about 1.5 if the lamps were to have been connected directly across the tank capacitor. However, had the lamps instead been connected directly across the tank inductor, the lamp current crest factor would have been far in excess of 1.7; which is normally considered the maximum permissible level for lamp current crest factor. On the other hand, with only one third of the lamp voltage derived from the tank inductor voltage, and with the remaining two thirds being derived from the tank capacitor voltage, the net resulting lamp current crest factor is kept just below 1.7—as is indeed the case in the arrangement of FIG. 4.

Additional Comments

(a) The reason that the AW winding is loosely coupled with tank inductor L2 is related to minimizing detrimental effects on lamp current crest factor due to powering the lamps in part by the tank inductor voltage. By effectively providing the tank inductor voltage to the fluorescent lamp via a series inductance, the detrimental effects on lamp current crest factor are indeed reduced. However, the same result can be obtained by placing an inductor in series with the fluorescent lamps.

(b) Light-emitting means LEM is likely not to be totally linear in terms of light output as function of the magnitude of the driving DC current. Likewise, photo-sensitive resistive means is likely not to be totally linear with respect to its effective conductance versus amount of light received. In addition, the symmetry-modulation and frequency-modulation resulting from changes in the conductance of the photo-sensitive resistive means are not likely to be totally linear.

However, it is not necessary that these various relationships be totally linear. Instead, the largest part of the sought-after effect—namely power factor correction and reduction of power line harmonics—will result even if the various relationships be quite non-linear.

Of course, by carefully selecting and matching the non-linearities of the different effects, as well as by introducing various linearizing means, nearly any desired degree of final power factor correction can be achieved.

(c) The time constant associated with bias capacitor BC1b and its associated charge leakage means—namely resistor R1b2 and photo-sensitive resitive means PSRM—must be long with respect to a complete cycle of the 30 kHz inverter frequency. However, this time constant should be short compared with a complete cycle of the 120 Hz ripple voltage on the DC voltage present between the DC– bus and the DC+ bus.

(d) The waveshape of the high-frequency current flowing from tank capacitor C1 of FIG. 4 is substantially sinusoidal— with the positive halves flowing though rectifier HSRa and the negative halves flowing through rectifier HSRb.

(e) The absolute instantaneous magnitude of the high-frequency current flowing from junction J1 and through rectifiers HSRa and HSRb is—except for any imperfections in the HSRa/HSRb rectifiers—equal to that of the net DC current flowing into energy-storing capacitors ESCa and ESCb.

(f) The waveshape of the current flowing through the three series-connected fluorescent lamps is also nearly sinusoidal; except that a modest degree of distortion is introduced by the harmonics of the component of high-frequency voltage provided by auxiliary winding AW. However, as long as the magnitude of this component is kept relatively small compared with the magnitude of the voltage provided from the tank capacitor— say, no higher than about 50% thereof—the distortion of the lamp current is insufficient to cause significant deterioration of the lamp current crest factor.

(g) Each of transistors Q1a and Q1b (as combined with their respective commutating rectifiers CR1a and CR1b) acts as a rapidly operating ON-OFF switch—current flowing through one or the other transistor, but never through both transistors at the same time. The forward conduction time of transistor Q1b decreases as the absolute magnitude of the negative bias voltage on bias capacitor BC1b increases. Thus, with the instantaneous magnitude of the DC voltage present between the DC– bus and the DC+ bus being substantially equal to that of the AC power line voltage provided from source S, the absolute magnitude of this bias voltage varies synchronously with that of the AC power line voltage; thereby, in turn, causing the effective ON-time of transistor Q1b to vary synchronously with the absolute magnitude of the AC power line voltage as well.

(h) It is important that the natural resonance frequencies of the L1-C1 and the L2-C2 series-tuned circuits of FIG. 4 are lower than (or at least not higher than) the operating (or switching) frequencies of the PFCI and the BI inverters, respectively.

(i) It is believed that the present invention and its several attendant advantages and features will be understood from the preceeding description. However, without departing from the spirit of the invention, many changes may be made in its form and in the selection, construction and interrelationships of is constituent parts, the form herein presented merely representing the presently preferred embodiment.

I claim:

1. An arrangement comprising:

first rectifier means connected with a source of power line voltage and operative to provide a first DC voltage at a first set of DC terminals;

power conditioner means connected with the first DC terminals and operative to provide a high-frequency output voltage at a high-frequency output; the high-frequency output voltage having a fundamental frequency substantially higher than that of the power line voltage; the power conditioner means including tuned L-C circuit means having a tank inductor and a tank capacitor; a capacitor voltage existing across the tank capacitor; the tank inductor having an auxiliary winding; an auxiliary voltage existing across the auxiliary winding; the high-frequency output voltage being the sum of the capacitor voltage and the auxiliary voltage; the magnitude of the high-frequency output voltage thereby being substantively higher than that of the capacitor voltage by itself; and

gas discharge lamp means connected with the high-frequency output and operative to be powered by the high-frequency output voltage provided thereat.

2. An arrangement comprising:

rectifier means connected with a source of power line voltage and operative to provide a DC voltage at a set of DC terminals;

power conditioner means connected with the DC terminals and operative to provide a high-frequency output voltage at a high-frequency output; the high-frequency output voltage having a fundamental frequency substantially higher than that of the power line voltage; the power conditioner means including a tuned L-C circuit having a tank inductor and a tank capacitor; a capacitor voltage existing across the tank capacitor; the tank inductor having an auxiliary winding; the auxiliary winding being characterized by not being connected across the terminals of a thermionic cathode; an auxiliary voltage existing across the auxiliary winding; the high-frequency output voltage being the sum of the capacitor voltage and the auxiliary voltage; and

gas discharge lamp operative to connect with the high-frequency output and to be powered by the high-frequency output voltage provided thereat.

3. The arrangement of claim 2 wherein the RMS magnitude of the high-frequency output voltage is substantially higher than the RMS magnitude of the capacitor voltage.

4. The arrangement of claim 3 where the RMS magnitude of the high-frequency output voltage is at least 25 percent higher than the RMS magnitude of the capacitor voltage.

5. The arrangement of claim 2 wherein the tuned L-C circuit has a natural resonance frequency equal to or lower than the fundamental frequency of the high-frequency output voltage.

6. An arrangement comprising:

first rectifier means connected with a source of power line voltage and operative to provide a first DC voltage at a first set of DC terminals;

power conditioner means connected with the first DC terminals and operative to provide a high-frequency output voltage at a high-frequency output; the high-frequency output voltage having a fundamental frequency substantially higher than that of the power line voltage; the power conditioner means including inverter means connected with the first set of DC terminals and operative to provide the high-frequency output voltage; the inverter means including tuned L-C circuit means having a tank inductor series-connected with a tank capacitor; a capacitor voltage existing across the tank capacitor; the tank inductor having an auxiliary winding; an auxiliary voltage existing across the auxiliary winding; the high-frequency output voltage being the sum of the capacitor voltage and the auxiliary voltage, thereby to attain an RMS magnitude substantially higher than that of the capacitor voltage by itself; and

gas discharge lamp operative to connect with the high-frequency output and to be powered by the high-frequency output voltage provided thereat.

7. An arrangement comprising:

a first sub-assembly having power input terminals and DC output terminals; the power input terminals being connected with a pair of power line terminals across which exists an AC power line voltage; the DC output terminals being connected with the power input terminals by way of a rectifier and a waveshaping circuit; the waveshaping circuit being operative to cause a substantially sinusoidal current to be drawn by the power input terminals from the power line terminals; the substantially sinusoidal current having a waveform characterized by exhibiting less than 10% total harmonic distortion;

a second sub-assembly connected with the DC terminals and operative to provide a high-frequency voltage at a pair of inverter terminals; the high-frequency voltage

having a fundamental frequency substantially higher than that of the AC power line voltage; and

a third sub-assembly connected with the inverter terminals and operative to supply a high-frequency output voltage at a pair of output terminals; at least three gas discharge lamps being series-connected across the output terminals; the third sub-assembly being operative to cause the three series-connected gas discharge lamps to be ignited and subsequently to be supplied with high-frequency lamp current.

8. The arrangement of claim 7 wherein (i) each gas discharge lamp includes a thermionic cathode with a pair of cathode terminals, and (ii) the third sub-assembly is further characterized by having terminals connected with each of the cathode terminals.

9. The arrangement of claim 7 wherein the second sub-assembly is further characterized by including an energy-storing inductor connected in circuit between the power input terminals and the DC output terminals.

10. The arrangement of claim 7 wherein the rectifier has a first pair and a second pair of rectifier terminals; the first pair of rectifier terminals being connected with the power input terminals; the second pair of rectifier terminals being connected with the waveshaping circuit.

11. The arrangement of claim 10 wherein the first sub-assembly is additionally characterized in that (i) a unidirectional voltage exists between the second pair of rectifier terminals, and (ii) the absolute instantaneous magnitude of this unidirectional voltage is substantially equal to that of the AC power line voltage.

12. The arrangement of claim 7 wherein the first subassembly is further characterized by including an intermittently conducting transistor connected in circuit between the power input terminals and the DC output terminals.

13. The arrangement of claim 12 wherein an energy-storing inductor is connected in circuit with the intermittently conducting transistor.

14. The arrangement of claim 7 wherein the first subassembly is further charactized by including a pair of terminals across which, at least under some conditions, exists a non-symmetrical squarewave voltage.

* * * * *

EXHIBIT F

US005550439A

United States Patent [19]

Nilssen

[11] **Patent Number: 5,550,439**

[45] **Date of Patent: Aug. 27, 1996**

[54] **ELECTRONIC BALLAST HAVING PULSATING OUTPUT VOLTAGE**

[76] Inventor: **Ole K. Nilssen**, Caesar Dr., Barrington, Ill. 60010

[21] Appl. No.: **772,547**

[22] Filed: **Oct. 4, 1991**

Related U.S. Application Data

[63] Continuation of Ser. No. 346,321, May 1, 1989, abandoned, which is a continuation of Ser. No. 686,275, Dec. 26, 1984, abandoned, which is a continuation-in-part of Ser. No. 677,562, Dec. 3, 1984, Pat. No. 4,698,553, and a continuation-in-part of Ser. No. 456,276, Feb. 22, 1983, Pat. No. 4,503,363.

[51] **Int. Cl.**[6] **H05B 37/00**

[52] **U.S. Cl.** **315/244**; 315/127; 315/225; 315/DIG. 7

[58] **Field of Search** .... 315/244, 242, 315/127, 125, 119, 225, DIG. 7

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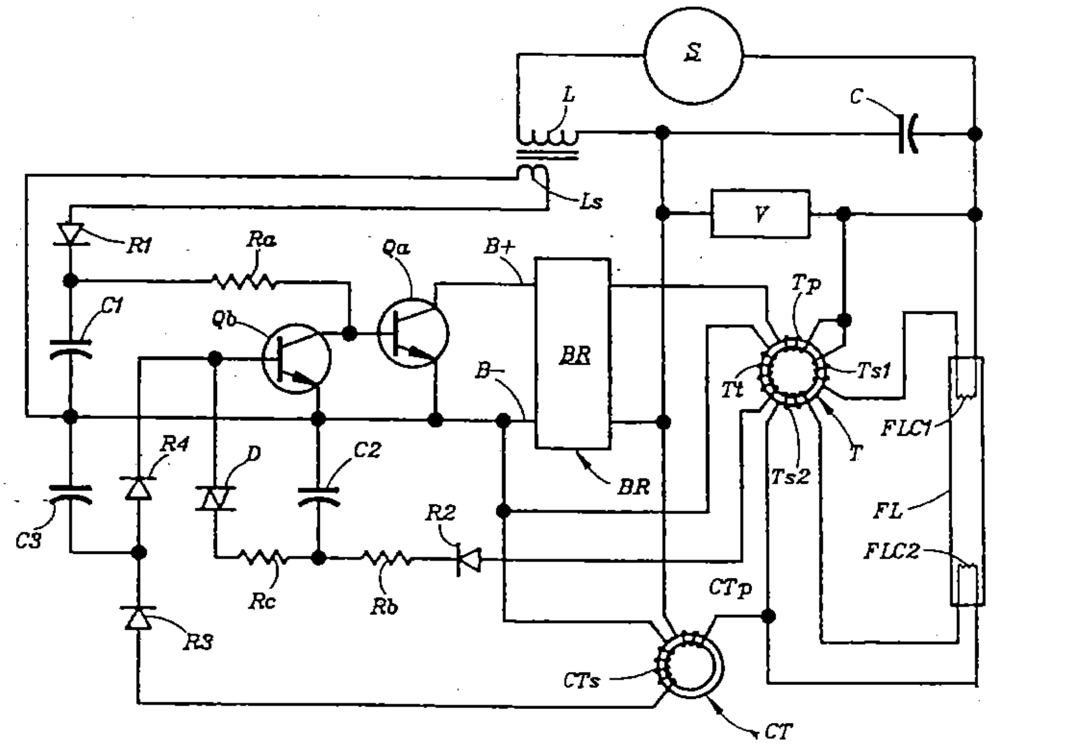
Primary Examiner—Robert J. Pascal
Assistant Examiner—Reginald A. Ratliff

[57] **ABSTRACT**

In a fluorescent lamp ballast, a source of high-frequency voltage is applied directly across a series-resonant L-C circuit. The fluorescent lamp is connected in parallel with the capacitor of the L-C circuit and a voltage-limiting means is operative to prevent the series-resonant L-C circuit from overloading the voltage source during any period when the lamp is not effective in providing circuit loading. When power is initially applied to the series-resonant L-C circuit, a control means provides a short circuit across the capacitor, and, by way of a first current transformer, the resulting short circuit current is used for pre-heating the fluorescent lamp cathodes. After about 1.5 second, the control means provides for removal of the short circuit for a period of about 25 milli-seconds, thereby permitting the voltage across the capacitor to grow to a magnitude sufficient to ignite and operate the lamp. If the lamp ignites, the resulting lamp current is then used, by way of a second current transformer, to prevent the control means from re-providing the short circuit. If lamp current does not flow, or if it at any time ceases to flow, the control means will re-provide the short circuit within about 25 milli-seconds. Thereafter, until power is removed or until an operable lamp is connected, the control means will continuously repeat the cycle of 1.5 second short circuit and 25 milli-seconds open circuit.

27 Claims, 1 Drawing Sheet



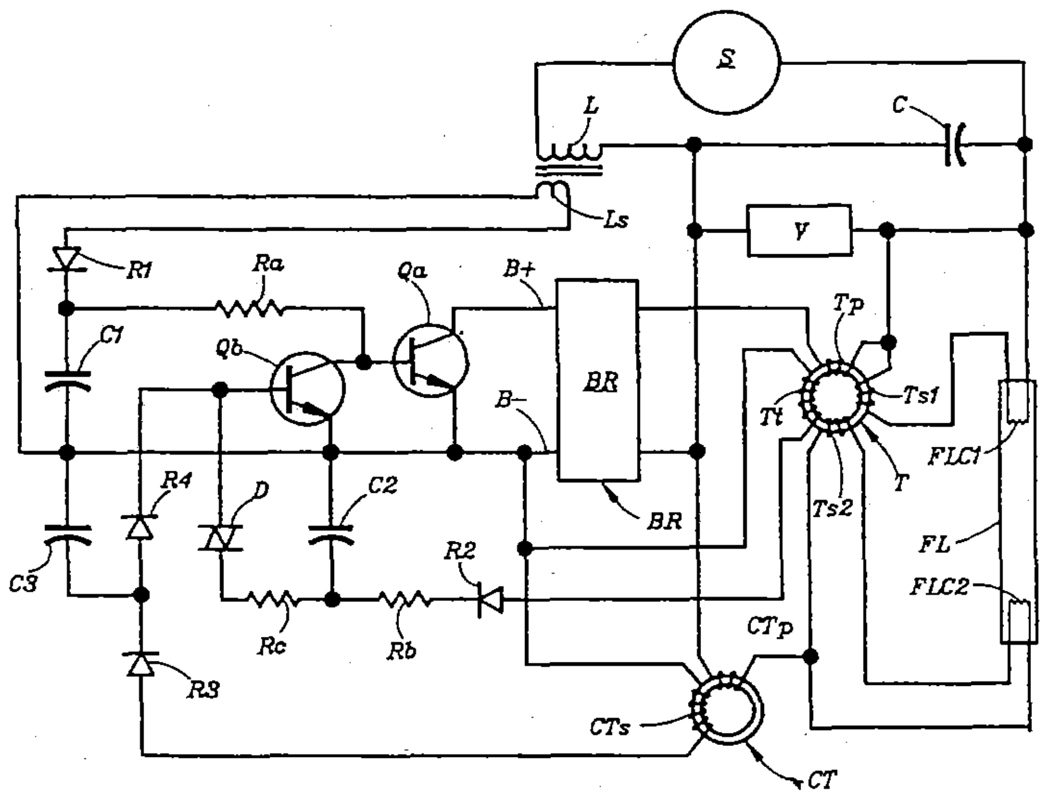

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Fig. 1

ELECTRONIC BALLAST HAVING PULSATING OUTPUT VOLTAGE

Instant application is a continuation of Ser. No. 346,321 filed May 1, 1989 now abandoned; which is a continuation of Ser. No. 06/686,275 filed Dec. 26, 1984, abandoned; which was a continuation-in-part of Ser. No. 677,562 filed Dec. 3, 1984, U.S. Pat. No. 4,698,553, as well as of Ser. No. 456,276 filed Feb. 22, 1983, U.S. Pat. No. 4,503,363.

FIELD OF THE INVENTION

The present invention relates to high-frequency series-resonant ballasts for fluorescent lamps.

PRIOR ART AND BACKGROUND CONSIDERATIONS

High-frequency series-resonant fluorescent lamp ballasts have been previously described, such as in U.S. Pat. No. 3,710,177 to Ward and U.S. Pat. No. 4,370,600 to Zansky. However, these previously described ballasts do not provide solutions to several basic problems associated with practical applications of such ballasts. These problems relate to the excessive power drain by and the self-destructive nature of the series-resonant ballast under the condition of being connected to an inoperative lamp.

In powering a fluorescent lamp by way of a high-frequency series-resonant ballast, where the ballast constitutes a high-Q resonant L-C circuit series-excited from an AC voltage source and parallel-loaded by the fluorescent lamp, there is a serious problem associated with the situation where the fluorescent lamp for one reason or another ceases to constitute an effective load for this high-Q series-excited L-C circuit. In such a situation, which is most apt to occur toward the end of normal lamp life, the power drawn by the high-Q resonant L-C circuit from its AC voltage source is so excessively high as to cause damage to or even destruction of the L-C circuit and/or the AC voltage source. If, contrary to expectations, destruction of the circuit and/or the source does not occur, the amount of power drawn from the source will be so large as to represent an unacceptable level of energy waste—recognizing that it may often take a long time before a worn-out fluorescent lamp is replaced.

SUMMARY OF THE INVENTION

Objects of the Invention

An object of instant invention is that of providing safe and efficient high-frequency series-resonant ballasts for fluorescent lamps. Other objects and advantages of the present invention will become apparent from the following description.

BRIEF DESCRIPTION

An L-C circuit with an unloaded Q-factor of about 50 is series-connected directly across the output of a 100 Volt/30 kHz voltage source. This L-C circuit is resonant at 30 kHz, which means that it is series-resonant at the very frequency of the high-frequency voltage source.

A regular 40 Watt fluorescent lamp and a voltage-limiting means are both connected in parallel with the tank-capacitor of the L-C circuit, the voltage-limiting means being operative to limit the voltage developed across the tank-capacitor to a magnitude suitable for proper lamp starting, which magnitude is about 300 Volt.

Without the voltage-limiting means, with an unloaded Q-factor of 50 and linear circuit operation, the magnitude of the voltage developing across the tank-capacitor would have been 5000 Volt.

Under normal operation, the fluorescent lamp limits the magnitude of the voltage developing across the tank-capacitor to about 100 Volt; and at that point the loaded L-C circuit draws approximately 40 Watt of power from the source. Thus, with a loaded Q-factor of about unity and an unloaded Q-factor of 50, the implication is that the losses in the L-C circuit amount to about 2% of the total power drawn from the source under normal operating conditions.

If for some reason the fluorescent lamp should fail to constitute an effective load for the L-C circuit, the magnitude of the voltage across the tank-capacitor would increase to about 300 Volt, which implies that the power drawn from the source at that point would be about 120 Watt, with substantially all of it being dissipated in the voltage-limiting means.

If there were no voltage-limiting means present, however, the power drawn by the L-C circuit from the source-assuming no breakdown—would be about 2000 Watt, with all of it being dissipated within the L-C circuit itself.

The present invention provides for means to prevent the L-C circuit from operating in its resonant mode—and thereby to prevent it from drawing excessive power—in case the fluorescent lamp should fail even for a brief period to constitute a proper load for the L-C circuit. This effect is accomplished by a transistor operative, by way of a rectifier bridge, to provide a short circuit across the tank-capacitor whenever lamp current fails to flow for about 25 milli-seconds. With a short-circuited tank-capacitor, the amount of power drawn by the L-C circuit is negligibly small.

More particularly, a control means is connected with the L-C circuit and is operative to provide for the following functions.

a) Upon initially providing power to the L-C circuit, the control means provides for a short circuit across the tank-capacitor for an initial period of about 1.5 second; which is the length of time normally required for the cathodes of the fluorescent lamp to become fully thermionic.

b) After this initial period, the control means removes the short circuit for a period of about 25 milli-seconds; which period is long enough to provide for proper lamp starting under normal circumstances.

c) If lamp current starts to flow within this initial period, the control means operates to keep the short circuit removed for as long as lamp current flows.

d) If lamp current fails to flow within this initial period, the control means re-imposes the short circuit and keeps it so imposed for a period of about 1.5 second; whereafter it again removes the short circuit for a period of about 25 milli-seconds.

e) It the fluorescent lamp is removed from the L-C circuit, or if for other reasons it fails to continue to operate, the control means operates to provide a short circuit across the tank-capacitor within a period of about 25 milli-seconds.

f) As long as there is no lamp loading the L-C circuit, the control means tries every 1.5 second or so to start the lamp by removing the short circuit for a period of about 25 milliseconds. Thus, with a duty-cycle of about 25 milli-seconds out of 1500 milli-seconds (1.67%), the average power dissipation of the unloaded L-C circuit will be only about 2.0 Watt.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 provides a schematic circuit diagram of the preferred embodiment of the invention.

DESCRIPTION OF THE PREFERRED EMBODIMENT

DETAILS OF CONSTRUCTION

FIG. 1 shows an AC voltage source S, which in reality is a power-line-operated frequency converter providing an output voltage of 100 Volt RMS magnitude and 30 kHz frequency.

Connected directly across S is a series-combination of an inductor L and a capacitor C. Inductor L has a tightly coupled secondary winding Ls.

A Varistor V is connected directly across capacitor C.

A fluorescent lamp FL, having cathodes FLC1 and FLC2, is connected in series with the primary winding CTp of control transformer CT, and this series-combination of FL and CTp is connected across capacitor C.

A bridge rectifier BR, having a B+ output bus terminal and a B– output bus terminal, is connected in series with the primary winding Tp of transformer T, and this series-combination of BR and Tp is connected across capacitor C.

A transistor Qa is connected with its collector to the B+ bus and with its emitter to the B– bus.

A series-combination of a capacitor C1 and a rectifier R1, with R1 being connected with the capacitor by way of its cathode, is connected directly across the output of secondary winding Ls of inductor L. The cathode of rectifier R1 is connected to the base of transistor Qa by way of a resistor Ra.

A transistor Qb is connected with its collector and emitter to the base and emitter, respectively, of transistor Qa.

Transformer T has a first secondary winding Ts1 connected with cathode FLC1 of fluorescent lamp FL, and a second secondary winding Ts2 connected with fluorescent lamp cathode FLC2.

Transformer T also has a tertiary winding Tt, which tertiary winding is connected between the B– bus and the anode of a rectifier R2.

A series-combination of a resistor Rb and a capacitor C2 is connected between the cathode of rectifier R2 and the B– bus, with one terminal of resistor Rb being connected with the cathode of R2.

A series-combination of a resistor Rc and a Diac D is connected between the base of transistor Qb and the junction between resistor Rb and capacitor C2.

Control transformer CT has a secondary winding CTs connected between the B– bus and the anode of a rectifier R3. A rectifier R4 is connected between the cathode of rectifier R3 and the base of transistor Qb, the anode of rectifier R4 being connected with the cathode of rectifier R3.

A capacitor C3 is connected between the cathode of rectifier R3 and the B– bus.

DETAILS OF OPERATION

In FIG. 1, when the 100 Volt/30 kHz voltage from source S is initially applied to the L-C series-circuit, a voltage is developed across secondary winding Ls. This voltage is rectified and filtered by rectifier R1 and capacitor C1, and then applied to the base of transistor Qa by way of a current-limiting resistor Ra. The value of Ra is so chosen that the resulting DC current provided to the base of transistor Qa is adequate to make this transistor conduct in a substantially saturated mode and thereby to constitute an effective short circuit between the B+ bus and the B– bus.

With transistor Qa so conducting, and since transformer T is a current transformer with a very low-impedance primary winding, and since a short circuit at the output of the bridge rectifier is essentially equivalent to a short circuit at its input, there is in effect a short circuit provided across tank-capacitor C. Thus, as long as transistor Qa is an effective short circuit, the magnitude of the current drawn from the source is limited by the reactance of L, thereby in effect representing a non-dissipative load.

As long as transistor Qa conducts, current is forced through the primary winding Tp of transformer T. By transformer action, this current is provided to the effective parallel connection of the two fluorescent lamp cathodes, thereby providing to these cathodes the modest amount of power required to bring about thermionic emission. The tertiary winding Tt provides a voltage output that is used for charging capacitor C2 with a current that is limited by resistor Rb. Eventually, the magnitude of the voltage on C2 gets to be high enough to cause Diac D to break down, at which point the charge that had accumulated on capacitor C2 gets discharged into the base of transistor Qb—with the magnitude of the discharge current being principally determined by the resistance of Rc. This magnitude is so chosen that—as soon as the Diac breaks down—transistor Qb becomes conductive to the point of shunting away the base current provided to transistor Qa by way of resistor Ra.

In other words, as soon as the voltage on capacitor C2 has increased to some pre-determined magnitude, the Diac breaks down and immediately renders transistor Qa non-conductive. The time it takes for the voltage on capacitor C2 to reach this predetermined magnitude is a function of the time-constant associated with C2 and Rb as well as of the magnitude of the voltage being provided by the tertiary winding Tt. In the circuit of FIG. 1, this time was chosen to be about 1.5 second; which is the length of time normally required by fluorescent lamp cathodes to reach the point of thermionic emission.

For as long as capacitor C2 is providing base current for transistor Qb, this transistor is operative to prevent current from being applied to the base of transistor Qa, thereby making Qa non-conductive. The length of time during which Qa is thereby kept non-conductive is determined by the parameters of capacitor C2 and resistor Rc. In the circuit of FIG. 1, these parameters were so chosen as to make this length of time about 25 milli-seconds.

Thus, after the initial period of about 1.5 second, during which Qa represented a short circuit and the fluorescent lamp cathodes were provided with heating power, Qa is switched off and becomes an open circuit for about 25 milli-seconds. During this 25 milli-second period, the voltage across capacitor C increases in magnitude to the point where the lamp starts. With already pre-heated cathodes, the time required for the lamp to start is normally less than 25 milli-seconds.

As soon as lamp current starts to flow, control transformer CT, by way of rectifier R3 and filter capacitor C3, provides a DC current to the base of transistor Qb, thereby causing transistor Qb to continue to shunt away the base current for Qa. Thus, as long as the lamp starts to draw current within the 25 milli-second period, transistor Qa will continue to be non-conductive, and the lamp will continue to operate.

However, if lamp current does not start to flow within the 25 milli-second period, base current for Qb will cease, which means that Qb will cease shunting away the base current for Qa. Thus, after about 25 milli-seconds, if lamp current fails to flow, transistor Qa will again become conductive and operative to provide a short circuit across capacitor C.

Now, with Qa conductive, current will again flow through the primary winding Tp of transformer P, and again will charge C2 to the point of breaking down the Diac; which then again starts a 25 milli-second period of shunting away the base current for transistor Qa.

In other words, with the fluorescent lamp inoperative or disconnected, the circuit of FIG. 1 operates in a cyclical fashion, with each cycle consisting of a 1.5 second period during which transistor Qa is conductive—which implies that capacitor C is shorted—and a 25 milli-second period during which transistor Qa is non-conductive. With the lamp operating, on the other hand, this cyclical circuit operation is prevented by the flow of lamp current.

The lamp cathodes are supplied with heating power only as long as Qa conducts. After the lamp has ignited, however, heating power is no longer needed and is no longer supplied, thereby providing for noticably improved lamp efficacy.

It is believed that the present invention and its several attendant advantages and features will be understood from the preceeding description. However, without departing from the spirit of the invention, changes may be made in its form and in the construction and interrelationships of its component parts, the form herein presented merely representing the presently preferred embodiment.

I claim:

1. A ballast for a gas discharge lamp; the ballast being adapted to operate from a source of high-frequency AC voltage; the high-frequency AC voltage being provided by a frequency converter powered by the power line voltage provided by an ordinary electric utility power line; the ballast comprising:

an L-C circuit series-connected across the source; the L-C circuit being series-resonant at or near the frequency of the high-frequency AC voltage and having at least one inductive and one capacitive reactance means;

connect means operable to connect the lamp in parallel-circuit with one of said reactance means, thereby to provide for proper lamp starting and operating voltage; and

control means connected in circuit with the reactance means and operative, but only when there is no current flowing through the lamp, to cause a repetitive alternation between two states: (i) a first state characterized by the provision of an effective short circuit across one of said reactance means, and (ii) a second state characterized by not providing such a short circuit; the repetitive alternation occurring at a repetition frequency substantially lower than the fundamental frequency of the high-frequency AC voltage; the magnitude of any voltage present across one of said reactance means being substantially negligible throughout the duration of the first state.

2. The ballast of claim 1 wherein the first state is further characterized by being substantially longer in duration than the second state.

3. The ballast of claim 1 wherein: (i) the frequency of the high-frequency AC voltage is substantially higher than that of the power line voltage; and (ii) the repetition frequency is substantially lower than the frequency of the power line voltage.

4. The ballast of claim 1 wherein a voltage-limiting non-linear resistance means is connected in circuit across one of said reactance means.

5. The ballast of claim 1 wherein the control means is operative as specified even when the lamp is not connected in parallel-circuit with one of said reactance means.

6. A ballast for a gas discharge lamp; the ballast being adapted to operate from a source of AC voltage and comprising:

an L-C circuit series-connected across the source; the L-C circuit being series-resonant at or near the frequency of the AC voltage and having at least one inductive and one capacitive reactance means;

connect means operable via a set of connect terminals to connect the lamp in parallel-circuit with one of the reactance means, thereby to provide across the connect terminals an output voltage suitable for proper lamp starting and operation; and

control means connected in circuit with the lamp and responsive to current flowing therethrough; the control means being operative, but only when there is no current flowing through the lamp, to impose a short circuit intermittently and periodically across the connect terminals; the short circuit being so imposed at a repetition frequency substantially lower than that of the AC voltage; the repetition frequency therefore having a repetition period substantially longer than the period of the AC voltage;

whereby, as long as no current flows through the lamp, the output voltage is modulated in magnitude at said repetition frequency; the output voltage being of a relatively low magnitude for a relatively long period of time and being of a relatively high magnitude for a relatively short period of time; the relatively long period of time being at least twice as long as the relatively short period of time.

7. The ballast of claim 6 wherein the control means is functional as specified even when the lamp is not connected with the connect terminals.

8. The ballast of claim 6 wherein: (i) the frequency of the AC voltage is substantially higher than the frequency of the power line voltage normally present on an ordinary electric utility power line; and (ii) the repetition frequency being substantially lower than the frequency of the power line voltage.

9. A ballast for a gas discharge lamp; the ballast being adapted to operate from a source of AC voltage; the fundamental frequency of the AC voltage being substantially higher than that of the power line voltage normally present on an ordinary electric utility power line; the ballast comprising:

an L-C circuit adapted to series-connect across said source; the L-C circuit being resonant at or near the frequency of the AC voltage and having at least one inductive and one capacitive reactance means;

connect means operative to permit connection of said lamp in parallel-circuit with one of the reactance means; and

shorting means operative to provide an intermittently interrupted short circuit across one of the reactance means; the shorting means causing any voltage existing across said one of the reactance means to be of negligible magnitude whenever a short circuit is indeed being provided; the shorting means being operative as described even when the lamp is not connected in parallel-circuit with one of the reactance means.

10. The ballast of claim 9 including voltage-limiting non-linear resistance means connected in parallel-circuit with one of the reactance means.

11. An arrangement comprising:

a gas discharge lamp means having a first and a second thermionic cathode; and an AC source operative to provide an AC voltage between a first and a second AC output terminal; the frequency of the AC voltage being about 20 kHz or higher; the first and the second AC output terminal being connected with the first and the second thermionic cathode, respectively; the AC source being characterized by including: (i) internal impedance operative to limit the magnitude of any current flowing from its AC output terminals; and (ii) control means operative, except when the gas discharge lamp draws more than a certain minimum amount of current from the AC output terminals, to cause the magnitude of the AC voltage to vary periodically at a low frequency between a relatively high level and a relatively low level; the low frequency being substantially lower than the fundamental frequency of the power line voltage;

whereby, if the gas discharge lamp were to be disconnected, or if it otherwise were to fail to draw at least said certain minimum amount of current from the AC output terminals, the RMS magnitude of the AC voltage would vary periodically between the relatively high level and the relatively low level.

12. The arrangement of claim 11 wherein: (i) at least one of the thermionic cathodes has a pair of cathode terminals; and (ii) the AC source has a transformer with at least one transformer winding connected across the cathode terminals.

13. An arrangement comprising:

an ordinary electric utility power line operative to provide an AC power line voltage at a pair of power line terminals;

circuitry connected with the power line terminals and operative to provide a high-frequency output voltage between a first pair of output terminals and a second pair of output terminals; the circuitry including control means functional to control the magnitude of the high-frequency output voltage; the fundamental frequency of the high-frequency output voltage being substantially higher than that of the AC power line voltage; and

gas discharge lamp means having a first pair of lamp terminals and a second pair of lamp terminals; the first pair of lamp terminals being operable to connect with the first pair of output terminals; the second pair of lamp terminals being operable to connect with the second pair of output terminals;

the arrangement being characterized by functioning such that:

(a) whenever the lamp terminals are connected with the output terminals and a lamp current is flowing through the gas discharge lamp, the magnitude of the output voltage remains substantially constant; but

(b) in case the lamp current were to cease to flow through the gas discharge lamp, or if the lamp terminals were to be disconnected from the output terminals, the magnitude of the high-frequency output voltage would vary cyclically at a relatively low frequency between a relatively low level and a relatively high level; the relatively low frequency being substantially lower than the frequency of the AC power line voltage; the relatively low frequency being characterized by having a cycle period.

14. The arrangement of claim 13 wherein, whenever the magnitude of the output voltage be varying cyclically, the magnitude of the output voltage will be at the relatively low level for a relatively large fraction of the cycle period and at the relatively high level for a relatively small fraction of the cycle period.

15. The arrangement of claim 14 wherein the cycle period has a duration between approximately 0.4 and approximately 4.0 seconds.

16. The arrangement of claim 14 wherein the relatively large fraction corresponds to a time duration between approximately 0.5 and approximately 2.5 seconds.

17. The arrangement of claim 16 wherein the relatively small fraction corresponds to a time duration between 10 and 250 milli-seconds.

18. A ballast for a gas discharge lamp; the gas discharge lamp having a first pair of lamp terminals connected with a first thermionic cathode and a second pair of lamp terminals connected with second thermionic cathode; the ballast comprising:

frequency-conversion circuit connected with the power line terminals and operative to provide a high-frequency output voltage between a first pair of output terminals and a second pair of output terminals; the frequency-conversion circuit including control circuitry functional to control the magnitude of the high-frequency output voltage; the fundamental frequency of the high-frequency output voltage being substantially higher than that of the AC power line voltage; and

means to permit connection of the first pair of output terminals with the first pair of lamp terminals and the second pair of output terminals with the second pair of lamp terminals; thereby to cause the high-frequency output voltage to be applied between the first and the second thermionic cathode whenever the lamp terminals be connected with the output terminals;

the ballast being characterized by functioning such that, prior to connecting the output terminals with the lamp terminals, the magnitude of the high-frequency output voltage varies cyclically at a relatively low frequency between a relatively low level and a relatively high level; the relatively low frequency being substantially lower than the frequency of the AC power line voltage.

19. An arrangement comprising:

a source operative, when connected with a power line, to provide an AC voltage at a set of source terminals; the frequency of this AC voltage being substantially higher than that of the voltage on the power line;

a fluorescent lamp having a first and a second lamp terminal; and

a circuit having a set of input terminals and a set of output terminals; the input terminals being connected with the source terminals; the output terminals being connected with the lamp terminals; the circuit being operative to cause a lamp voltage to be provided at the output terminals and therefore, as long as the lamp terminals are connected therewith, also between the first and the second lamp terminals;

the circuit being further characterized functioning such that:

(a) during an initial brief period commencing after the source has been connected with the power line, the RMS magnitude of the lamp voltage is maintained at a relatively low but substantially constant level; and

(b) after this initial brief period, the RMS magnitude of the lamp voltage is increased toward a relatively high level, thereby normally to cause the lamp to ignite and to provide light.

20. The arrangement of claim 19 wherein, in case the lamp were to fail to ignite, the RMS magnitude of the lamp voltage would cycle periodically between the relatively high level and the relatively low level.

21. The arrangement of claim 20 wherein the RMS magnitude of the lamp voltage would cycle at a frequency substantially lower than that of the power line voltage.

22. The arrangement of claim 19 wherein the duration of the initial period is between one tenth of one second and about two seconds.

23. The arrangement of claim 19 wherein, in case the lamp terminals were to be disconnected from the output terminals, the RMS magnitude of the lamp voltage would cycle periodically between the relatively high level and the relatively low level.

24. An arrangement comprising:

a fluorescent lamp having a first and a second cathode; the first cathode having a first lamp terminal; the second cathode having a second lamp terminal; and

a fluorescent lamp power source having first and second output terminals respectively connected with the first and second lamp terminals; the fluorescent lamp power source being operative, when connected with a power line having a power line voltage, to provide a lamp voltage between the first and the second output terminal; the fluorescent lamp power source including control circuitry functional to control the magnitude of the lamp voltage; the lamp voltage having a frequency and an RMS magnitude; the frequency being substantially higher than that of the power line voltage;

the fluorescent lamp power source being further characterized by functioning such that:

(a) during an initial brief period after having connected the fluorescent lamp power source with the power line, the RMS magnitude is maintained at a relatively low but substantially constant level; and

(b) after this initial brief period, the RMS magnitude is increased toward a relatively high level, thereby normally to cause the lamp to ignite and to provide light.

25. The arrangement of claim 24 wherein, in case the fluorescent lamp were to fail to ignite, the RMS magnitude will cycle periodically between the relatively low substantially constant level and the relatively high level.

26. The arrangement of claim 24 wherein, in case the fluorescent lamp were to be disconnected, the fluorescent lamp power source is characterized by: (i) continuing to provide the lamp voltage between the first and the second pair of output terminals; and (ii) making the RMS magnitude of the lamp voltage cycle, at a frequency substantially lower than that of the power line voltage, between the relatively low substantially constant level and the relatively high level.

27. An arrangement comprising:

a gas discharge lamp having lamp terminals; and

circuitry connected with the power line voltage of an ordinary electric utility power line and functional to provide a high-frequency output voltage between a pair of output terminals operable to connect with the lamp terminals; the circuitry including sub-circuitry functional to control the RMS magnitude of the high-frequency output voltage; the high-frequency output voltage being of frequency substantially higher than that of the power line voltage; the circuitry functioning such as to cause the RMS magnitude of the high-frequency output voltage: (i) to remain substantially constant whenever the gas discharge lamp is indeed connected to and drawing power from the output terminals, and (ii) to vary periodically at a frequency lower than that of the power line voltage whenever the gas discharge lamp is not connected with the output terminals.

* * * * *

EXHIBIT G

US005710488A

United States Patent [19]

Nilssen

[11] **Patent Number: 5,710,488**

[45] **Date of Patent: Jan. 20, 1998**

[54] **LOW-FREQUENCY HIGH-EFFICACY ELECTRONIC BALLAST**

[76] Inventor: Ole K. Nilssen, 408 Caesar Dr., Barrington, Ill. 60010

[21] Appl. No.: **859,616**

[22] Filed: **Mar. 23, 1992**

Related U.S. Application Data

[63] Continuation of Ser. No. 503,094, Apr. 2, 1990, abandoned, which is a continuation of Ser. No. 944,191, Dec. 22, 1986, abandoned.

[51] **Int. Cl.**[6] **H05B 41/36**

[52] **U.S. Cl.** .... **315/224**; 315/200 R; 315/307; 315/DIG. 4; 315/DIG. 5; 315/DIG. 7

[58] **Field of Search** .... 315/DIG. 7, DIG. 5, 315/DIG. 4, 158, 157, 200 R, 151, 163, 165, 166, 307, 310, 224

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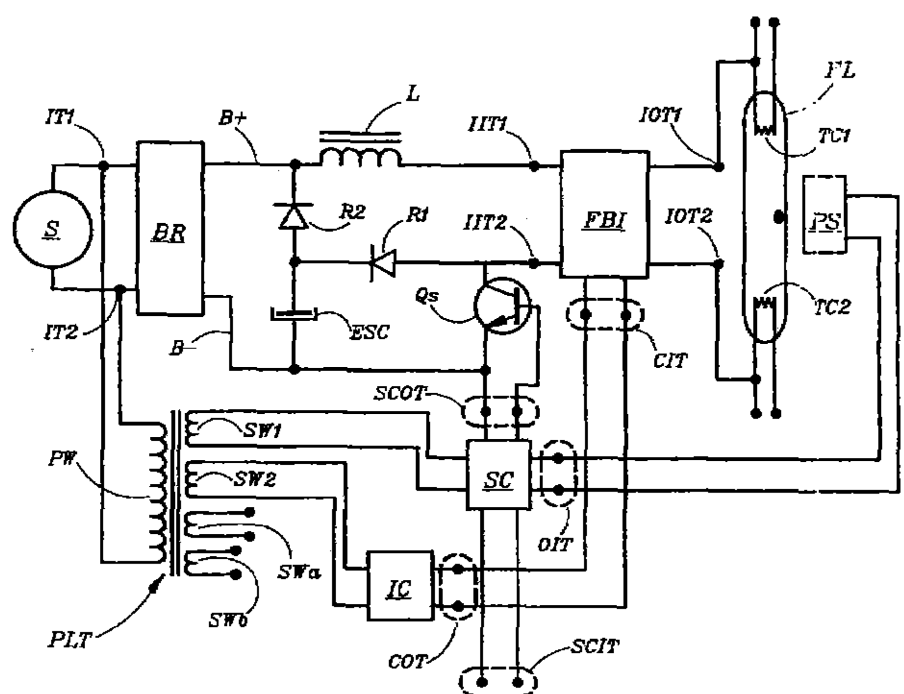
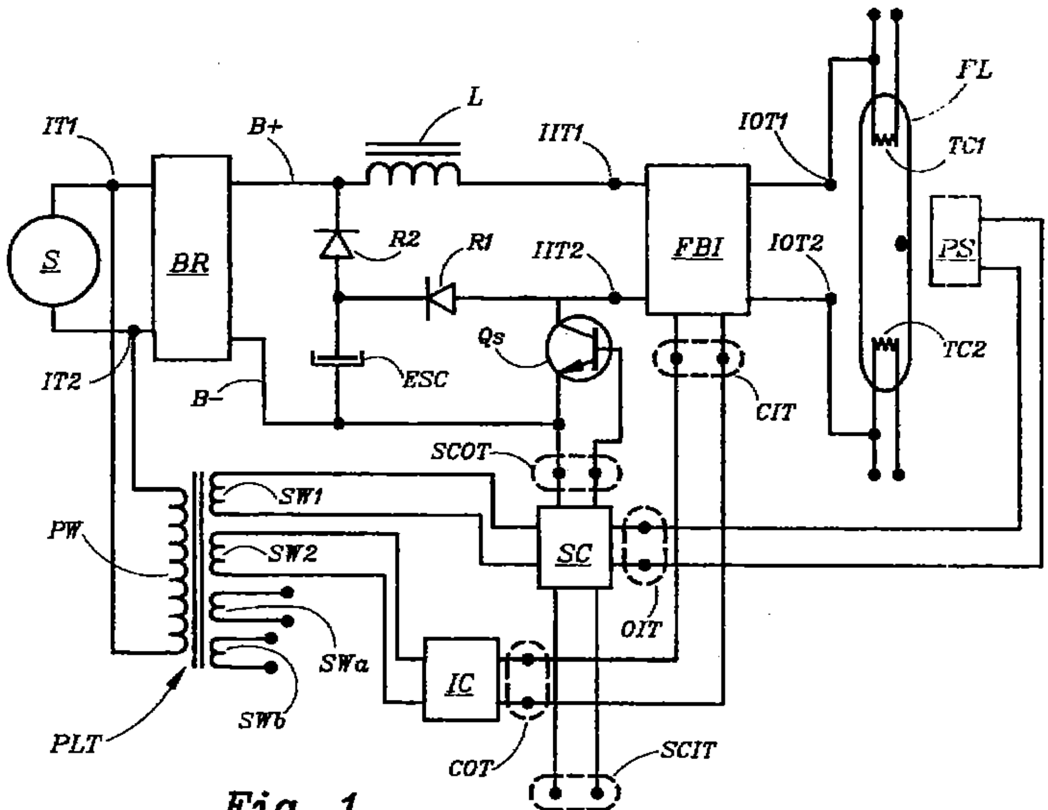
Primary Examiner—Do Hyun Yoo

[57] **ABSTRACT**

A bridge rectifier is connected with a 277Volt/60 Hz power line and provides full-wave-rectified unfiltered DC voltage to a series-combination of: i) an inductor, ii) a full bridge inverter switched in synchronism with the 60 Hz power line voltage, and iii) an electronic switching device. A fluorescent lamp is connected with the inverter's output and receives 60 Hz current of exceptionally low crest-factor, thereby operating at an exceptionally high efficacy. The electronic switching device is normally in a fully conductive state. However, it is controlled—by a photo sensor responsive to the light output of the fluorescent lamp—in such a manner that whenever the instantaneous light output exceeds a certain adjustably predetermined upper level, it switches into a non-conductive state where it remains until the instantaneous light output level diminishes to a certain adjustably predetermined lower level. Whenever the electronic switching device is switched off while current flows through the inductor, a flywheel diode shunts the current away from the switching device and into an energy-storing capacitor, the DC voltage on which is used for filling-in the valleys between the individual 120 Hz DC pulses on the unfiltered power supply.

12 Claims, 2 Drawing Sheets





IT1
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ESC
TC2
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Fig. 1

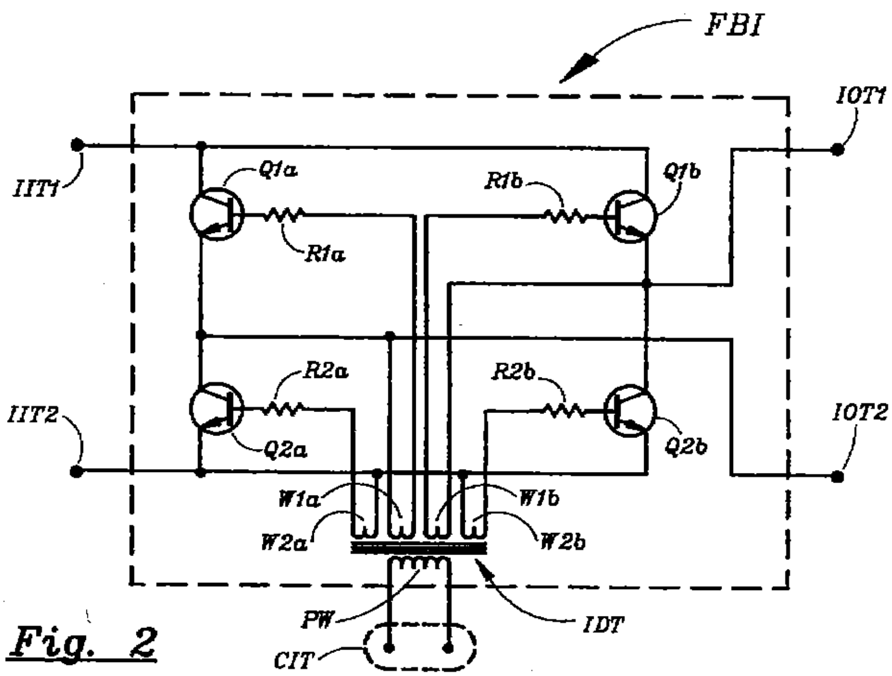

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Fig. 2

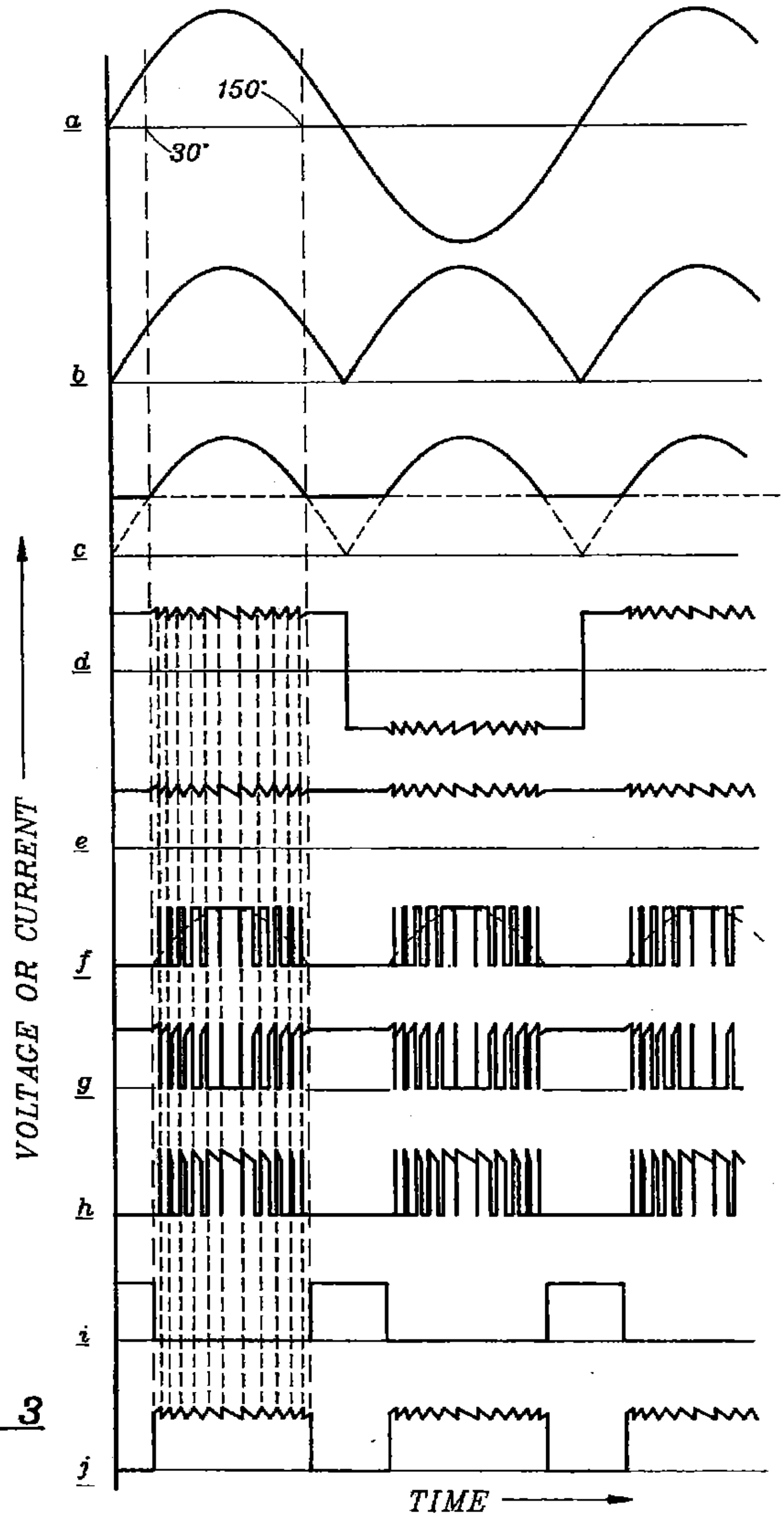

Fig. 3

LOW-FREQUENCY HIGH-EFFICACY ELECTRONIC BALLAST

This application is a continuation of parent application Ser. No. 07/503,094, filed 2 Apr. 1990, now abandoned, which is a continuation of Ser. No. 06/944,191, filed 22 Dec. 1986, now abandoned.

BACKGROUND OF THE INVENTION

1. Field of Invention

The invention relates to ballasts for gas discharge lamps, particularly of a kind wherein: i) the lamps are powered with a relatively low frequency current, and ii) the instantaneous lamp light output flux is maintained substantially constant.

2. Elements of Prior Art

It is well known that significant improvements in overall cost-effectivity of the lighting function can result from appropriately controlling the level of light output from lighting fixtures used for general lighting in offices and the like.

Fluorescent lamp ballasting systems adapted to permit control of light output level on a systems basis presently do exist—as for instance in accordance with U.S. Pat. Nos. 4,207,498 and 4,350,935 to Spira et al.

However, there are significant complexities associated with practical applications of such light level control systems; and, in spite of the very significant improvements potentially available in overall lighting efficacy, such light control systems have not gained wide acceptance.

3. Inventive Rationale

Much of the value available from a light control system may be attained by control of each individual lamp. That way, for instance, light output from each fixture could be kept constant irrespective of any variations in the magnitude of the power line voltage and/or regardless of changes in luminous efficacy of the fluorescent lamp(s).

To make this kind of approach commercially feasible, the present invention provides for a ballast comprising its own individual light sensing means which is so positioned and arranged that, when this ballast is built into a lighting fixture, its light sensor intercepts a part of the light produced by the lamp(s) powered by the ballast and then causes the lamp current to be controlled in such manner as to maintain the lamp light output at a desired level.

Moreover, additional efficacy improvement is attained by powering the lamps in such manner as to keep the instantaneous light flux output from each individual lamp at a substantially constant level; which is to say, by minimizing the amount of flicker—even if that flicker is non-perceivable to the normal human eye.

SUMMARY OF THE INVENTION

Objects of the Invention

A first object of the present invention is that of providing means whereby the light output level of a gas discharge lamp means may be effectively controlled.

A second object is that of providing a ballast comprising means for sensing the light output produced by the gas discharge lamp powered by that ballast, thereby automatically to control that light output in accordance with a desired purpose.

A third object is that of providing means by which to control the magnitude of the current in a gas discharge lamp

such as to maintain its absolute magnitude at an adjustably presetable substantially constant level.

A fourth object is that of providing a ballast operable to power a fluorescent lamp with a current having a particularly low crest-factor.

A fifth object is that of providing a power-line-operated electronic ballast operable to power a gas discharge lamp with 60 Hz current, yet providing improved lamp efficacy.

These as well as several other objects, features and advantages of the present invention will become apparent from the following description and claims.

Brief Description

In its preferred embodiment, the present invention comprises a rectifier means connected with a 277Volt/60 Hz power line and operative to provide full-wave-rectified unfiltered DC voltage to a series-combination of: i) an inductor, ii) a full bridge inverter switched in synchronism with the 60 Hz power line voltage, and iii) an electronic switching means.

A fluorescent lamp is connected with the inverter's output and receives 60 Hz current of exceptionally low crest-factor, thereby operating at an exceptionally high efficacy.

The electronic switching means is normally in a fully conductive state. However, it is controlled—by a photo sensor responsive to the light output of the fluorescent lamp—in such a manner that whenever the instantaneous light output exceeds a certain adjustably predetermined upper level, it switches into a non-conductive state where it remains until the instantaneous light output level diminishes to a certain adjustably predetermined lower level, at which point it switches back into its normally fully conductive state.

Whenever the electronic switching means is switched off while current flows through the inductor, a flywheel diode shunts the current away from the switch means and into an energy-storing capacitor, the DC voltage on which is used for filling-in the valleys between the individual 120 Hz DC pulses on the unfiltered power supply.

By suitable choice of lamp operating voltage, the DC voltage on the energy-storing capacitor can be arranged to be about half the peak mangitude of the power line voltage, in which case power is drawn from the power line with both high power factor as well as good suppression of third harmonics.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 schematically illustrates the preferred embodiment of the invention.

FIG. 2 provides details of the full bridge inverter used in the arrangement of FIG. 1.

FIG. 3 illustrates various voltage and current waveforms associated with the operation of the preferred embodiment.

DESCRIPTION OF THE PREFERRED EMBODIMENT

Description of the Drawings

In FIG. 1, a source S of 277Volt/60 Hz voltage is applied to input terminals IT1 and IT2 of a bridge rectifier BR, the unidirectional voltage output of which is applied directly between a B+ bus and a B− bus, with the positive voltage being connected to the B+ bus.

Between the B+ bus and a first inverter input terminal IIT1 of a full bridge inverter FBI is connected an inductor L.

A switching transistor Qs is connected with its collector to a second inverter input terminal IIT2 and with its emitter to the B− bus. A first rectifier R1 is connected with its anode to inverter input terminal IIT2 and with its cathode to the anode of a second rectifier R2. The cathode of rectifier R2 is connected with the B+ bus. An energy-storing capacitor ESC is connected between the cathode of rectifier R1 and the B− bus.

The output from inverter FBI is provided across inverter output terminals IOT1 and IOT2; which are respectively connected with thermionic cathodes TC1 and TC2 of a fluorescent lamp FL. Inverter FBI has a pair of control input terminals CIT connected with control output terminals COT of an inverter controller IC.

A switch controller SC has: i) a pair of switch controller output terminals SCOT connected between the base and the emitter of switching transistor Qs, ii) a pair of switch controller input terminals SCIT, and iii) a pair of opto-input terminals OIT connected with a photo-sensor PS.

Switch controller SC and inverter controller IC are respectively connected with a first secondary winding SW1 and a second secondary winding SW2 of a power line transformer PLT; which power line transformer has two additional secondary windings SWa and SWb which, each connected with one of thermionic cathodes. TC1 and TC2 of fluorescent lamp FL. Power line transformer PLT has a primary winding PW connected between input terminals IT1 and IT2 of bridge rectifier BR.

FIG. 2 illustrates key details of full bridge inverter FBI.

In FIG. 2, a first transistor Q1*a* is connected with its collector to inverter input terminal IIT1 and with its emitter to the collector of a second transistor Q2*a*, whose emitter is connected with inverter input terminal IIT2. A third transistor Q1*b* is similarly connected with its collector to inverter input terminal IIT1 and with its emitter to the collector of a fourth transistor Q2*b*, whose emitter is connected with inverter input terminal IIT2.

Control input terminals CIT of inverter FBI are connected with primary winding PW of an inverter drive transformer IDT. This transformer has four secondary windings W1*a*, W2*a*, W1*b* and W2*b*; which windings are connected with the base-emitter junctions of transistors Q1*a*, Q2*a*, Q1*b* and Q2*b* by way of resistors R1*a*, R2*a*, R1*b* and R2*b*; all respectively.

Details of Operation

The operation of the ballast arrangement of FIG. 1 may best be understood when reading the following explanation in light of the waveforms illustrated by FIG. 3.

FIG. 3*a* illustrates the waveform of the power line voltage present between input terminals IT1 and IT2; which waveform is identical to the waveform of the voltage applied across fluorescent lamp FL before lamp ignition.

FIG. 3*b* illustrates the pulsed DC voltage resulting from full-wave-rectification of the power line voltage of FIG. 3*a*.

FIG. 3*c* illustrates the waveform of the net DC voltage present between the B− bus and the B+ bus after the fluorescent lamp has ignited and is in stable operation.

FIG. 3*d* illustrates the waveform of the current flowing through the fluorescent lamp during normal operation.

FIG. 3*e* indicates the instantaneous magnitude of the light flux emitted from fluorescent lamp FL. In addition, in rough approximation, FIG. 3*e* indicates the absolute value of the magnitude of the voltage across fluorescent lamp FL.

FIG. 3*f* indicates the waveform of the voltage present across switching transistor Qs.

FIG. 3*g* indicates the current flowing through switching transistor Qs.

FIG. 3*h* indicates the current flowing into energy-storing capacitor ESC through R1.

FIG. 3*i* indicates the current drawn from energy-storing capacitor ESC through rectifier R2.

FIG. 3*j* indicates the waveform of the current drawn from the power line by bridge rectifier BR.

The details of operation of the circuit of FIG. 1 may now be explained as follows.

In FIG. 1, the source S represents an ordinary electric utility power line, the 277Volt/60 Hz power line voltage from which (see FIG. 3*a*) is applied directly to the bridge rectifier (BR). This bridge rectifier is of conventional construction and provides for the full-wave-rectified power line voltage (see FIG. 3*b*) to be applied to the circuit by way of the B+ bus and the B− bus.

As soon as the power line voltage is connected with input terminals IT1 and IT2, cathode heating voltages are applied to thermionic cathodes TC1 and TC2, thereby bringing these cathodes to incandescence within about 1.5 seconds. Hence, the fluorescent lamp is ready to be ignited in rapid-start manner within about 1.5 second after initial application of power line voltage.

Also, as long as power line voltage is provided to input terminals IT1/IT2, power is provided to switch controller SC and inverter controller IC by way of secondary windings SW1 and SW2, respectively, of power line transformer PLT.

The inverter controller (IC) is operative to convert the 60 Hz sinusoidal voltage received from secondary winding SW2 to a 60 Hz squarewave voltage, which is provided at its output terminals COT and thereby to the primary winding of inverter drive transformer IDT. In turn, by way of transformer IDT, the base-emitter junctions of transistors Q1*a*, Q2*a*, Q1*b* and Q2*b* is provided with a squarewave current-limited voltage drive—with resistors R1*a*, R2*a*, R1*b* and R2*b* acting as the current-limiting means. Thus, as long as the arrangement of FIG. 1 is connected with the power line, inverter FBI operates to invert in complete synchrony with the frequency of the power line voltage.

The switch controller (SC) is operable to provide a control voltage to switching transistor Qs such as to cause it to enter its fully conductive state, where it will remain until the output from the photo sensor (PS) reaches a certain predetermined upper magnitude. At that point, the switch controller abruptly reverses the control voltage supplied to the switching transistor such as to cause it to enter its non-conductive state, where it will remains until the output from the photo sensor decreases by at least a relatively small percentage from this certain predetermined upper level to a certain predetermined lower level. The certain predetermined upper level is adjustably controllable (i.e., setable) by provision of a control signal to switch controller input terminals SCIT; the magnitude-ratio between the certain upper level and the certain lower level remaining approximately constant.

Before the fluorescent lamp ignites (which is to say, before significant lamp current flows), the voltage on energy-storing capacitor ESC is zero. Moreover, since the lamp then provides no light output, the switching transistor (Qs) exists in its fully conductive state. Thus, with the inverter (FBI) providing for full-wave inversion of the voltage applied to it, the starting voltage applied to the fluorescent lamp (FL) is substantially identical to the power line voltage applied between input terminals IT1 and IT2 (see FIG. 3*a*).

As the fluorescent lamp ignites, lamp current starts flowing and light starts being provided by the lamp. After a few milliseconds (the exact length of time being principally determined by the magnitude of the supply voltage and the inductance of the current-limiting inductor (L), the lamp's light output level reaches the certain predetermined upper level, at which point switching transistor Qs switches into its non-conductive state. After this point, the lamp current continues to flow through rectifier R1 and into the energy-storing capacitor (ESC); which then starts to charge up, eventually reaching the point at which its voltage becomes so high as to cause the magnitude of the lamp current to diminish—eventually to reach the certain predetermined lower level, thereby to cause switching transistor Qs to switch back into its fully conductive state.

After the above-described initial starting period, during which light output from the fluorescent lamp will have exceeded its normally maximum instantaneous light output level for a brief period, operation of the circuit arrangement of FIG. 1 settles into a steady state characterized by the waveforms of FIG. 3 and otherwise explained as follows.

1. The magnitude of the DC voltage on energy-storing capacitor ESC will be substantially constant and approximately equal to the difference between: i) the peak magnitude of the voltage provided from bridge rectifier BR (see FIG. 3*b*), and ii) the average of the absolute magnitude of the voltage present across the fluorescent lamp. In the preferred embodiment, the fluorescent lamp actually consists of a special 96"/T12 rapid-start fluorescent lamp, and the average absolute magnitude of the lamp voltage is about 196 Volt. With the power line voltage being 277Volt/60 Hz, the peak magnitude of the voltage provided from the bridge rectifier is about 392 Volt; which means that the magnitude of the substantially constant DC voltage on energy-storing capacitor ESC is also about 196 Volt. Thus, as indicated in FIG. 3*c*, the DC voltage actually provided between the B− bus and the B+ bus is the higher of: i) the instantaneous magnitude of the full-wave-rectified power line voltage, and ii) the substantially constant magnitude of the DC voltage on energy-storing capacitor ESC.

2. The current flowing through the fluorescent lamp will be as indicated in FIG. 3*d*; which waveform, in terms of absolute magnitude, correlates closely with the instantaneous magnitude of the luminous flux emitted from the fluorescent lamp, as indicated in FIG. 3*e*. Moreover, the details of the waveform of the luminous flux emitted from the fluorescent lamp correlates with the waveform of the voltage across switching transistor Qs as interpreted in correlation with the waveform of the inverter's DC supply voltage of FIG. 3*c*.

3. The waveform of the current drawn from the power line is indicated in FIG. 3*j* and is seen to be of substantially constant magnitude between 30 degrees and 150 degrees of each half-cycle of the power line voltage. Consequently, there is substantially no third harmonic content in the current waveform; which fact is important in most installations of fluorescent lighting systems. The reason that current is drawn from the power line only during this particular interval relates to the fact that the magnitude of the DC voltage on energy-storing capacitor ESC is about half that of the peak magnitude of the power line voltage. With that being the case, the instantaneous magnitude of the full-wave-rectified power line voltage starts exceeding the magnitude of the voltage on capacitor ESC at about 30 degrees; and it starts falling below the magnitude of the capacitor voltage at about 150 degrees. Moreover, with the particular waveform of FIG. 3*j*, the power factor with which power is drawn from the power line is relatively high at about 85%.

From an overall functional viewpoint, the steady-state operation of the circuit of FIG. 1 may be explained as follows.

Whenever the magnitude of the DC voltage applied between the B− bus and the B+ bus exceeds the magnitude of the voltage across the fluorescent lamp, and as long as switching transistor Qs is in its fully conductive state, there is a net forward voltage present across inductor L; which means that the current through inductor L (and thereby through the fluorescent lamp) will increase. As this inductor/lamp current increases, so—within a few micro-seconds—does the lamp light output; and a point is soon reached at which the lamp light output becomes large enough to make the output from photo sensor PS such as to cause the switch controller to cause the switching transistor to switch into its non-conductive state.

After that point, the inductor/lamp current continues to flow, except that this current now has to flow into capacitor ESC. In doing so, the current must overcome both the lamp voltage as well as the DC voltage on capacitor ESC, the sum of whose magnitudes exceeds the instantaneous magnitude of the voltage present between the B− bus and the B+ bus. Thus, there is a net reverse voltage present across inductor L, which means that the magnitude of the inductor/lamp current will now start to decrease. As this inductor/lamp current decreases, so—within a few micro-seconds—does the lamp light output; and a point is soon reached at which the lamp light output becomes low enough to make the output from photo sensor PS such as to cause the switch controller to cause the switching transistor to switch back into its fully conductive state.

Thereafter, the cycle will repeat with a repetition rate depending on: i) the instantaneous magnitude of the DC voltage between the B− bus and the B+ bus, ii) the degree of hysteresis associated with the photo sensor and the switch controller, iii) the absolute magnitude of the voltage across the lamp, iv) the magnitude of the inductance of inductor L, and v) the delay between an increase/decrease in lamp current versus the corresponding increase/decrease in lamp light output.

Since the delay between the increase/decrease of lamp current versus the corresponding increase/decrease in lamp light output is less than about 25 micro-seconds for most ordinary fluorescent lamps, it is clearly necessary to make the time-period of each increase/decrease of inductor/lamp current substantially longer than about 25 micro-seconds; which means that it is necessary to make the inductance of inductor L large enough to cause detectable changes in current magnitude to occur over time-periods substantially longer than 25 micro-seconds.

Of course, the detectable changes in current magnitude depends directly on the detectable changes in the level of light flux output; which, in turn, depends on the specifications of the switch controller and particularly on the amount of hysteresis built thereinto. In the preferred embodiment, the sensitivity has been so arranged that the relative hysteresis-gap is about plus/minus 10%.

Thus, with reference to FIGS. 3*d*, 3*e*, and 3*j*, the indicated variations in magnitude stays within the band of ±10%.

Additional Comments

a) The waveforms of FIG. 3 illustrate steady-state operation of the ballasting arrangement of FIG. 1 under the particular condition where the magnitude of the voltage-drop across the fluorescent lamp is approximately half the peak magnitude of the power line voltage; which condition represents a highly desirable situation and is in fact apporoximately attainable in many actual applications.

If the magnitude of the voltage-drop across the fluorescent lamp were to be substantially less than half the peak magnitude of the power line voltage, the result would be that: i) the magnitude of the DC voltage on capacitor ESC would increase, and ii) switching transistor Qs would be activated more frequently and even during the period when lamp power is being provided by the energy-storing capacitor, which is in contrast with the situation illustrated in FIG. 3.

On the other hand, if the magnitude of the voltage-drop across the fluorescent lamp were to be somewhat larger than half the peak magnitude of the power line voltage, the result would be that: i) the magnitude of the DC voltage on capacitor ESC would decrease, and ii) switching transistor Qs would be activated less frequently. However, for the ballast circuit to work at all, it is necessary that the magnitude of the voltage-drop across the lamp be no higher than the average magnitude of the voltage provided between the B− bus and the B+ bus; which, in the limiting case, means that the magnitude of the voltage-drop across the lamp can not exceed about 63% of the peak magnitude of the power line voltage.

In any case, as long as the peak magnitude of the power line voltage exceeds the magnitude of the voltage-drop across the fluorescent lamp by at least 58%, the ballasting circuit of FIG. 1 will automatically operate to properly power the fluorescent lamp.

b) When the magnitude of the voltage-drop across the lamp is significantly less than half the peak magnitude of the power line voltage, the conduction angle of the current drawn from the power line gets reduced; and the power factor with which the ballast draws power from the power line gets correspondingly reduced.

c) It is of course a simple matter to increase or decrease the magnitude of the voltage applied to the ballast input terminals IT1 and IT2. This can be done by auto-transformer action, using therefor a tapped primary winding on power line transformer PLT.

d) One of the major values provided by the ballasting arrangement of FIG. 1 is that of providing for gas in the fluorescent lamp to operate at an essentially constant level of ionization; which, in turn, results in several important values, such as: i) higher luminous efficacy, ii) longer lamp life, and iii) reduced flicker

e) The degree of hysteresis built into the switch controller can be chosen at will over a wide range. However, in view of practical considerations, in the preferred embodiment, a relative hysteresis range of plus/minus 10% was chosen. This value is readily attainable by use of commonly available electronic components, such as the opto-actuated Schmitt trigger used in Motorola's H11L1 opto coupler/isolator.

f) Adjustment of the light level about which the automatic control takes place can readily be accomplished in Several ways.

For instance, the positioning of photo sensor PS relative to the fluorescent lamp determines how much of the lamp light flux it receives, thereby determining its control threshold.

Or, a shade can be used to block off more or less of the light flux reaching the photo sensor.

A more practical arrangement, however, is that of providing an adjustable bias to the trigger means (or hysteresis means) comprised within switch controller SC; which is indeed the arrangement used in the preferred embodiment.

g) Ordinarily, when a fluorescent lamp is initially provided with power, its light output will be substantially lower than it will be once the lamp has warmed up to proper operating temperature. The ballast of FIG. 1 provides compensation for this effect, in that the lamp will automatically be provided with substantially higher current as long as the light output is not up to the desired level.

h) An important value associated with providing automatic light output control as herein described relates to energy-efficiency beyond the point of simply making the lamp itself operate at a higher efficacy. For a specified level of light output, by automatically compensating for line voltage fluctuations and the naturally-occuring lamp light output deterioration over time, an overall additional efficiency-advantage of nearly 20% is attained.

i) It is believed that the present invention and its several attendant advantages and features will be understood from the preceeding description. However, without departing from the spirit of the invention, changes may be made in its form and in the construction and interrelationships of its component parts, the form herein presented merely representing the presently preferred embodiment.

I claim:

1. An arrangement comprising:

a source operative to provide an AC voltage at a pair of AC terminals; and

a circuit connected with the AC terminals and operable to power a gas discharge lamp; the circuit being characterized by:

(i) providing a conditioned voltage at a pair of terminals; and

(ii) including a transistor conducting intermittently at a time-varying switching frequency, thereby to maintain the absolute magnitude of the conditioned voltage substantially constant.

2. The arrangement of claim 1 wherein the switching frequency is:

(i) many times higher than the frequency of the AC voltage; and

(ii) time-varying at a frequency equal to twice the frequency of the AC voltage.

3. The arrangement of claim 1 wherein the source is an ordinary electric utility power line.

4. The arrangement of claim 1 wherein the conditioned voltage is:

(i) an alternating voltage; and

(ii) of frequency substantially lower than the switching frequency.

5. An arrangement comprising:

a source operative to provide an AC voltage at a pair of AC terminals; and

a circuit connected with the AC terminals and operable to power a gas discharge lamp; the circuit being characterized by:

(i) providing a conditioned voltage at a pair of terminals; and

(ii) including a transistor conducting intermittently with a time-varying duty-cycle at a switching frequency, thereby being operative to maintain the absolute magnitude of the conditioned voltage substantially constant.

6. The arrangement of claim 5 wherein:

(i) the switching frequency is many times higher than the frequency of the AC voltage; and

(ii) the duty-cycle varies at a frequency equal to twice the frequency of the AC voltage.

7. The arrangement of claim 5 wherein the conditioned voltage is:

(i) an alternating voltage; and

(ii) of frequency substantially lower than the switching frequency.

8. An arrangement comprising:

a source providing an AC voltage at a pair of AC terminals; and

an assembly of electrical components connected with the AC terminals and characterized by:

(i) including a gas discharge lamp;

(ii) providing an output voltage from a pair of output terminals; and

(iii) including a transistor conducting intermittently at a time-varying switching frequency, thereby to maintain the absolute magnitude of the output voltage substantially constant.

9. The arragement of claim 8 wherein an electrical conduction path exists, at least at certain times, between one of the output terminals and one of the AC terminals.

10. The arrangement of claim 8 wherein the assembly is additionally characterized by including an energy-storing inductor.

11. The arrangement of claim 8 wherein the gas discharge lamp is a fluorescent lamp operating at a substantially constant level of ionization.

12. An arrangement comprising:

a source providing an AC voltage at a pair of AC terminals; and

an assembly of electrical components and parts connected with the AC terminals and characterized by:

(i) including a gas discharge lamp;

(ii) being operable to power the gas discharge lamp at a substantially constant level of ionization, thereby causing the lamp to emit a substantially constant level of light output;

(iii) providing an output voltage at a pair of output terminals; and

(iv) including a transistor conducting intermittently at a time-varying switching frequency, thereby to maintain the absolute magnitude of the output voltage substantially constant.

* * * * *

EXHIBIT H

US005710489A

United States Patent [19]

Nilssen

[11] Patent Number: 5,710,489

[45] Date of Patent: Jan. 20, 1998

[54] OVERVOLTAGE AND THERMALLY PROTECTED ELECTRONIC BALLAST

[76] Inventor: Ole K. Nilssen, 408 Caesar Dr., Barrington, Ill. 60010

[21] Appl. No.: 348,327

[22] Filed: Dec. 2, 1994

Related U.S. Application Data

[63] Continuation-in-part of Ser. No. 771,801, Oct. 7, 1991, abandoned, which is a continuation of Ser. No. 88,592, Aug. 24, 1987, abandoned, which is a continuation-in-part of Ser. No. 720,387, Apr. 5, 1985, Pat. No. 4,663,571, and a continuation-in-part of Ser. No. 720,386, Apr. 5, 1985, Pat. No. 4,675,576, and a continuation-in-part of Ser. No. 691,171, Jan. 14, 1985, Pat. No. 4,644,228, and a continuation-in-part of Ser. No. 686,447, Dec. 26, 1984, Pat. No. 4,638,395, and a continuation-in-part of Ser. No. 677,562, Dec. 3, 1984, Pat. No. 4,698,553, and a continuation-in-part of Ser. No. 612,058, May 18, 1984, Pat. No. 4,667,131, and a continuation-in-part of Ser. No. 605,479, Apr. 30, 1984, Pat. No. 4,626,953, which is a continuation-in-part of Ser. No. 720,387, Apr. 5, 1985, Pat. No. 4,663,571, which is a continuation-in-part of Ser. No. 720,386, Apr. 5, 1985, Pat. No. 4,675,576, which is a continuation-in-part of Ser. No. 691,171, Jan. 14, 1985, Pat. No. 4,644,228, which is a continuation-in-part of Ser. No. 686,447, Dec. 26, 1984, Pat. No. 4,638,395, which is a continuation-in-part of Ser. No. 677,562, Dec. 3, 1984, Pat. No. 4,698,553, which is a continuation-in-part of Ser. No. 612,058, May 18, 1984, Pat. No. 4,667,131, which is a continuation-in-part of Ser. No. 605,479, Apr. 30, 1984, Pat. No. 4,626,953, and a continuation-in-part of Ser. No. 640,240, Aug. 13, 1984, Pat. No. 4,563,719, and a continuation-in-part of Ser. No. 506,420, Jun. 21, 1983, Pat. No. 4,581,562, and a continuation-in-part of Ser. No. 500,841, Jun. 3, 1983, Pat. No. 4,538,095, and a continuation-in-part of Ser. No. 495,540, May 17, 1983, Pat. No. 4,554,487, and a continuation-in-part of Ser. No. 481,714, Apr. 4, 1983, Pat. No. 4,507,698, and a continuation-in-part of Ser. No. 456,276, Feb. 22, 1983, Pat. No. 4,503,363, and a continuation-in-part of Ser. No. 411,263, Aug. 25, 1982, Pat. No. 4,461,980.

[51] Int. Cl.[6] G05F 1/00

[52] U.S. Cl. .... **315/309**; 315/291; 315/225; 315/209 R; 315/DIG. 7

[58] Field of Search .... 315/291, 309, 315/224, 225, 244, 241 R, 276, 209 R, 206, DIG. 7

[56] **References Cited**

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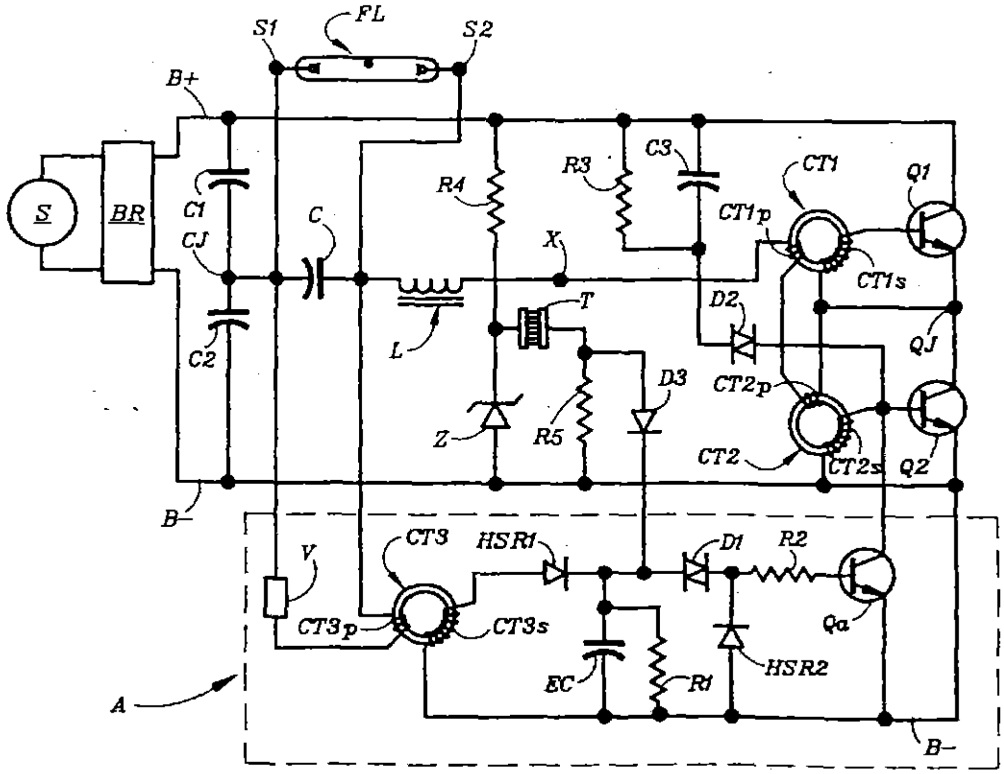
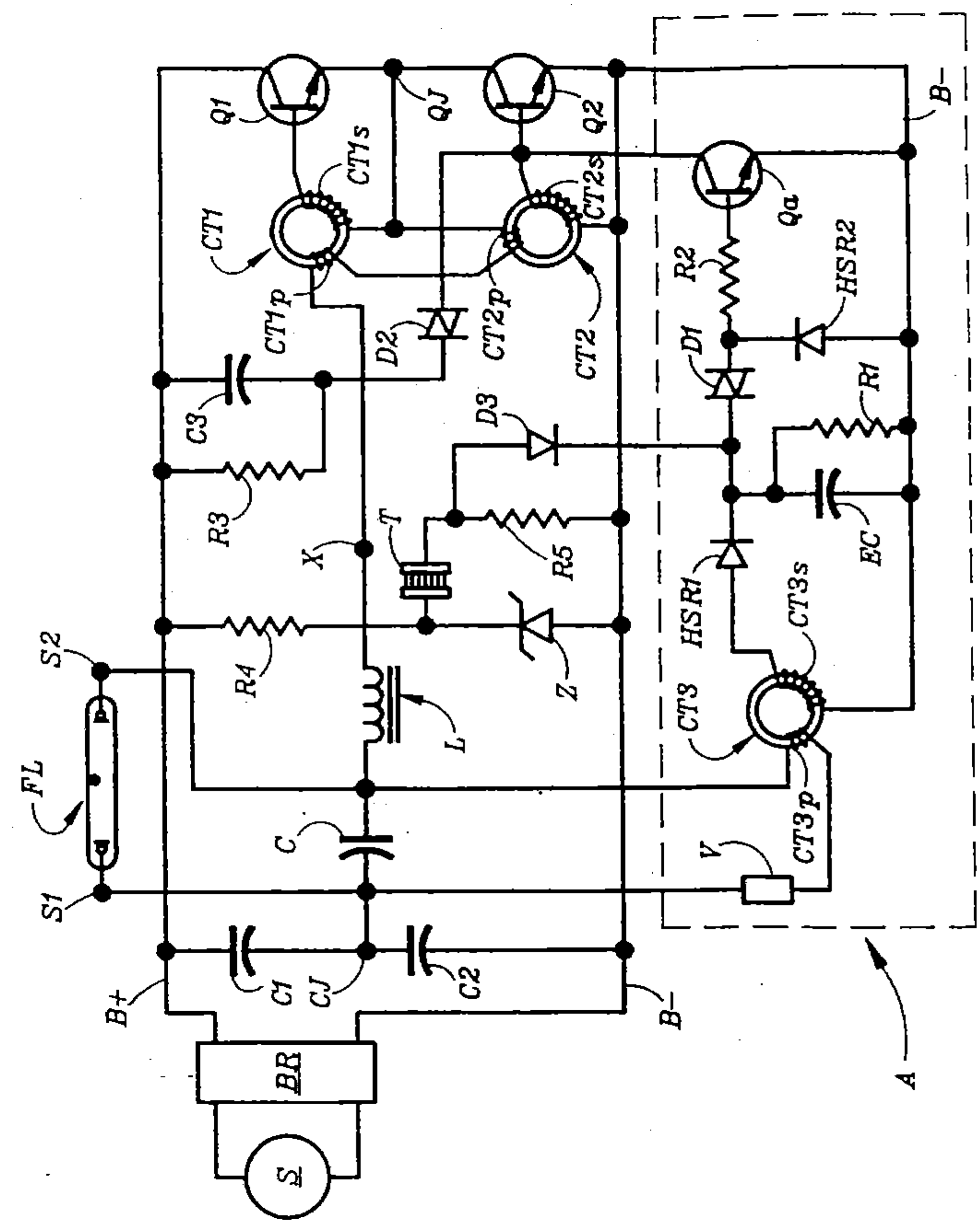
Primary Examiner—Robert Pascal
Assistant Examiner—Haissa Philogene

[57] **ABSTRACT**

An inverter-type electronic fluorescent lamp ballast has means to disable inverter operation in case of an overvoltage condition and/or in case its internal temperature were to exceed a safe level, thereby to provide automatic protection against damage that might otherwise result from excessive voltage and/or temperatures.

15 Claims, 1 Drawing Sheet

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FL
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CT3p
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S1
C1
CJ
C2
B+
BR
S
A

Fig. 1

OVERVOLTAGE AND THERMALLY PROTECTED ELECTRONIC BALLAST

RELATED APPLICATIONS

Instant application is a continuation-in-part of Ser. No. 07/771,801 filed Oct. 7, 1991 now abandoned; which is a continuation of Ser. No. 07/088,592 filed Aug. 24, 1987 now abandoned; which application Ser. No. 07/088,592 is also a contination-in-part of the following seven applications:

1. Ser. No. 06/720,387 filed Apr. 5, 1985, now U.S. Pat. No. 4,663,571;
2. Ser. No. 06/720,386 filed Apr. 5, 1985, now U.S. Pat. No. 4,675,576;
3. Ser. No. 06/691,171 filed Jan. 14, 1985, now U.S. Pat. No. 4,644,228;
4. Ser. No. 06/686,447 filed Dec. 26, 1984, now U.S. Pat. No. 4,638,395;
5. Ser. No. 06/677,562 filed Dec. 3, 1984, now U.S. Pat. No. 4,698,553;
6. Ser. No. 06/612,058 filed May 18, 1984, now U.S. Pat. No. 4,667,131;
7. Ser. No. 06/605,479 filed Apr. 30, 1984, now U.S. Pat. No. 4,626,953;

each one if which seven applications is a continuation-in-part of: (i) Ser. No. 06/640,240 filed Aug. 13, 1984, now U.S. Pat. No. 4,563,719; (ii) Ser. No. 06/506,420 filed Jun. 21, 1983, now U.S. Pat. No. 4,581,562; (iii) Ser. No. 06/500,841 filed Jun. 3, 1983, now U.S. Pat. No. 4,538,095; (iv) Ser. No. 06/495,540 filed May 17, 1983, now U.S. Pat. No. 4,554,487; (v) Ser. No. 06/481,714 filed Apr. 4, 1983, now U.S. Pat. No. 4,507,698; (vi) Ser. No. 06/456,276 filed Feb. 22, 1983, now U.S. Pat. No. 4,503,363; and (vii) Ser. No. 06/411,263 filed Aug. 25, 1982, now U.S. Pat. No. 4,461,980.

BACKGROUND OF THE INVENTION

Field of Invention

Instant invention relates to means for automatically protecting inverter-type electronic ballasts from excessive voltages and temperatures, particularly as accomplished by automatic inverter disablement.

SUMMARY OF THE INVENTION

Objects of the Invention

An object of the present invention is that of providing a cost-effective means for preventing an inverter-type electronic ballast from becoming a potential electric shock hazard and/or fire initiation hazard.

These as well as other objects, features and advantages of the present invention will become apparent from the following description and claims.

Brief Description

In its preferred embodiment, subject invention constitutes a series-excited parallel-loaded fluorescent lamp ballast comprising the following key component parts:

a source of DC voltage, which DC voltage is derived by rectification of the AC voltage from a regular 60 Hz power line;

an inverter connected with the source of DC voltage and operative to provide across an output a relatively high-frequency squarewave voltage, the inverter comprising a

disable-means operative on receipt of a disable-signal to disable the inverter and thereby to remove the squarewave voltage from the output while also substantially reducing the power drawn by the inverter from the source of DC voltage;

a series LC circuit connected across the output, the LC circuit being substantially series-resonant at the fundamental frequency of the squarewave voltage;

a fluorescent lamp connected across the tank-capacitor of the LC circuit;

temperature sensor means operative to provide the disable signal whenever its temperature exceeds a predetermined level;

overvoltage sensor means operative to provide the disable signal whenever the magnitude of the voltage across the tank-capacitor exceeds a pre-established level; and

whereby, if the magnitude of the voltage across the lamp and/or the temperature of the ballast were to become excessive, the inverter be disabled.

BRIEF DESCRIPTION OF THE DRAWING

FIG. 1 schematically illustrates the preferred embodiment of the invention.

DESCRIPTION OF THE PREFERRED EMBODIMENT

Description of the Drawing

In FIG. 1, a source S of 120 Volt/60 Hz voltage is applied to a full-wave bridge rectifier BR, the unidirectional voltage output of which is applied directly between a B+ bus and a B– bus, with the positive voltage being connected to the B+ bus.

Between the B+ bus and the B– bus are connected a series-combination of two transistors Q1 and Q2 as well as a series-combination of two energy-storing capacitors C1 and C2.

The secondary winding CT1s of positive feedback current transformer CT1 is connected directly between the base and the emitter of transistor Q1; and the secondary winding CT2s of positive feedback current transformer CT2 is connected directly between the base and the emitter of transistor Q2.

The collector of transistor Q1 is connected directly with the B+ bus; the emitter of transistor Q2 is connected directly with the B– bus; and the emitter of transistor Q1 is connected directly with the collector of transistor Q2, thereby forming junction QJ.

One terminal of capacitor C1 is connected directly with the B+ bus, while the other terminal of capacitor C1 is connected with a junction CJ. One terminal of capacitor C2 is connected directly with the B– bus, while the other terminal of capacitor C2 is connected directly with junction CJ.

An inductor L and a capacitor C are connected in series with one another and with the primary windings CT1p and CT2p of current transformers CT1 and CT2.

The series-connected primary windings CT1p and CT2p are connected directly between junction QJ and a point X. Inductor L is connected with one of its terminals to point X and with the other of its terminals to one of the terminals of capacitor C. The other terminal of capacitor C is connected directly with junction CJ.

A fluorescent lamp FL is connected, by way of lamp sockets S1 and S2, in parallel-circuit across capacitor C.

A Varistor V and primary winding CT3p of current transformer CT3 are connected in series across capacitor C.

One terminal of the secondary winding CT3s of transformer CT3 is connected with the B– bus; the other terminal of this secondary winding is connected with the anode of a high speed rectifier HSR1. The cathode of rectifier HSR1 is connected to the positive terminal of an energy-storing capacitor EC. The negative terminal of capacitor EC is connected directly to the B– bus. A bleeding resistor R1 is connected directly across capacitor EC.

A Diac D1 is connected between the cathode of rectifier HSR1 and the cathode of another high speed rectifier HSR2. The anode of rectifier HSR2 is connected to the B– bus.

Between the cathode of rectifier HSR2 and the base of an auxiliary transistor Qa is connected a resistor R2.

The collector of transistor Qa is connected directly to the base of transistor Q2, and the emitter of transistor Qa is connected directly to B– bus.

The combination of varistor V, current transformer CT3, rectifier HSR1, capacitor EC, Resistor R1, Diac D1, rectifier HSR2, resistor R2 and transistor Qa is referred to as subassembly A.

A series-combination of a capacitor C3 and a Diac D2 is connected between the B+ bus and the base of transistor Q2. A resistor R3 is connected in parallel with capacitor C3.

A resistor R4 is connected between the B+ bus and the cathode of a Zener diode Z, whose anode is connected with the B– bus. A thermistor T is connected between the cathode of Zener diode Z and the anode of a diode D3, whose cathode is connected with the cathode of rectifier HSR1. A resistor R5 is connected between the anode of diode D3 and the B– bus.

Details of Operation

In FIG. 1, the source S represents an ordinary electric utility power line, the voltage from which is applied directly to the bridge rectifier identified as BR. This bridge rectifier is of conventional construction and provides for the rectified line voltage to be applied to the inverter circuit by way of the B+ bus and the B– bus.

The two energy-storing capacitors C1 and C2 are connected directly across the output of the bridge rectifier BR and serve to filter the rectified line voltage, thereby providing for the voltage between the B+ bus and the B– bus to be substantially constant. Junction CJ between the two capacitors serves to provide a power supply center tap.

The inverter circuit of FIG. 1, which represents a so-called half-bridge inverter, operates in a manner that is analogous with circuits previously described in published literature, as for instance in U.S. Pat. No. 4,184,128 entitled High Efficiency Push-Pull Inverters.

Upon initial application of power to the circuit, inverter oscillation is initiated by one or a few trigger pulses applied to the base of transistor Q2 by way of the combination of Capacitor C3 and Diac D2. Once the magnitude of the B+ voltage has stabilized, due to the effect of R3, periodic additional trigger pulses will be provided.

Under normal circumstances, these additional trigger pulses will have substantially no effect on the circuit. However, if for one reason or other inverter oscillations were to be interrupted, the trigger pulses will serve to restart the oscillations.

The output of the half-bridge inverter is a substantially squarewave 33 kHz AC voltage provided between point X and junction CJ. Directly across this output is connected a

resonant or near-resonant L-C series circuit—with the fluorescent lamp connected in parallel with the tank-capacitor thereof.

The resonant or near-resonant action of the L-C series circuit provides for appropriate lamp starting and operating voltages, as well as for proper lamp current limiting; which is to say that it provides for appropriate lamp ballasting.

When the inverter is operating, the voltage developed across the tank-capacitor is essentially only limited by the voltage-clamping characteristics of either the fluorescent lamp FL or the Varistor V—i.e., by the one which clamps at the lower voltage. If the lamp is inoperable, or if the lamp is removed from the circuit, or during the brief period before the lamp ignites, the Varistor acts as the principal voltage-clamping means; and the circuit load current then flows through this Varistor. As soon as the lamp gets into operation, however, the voltage across the tank-capacitor (and thereby across the Varistor) falls to a magnitude that is so low that current will no longer flow through the Varistor.

In the arrangement of FIG. 1, the various relevant voltage and current magnitudes are approximately as follows: i) maximum required lamp starting voltage: 500 Volt RMS for not more than about 50 milli-Second; ii) Varistor RMS and peak clamping voltage, as well as energy-handling capability: 511 Volt RMS, 750 Volt and 40 Joules, respectively; lamp operating voltage and current: 140 Volt RMS and 0.2 Amp RMS, respectively.

In an LC series-resonant circuit, the power provided to a resistive load connected in parallel with the circuit tank-capacitor is approximately proportional to the magnitude of the load resistance. Hence, in FIG. 1, as long as the parameters of the LC circuit have been arranged to provide the fluorescent lamp with its required 0.2 Amp operating current at 140 Volt RMS (which corresponds to 28 Watt), the load power resulting at higher voltages will be roughly proportionately larger. Thus, at the point where the Varistor is clamping (at about 511 Volt RMS), the power provided to the Varistor is on the order of 100 Watt. However, since the fluorescent lamp is supposed to start within 500 milli-Second, the total cumulated energy dissipation in the Varistor is limited by the lamp to about 5 Joule.

That is, under normal conditions, current will flow through the Varistor for but a very brief period of time. Thereafter, the lamp starts and the Varistor in effect gets disconnected.

However, if the lamp is inoperative or not connected, the amount of energy that would be dissipated in the Varistor would rapidly exceed its energy-handling capability. In particular, for the parameters indicated above, the maximum energy capable of being absorbed by the Varistor would be reached in only 0.4 Second.

As long as current is flowing through the Varistor, it also flows through the primary winding CT3p of current-transformer CT3; which roughly implies that a corresponding output current can be obtained from the secondary winding CT3s. By way of rectifier HSR1, the positive component of this output current is used for charging energy-storing capacitor EC; which, after a brief period, accumulates a charge and develops a corresponding voltage. After this capacitor voltage has reached a magnitude high enough to cause the Diac D1 to break down, the accumulated charge on the capacitor is discharged into the base of transistor Qa—the magnitude of the discharge current being limited by the resistance of R2.

With a Diac breakdown voltage of about 30 Volt and a capacitance value of 33 uF for the energy-storing capacitor

EC, the amount of charge accumulated at the point of breakdown is about 1 milli-Coulomb. Thus, if the breakdown is to occur in a time period of about 250 milli-Second (which is chosen as being a suitable value), the magnitude of the current supplied to the capacitor would have to be about 10 milli-Amp; which is indeed what is approximately provided in the circuit of FIG. 1.

Now, as the Diac breaks down, the 1 milli-Coulomb charge on capacitor EC discharges into the base of Qa—limited mainly by the resistance of R2. With the Qa transistor being thusly switched into a conductive state, albeit for just a brief moment, a very low impedance path is provided between the base and the emitter of transistor Q2. As a result, the inverter feedback path is broken and the inverter stops oscillating.

And, of course, once it has stopped oscillating, the inverter will not restart until trigger pulses are provided by way of Diac D2; which pulses will be provided periodically due to the effect of resistor R3. Thus, after a predetermined period, the inverter will restart; but, except if now operating properly, it will be disabled again almost immediately.

However, the key aspect of the present invention is associated with the effects of elements R4, Z, T, R5 and D3.

Resistor R4 is operative to cause a current to flow through Zener diode Z, which has a Zenering-voltage of about 50 Volt, thereby establishing a substantially constant-magnitude 50 Volt DC voltage thereacross.

The 50 Volt DC voltage is voltage-divided by way of thermistor T and resistor R5; and the divided voltage is applied to energy-storing capacitor EC by way of diode D3. Thus, when the divided voltage reaches a magnitude of about 30 Volt, EC will eventually reach a voltage large enough to cause Diac D1 to break over, thereby disabling the inverter.

Due to the basic nature of a Thermistor, its resistance will decrease gradually though significantly with temperature; which means that the magnitude of the divided voltage will correspondingly increase with temperature.

Values of Thermistor T and resistor R5 are so chosen that the divided voltage will reach 30 Volt at a temperature of about 90 degrees Centigrade; which means that a disable signal will then be provided, thereby to disable the inverter.

After a predetermined time, the inverter will be triggered into oscillation again, but only to be disabled immediately except if the temperature has decreased substantially.

Additional Comments

(a) As the temperature of Thermistor T increases, the magnitude of the divided voltage increases correspondingly, thereby providing for a temperature-dependent voltage-bias on energy-storing capacitor EC. In turn, this implies that the time it takes for the Varistor current (indirectly) to charge capacitor EC to the point of Diac breakover will be shorter at higher temperatures—a feature that is generally advantageous: preventing the fluorescent lamp from getting ignited until the temperature has fallen substantially below the point at which the divided voltage is by itself is large enough to cause Diac breakover.

(b) In the circuit of FIG. 1, if fluorescent lamp FL were to be removed or otherwise fail to provide a proper loading for tank-capacitor C, the circuit will enter a mode whereby it will provide between its output terminals (i.e., the terminals of sockets S1 and S2) a high-frequency voltage that alternates at la relatively low frequency (e.g., one cycle per second) between being of a relatively high magnitude (e.g. 500 Volt RMS) and being of a relatively low magnitude (e.g., zero), spending a relatively brief amount of time (e.g., 100 milli-seconds) being at the relatively high magnitude and a relatively long time (e.g., 900 milli-seconds) being at the relatively low magnitude.

Thus, with the lamp removed from its sockets, the ballast output voltage is a high-frequency (e.g., 33 kHz) voltage amplitude modulated at a low frequency (e.g., 1 Hz); whereas, with the lamp in its sockets and drawing a proper amount of power, the ballast output voltage is a high-frequency voltage of substantially constant amplitude.

(c) The value of resistor R1 is such as to provide for a slow discharge of energy-storing capacitor C3.

In accurately calculating the voltage-division-ratio associated with the Thermistor and R5, it is necessary to consider the extra loading caused by leakage resistor R1.

In fact, in many cases, R5 may be eliminated and its effect be provided by a properly chosen R1.

(d) The RMS magnitude of the (high-frequency) voltage required to properly instant-start a fluorescent lamp is usually at least 3.3 times as high as that of the voltage developing across the lamp under normal full-power operation. Thus, an open circuit ballast output voltage of 500 Volt RMS would be appropriate for a lamp with normal full-power operating voltage of about 150 Volt RMS.

Thus, with a lamp having a full-power operating voltage of 150 Volt RMS, the circuit arrangement of FIG. 1 would appropriately have the following characteristics:

(i) with fluorescent lamp FL connected and drawing full power, the RMS magnitude of the voltage across lamp FL, and thereby between the socket terminals of sockets S1 and S2, is about 150 Volt RMS and substantially constant;

(ii) with lamp FL not connected, the RMS magnitude of the voltage between the socket terminals periodically varies between zero and 500 Volt RMS: being at zero magnitude for about 90% of the time and at 500 Volt RMS for about 10% of the time, thereby having an average RMS magnitude of about 50 Volt RMS;

(iii) which means that, with the lamp not connected, the high-frequency (33 kHz) voltage existing between the socket terminals is amplitude-modulated at a relatively low frequency (e.g., or 1 Hz), in that the amplitude of this high-frequency voltage periodically (at 1 Hz) alternates between having zero magnitude and having a 500 Volt RMS magnitude;

(iv) which further means that, with the lamp not connected, the RMS magnitude of the amplitude-modulated voltage present between the socket terminals—as measured over an integrating period of a complete modulation cycle—is the square root of: (the square of 500 Volt RMS divided by 10), which is the same as 500 Volt RMS divided by the square root of 10; which is to say, about 158 Volt RMS. In other words, in terms of net RMS value (i.e., net power generating effect), a voltage of 500 Volt RMS magnitude existing for 10% of the time during a period where the magnitude is zero during the remaining 90% of the period, is equivalent to a voltage of 158 Volt RMS existing throughout the whole period. Thus, since the magnitude of the voltage existing between the socket terminals when the lamp is connected is about 150 Volt RMS, it is noted that the long term RMS magnitude of the voltage existing between the socket terminals is about the same (i.e., 158 Volt RMS) when the lamp is not connected. However, by making the duty-cycle of the modulation a little smaller (e.g., more than about 12), the RMS magnitude of the voltage between the socket terminals could be made to be smaller when the lamp is not connected versus what it is when the lamp is connected.

Thus, with particular reference to FIG. 1 and Items (ii) through (iv) above, it is seen that, whenever the lamp is not connected, the voltage provided between the socket terminals is a high frequency AC voltage that is amplitude-modulated at a relatively low frequency (e.g., 1 Hz), where the relatively low frequency is low relative to the frequency of the high frequency (e.g., 33 kHz) AC voltage. In the particular embodiment herein disclosed, with the lamp not connected, the AC voltage between the socket terminals is represented by periodic bursts of high-frequency AC voltage.

Also, with particular reference to FIG. 1 and Items (i) and (ii) above, it is seen that: whenever the lamp is connected with the socket terminals, the AC voltage thereacross exhibits a first magnitude, namely a steady or non-varying RMS magnitude (e.g., 150 Volt) with a corresponding peak-to-peak magnitude (e.g., 424 Volt); and whenever the lamp is not so connected, the AC voltage across the lamp terminals exhibits a second magnitude. This second magnitude is characterized by varying between zero and a very high maximum RMS or peak-to-peak level (e.g., 500 Volt RMS or 1400 Volt peak-to-peak), with an average RMS magnitude far lower than the maximum level (e.g., an average of 50 Volt RMS). Thus, it is seen that the AC voltage across the lamp terminals exhibits a first peak-to-peak magnitude whenever the lamp is connected between the lamp terminals, and a second peak-to-peak magnitude whenever the lamp is not so connected, with the second peak-to-peak magnitude being distinctly higher than the first peak-to-peak magnitude. Further, it is seen that the RMS magnitude of the AC voltage between the lamp terminals is distinctly higher when the lamp is connected thereacross as compared with a situation when the lamp is not so connected.

(e) It is believed that the present invention and its several attendant advantages and features will be understood from the preceeding description. However, without departing from the spirit of the invention, changes may be made in its form and in the construction and interrelationships of its component parts, the form herein presented merely representing the presently preferred embodiment.

I claim:

1. An arrangement comprising:

a source operative to provide a DC voltage across a pair of DC terminals;

a lamp holder having a pair of lamp sockets operative to receive and hold a gas discharge lamp; the lamp holder having a pair of socket terminals; and

an assembly connected with the DC terminals and the socket terminals; the assembly being functional: (i) as long as the lamp is connected with the socket terminals, to deliver a lamp current thereto; and (ii) whenever the lamp is not so connected, to provide across the socket terminals a high frequency AC voltage that is amplitude-modulated at a low frequency; high frequency being defined as a frequency higher than 10 kHz; low frequency being defined as a frequency lower than 100 Hz; the assembly being further characterized by not including a thermal circuit breaker functional to make and break connection between two terminals by way of a thermally activated mechanical contactor means.

2. The arrangement of claim 1 wherein the assembly is further characterized by including a control sub-assembly operative, on receipt of a first control action at a control

action input, to cause a substantial reduction of the magnitude of any voltage present across the socket terminals.

3. The arrangement of claim 2 wherein the assembly is still further characterized by including a restoring sub-assembly operative, whenever the magnitude of any voltage present across the socket terminals has remained below a certain low level for longer than a certain length of time, to cause a substantial increase in this magnitude.

4. The arrangement of claim 1 wherein the assembly is further characterized by including a control sub-assembly operative, on receipt of a control action at a control action input, to stop delivery of the lamp current even under a condition when the lamp is connected with its socket terminals.

5. The arrangement of claim 4 wherein the control sub-assembly is further characterized by including a temperature sensing sub-circuit operative to supply said control action in case a temperature associated with the assembly were to exceed a predetermined level.

6. An arrangement comprising:

a source operative to provide a DC voltage across a pair of DC terminals;

a lamp holder having a pair of lamp sockets operative to receive and hold a gas discharge lamp; the lamp holder having a pair of socket terminals; and

an assembly connected in circuit with the DC terminals and the socket terminals; the assembly being functional to provide an AC output voltage to the socket terminals; the assembly including a control sub-assembly connected in circuit therewithin and being operative to affect the RMS magnitude of the AC voltage in response to a control action received at a control action input; the assembly being further characterized by being operative, whenever the lamp is connected with the socket terminals, to cause a lamp current to flow through the lamp and to cause the RMS magnitude of the AC voltage to assume a first average level; and (ii) whenever the lamp is not connected with the socket terminals, to cause provision of said control action to the control action input, thereby to cause the RMS magnitude of the AC voltage to assume a second average level; the second average level being distinctly lower than the first average level.

7. The arrangement of claim 6 wherein the control sub-assembly is further characterized by causing, whenever the lamp is disconnected from the socket terminals, the AC voltage to be amplitude-modulated at a low frequency; the low frequency being lower than the frequency of the AC voltage by a factor higher than one hundred.

8. The arrangement of claim 7 wherein, whenever the lamp is disconnected from the socket terminals, the AC voltage comprises periodic bursts of high-frequency AC voltage; each burst of high-frequency AC voltage having a certain high-level RMS magnitude and being followed by a period during which the RMS magnitude of any voltage then present at the socket terminals is substantially lower than said high-level RMS magnitude.

9. The arrangement of claim 8 wherein the assembly is further characterized in that whenever the lamp is: (i) connected with the socket terminals, the AC voltage exhibits a first peak-to-peak magnitude; and (ii) not connected with the socket terminals, the AC voltage exhibits a second peak-to-peak magnitude; the second peak-to-peak magnitude being distinctly higher than the first peakto-peak magnitude.

10. An arrangement comprising:

source means operative to provide a DC voltage across a pair of DC terminals;

a lamp holder having a pair of lamp sockets operative to receive and hold a gas discharge lamp; the lamp holder having a pair of socket terminals; and

inverter power supply means connected in circuit between the DC terminals and the socket terminals; the inverter power supply means: (i) being operable to provide an inverter output voltage between the socket terminals; (ii) being operable to properly power the gas discharge lamp as long as this gas discharge lamp is indeed being held by the lamp sockets; and (iii) having control means functional to control the magnitude of the inverter output voltage such as to prevent its RMS magnitude from exceeding a predetermined level; the control means being further characterized by not including a thermal circuit breaker functional to make and break connection between two terminals by way of a thermally activated mechanical contactor means.

11. An arrangement comprising:

rectifier means adapted to connect with an ordinary electric utility power line and, when so connected, to provide a DC voltage at a DC output;

self-oscillating inverter means connected with the DC output and, except after having been provided with a disable signal at a disable input, operable to convert the DC voltage to a high frequency voltage provided at an inverter output;

connect and matching means connected with the inverter output and operative to connect with a gas discharge lamp means, thereby to properly power that lamp means; and

temperature sensor means responsive to a temperature associated with the inverter means; the temperature sensor means being connected with the disable input and operative, whenever the temperature exceeds a predetermined level, to provide the disable signal; the temperature sensor means being further characterized by including an element whose electrical characteristics changes gradually but substantively as a result of gradual changes in said temperature;

restart means connected with the inverter means and operative, after a predetermined period, to cause the inverter means to resume converting the DC voltage to the high frequency voltage;

thereby, whenever the temperature has exceeded a predetermined level: (i) to disable the inverter means, therefore preventing it from converting the DC voltage to the high frequency voltage; (ii) substantially to cease supplying power to the lamp means; and (iii) substantially to stop the rectifier means from drawing power from the power line; and (iv) after the predetermined period, to cause the inverter means to attempt but fail to resume supplying power to the lamp means.

12. An arrangement comprising:

rectifier connected with an ordinary electric utility power line and operative to provide a DC voltage at a DC output;

self-oscillating inverter connected with the DC output and, except after having been provided with a disable signal at a disable input, operative to provide a high frequency voltage at an inverter output;

connect and matching means connected with the inverter output and operative to connect with a gas discharge lamp means, thereby to provide power to the lamp means;

temperature sensor means responsive to a temperature associated with the inverter and operative whenever that temperature exceeds a predetermined level, to provide the disable signal, thereby to disable the inverter and to cease to provide power to the lamp means; the temperature sensor means being further characterized by including an element whose electrical characteristics changes gradually but substantively as a result of gradual changes in said temperature; and

restart means connected in circuit with the inverter and operative to cause the inverter to re-start its operation some time after it has been disabled.

13. An improvement for a power supply for a lamp; the power supply comprising self-oscillating inverter connected with a DC voltage and operable to convert the DC voltage to a high frequency voltage, therewith to power the lamp; the improvement comprising:

disable means connected with the self-oscillating inverter; the disable means comprising an element whose electrical characteristics changes gradually but substantively as a result of gradual changes in a temperature associated with the inverter; the disable means being operable, whenever the temperature exceeds a predetermined level, to disable the self-oscillating inverter and thereby to prevent it from converting the DC voltage to the high frequency voltage; the disable means including automatic restart circuitry operative to cause the inverter to re-initiate its oscillation and to resume providing power to the lamp.

14. An arrangement comprising:

a source operative to provide a DC voltage of substantially constant magnitude across a pair of DC terminals;

a lamp holder having a pair of lamp sockets operative to receive and hold a gas discharge lamp; the lamp holder having a pair of socket terminals; and

an assembly connected in circuit with the DC terminals as well as with the socket terminals; the assembly being operative, whenever the lamp is not connected with the socket terminals, to provide thereacross an AC voltage amplitude-modulated at a frequency no higher than one hundredth that of the AC voltage; the assembly being further characterized by not including a thermal circuit breaker functional to make and break connection between two terminals by way of a thermally activated contactor means.

15. An arrangement comprising:

a source operative to provide a DC voltage of substantially constant magnitude across a pair of DC terminals;

a lamp holder having a pair of lamp sockets operative to receive and hold a gas discharge lamp; the lamp holder having a pair of socket terminals; and

an assembly connected in circuit with the DC terminals as well as with the socket terminals; the assembly being operative: (i) whenever the lamp is connected with the socket terminals, to supply to the lamp an alternating current of substantially constant amplitude; and (ii) whenever the lamp is not so connected, to provide across the socket terminals an AC voltage whose RMS magnitude varies periodically at a relatively low frequency; the frequency of the AC voltage being at least one hundred times higher than that of the relatively low frequency; the assembly being further characterized by not including a thermal circuit breaker functional to make and break connection between two terminals by way of a thermally activated contactor means.

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EXHIBIT I

US005757140A

United States Patent [19]

Nilssen

[11] Patent Number: 5,757,140

[45] Date of Patent: May 26, 1998

[54] **ELECTRONIC BALLAST WITH FREQUENCY CONTROL**

[76] Inventor: **Ole K. Nilssen**, Caesar Dr., Barrington, Ill. 60010

[21] Appl. No.: **851,887**

[22] Filed: **Mar. 16, 1992**

Related U.S. Application Data

[63] Continuation of Ser. No. 819,655, Jan. 13, 1992, Pat. No. 5,191,262, which is a continuation of Ser. No. 643,023, Jan. 18, 1991, abandoned, which is a continuation-in-part of Ser. No. 787,692, Oct. 15, 1985, abandoned, which is a continuation of Ser. No. 644,155, Aug. 27, 1984, abandoned, which is a continuation of Ser. No. 555,426, Nov. 23, 1983, abandoned, which is a continuation of Ser. No. 178,107, Aug. 14, 1980, abandoned, said Ser. No. 555,426, Nov. 23, 1983, is a continuation-in-part of Ser. No. 330,159, Dec. 14, 1981, Pat. No. 4,430,628, which is a division of Ser. No. 973,741, Dec. 28, 1978, abandoned, which is a continuation-in-part of Ser. No. 890,586, Mar. 20, 1978, Pat. No. 4,184,128.

[51] **Int. Cl.^{6}** **H05B 41/36**

[52] **U.S. Cl.** .... **315/209 R**; 315/219; 315/DIG. 4

[58] **Field of Search** .... 315/209 R, 219, 315/DIG. 4, DIG. 5, DIG. 7, 220, 224, 240, 53, 276, 278, 279, 307; 331/113 A

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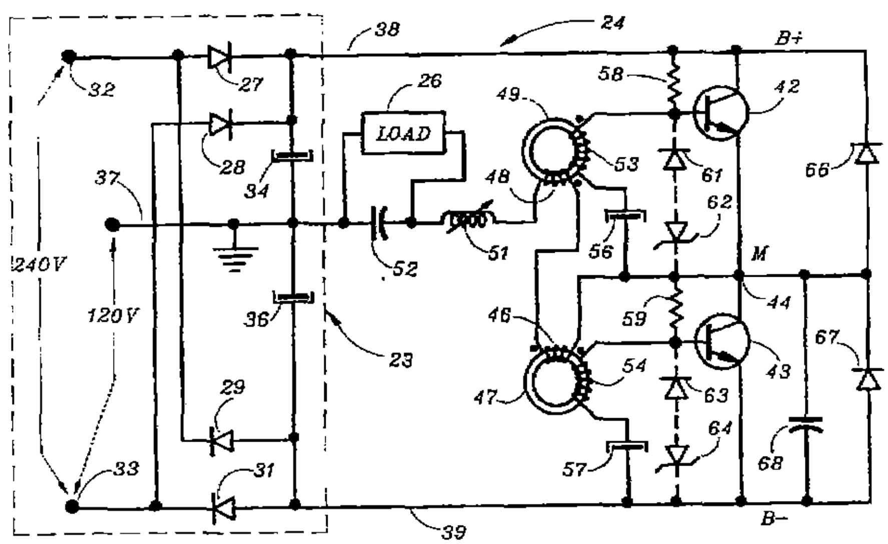
Primary Examiner—Ali Neyzari

[57] **ABSTRACT**

A half-bridge inverter is powered from a constant DC voltage and loaded by way of an L-C circuit that has a natural resonance frequency equal to or lower than the inverter's operating frequency. A gas discharge lamp is disconnectably connected across the tank capacitor of the L-C circuit and, when indeed so connected, is provided with a current-limited high-frequency (30 kHz) voltage. The magnitude of the resulting lamp current decreases as the inverter's operating frequency is increased; which operating frequency is controlled by way of a negative feedback arrangement that causes the frequency to increase as a function of increasing magnitude of the current flowing through the tank capacitor. Thus, particularly with the lamp disconnected, the magnitude of the current flowing through the capacitor—and therefore also the magnitude of the voltage existing across it—will be regulated so as to be lower than it would be in the absence of the negative feedback.

29 Claims, 5 Drawing Sheets



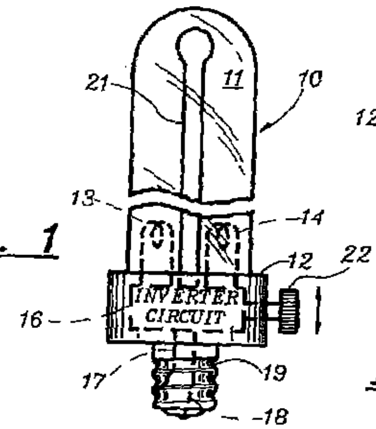


Fig. 1

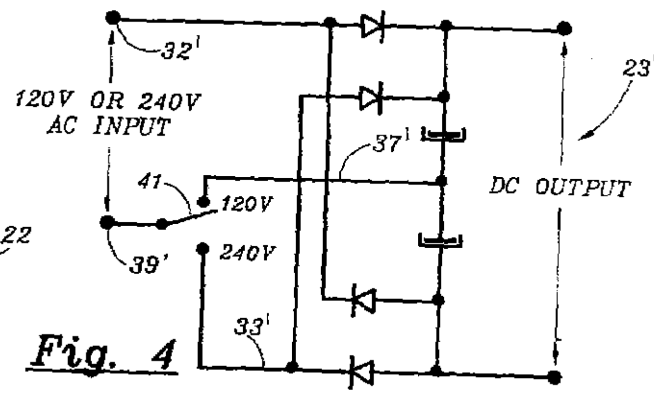


Fig. 4

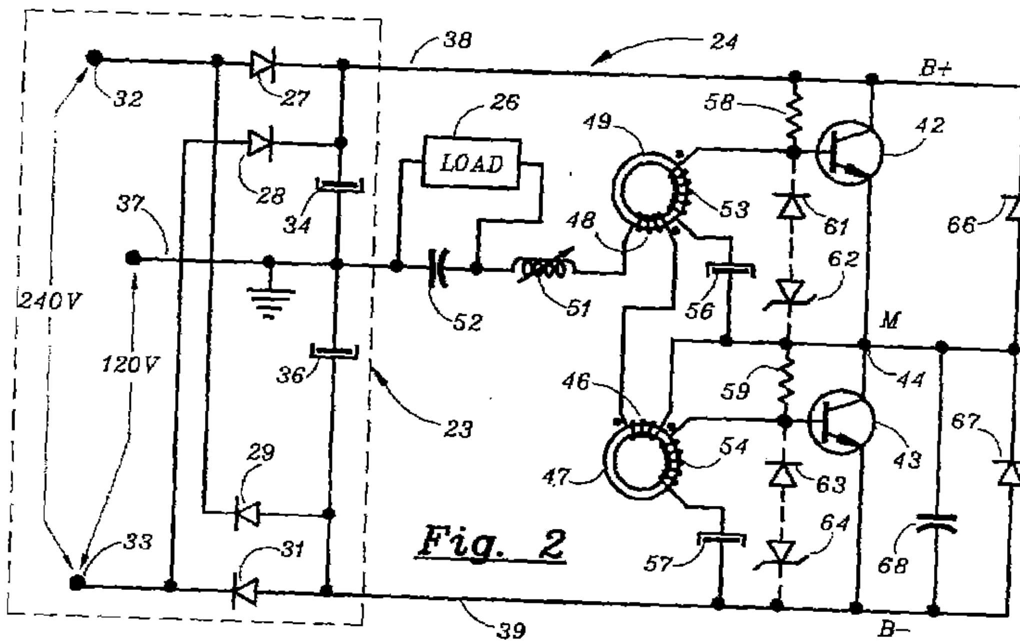


Fig. 2

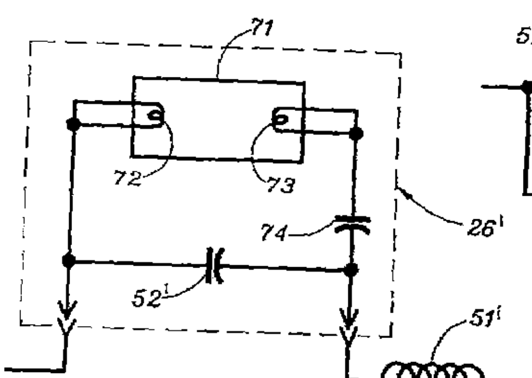


Fig. 5

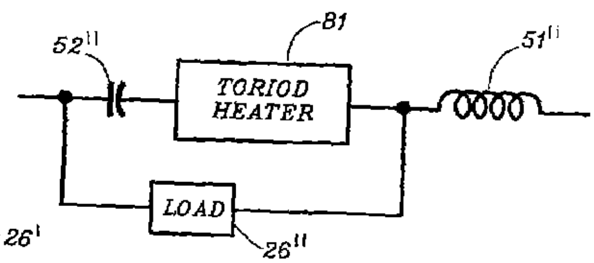


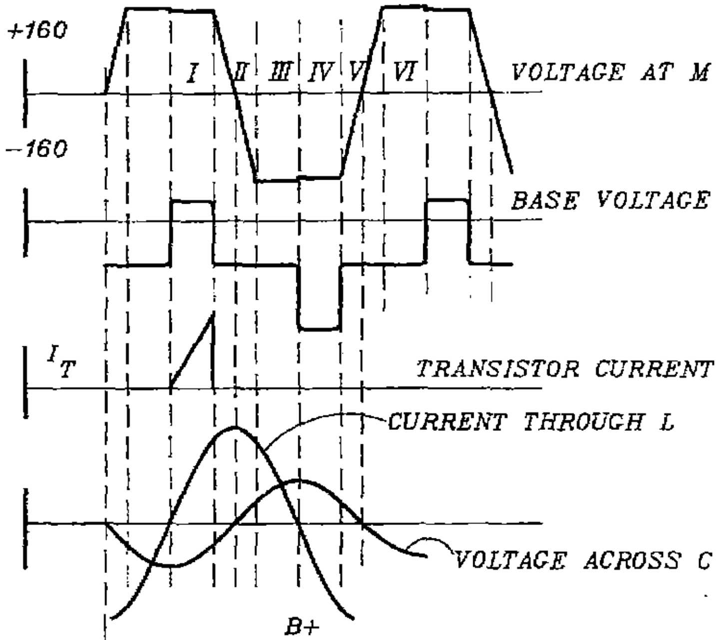
Fig. 6

Fig. 3A

Fig. 3B

Fig. 3C

Fig. 3D



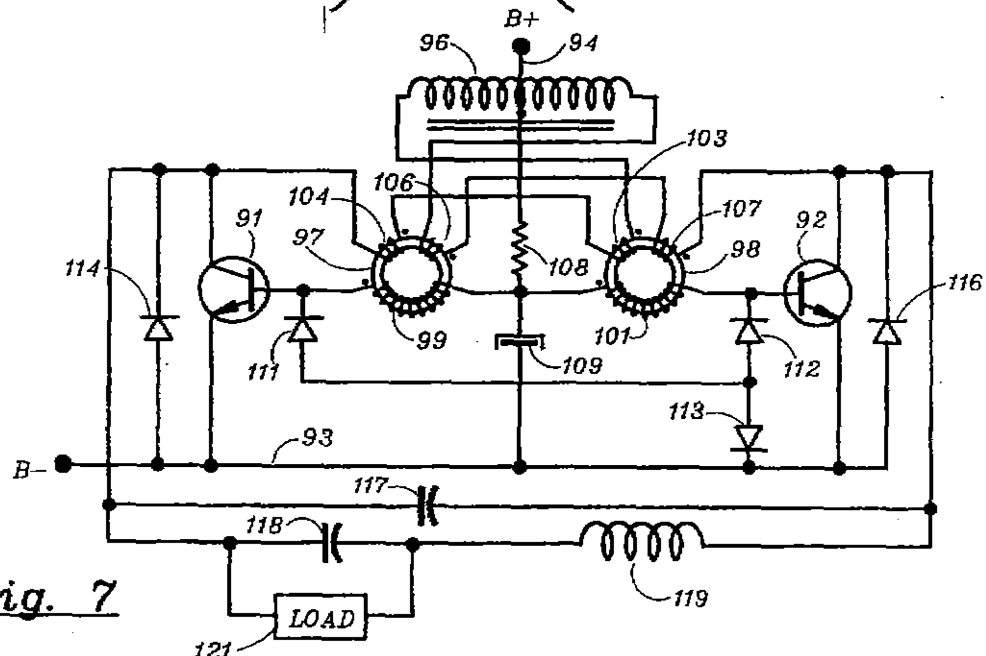


Fig. 7

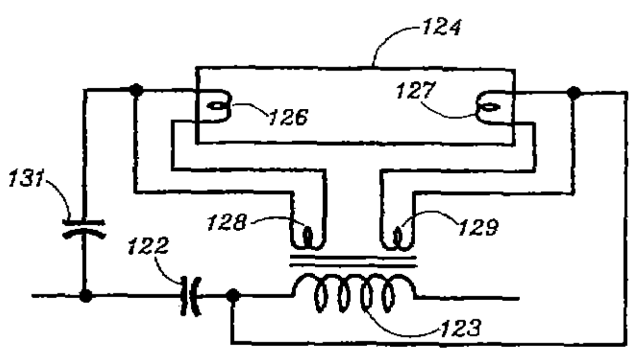


Fig. 8

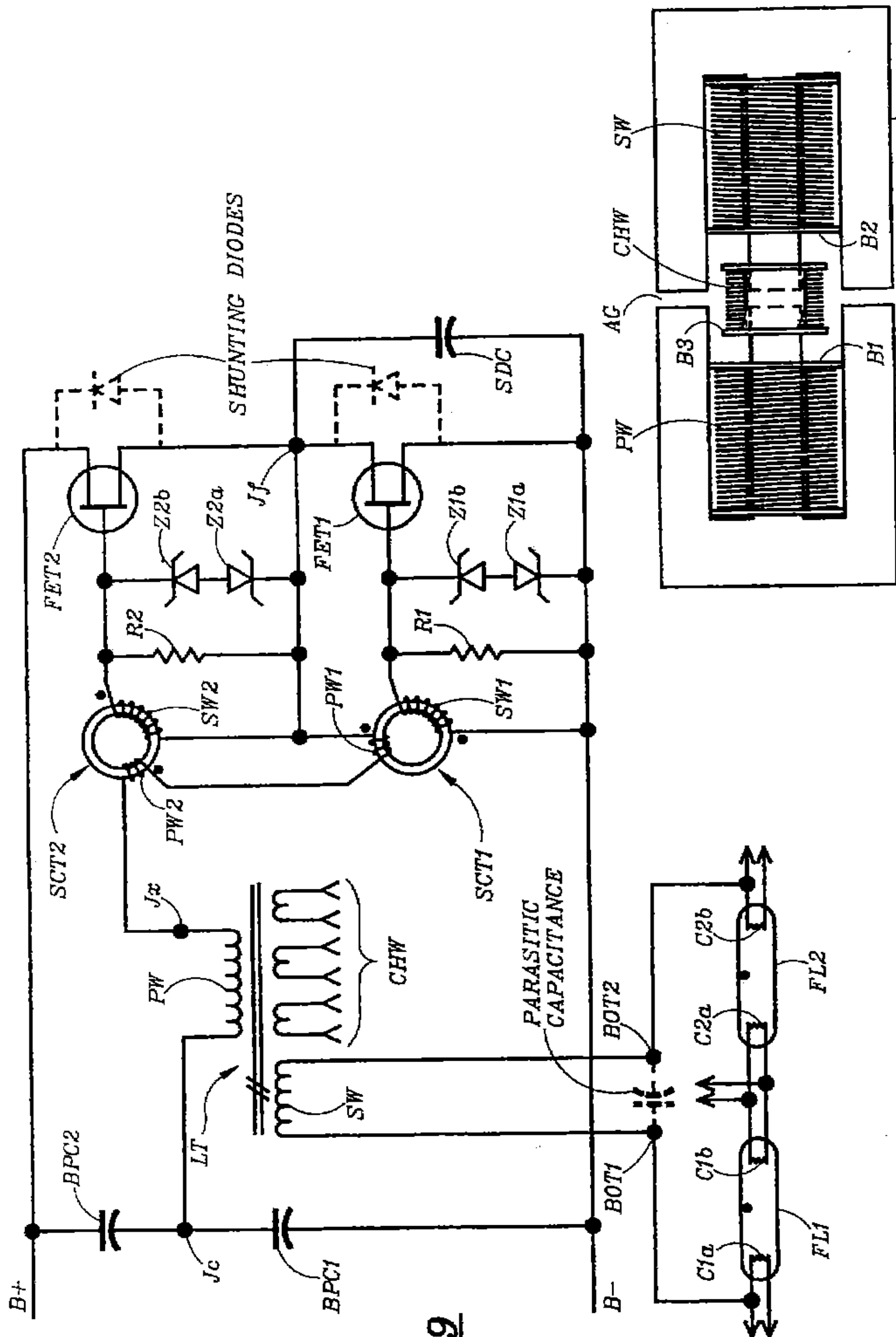

SW
CHW
AG
B3
PW
FC2
B2
B1
FC1
SHUNTING DIODES
SDC
FET2
Z2b
Z2a
Jf
FET1
Z1b
Z1a
R2
R1
SW2
PW1
SW1
SCT2
PW2
SCT1
Jx
CHW
PW
PARASITIC CAPACITANCE
BOT2
SW
LT
BPC2
BOT1
B+
Jc
BPC1
B–
C2b
FL2
C2a
C1b
C1a
FL1

Fig. 10

Fig. 9

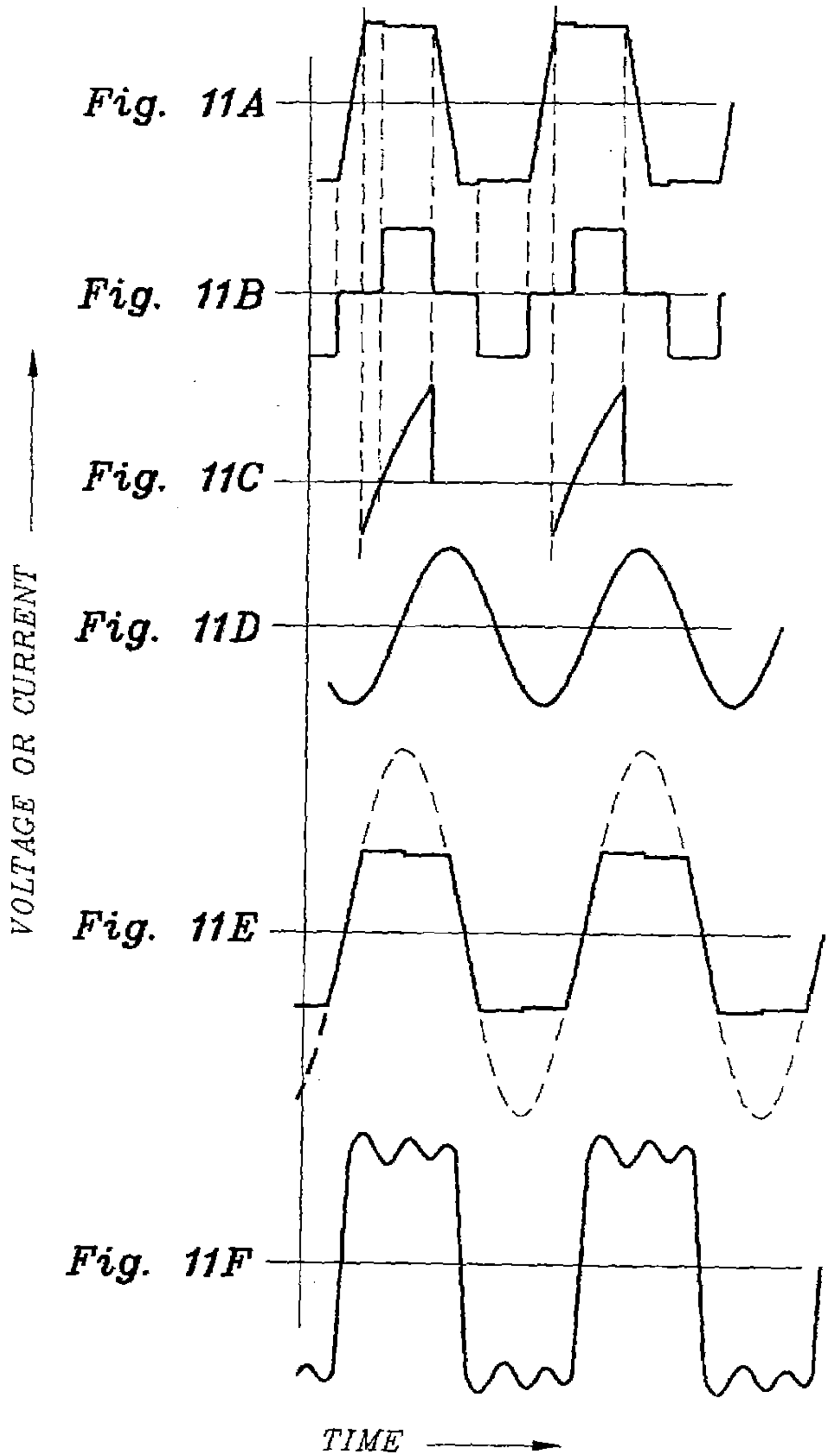
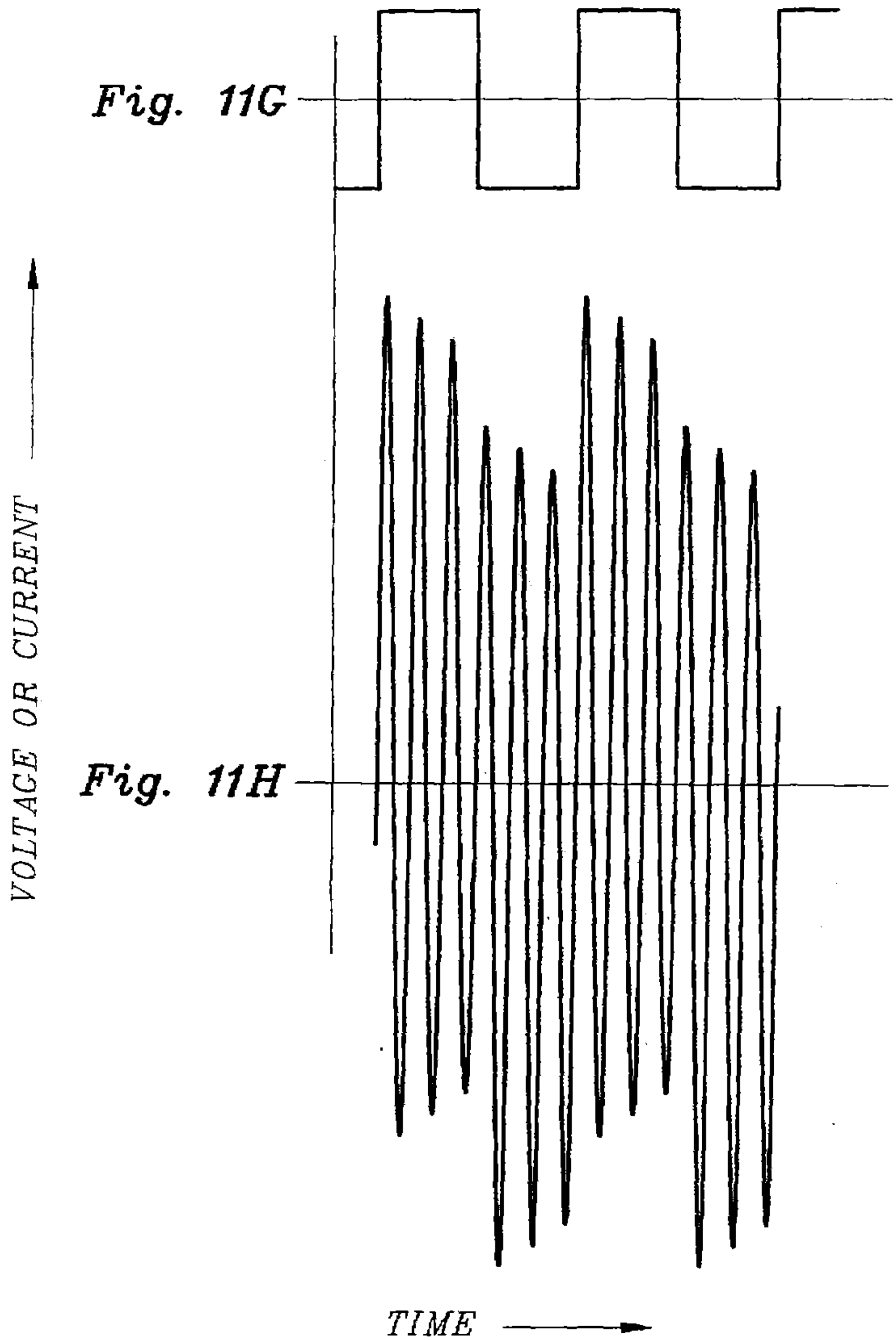


Fig. 11A
Fig. 11B
Fig. 11C
Fig. 11D
Fig. 11E
Fig. 11F
VOLTAGE OR CURRENT
TIME

Fig. 11G

Fig. 11H



ELECTRONIC BALLAST WITH FREQUENCY CONTROL

RELATED APPLICATIONS

The present application is a Continuation of Ser. No. 07/819,655 filed Jan. 13, 1992 now U.S. Pat. No. 5,191,262; which is a Continuation of Serial No. 07/643,023 filed Jan. 18, 1991 now abandoned; which is a Continuation-in-Part of Ser. No. 06/787,692 filed Oct. 15, 1985 now abandoned; which is a Continuation of Ser. No. 06/644,155 filed Aug. 27, 1994, now abandoned; which is a Continuation of Ser. No. 06/555,426 filed Nov. 23, 1983, now abandoned; which is a Continuation of Ser. No. 06/178,107 filed Aug. 14, 1980, now abandoned; which Ser. No. 06/555,426 is also a Continuation-in-Part of Ser. No. 06/330,159 filed Dec. 14, 1981, now U.S. Pat. No. 4,430,628; which is a Division of Ser. No. 05/973,741 filed Dec. 28, 1978, now abandoned; which is a Continuation-in-Part of Ser. No. 05/890,586 filed Mar. 20, 1978, now U.S. Pat. No. 4,184,128.

BACKGROUND OF THE INVENTION

1. Field of Invention

This invention relates to ballasting means for gas discharge lighting means.

2. Description of Prior Art

For a description of pertinent prior art, reference is made to U.S. Pat. No. 4,677,345 to Nilssen; which patent issued from a Division of application Ser. No. 06/178,107 filed Aug. 14, 1980; which application is the original in-part progenitor of instant application. Also, reference is made to U.S. Pat. No. 3,263,122 to Genuit; No. 3,320,510 to Locklair; No. 3,996,493 to Davenport et al.; No. 4,100,476 to Ghiringhelli; No. 4,262,327 to Kovacik et al.; No. 4,370,600 to Zansky; No. 4,634,932 to Nilssen; and No. 4,857,806 to Nilssen.

SUMMARY OF THE INVENTION

Objects of the Invention

A main object of the present invention is that of providing a cost-effective ballasting means for gas discharge lamps.

This as well as other objects, features and advantages of the present invention will become apparent from the following description and claims.

Brief Description of the Invention

A half-bridge inverter is powered from a constant DC voltage and loaded by way of an L-C circuit that has a natural resonance frequency equal to or lower than the inverter's operating frequency. A gas discharge lamp is disconnectably connected across the tank capacitor of the L-C circuit and, when indeed so connected, is provided with a current-limited high-frequency (30 kHz) voltage. The magnitude of the resulting lamp current decreases as the inverter's operating frequency is increased; which operating frequency is controlled by way of a negative feedback arrangement that causes the frequency to increase as a function of increasing magnitude of the current flowing through the tank capacitor. Thus, particularly with the lamp disconnected, the magnitude of the current flowing through the capacitor—and therefore also the magnitude of the voltage existing across it—will be regulated so as to be lower than it would be in the absence of the negative feedback.

Otherwise, in the form of another embodiment, in an electronic inverter-type ballast for a gas discharge lamp, a basic and significant problem associated with powering the lamp by way of a current-limiting inductance means more-or-less directy from the inverter's high-frequency (e.g. 30 kHz) squarewave voltage (as opposed to first shaping this squarewave voltage into a sinusoidal voltage by way of a tuned circuit) is that of spurious resonances occurring due to resonant interactions (at harmonic components of the squarewave voltage) between the effective output inductance represented by the current-limiting inductance means and the unavoidable stray capacitance associated with the output wiring means used for connecting between the ballast's output and the lamp. This problem is mainly significant during periods of open circuit operation (such as prior to lamp ignition); but during those periods, the spurious resonances are apt to cause excessive power dissipations within the ballast, thereby potentially causing damage to the ballast. Since the particular capacitance value associated with the output wiring means is an unknown—being dependent on some unknown end-use situation—it is not feasible in a straight forward manner simply to tune the ballast output inductance and/or the inverter's operating frequency such as to avoid these spurious resonances.

Of course, the reason these spurious resonances occur in the first place is that the inverter's squarewave voltage contains a substantial amount of odd harmonic components. In particular, it contains one third (i.e. 33.3%) third harmonics, one fifth (i.e. 20.0%) fifth harmonics, etc.

The usual approach to avoiding the above-mentioned problem of uncontrollable spurious resonances is that of powering the lamp by way of a tuned circuit resonantly tuned to the fundamental component of the inverter's squarewave voltage; and, as a result of this tuning, the problems associated with the harmonic components are substantially eliminated.

In an initial preferred ambodiment of the present invention, a half-bridge inverter is powered from a constant DC voltage and provides an AC output voltage that is—in contrast with the usual squarewave voltage—describable as being a sinusoidal waveform with the tops clipped off at some fixed magnitude; or, described differently, a waveform composed of truncated sinusoidal waves; or, described still differently, a waveform having trapezoidally shaped half-cycles. This AC voltage is applied across the primary winding of a so-called reactance transformer, whose loosely coupled secondary winding is connected across a gas discharge lamp. The internal inductive reactance of the secondary winding constitutes a lamp ballasting means by way of limiting the magnitude of the resulting lamp current to a pre-established desired level. Potentially damaging spurious or parasitic resonances—which are very likely to occur under actual operational circumstances with an unloaded secondary winding when the primary winding is supplied with a squarewave voltage—are avoided because of the truncated sinusoidal waveshape of the AC voltage; which truncated sinusoidal shape is efficiently attained by a combination of three factors: (i) using rapidly switching transistors in the inverter; (ii) having the transformer's primary winding exhibit a substantial shunt inductance; and (iii) providing for a slow-down capacitor coupled directly across the primary winding, thereby to substantially slow down the rise time of the inverter's output voltage as conpared with what it would have been if it were to have been determined solely by the high switching speed of the transistors.

Otherwise and more generally, the present invention is directed to providing improved gas discharge lighting means and inverter circuits for powering and controlling gas discharge lamps. The inverter circuits according to the present invention are highly efficient, can be compactly constructed and are ideally suited for energizing gas discharge lamps, particularly compact folded "instant-start" "self-ballasted" fluorescent lamps.

According to one feature of the present invention, a series-connected combination of an inductor and a capacitor is provided in circuit with the inverter transistors to be energized upon periodic transistor conduction. Transistor drive current is preferably provided through the use of at least one saturable inductor to control the transistor inversion frequency to be equal to or greater than the natural resonant frequency of the inductor and capacitor combination. The high voltages efficiently developed by loading the inverter with the inductor and capacitor are ideally suited for energizing external loads such as gas discharge lamps. In such an application, the use of an adjustable inductor permits control of the inverter output as a means of adjusting the level of lamp illumination.

According to another feature of the present invention, reliable and highly efficient half-bridge inverters include a saturable inductor in a current feedback circuit to drive the transistors for alternate conduction. The inverters also include a load having an inductance sufficient to effect periodic energy storage for self-sustained transistor inversion. Importantly, improved reliability is achieved because of the relatively low and transient-free voltages across the transistors in these half-bridge inverters.

Further, according to another feature of the present invention, novel and economical power supplies particularly useful with the disclosed inverter circuits convert conventional AC input voltages to DC for supplying to the inverters.

Yet further, according to still another feature of the invention, a rapid-start fluorescent lamp is powered by way of a series-resonant LC circuit; while heating power for the lamp's cathodes is provided via loosely-coupled auxiliary windings on the tank inductor of the LC circuit. Alternatively, cathode heating power is provided from tightly-coupled windings on the tank inductor; in which case output current-limiting is provided via a non-linear resistance means, such as an incandescent filament in a light bulb, connected in series with the output of each winding.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 is a front elevation of a folded fluorescent lamp unit adapted for screw-in insertion into a standard Edison incandescent socket;

FIG. 2 is a schematic diagram illustrating the essential features of a push-pull inverter circuit particularly suitable for energizing the lamp unit of FIG. 1;

FIG. 3A–3D is a set of waveform diagrams of certain significant voltages and currents occurring in the circuit of FIG. 2;

FIG. 4 is a schematic diagram of a DC power supply connectable to both 120 and 240 volt AC inputs;

FIG. 5 is a schematic diagram which illustrates the connection of a non-self-ballasted gas discharge lamp unit to the FIG. 2 inverter circuit;

FIG. 6 is a schematic diagram which illustrates the use of a toroid heater for regulation of the inverter output;

FIG. 7 is an alternate form of push-pull inverter circuit accordind to the present invention;

FIG. 8 is a schematic diagram showing the connection of a gas discharge lamp of the "rapid-start" type to an inductor-capacitor-loaded inverter according to the present invention;

FIG. 9 is a schematic diagram illustrating an inverter ballast circuit arrangement wherein a pair of series-connected fluorescent lamps is powered, by way of a reactance transformer, from an inverter output voltage having a trapezoidal (i.e. truncated sinewave) waveform like that of FIG. 3A.

FIG. 10 is a schematic illustration of the reactance transformer used in the circuit arrangement of FIG. 9.

FIG. 11A–11H show various voltage and current waveforms associated with the circuit arrangement of FIG. 9.

DESCRIPTION OF THE PREFERRED EMBODIMENTS

FIG. 1 illustrates a screw-in gas discharge lamp unit **10** comprising a folded fluorescent lamp **11** suitably secured to an integral base **12**. The lamp comprises two cathodes **13**, **14** which are supplied with the requisite high operating voltage from a frequency-converting power supply and ballasting circuit **16**; which, because of its compact size, conveniently fits within the base **12**. The inverter circuit **16** is connected by leads **17**, **18** to a screw-type plug **19** adapted for screw-in insertion into a standard Edison-type incandescent lamp socket at which ordinary 120Volt/60Hz power line voltage is available. A ground plane comprising a wire or metallic strip **21** is disposed adjacent a portion of the fluorescent lamp **11** as a starting aid. Finally, a manually rotatable external knob **22** is connected to a shaft for mechanical adjustment of the air gap of a ferrite core inductor to vary the inductance value thereof in order to effect adjustment of the inverter voltage output connected to electrodes **13**, **14** for controlled variation of the lamp illumination intensity.

With reference to FIG. 2, a power supply **23**, connected to a conventional AC input, provides a DC output for supplying a high-efficiency inverter circuit **24**. The inverter is operable to provide a high voltage to an external load **26**, which may comprise a gas discharge device sich as the fluorescent lamp **11** of FIG. 1.

The power supply **23** comprises bridge rectifier having four diodes **27**, **28**, **29** and **31** connectable to a 240 volt AC supply at terminals **32**, **33**. Capacitors **34**, **36** are connected between a ground line **37** (in turn directly connected to the inverter **24**) and to a B+ line **38** and a B− line **39**, respectively. The power supply **23** also comprises a voltage doubler and rectifier optionally connectable to a 120 volt AC input taken between the ground line **37** and terminal **33** or **32**. The voltage doubler and rectifier means provides a direct electrical connection by way of line **37** between one of the 120 volt AC power input lines and the inverter **24**, as shown in FIG. 2. The bridge rectifier and the voltage doubler and rectifier provide substantially the same DC output voltage to the inverter **24** whether the AC input is 120 or 240 volts. Typical voltages are +160 volts on the B+ line **38** and −160 volts on the B− line **39**.

With additional reference to FIG. **4**, which shows an alternate power supply **23'**, the AC input, whether 120 or 240 volts, is provided at terminals **32'** and **39**. Terminal **39** is in turn connected through a single-pole double-throw selector switch **41** to terminal **37'** (for 120 volt operation) or terminal **33'** (for 240 volt operation) . In all other respects, power supplies **23** and **23'** are identical.

The inverter circuit **24** of FIG. **2** is a half-bridge inverter comprising transistors **42**, **43** connected in series across the DC voltage output of the power supply **23** on B+ and B− lines **38** and **39**, respectively. The collector of trasistor **42** is connected to the B+ line **38**, the emitter of transistor **42** and the collector of transistor **43** are connected to a midpoint line

44 (designated "M") and the emitter of transistor 43 is connected to the B− line 39. The midpoint line 44 is in turn connected to the ground line 37 through primary winding 46 of a toroidal saturable core transformer 47, a primary winding 48 on an identical transformer 49, an inductor 51 and a series-connected capacitor 52. The inductor 51 and capacitor 52 are energized upon alternate transistor conduction in a manner to be described later.

An external load 26 is preferably taken off capacitor 52, as shown in FIG. 2. The inductor 51, preferably a known ferrite core inductor, has an inductance variable by mechanical adjustment of the air gap in order to effect variation in the level of the inductor and capacitor voltage and hence the power available to the load, as will be described. When the load is a gas discharge lamp such as lamp 11 in FIG. 1, variation in this inductance upon rotation of knob 22 accomplishes a lamp dimming effect.

Drive current to the base terminals of transistors 42 and 43 is provided by secondary windings 53, 54 of transformers 49, 47, respectively. Winding 53 is also connected to midpoint lead 44 through a bias capacitor 56, while winding 54 is connected to the B− lead 39 through an identical bias capacitor 57. The base terminals of transistors 42 and 43 are also connected to lines 38 and 44 through bias resistors 58 and 59, respectively. For a purpose to be described later, the base of transistor 42 can be optionally connected to a diode 61 and a series Zener diode 62 in turn connected to the midpoint line 44; similarly, a diode 63 and series Zener diode 64 in turn connected to the B− line 39 can be connected to the base of transistor 43. Shunt diodes 66 and 67 are connected across the collector-emitter terminals of transistors 42 and 43, respectively. Finally, a capacitor 68 is connected across the collector-emitter terminals of transistor 43 to restrain the rate of voltage rise across those terminals, as will be seen presently.

The operation of the circuit of FIG. 2 can best be understood with additional reference to FIG. 3, which illustrates significant portions of the waveforms of the voltage at midpoint M (FIG. 3A), the base-emitter voltage on transistor 42 (FIG. 3B), the current through transistor 42 (FIG. 3C), and the capacitor 52 voltage and the inductor 51 current (FIG. 3D).

Assuming that transistor 42 is first to be triggered into conduction, current flows from the B+ line 38 through windings 46 and 48 and the inductor 51 to charge capacitor 52 and returns through capacitor 34 (refer to the time period designated I in FIG. 3). When the saturable inductor 49 saturates at the end of period I, drive current to the base of transistor 42 will terminate, causing voltage on the base of the transistor to drop to the negative voltage stored on the bias capacitor 56 in a manner to be described, causing this transistor to become nonconductive. As shown in FIG. 3c, current-flow in transistor 43 terminates at the end of period I.

Because the current through inductor 51 cannot change instantaneously, current will flow from the B− bus 39 through capacitor 68, causing the voltage at midpoint line 44 to drop to −160 volts (period II in FIG. 3). The capacitor 68 restrains the rate of voltage change across the collector and emitter terminals of transistor 42. The current through the inductor 51 reaches its maximum value when the voltage at the midpoint line 44 is zero. During period III, the current will continue to flow through inductor 51 but will be supplied from the B− bus through the shunt diode 67. It will be appreciated that during the latter half of period II and all of period III, positive current is being drawn from a negative

voltage; which, positive current is being drawn from a negative voltage; which, in reality, means that energy is being returned to the power supply through a path of relatively low impedance.

When the inductor current reaches zero at the start of period IV, the current through the primary winding 46 of the saturable inductor 47 will cause a current to flow out of its secondary winding 54 to cause transistor 43 to become conductive, thereby causing a reversal in the direction of current through inductor 51 and capacitor 52. When transformer 47 saturates at the end of period IV, the drive current to the base of transistor 43 terminates and the current through inductor 51 will be supplied through capacitor 68, causing the voltage at midpoint line 44 to rise (period V). When the voltage at the midpoint line M reaches 160 volts, the current will then flow through shunt diode 66 (period VI). The cycle is then repeated.

As seen in FIG. 3, saturable transformers 47, 49 provide transistor drive current only after the current through inductor 51 has diminished to zero. Further, the transistor drive current is terminated before the current through inductor 51 has reached its maximum amplitude. This coordination of base drive current and inductor current is achieved because of the series-connection between the inductor 51 and the primary windings 46, 48 of saturable transformers 47, 49, respectively.

The series-connected combination of the inductor 51 and the capacitor 52 is energized upon the alternate conduction of transistors 42 and 43. With a large value of capacitance of capacitor 52, very little voltage will be developed across its terminals. As the value of this capacitance is decreased, however, the voltage across this capacitor will increase. As the value of the capacitor 52 is reduced to achieve resonance with the inductor 51, the voltage on the capacitor will rise and become infinite in a loss-free circuit operating under ideal conditions.

It has been found desirable to regulate the transistor inversion frequency, determined mainly by the saturation time of the saturable inductors 47, 49, to be equel to or higher than the natural resonance frequency of the inductor and capacitor combination in order to provide a high voltage output to external load 26. A high voltage across capacitor 52 is efficiently developed as the transistor inversion frequency approaches the natural resonant frequency of the inductor 51 and capacitor 52 combination. Stated another way, the conduction period of each transistor is desirably shorter in duration than one quarter of the full period corresponding to the natural resonant frequency of the inductor and capacitor combination. When the inverter 24 is used with a self-ballasted gas discharge lamp unit, it has been found that the inversion frequency can be at least equal to the natural resonant frequency of the tank circuit. If the capacitance value of capacitor 52 is reduced still further beyond the resonance point, unacceptably high transistor currents will be experienced during transistor switching and transistor burn-out will occur.

It will be appreciated that the sizing of capacitor 52 is determined by the application of the inverter circuit 24. Variation in the values of the capacitor 52 and the inductor 51 will determine the voltages developed in the inductor-capacitor tank circuit. The external load 26 may be connected in circuit with the inductor 51 (by a winding on the inductor, for example) and the capacitor may be omitted entirely. If the combined circuit loading of the inductor 51 and the external load 26 has an effective inductance of value sufficient to effect periodic energy storage for self-sustained transistor inversion, the current feedback provided by the saturable inductors **47**, **49** will effect alternate transistor conduction without the need for additional voltage feedback. When the capacitor **52** is omitted, the power supply **23** provides a direct electrical connection between one of the AC power input lines and the inverter load circuit.

Because the voltages across transistors **42**, **43** are relatively low (due to the effect of capacitors **34**, **36**), the half-bridge inverter **24** is very reliable. The absence of switching transients minimizes the possibility of transistor burn-out.

The inverter circuit **24** comprises means for supplying reverse bias to the conducting transistor upon saturation of its associated saturable inductor. For this purpose, the capacitors **56** and **57** are charged to negative voltages as a result of reset current flowing into secondary windings **53**, **54** from the bases of transistors **42**, **43**, respectively. This reverse current rapidly turns off a conducting transistor to increase its switching speed and to achieve inverter circuit efficiency in a manner described more fully in my co-pending U.S. patent application Ser. No. 103,624 filed Dec. 14, 1979 and entitled "Bias Control for High Efficiency Inverter Circuit" (now U.S. Pat. No. 4,307,353). The more negative the voltage on the bias capacitors **56** and **57**, the more rapidly charges are swept out of the bases of their associated transistors upon transistor turn-off.

When a transistor base-emitter junction is reversely biased, it exhibits the characteristics of a Zener diode having a reverse breakdown voltage on the order of 8 to 14 Volt for transistors typically used in high-voltage inverters. As an alternative, to provide a negative voltage smaller in magnitude on the base lead of typical transistor **42** during reset operation, the optional diode **61** and Zener diode **62** combination can be used. For large values of the bias capacitor **56**, the base voltage will be substantially constant.

If the load **26** comprises a gas discharge lamp, the voltage across the capacitor **52** will be reduced once the lamp is ignited to prevent voltages on the inductor **51** and the capacitor **52** from reaching destructive levels. Such a lamp provides an initial time delay during which a high voltage, suitable for instant starting, is available.

FIG. **5** illustrates the use of an alternate load **26'** adapted for plug-in connection to an inverter circuit such as shown in FIG. **2**. The load **26'** consists of a gas discharge lamp **71** having electrodes **72**, **73** and connected in series with a capacitor **74**. The combination of lamp **71** and capacitor **74** is connected in parallel with a capacitor **52'** which serves the same purpose as capacitor **52** in the FIG. **2** circuit. However, when the load **26'** is unplugged from the circuit, the inverter stops oscillating and the development of high voltages in the inverter is prevented. The fact that no high voltages are generated by the circuit if the lamp is disconnected while the circuit is oscillating is important for safety reasons.

FIG. **6** illustrates a capacitor **52"** connected in series with an inductor **51"** through a heater **81** suitable for heating the toroidal inductors **47**, **49** in accordance with the level of output. The load **26"** is connected across the series combination of the capacitor **52"** and the toroid heater. The heater **81** is preferably designed to controllably heat the toroidal saturable inductors in order to decrease their saturation flux limit and hence their saturation time. The result is to decrease the periodic transistor conduction time and thereby increase the transistor inversion frequency. When a frequency-dependent impedance means, that is, an inductor or a capacitor, is connected in circuit with the AC voltage output of the inverter, change in the transistor inversion frequency will modify the impedance of the frequency-dependent impdance means and correspondingly modify the inverter output. Thus as the level of the output increases, the toroid heater **81** is correspondingly energized to effect feedback regulation of the output. Further, transistors **42**, **43** of the type used in high voltage inverters dissipate heat during periodic transistor conduction. As an alternative, the toroid heater **81** can use this heat for feedback regulation of the output or control of the temperature of transistors **42**, **43**.

The frequency dependent impedance means may also be used in a circuit to energize a gas discharge lamp at adjustable illumination levels. Adjustment in the inversion frequency of transistors **42**, **43** results in control of the magnitude of the AC current supplied to the lamp. This is preferably accomplished where saturable inductors **47**, **49** have adjustable flux densities for control of their saturation time.

FIG. **7** schematically illustrates an alternate form of inverter circuit, shown without the AC to DC power supply connections for simplification. In this Figure, the transistors are connected in parallel rather than in series but the operation is essentially the same as previously described.

In particular, this circuit comprises a pair of alternately conducting transistors **91**, **92**. The emitter terminals of the transistors are connected to a B− line **93**. A B+ lead **94** is connected to the center-tap of a transformer **96**. In order to provide drive current to the transistors **91**, **92** for control of their conduction frequency, saturable inductors **97**, **98** have secondary windings **99**, **101**, respectively, each secondary winding having one end connected to the base of its associated transistor; the other ends are connected to a common terminal **102**. One end of transformer **96** is connected to the collector of transistor **91** through a winding **103** on inductor **98** in turn connected in series with a winding **104** on inductor **97**. Likewise, the other end of transformer **96** is connected to the collector of transistor **92** through a winding **106** on inductor **97** in series with another winding **107** on inductor **98**.

The B+ terminal is connected to terminal **102** through a bias resistor **108**. A bias capacitor **109** connects terminal **102** to the B− lead **93**. This resistor and capacitor serve the same function as resistors **58**, **59** and capacitors **56**, **57** in the FIG. **2** circuit.

The bases of transistors **91**, **92** are connected by diodes **111**, **112**, respectively, to a common Zener diode **113** in turn connected to the B− lead **93**. The common Zener diode **113** serves the same function as individual Zener diodes **62**, **64** in FIG. **2**.

Shunt diodes **114**, **116** are connected across the collector-emitter terminals of transistors **91**, **92**, respectively.

A capacitor **117** connecting the collectors of transistors **91**, **92** restrains the rate of voltage rise on the collectors in a manner similar to the collector-emitter capacitor **68** in FIG. **2**.

Inductive-capacitive loading of the FIG. **7** inverter is accomplished by a capacitor **118** connected in series with with an inductor **119**, the combination being connected across the collectors of the transistors **91**, **92**. A load **121** is connected across the capacitor **118**.

FIG. **8** illustrates how an inverter loaded with a series capacitor **122** and inductor **123** can be used to energize a "rapid-start" fluorescent lamp **124** (the details of the inverter circuit being omitted for simplication). The lamp **124** has a pair of cathodes **126**, **127** connected across the capacitor **122** for supply of operating voltage in a manner identical to that previously described. In addition, the inductor **123** comprises a pair of magnetically-coupled auxiliary windings 128, 129 for electrically heating the cathodes 126, 127, respectively. A small capacitor 131 is connected in series with lamp 124.

FIG. 9 shows an embodiment of the present invention that is expressly aimed at an alternative way of taking advantage of the fact that the inverter output voltage of the inverter circuit arrangement of FIG. 2 has the particular trapezoidal waveshape illustrated by FIG. 3A.

In FIG. 9, a DC supply voltage of about 320 Volt is assumed to be provided between a B– bus and a B+ bus.

A first high-frequency bypass capacitor BPC1 is connected between the B– bus and a junction Jc; and a second high-frequency bypass capacitor BPC2 is connected between junction Jc and the B+ bus. The source of a first field effect transistor FET1 is connected with the B– bus, while the drain of this same transistor is connected with a junction Jf. The source of a second field effect transistor FET2 is connected with junction Jf, while the drain of this same transistor is connected with the B+ bus. As shown in dashed outline, each field effect transistor has a commutating diode built-in between its drain and source. A slow-down capacitor SDC is connected between junction Jf and the B– bus.

The primary winding PW of a leakage transformer LT is connected between junction Jc and a junction Jx; the primary winding PW1 of a first saturable current transformer SCT1 is series-connected with the primary winding PW2 of a second saturable current transformer SCT2 between junctions Jf and Jx.

A secondary winding SW1 of transformer SCT1 is connected between the source and gate terminals of FET1; and a secondary winding SW2 of transformer SCT2 is connected between the source and gate terminals of FET2. A resistor R1 is connected across secondary winding SW1; and a resistor R2 is connected across secondary winding SW2. A Zener diode Z1a is connected with its cathode to the source of FET1 and with its anode to the anode of a Zener diode Z1b, whose cathode is connected with the gate of FET1. A Zener diode Z2a is connected with its cathode to the source of FET2 and with its anode to the anode of a Zener diode Z2b, whose cathode is connected with the gate of FET2.

A secondary winding SW of leakage transformer LT is connected between ballast output terminals BOT1 and BOT2.

A first fluorescent lamp FL1 is series-connected with a second fluorescent lamp FL2 to form a series-combination; which series-combination is connected between ballasts output terminals BOT1 and BOT2. Lamp FL1 has a first cathode C1*a* and a second cathode C1*b*; while lamp FL2 has a first cathode C2*a* and a second cathode C2*b*. Each cathode has two cathode terminals. Each of the terminals of cathode C1*b* is connected with one of the terminals of cathode C2*a*. Each cathode's terminals are connected with the terminals of one of three separate cathode heater windings CHW.

The leakage transformer of FIG. 9 is illustrated in further detail in FIG. 10. In particular and by way of example, leakage transformer LT includes a first and a second ferrite core element FC1 and FC2, each of which is an extra long so-called E-core; which E-cores abut each other across an air gap AG. Primary winding PW is wound on a first bobbin B1; and secondary winding SW is wound on a second bobbin B2. Cathode heating windings CHW are wound on a small third bobbin B3; which bobbin B3 is adjustably positioned between bobbins B1 and B2.

The operation of the circuit arrangement of FIG. 9 may best be understood by referring to the voltage and current waveforms of FIGS. 11A to 11F.

FIG. 11A shows the waveform of the voltage provided at the output of the half-bridge inverter of FIG. 9 during a situation where lamps FL1 and FL2 are being fully powered. In particular, FIG. 11A shows the waveform of the voltage provided at junction Jf as measured with reference to junction Jc. (The voltage at Jx is substantially equal to the voltage at Jf).

This waveform is substantially equal to that of FIG. 3A.

FIG. 11B shows the corresponding waveform of the gate-to-source voltage (i.e. the control voltage) of FET2.

FIG. 11C shows the corresponding drain current flowing through FET2; which is the current drawn by the upper half of the half-bridge inverter from the DC supply voltage (i.e., from the B+ bus).

FIG. 11D shows the corresponding current flowing through fluorescent lamps FL1 and FL2.

FIG. 11E shows the waveform of the voltage provided at the output of the half-bridge inverter of FIG. 9 for a situation where ballast output terminals BOT1/BOT2 are unloaded except for stray (or parasitic) capacitance associated with the wiring extending between ballast output terminals BOT1/BOT2 and lamp cathodes C1*a* and C2*b*.

The waveform of FIG. 11E is substantially equal to that of FIG. 11A except for an increase in the duration of each cycle period.

FIG. 11F shows the corresponding open circuit output voltage present across ballast output terminals BOT1 and BOT2.

FIG. 11G shows the waveform of the voltage provided at the output of the half-bridge inverter of FIG. 9 for a situation where: (i) slowdown capacitor SDC has been removed; and (ii) ballast output terminals BOT1/BOT2 are unloaded except for stray (or parasitic) capacitance associated with the wiring extending between ballast output terminals BOT1/BOT2 and lamp cathodes C1a and C2b.

It is noted that the waveform of FIG. 11G is substantially a true squarewave as opposed to the trapezoidal (or truncated sinusoidal) waveforms of FIGS. 11A and 11E.

FIG. 11H shows the waveform of the corresponding voltage present across ballast output terminals BOT1 and BOT2.

The basic inverter part of FIG. 9 operates much like the inverter part of FIG. 2, except that the switching transistors are field effect transistors instead of bi-polar transistors.

The loading of the inverter, however, is different. In the circuit of FIG. 9, the inverter's output voltage is applied to the primary winding of a leakage transformer (LT); and the output is drawn from a primary winding of this leakage transformer. In this connection, it is important to notice that a leakage transformer is a transformer wherein there is substantial leakage of magnetic flux between the primary winding and the secondary winding; which is to say that a substantial part of the flux generated by the transformer's primary winding does not link with the transformer's secondary winding.

The flux leakage aspect of transformer LT is illustrated by the structure of FIG. 10. Magnetic flux generated by (and emanating from) primary winding PW passes readily through the high-permeability ferrite of ferrite core FC1. However, as long as secondary winding SW is connected with a load at its output (and/or if there is an air gap, as indeed there is), the flux emanating from the primary winding has to overcome magnetic impedance to flow through the secondary winding; which implies the development of a magnetic potential difference between the legs of the long

E-cores—especially between the legs of ferrite core FC1. In turn, this magnetic potential difference causes some of the magnetic flux generated by the primary winding to flow directly between the legs of the E-cores (i.e. directly across the air gap between the legs of the E-cores), thereby not linking with (i.e. flowing through) the secondary winding. Thus, the longer the legs of the E-cores and/or the larger the air gap, the less of the flux generated by the primary winding links with the secondary winding—and conversely. As a result, the magnitude of the current available from the secondary winding is limited by an equivalent internal inductance.

Due to the substantial air gap (AG), the primary winding of leakage transformer LT is capable of storing a substantial amount of inductive energy (just as is the case with inductor 51 of FIG. 2). Stated differently but equivalently, leakage transformer LT has an equivalent input-shunt inductance (existing across the input terminals of its primary or input winding) capable of storing a substantial amount of energy. It also has an equivalent output-series inductance (effectively existing in series with the output terminals of its secondary or output winding) operative to limit the magnitude of the current available from its output. It is important to recognize that the input-shunt inductance is an entity quite separate and apart from the output-series inductance.

Just as in the circuit of FIG. 2, when one of the transistors is switched OFF, the current flowing through primary winding PW can not instantaneously stop flowing. Instead, it must continue to flow until the energy stored in the input-shunt inductance is dissipated and/or discharged. In particular and by way of example, at the moment FET2 is switched OFF, current flows through primary winding PW, entering at the terminal connected with junction Jx and exiting at the terminal connected with junction Jc. Just after the point in time where FET2 is switched OFF, this current will continue to flow, but—since it can not any longer flow through transistor FET2—it must now flow through slow-down capacitor SDC. Thus, the current drawn out of capacitor SDC will cause this capacitor to change its voltage: gradually causing it to decrease from a magnitude of about +160 Volt (which is the magnitude of the DC supply voltage present at the B+ bus as referenced-to junction Jc) to about −160 Volt (which is the magnitude of the DC supply voltage present at the B− bus as referended-to junction Jc). Of course, as soon as it reaches about −160 Volt, it gets clamped by the commutating (or shunting, or clamping) diode built-into FET1; which built-in diode corresponds to shunting diode 67 of the FIG. 2 circuit.

The resulting waveform of the inverter's output voltage will be as illustrated by FIGS. 11A and 11E. The slope of the inverter output voltage as it alternatingly changes between −160 Volt and +160 Volt is determined by two principal factors: (i) the value of the input-shunt inductance of primary winding PW; and (ii) the magnitude of slow-down capacitor SDC. The lower the capacitance of the slow-down capacitor, the steeper the slope. The lower the inductance of the input-shunt inductance, the steeper the slope. Without any slow-down capacitor, the slope will be very steep: limited entirely by the basic switching speed of the inverter's transistors; which, for field effect transistors is particularly high (i.e. fast).

In particular, in the circuit of FIG. 9, the relatively modest up- and down- slopes of the inverter's output voltage (see waveforms of FIGS. 11A and 11E)—which are determined by the capacitance of the slow-down capacitor—are chosen to be far lower than the very steep slopes that result when the slow-down capacitor is removed; which latter situation is

illustrated by FIG. 11G. In fact, the slopes of the inverter's output voltage are chosen in such manner as to result in this output voltage having a particularly low content of harmonic components, thereby minimizing potential problems associated with unwanted resonances of the output-series inductance with parasitic capacitances apt to be connected with ballast output terminals BOT1/BOT2 by way of more-or-less ordinary wiring harness means used for connecting between these output terminals and the associated fluorescent lamps (FL1 and FL2).

With the preferred capacitance value of slow-down capacitor SDC, the inverter output voltage waveform will be as shown in FIGS. 11E, and the output voltage provided from secondary winding SW—under a condition of no load other than that resulting from a parasitic resonance involving a worst-value of parasitic output capacitance—will be as shown in FIG. 11F.

On the other hand, without having any slow-down capacitor, the inverter output voltage waveform will be as shown in FIG. 11G, and the output voltage provided from secondary winding SW—under a condition of no load other than that resulting from a parasitic resonance involving a worst-value of parasitic output capacitance—will be as shown in FIG. 11H. Under this condition, the power drawn by the inverter from its DC supply is more than 50 Watt; which power drain result from power dissipations within the inverter circuit and—if permitted to occur for more than a very short period—will cause the inverter to self-destruct.

On the other hand, the power drawn by the inverter under the same identical condition except for having modified the shape of the inverter's output voltage to be like that of FIG. 11E (instead of being like that of FIG. 11G) is only about 3 Watt; which amount of power drain is small enough not to pose any problem with respect to inverter self-destruction, nor even with respect to excessive power usage during extended periods where the inverter ballast is connected with its power source but without actually powering its fluorescent lamp load.

One difference between the circuit of FIG. 2 and that of FIG. 9 involves that fact that the FIG. 9 circuit uses field effect transistors. Never-the-less, the control of each transistor is effected by way of saturable current feedback transformers. However, instead of delivering its output current to a base-emitter junction, each current transformer now delivers its output current to a pair of series-connected opposed-polarity Zener diodes (as parallel-connected with a damping resistor and the gate-source input capacitance). The resulting difference in each transistor's control voltage is seen by comparing the waveform of FIG. 3B with that of FIG. 11B. In either case, however, the transistor is not switched into its ON-state until after the absolute magnitude of the voltage across its switched terminals (i.e. the source-drain terminals for a FET) has substantially diminished to zero.

In further contrast with the arrangement of FIG. 2, the inverter circuit of FIG. 9 is not loaded by way of a series-tuned L-C circuit. Instead, it is in fact loaded with a parallel-tuned L-C circuit; which parallel-tuned L-C circuit consists of the slow-down capacitor SDC as parallel-connected with the input-shunt inductance of primary winding PW. Yet, in complete contrast with other inverters loaded with parallel-tuned L-C circuits, the inverter of FIG. 9 is powered from a voltage source providing a substantially fixed-magnitude (i.e. non-varying) DC voltage.

Also in complete contrast with other inverters loaded with parallel-tuned L-C circuits, the inverter circuit of FIG. 9

provides for clamping (or clipping or truncating) of the naturally sinusoidal resonance voltage that would otherwise (i.e. in the absence of clamping) develop across the parallel-tuned L-C circuit; which naturally sinusoidal resonance voltage is illustrated by the dashed waveform of FIG. 11E.

In the FIG. 9 circuit, the indicated voltage clamping (or clipping or truncating) is accomplished by way of the commutating (or shunting) diodes built into each of the field effect switching transistors. In the FIG. 2 circuit, this clamping is accomplished by shunting diodes **66** and **67**.

As previously indicated, to minimize the spurious and potentially damaging resonances which might occur due to an unknown parasitic capacitance becoming connected with ballast output terminals BOT1 and BOT2, it is important to minimize the harmonic content of the inverter's output voltage (which harmonic content—for a symmetrical inverter waveform—consists of all the odd harmonics in proportionally diminishing magnitudes). To attain such harmonic minimization, it is important that the inverter's output voltage be made to match or fit as nearly as possible the waveform of a sinusoidal voltage; which "best fit" occurs when the duration of the up/down-slopes equals about 25% of the total cycle period; which, as can readily be seen by direct visual inspection, corresponds closely to the waveforms actually depicted by FIGS. 3A, 11A and 11E.

However, substantial beneficial effects actually results even if the total duration of the up/down slopes were to be less than 25% of the total duration of the inverter output voltage period. In fact, substantial beneficial effects are attained with up-down slopes constituting as little as 10% of the total cycle period.

Additional Explanations and Comments

(a) With reference to FIGS. 2 and 5, adjustment of the amount of power supplied to load **26'**, and thereby the amount of light provided by lamp **71**, may be accomplished by applying a voltage of adjustable magnitude to input terminals IP1 and IP2 of the Toroid Heater; which is thermally coupled with the toroidal ferrite cores of saturable transformers **47**, **49**.

(b) With commonly available components, inverter circuit **24** of FIG. 2 can be made to operate efficiently at any frequency between a few kHz to perhaps as high as 50 kHz. However, for various well-known reasons (i.e., eliminating audible noise, minimizing physical size, and maximizing efficiency), the frequency actually chosen is in the range of 20 to 40 kHz.

(c) The fluorescent lighting unit of FIG. 1 could be made in such manner as to permit fluorescent lamp **11** to be disconnectable from its base **12** and ballasting means **16**. However, if powered with normal line voltage without its lamp load connected, frequency-converting power supply and ballasting circuit **16** is apt to self-destruct.

To avoid such self-destruction, arrangements can readily be made whereby the very act of removing the load automatically establishes a situation that prevents the possible destruction of the power supply and ballasting means. For instance, with the tank capacitor (**52**) being permanently connected with the lamp load (**11**)—thereby automatically being removed whenever the lamp is removed—the inverter circuit is protected from self-destruction.

(d) At frequencies above a few kHz, the load represented by a fluorescent lamp—once it is ignited—is substantially resistive. Thus, with the voltage across lamp **11** being of a substantially sinusoidal waveform (as indicated in FIG. 3*d*), the current through the lamp will also be substantially sinusoidal in waveshape.

(e) In the fluorescent lamp unit of FIG. 1, fluorescent lamp **11** is connected with power supply and ballasting circuit **16** in the exact same manner as is load **26** connected with the circuit of FIG. 2. That is, it is connected in parallel with the tank capacitor (**52**) of the L-C series-resonant circuit. As is conventional in instant-start fluorescent lamps—such as lamp **11** of FIG. 1—the two terminals from each cathode are shorted together, thereby to constitute a situation where each cathode effectively is represented by only a single terminal. However, it is not necessary that the two terminals from each cathode be shorted together; in which case—for instant-start operation—connection from a lamp's power supply and ballasting means need only be made with one of the terminals of each cathode.

(f) In FIG. 9, a Parasitic Capacitance is shown as being connected across terminals BOT1 and BOT2. The value of this parasitic capacitance may vary over a wide range, depending on unpredictable details of the particular usage situation at hand. Values for the parasitic capacitance will expectedly vary between 100 and 1000 pico-Farad—depending on the nature of the wiring harness used for connecting between the output of secondary winding SW and the plural terminals of lamps FL1/FL2.

(g) The worst case of parasitic oscillation associated with the circuit arrangement of FIG. 9 is apt to occur when the value of the parasitic capacitance (i.e., the capacitance of the ballast-to-lamp wiring harness) is such as to cause series-resonance with the output-series inductance of secondary winding SW at the third harmonic component of the inverter's output voltage. The next worst case of parasitic oscillation is apt to occur when the value of the parasitic capacitance is such as to series-resonate with the output-series inductance at the fifth harmonic component of the inverter's output voltage. With the typical value of 5.4 milli-Henry for the output-series inductance, it takes a total of about 600 pico-Farad to resonate at the third harmonic component of the inverter's 30 kHz output voltage; and it takes about 220 pico-Farad to resonate at the fifth harmonic component of the inverter's output voltage. These capacitance values are indeed of such magnitudes that they may be encountered in an actual usage situation of an electronic ballast. Moreover, at higher inverter frequencies, the magnitudes of the critical capacitance values become even lower.

(h) FIG. **10** shows cathode heater windings CHW placed on a bobbin separate from that of primary winding PW as well as separate from that of secondary winding SW. However, in many situations, it would be better to place the cathode heater windings directly onto the primary winding bobbin B1. In other situations it would be better to place the cathode heater windings directly onto the secondary winding bobbin B2.

If the cathode heater windings are wound on bobbin B1 (i.e. in tight coupling with the primary winding), the magnitude of the cathode heating voltage will remain constant regardless of whether or not the lamp is ignited; which effect is conducive to maximixing lamp life. On the other hand, if the cathode heater windings are wound on bobbin B2 (i.e. in tight coupling with the secondary winding), the magnitude of the cathode heating voltage will be high prior to lamp ignition and low after lamp ignition; which effect is conducive to high luminous efficacy.

By placing the cathode heater windings in a location between primary winding PW and secondary winding SW, it is possible to attain an optimization effect: a maximization of luminous efficacy combined with only a modest sacrifice in lamp life. That is, by adjusting the position of bobbin B3, a corresponding adjustment of the ratio of pre-ignition to post-ignition cathode heater voltage magnitude may be accomplished.

(i) For easier lamp starting, a starting aid capacitor may be used in shunt across one of the fluorescent lamps FL1/FL2.

Also, a starting aid electrode (or ground plane) may advantageously be placed adjacent the fluorescent lamps; which starting aid electrode should be electrically connected with the secondary winding, such as via a capacitor of low capacitance value.

(j) To control (reduce) the degree of magnetic coupling between primary winding PW and secondary winding SW, a magnetic shunt may be positioned across the legs of the E-cores—in a position between bobbins B1 and B3.

(k) Considering the waveforms of FIGS. 1A, 11A and 11E each to include 360 degrees for each full and complete cycle: (i) each half-cycle would include 180 degrees; (ii) each total up-slope would include almost or about 60 degrees degrees; (iii) each total down-slope would include almost or about 60 degrees; and (iv) each horizontal segment would include about 120 degrees or more. Yet, as previously indicated, substantial utility may be attained even if each complete up-slope and down-slope were to include as little 18 degrees.

(l) In the FIG. 9 circuit, the inverter's operating frequency is not ordinarily (or necessarily) equal to the natural resonance frequency of the parallel-tuned L-C circuit that consists of slow-down capacitor SDC and the input-shunt inductance of primary winding PW. Rather, the inverter's actual operating frequency is ordinarily lower than would be this natural resonance frequency.

(m) In a trapezoidal waveform that constitutes a best fit for a sinusoidal waveform, the peak magnitude is lower than that of the sinusoidal waveform, and the up-slope and down-slope are each steeper that the corresponding slopes of the sinusoidal waveform.

(n) The FIG. 9 inverter arrangement has to be triggered into self-oscillation. A suitable automatic triggering means would include a resistor, a capacitor, and a so-called Diac. However, manual triggering may be accomplished by merely momentarily connecting a discharged capacitor (of relatively small capacitance value) between the gate of transistor FET1 and the B+ bus.

(o) Most switching-type field effect transistors have built-in commutating (or shunting) diodes, as indicated in FIG. 9. However, if such were not to be the case, such diodes should be added externally, as indicated in the FIG. 2 circuit.

(p) In ordinary inverter circuits, the inverter output voltage is effectively a squarewave voltage with very steep up-slopes and down-slopes. In inverters using field effect transistors, the time required for the inverter's squarewave output voltage to change between its extreme negative potential to its extreme positive potential is usually on the order of 100 nano-seconds or less. In inverters using bi-polar transistors, this time is usually on the order of 500 nano-seconds or less. In the inverter of the FIG. 9 circuit, however, this time has been extended—by way of the large-capacitance-value slow-down capacitor SDC—to be on the order of several micro-seconds, thereby achieving a substantial reduction of the magnitudes of the harmonic components of the inverter's (now trapezoidal) output voltage.

(q) In an actual prototype of the FIG. 9 ballast circuit—which prototype was designed to properly power two 48 inch 40 Watt T-12 fluorescent lamps—the following approximate parameters and operating results prevailed:

1. operating frequency: about 30 kHz;
2. slow-down capacitor: 0.02 micro-Farad;
3. shunt-input inductance: 1.4 milli-Henry;
4. up-slope duration: about 4 micro-seconds;
5. down-slope duration: about 4 micro-seconds;
6. series-output inductance: 5.4 milli-Henry;
7. parasitic capacitance across BOT1/BOT2 terminals; 800 pico-Farad;
8. power consumption when unloaded: about 4 Watt;
9. power consumption when loaded with two F40/T12 fluorescent lamps: about 70 Watt;
10. power consumption when unloaded but with slow-down capacitor removed: about 80 Watt.

It is be noted that the natural resonance frequency of the L-C circuit consisting of a slow-down capacitor of 0.02 micro-Farad as parallel-combined with a shunt-input inductance of about 1.4 milli-Henry is about 30 kHz. This means that—as far as the fundamental component of the 30 kHz inverter output voltage is concerned—the parallel-tuned L-C circuit represents a very high impedance, thereby constituting no substantive loading on the inverter's output.

(r) Of course, the FIG. 9 ballast circuit can be made in the form of a push-pull circuit such as illustrated by FIG. 7; in which case center-tapped transformer 96 would be modified in the sense of being made as a leakage transformer in full correspondence with leakage transformer LT of FIG. 9. Also, of course, inductor 119, capacitor 118, and load 121 would be removed. Instead, the load would be placed at the output of the secondary winding of the modified center-tapped transformer 96; which would be made such as to have appropriate values of input-shunt inductance and output-series inductance. Capacitor 117 would constitute the slow-down capacitor.

(s) It is thought that the present invention and many of its attendant advantages will be understood from the foregoing description and that many changes may be made in the form and construction of its components parts, the form described being merely a preferred embodiment of the invention.

I claim:

1. An arrangement comprising:

a pair of power line terminals at which is provided a power line voltage;

rectification and filtering circuitry connected with the power line terminals and operative to provide a DC supply voltage at a set of DC terminals; the absolute magnitude of the DC supply voltage being substantially constant; and

inverter and load circuitry connected with the DC terminals and operative to provide an AC voltage at a pair of AC terminals with which is connected a gas discharge lamp; thereby to cause an alternating lamp current to flow through the lamp; the alternating lamp current having a peak magnitude and an RMS magnitude; the AC voltage being of a frequency substantially higher than that of the power line voltage on an ordinary electric utility power line; the inverter and load circuitry being additionally characterized by causing the peak magnitude of the alternating lamp current to be distinctly lower than twice its RMS magnitude.

2. The arrangement of claim 1 wherein the absolute magnitude of the DC voltage is about twice as high as the peak absolute magnitude of the power line voltage.

3. The arrangement of claim 1 wherein:

(a) the gas discharge lamp (i) at certain times draws a lamp current from the AC terminals, (ii) at certain other times, such as if disconnected or otherwise non-functional, draws no lamp current; and

(b) the inverter and load circuitry is characterized by the fact that the frequency of the AC voltage is substantively higher during said certain other times as compared with during said certain times.

4. A ballasting arrangement for a gas discharge lamp, comprising:

a source providing a DC supply voltage at a set of DC terminals; and

an inverter circuit assembly connected with the DC terminals and operative to provide an inverter output current from a pair of inverter output terminals; the inverter output current being an alternating current of frequency substantially higher than that of the power line voltage on an ordinary electric utility power line; the inverter circuit assembly also including a pair of lamp output terminals adapted to connect with and to supply a lamp current to a gas discharge lamp; the inverter circuit assembly being otherwise characterized by having: (i) an inductor means connected between one of the inverter output terminals and one of the lamp output terminals, the inductor means having a frequency-dependent impedance in the sense that its impedance increases with increasing frequency of any current flowing through it; (ii) a frequency control sub-assembly operative, on receipt of a control action at a control action input, to control the frequency and thereby the magnitude of the inverter output current; and (iii) an action-generating element-operative, in response to at least a part of the inverter output current, to provide a control action at a control action output; the control action output being connected in communication with the control action input, thereby to supply thereto said control action, thereby to cause the frequency of the inverter output current to change in response to the magnitude of the inverter output current, thereby to cause the magnitude of the inverter output current to be different from what it would have been in the absence of providing the control action to the frequency control sub-assembly.

5. The ballasting arrangement of claim 4 wherein, by providing said control action to the frequency control sub-assembly, the magnitude of said part of the inverter output current is lower than it would have been in the absence of providing the control action to the frequency control sub-assembly.

6. The ballasting arrangement of claim 4 wherein the provision of said control action to the frequency control sub-asembly constitutes negative feedback in the sense that, if the magnitude of said part of the inverter output current were to change due to a change in the characteristics of a circuit element connected within the inverter circuit assembly, such a change in magnitude will be less because of the provision of said control action to the frequency control sub-assembly than it would have been without the provision of said control action to the frequency control sub-assembly.

7. The ballasting arrangement of claim 4 wherein, if the characteristics of a circuit element connected within the inverter circuit assembly were to be altered, such as by altering the inductance of the inductor means, the magnitude of said part of the inverter output current would change by a certain amount; which certain amount will be less when actually providing said control action to the frequency control sub-assembly than when not providing said control action to the frequency control sub-assembly.

8. An arrangement comprising:

a source providing a DC supply voltage at a set of DC terminals;

an inverter circuit connected with the DC terminals and operative to provide an AC inverter output voltage at a pair of inverter output terminals; the AC inverter output voltage being of frequency substantially higher than that of the power line voltage on an ordinary electric utility power line; the inverter circuit having a frequency-controlling sub-assembly having a control action input operative, on receipt of a control action signal, to control the frequency of the AC inverter output voltage;

an output circuit connected between the inverter output terminals and a pair of lamp current output terminals; the output circuit being characterized by including: (i) a frequency-dependent impedance element, such as an inductor; and (ii) a an action-signal-generating element having a control action output connected in communication with said control action input and operative to provide said control action signal in response to a current flowing in the output circuit, thereby to affect the frequency of the AC inverter output voltage and, therefore, the magnitude of said current; whereby the magnitude of said current is different from what it would have been if said control action signal had not been provided to the control action input; and

a gas discharge lamp connected with the lamp current output terminals.

9. The arrangement of claim 8 wherein the action-signal-generating element is operative to provide said control action signal to the control action input in such a manner that any change in the magnitude of said current that might occur due to a reason other than the provision of the control action signal itself will be reduced as a result of providing the control action signal.

10. A ballasting arrangement for a gas discharge lamp, comprising:

a source providing a DC supply voltage at a set of DC terminals; and

an assembly of components connected with the DC terminals and operative to provide an AC voltage at a pair of first terminals; the assembly being characterized by including various electrical components in addition to output circuitry having an action-generating element operative to provide a control action at a control action output in response to a current flowing in a certain conductor included as part of the output circuitry; the AC voltage being of frequency substantially higher than that of the power line voltage on an ordinary electric utility power line; the assembly also including a pair of second terminals connected with the first terminals by way of the output circuitry; the second terminals being adapted to connect with and to supply lamp current to a gas discharge lamp; the assembly being otherwise characterized by: (a) including an inductor as part of the output circuitry; (b) having a control sub-assembly operative, on receipt of a control action at a control action input, to control the frequency of the AC voltage, the control action input being functionally connected with the control action output; and (c) causing a current to flow through the certain conductor; thereby: (i) to cause said control action to be supplied from the control action output to the control action input, (ii) to cause the frequency of the AC voltage to be controlled in response to the current flowing through the certain conductor, and (iii) to cause the magnitude of this current to be different from what it would have been in the absence of providing the control action to the control action input.

11. The ballasting arrangement of claim 10 wherein the action-generating element is functional to provide the control action to the control action input in such a manner that any change in the magnitude of the current flowing in said conductor which might occur due to a reason other than the provision of the control action output itself will be reduced as a result of actually providing the control action.

12. An arrangement comprising:

a source providing a DC supply voltage at a set of DC terminals;

an inverter circuit assembly connected with the DC terminals and operative to provide an AC inverter voltage at a pair of AC inverter terminals; the AC inverter voltage being of a frequency substantially higher than that of the power line voltage on an ordinary electric utility power line; the inverter circuit assembly being characterized by including a control subassembly operative, on receipt of a control action at a control action input, to control the frequency of the AC inverter voltage; and

load circuitry connected with the AC inverter terminals; the load circuitry being characterized by: (a) including a disconnectable gas discharge lamp; (b) including an inductor; (c) including a conductor; (d) including an action-generating element connected in circuit with the conductor and operative to provide a control action at a control action output in response to a current flowing through the conductor; and (e) having the control action output connected in communication with the control action input, thereby: (i) to provide the control action such as to cause: (i) the frequency of the AC inverter voltage to be controlled in response to the magnitude of the current flowing through the conductor, and (ii) the magnitude of this current to be different from what it would have been in the absence of providing the control action to the control action input.

13. The arrangement of claim 12 wherein the effect of providing the control action is that of causing the frequency of the AC inverter voltage to increase as compared with a situation where the control action is absent.

14. The arrangement of claim 12 wherein the AC output voltage is characterized by having a waveform consisting of: (i) a first time-segment during which the instantaneous magnitude of the AC inverter voltage remains at a substantially constant negative level; (ii) a second time-segment during which the instantaneous magnitude of the AC inverter voltage increases in a substantially continuous manner; (iii) a third time-segment during which the instantaneous magnitude of the AC inverter voltage remains at a substantially constant positive level; and (iv) a fourth time-segment during which the instantaneous magnitude of the AC inverter voltage decreases in a substantially continuous manner.

15. An arrangement comprising:

a source providing a DC supply voltage at a set of DC terminals; and

inverter and load circuitry connected with the DC terminals and operative to provide an AC voltage at a pair of AC terminals; the AC voltage being of a frequency substantially higher than that of the power line voltage on an ordinary electric utility power line; the inverter and load circuitry being characterized by including (i) an inductor, (ii) a capacitor, (iii) a gas discharge lamp, and (iv) an intermittently conducting transistor; the AC voltage being characterized by having a waveform consisting of: (i) a first time-segment during which the instantaneous magnitude of the AC voltage remains at a substantially constant negative level, (ii) a second time-segment during which the instantaneous magnitude of the AC voltage increases in a substantially continuous manner, (iii) a third time-segment during which the instantaneous magnitude of the AC voltage remains at a substantially constant positive level, and (iv) a fourth time-segment during which the instantaneous magnitude of the AC voltage decreases in a substantially continuous manner; the inverter and load circuitry being further characterized in that substantially no current flows through the transistor during the second and fourth time-segments.

16. An arrangement comprising:

a source providing a DC supply voltage at a set of DC terminals; and

inverter and load circuitry connected with the DC terminals and operative to provide an AC voltage at a pair of AC terminals with which is connected a gas discharge lamp; the AC voltage being of a frequency substantially higher than that of the power line voltage on an ordinary electric utility power line; the gas discharge lamp being characterized by drawing a lamp current from the AC terminals at certain times and not at other times; the inverter and load circuitry being characterized in that the frequency of the AC voltage is different during the times when lamp current is being drawn as compared with times when lamp current is not being drawn.

17. The arrangement of claim 16 wherein the frequency of the AC voltage is lower during the times when lamp current is being drawn as compared with the times when lamp current is not being drawn.

18. The arrangement of claim 1 wherein the inverter and load circuitry is also characterized by including an L-C circuit electrically connected with the AC terminals and having a natural resonance frequency at or near the fundamental frequency of the AC voltage.

19. The arrangement of claim 1 wherein the inverter and load circuitry is also characterized by including a transistor having a pair of transistor terminals across which exists a transistor voltage and between which periodically flows a transistor current; the transistor current flowing only during periods when the absolute magnitude of the transistor voltage is substantially lower than that of the DC supply voltage.

20. The arrangement of claim 19 further defined in that current flows between the transistor terminals only during periods when the absolute magnitude of the transistor voltage is smaller than that of the DC supply voltage by a factor of ten or more.

21. The arrangement of claim 1 wherein the inverter and load circuitry is also characterized by including a certain terminal at which, as referenced to one of the DC terminals, exists a substantially sinusoidal voltage of frequency equal to that of the AC voltage.

22. The arrangement of claim 21 further defined by: (i) including a periodically conducting transistor having a transistor terminal; and (ii) having a conductive path between the transistor terminal and said certain terminal, the ohmic resistance of which conductive path being of negligible magnitude.

23. The arrangement of claim 1 wherein the inverter and load circuitry is also characterized by including a transistor having a pair of transistor terminals between which periodically flows a transistor current, which periodic transistor current has a period equal to that of the AC voltage and, for each complete cycle of the AC voltage, flows for a total time

duration that is distinctly shorter than 50% of the total duration of said complete cycle.

24. The arrangement of claim 1 wherein the inverter and load circuitry is also characterized by including a transistor having a pair of transistor terminals across which exists a periodic transistor voltage and between which flows a transistor current in the form of periodic unidirectional current pulses; the duration of each unidirectional current pulse being distinctly shorter than half the duration of a complete cycle of the AC voltage.

25. The arrangement of claim 1 wherein the inverter and load circuitry is further defined in that the frequency of the AC voltage is distinctly higher just prior to lamp ignition than it is after the lamp has ignited.

26. The arrangement of claim 1 wherein the inverter and load circuitry is further defined in that the frequency of the AC voltage would increase if the gas discharge lamp were to be disconnected.

27. The arrangement of claim 1 wherein the inverter and load circuitry is also characterized by including: (i) a transistor having a pair of transistor terminals; (ii) a first and a second conductor between which exists an alternating voltage of substantially sinusoidal waveform and with frequency equal to that of the AC voltage, the first conductor being characterized by having an AC potential substantially equal to that of one of the DC terminals, the second conductor being characterized by having a DC potential equal to one of the transistor terminals.

28. The arrangement of claim 27 yet additionally characterized by having a capacitor connected between said one of the transistor terminals and a termination point having an AC potential substantially equal to that of one of the DC terminals.

29. The arrangement of claim 1 wherein the inverter and load circuitry is also characterized by including: (i) a transistor having a transistor terminal; (ii) a first and a second conductor between which exists a substantially sinusoidal voltage of frequency equal to that of the AC voltage, the first conductor being characterized by having an AC potential substantially equal to that of one of the DC terminals, the second conductor being characterized by having a DC potential equal to the transistor terminal.

* * * * *

EXHIBIT J

US006121733A

United States Patent [19]

Nilssen

[11] **Patent Number: 6,121,733**

[45] **Date of Patent: Sep. 19, 2000**

[54] **CONTROLLED INVERTER-TYPE FLUORESCENT LAMP BALLAST**

[76] Inventor: **Ole K. Nilssen**, Caesar Dr., Barrington, Ill. 60010

[21] Appl. No.: **08/274,481**

[22] Filed: **Jul. 13, 1994**

Related U.S. Application Data

[63] Continuation-in-part of application No. 07/712,454, Jun. 10, 1991, abandoned.

[51] **Int. Cl.**[7] **H05B 41/24**

[52] **U.S. Cl.** **315/224**; 315/247; 315/205; 315/219; 315/224; 315/DIG. 7; 315/226

[58] **Field of Search** .... 315/247, 226, 315/205, DIG. 7, DIG. 5, 219, 224; 363/37, 101, 16, 65, 71; 323/205, 206, 207

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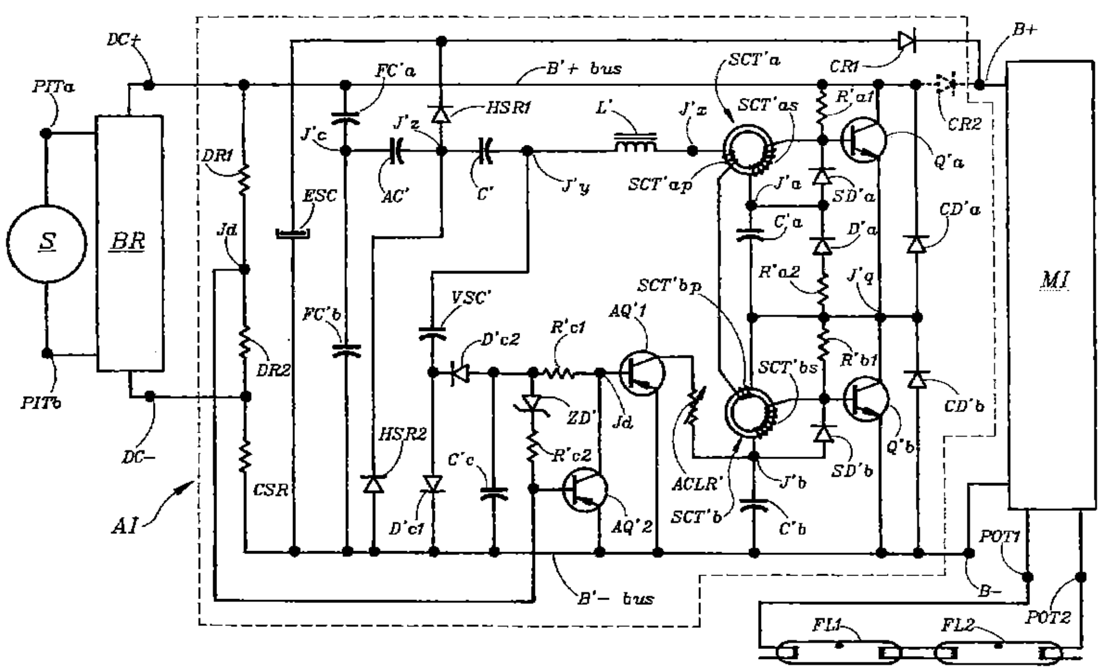
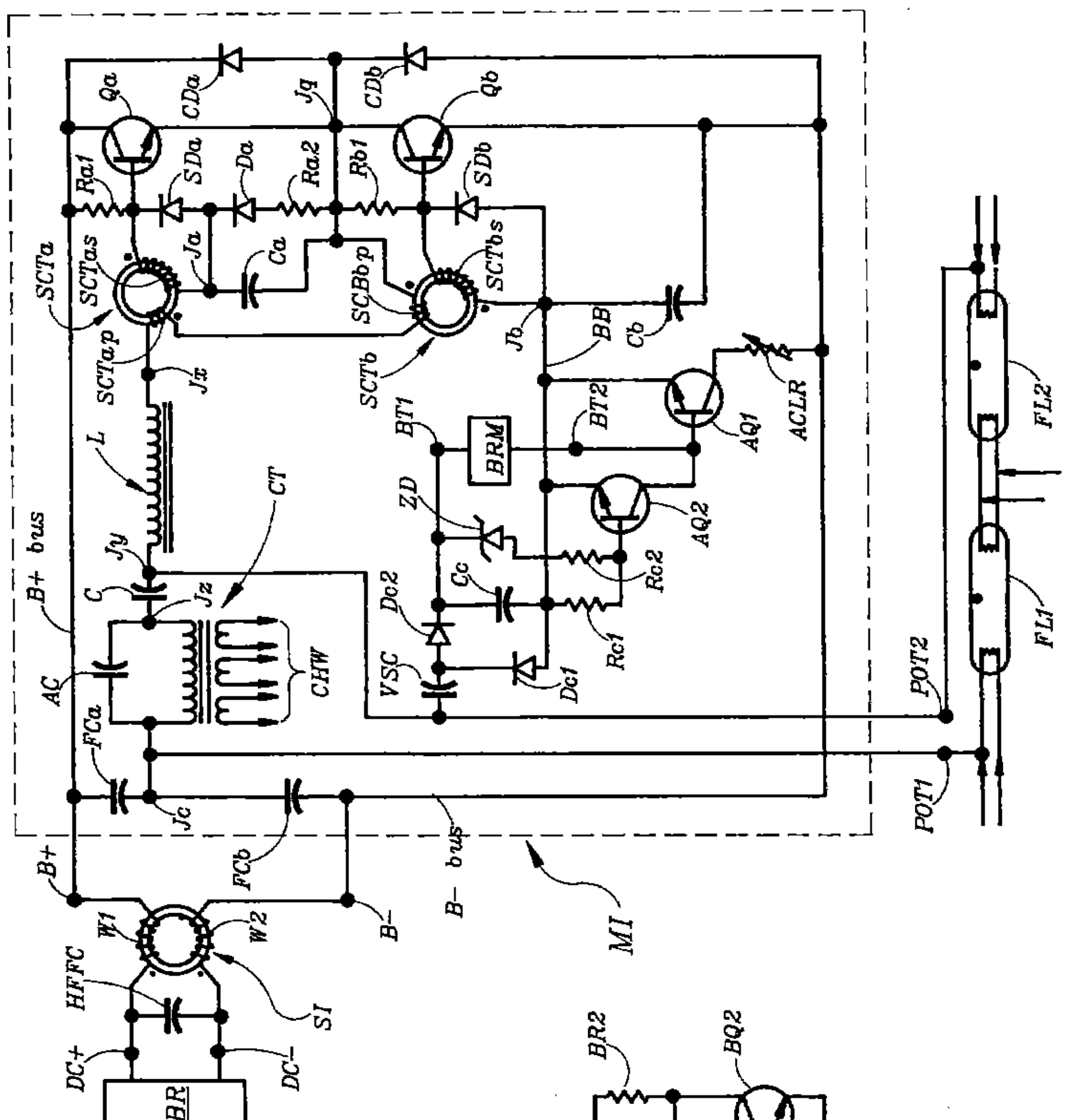
Primary Examiner—Arnold Kinkead

[57] **ABSTRACT**

Unfiltered full-wave-rectified 60 Hz power line voltage is supplied to a first self-oscillating series-resonance-loaded inverter, the high frequency output current from which is magnitude-controlled, rectified and used for maintaining a substantially constant-magnitude DC voltage on an energy-storing capacitor. This constant-magnitude DC voltage is used in combination with the unfiltered full-wave-rectified 60 Hz power line voltage for powering a second self-oscillating series-resonance-loaded inverter, the high frequency current output from which is used for powering a fluorescent lamp load. By frequency-modulating the first inverter at a 120 Hz rate, the current drawn from the power line is made to be of such waveshape as to result in a power factor well in excess of 90% while at the same time having a content of under 20% of odd triplets of third harmonic currents. The high frequency current provided to the fluorescent lamp load has a crest factor not in excess of 1.7.

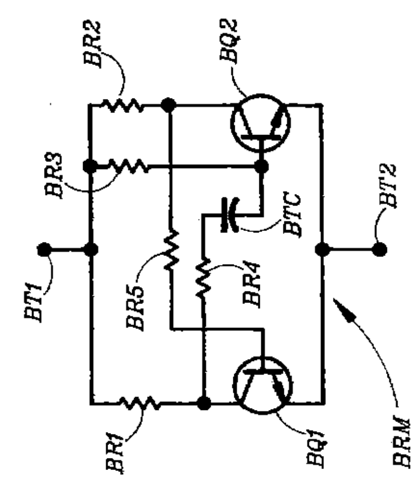
20 Claims, 5 Drawing Sheets





Qa
CDa
Jq
CDb
Qb
Ra1
SDa
Da
Ra2
Rb1
SDb
SCTa
SCTas
Ja
Ca
SCBbp
SCTbs
SCTap
Jx
SCTb
Jb
BB
Cb
BT2
AQ1
ACLR
FL2
L
BT1
BRM
CT
ZD
AQ2
B+ bus
Jy
C
Jz
Cc
Rc2
Dc2
AC
CHW
VSC
Dc1
Rc1
POT2
FL1
FCa
POT1
Jc
B+
FCb
B- bus
W1
W2
B-
MI
HFFC
SI
DC+
DC-
BR
PITa
S
PITb

Fig. 1



BR2
BQ2
BR3
BTC
BT2
BT1
BR5
BR4
BR1
BQ1
BRM

Fig. 2

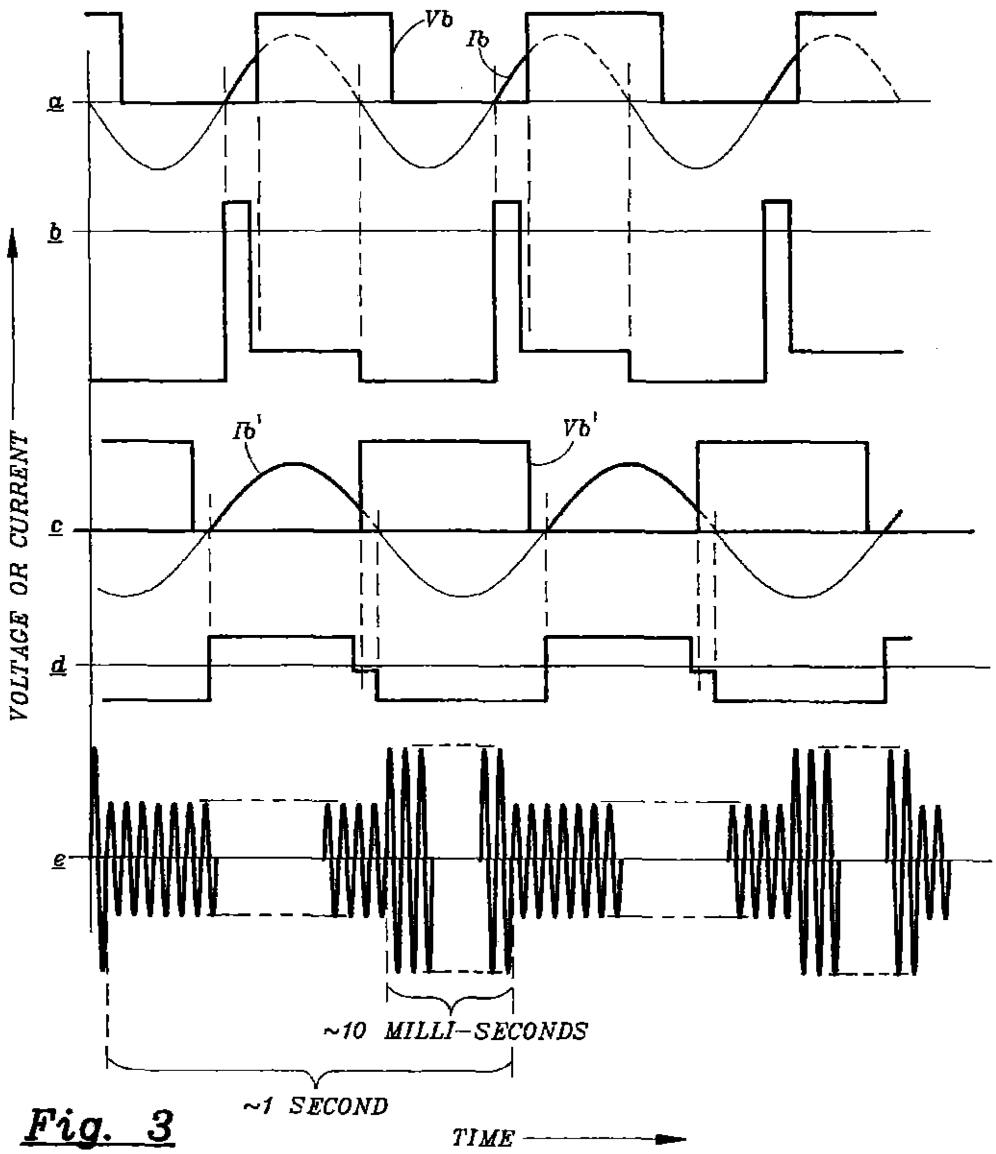

Vb
Ib
a
b
VOLTAGE OR CURRENT
Ib'
Vb'
c
d
e
~10 MILLI-SECONDS
~1 SECOND
TIME

Fig. 3

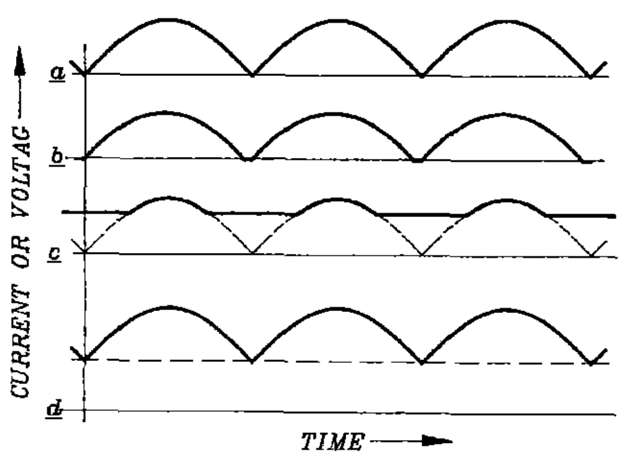

a
b
c
d
CURRENT OR VOLTAG
TIME

Fig. 7

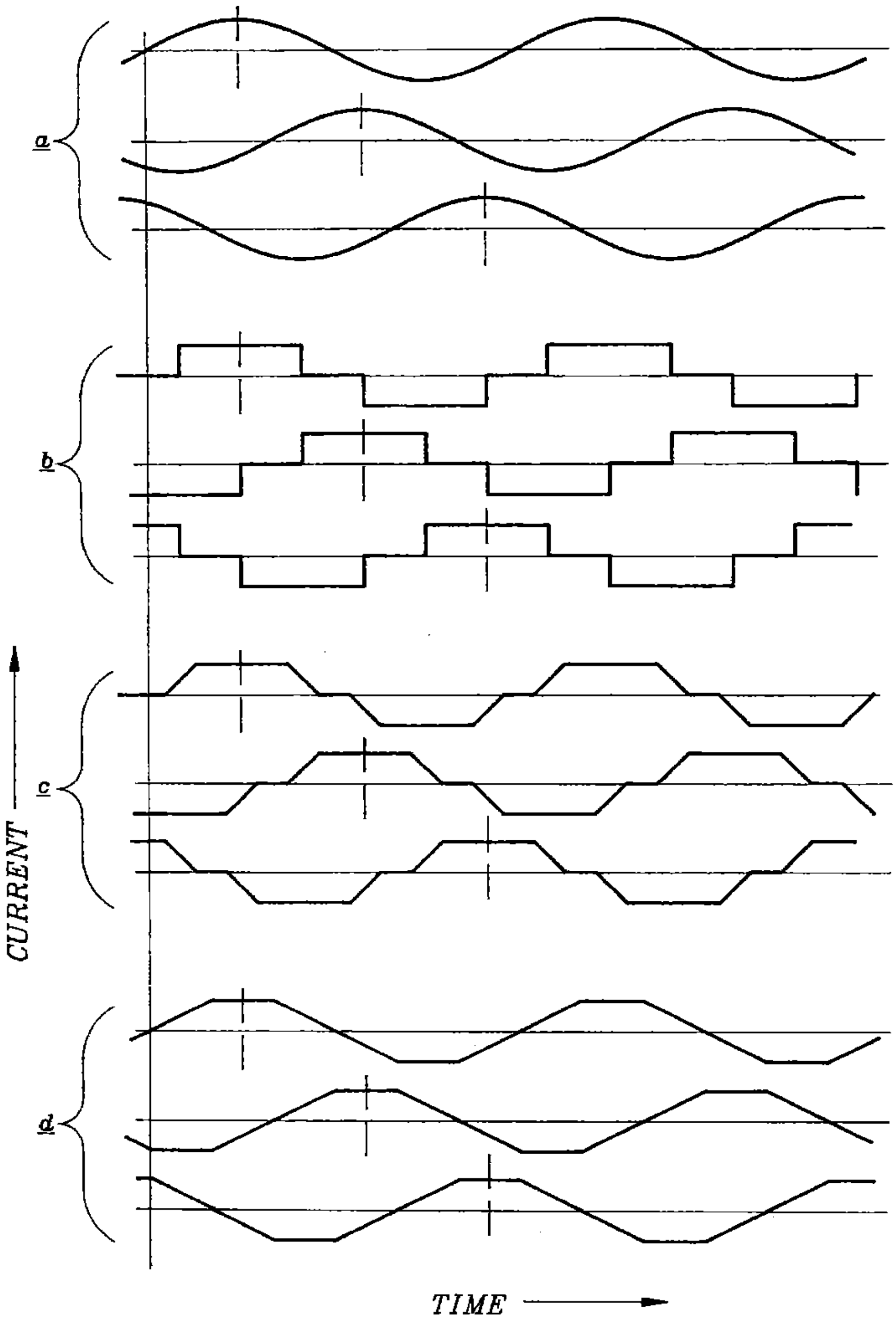


Fig. 4

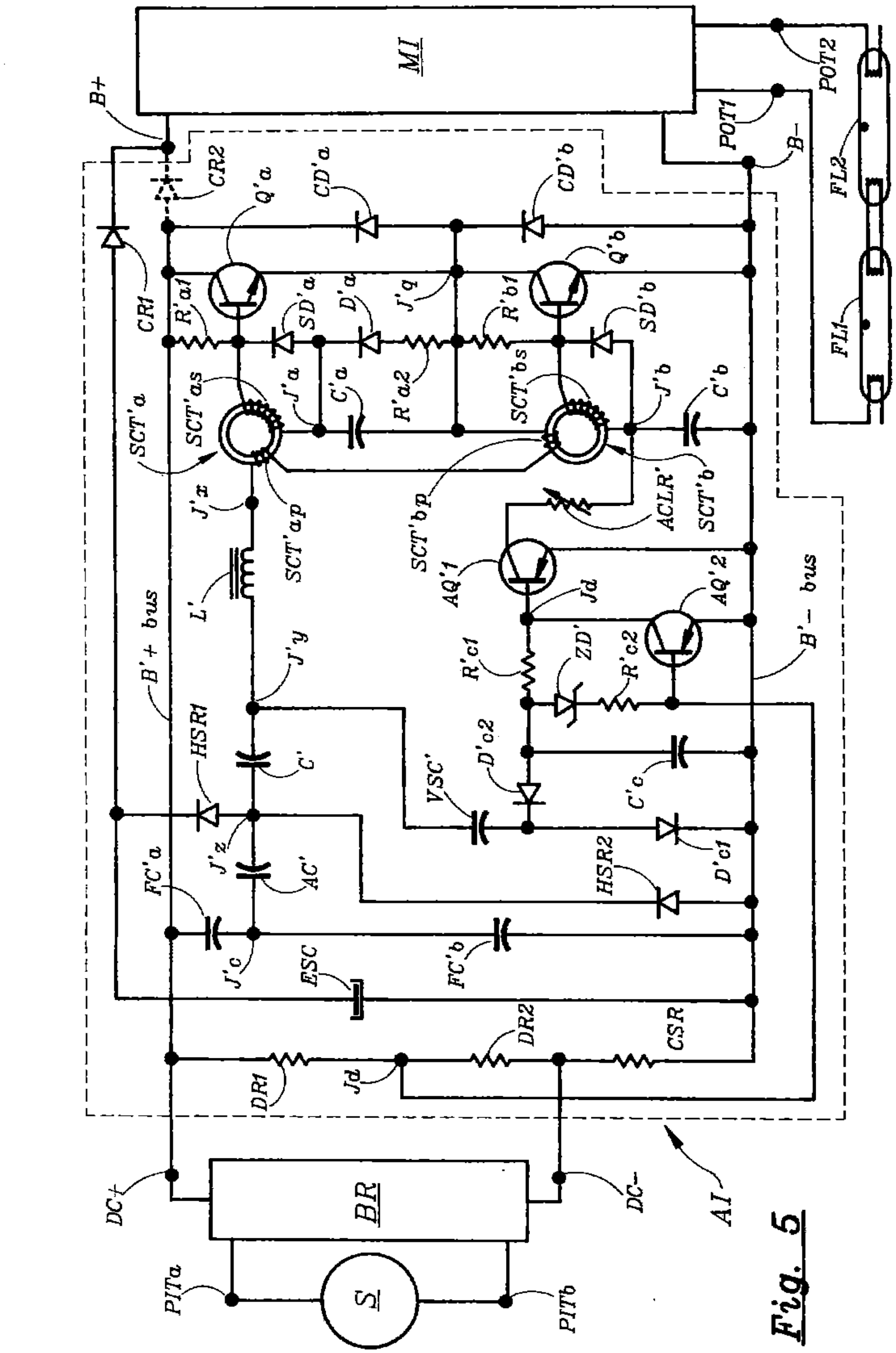

MI
B+
POT2
POT1
B–
FL2
FL1
CR2
CD'a
CD'b
Q'a
Q'b
CR1
R'a1
SD'a
D'a
J'q
R'b1
SD'b
J'a
C'a
R'a2
SCT'bs
J'b
C'b
SCT'as
SCT'a
J'x
SCT'ap
SCT'bp
ACLR'
SCT'b
L'
AQ'1
Jd
AQ'2
B'– bus
B'+ bus
J'y
R'c1
ZD'
R'c2
HSR1
C'
D'c2
VSC'
C'c
J'z
AC'
HSR2
D'c1
FC'a
J'c
ESC
FC'b
DR2
CSR
DR1
Jd
DC+
DC–
BR
AI
S
PITa
PITb

Fig. 5

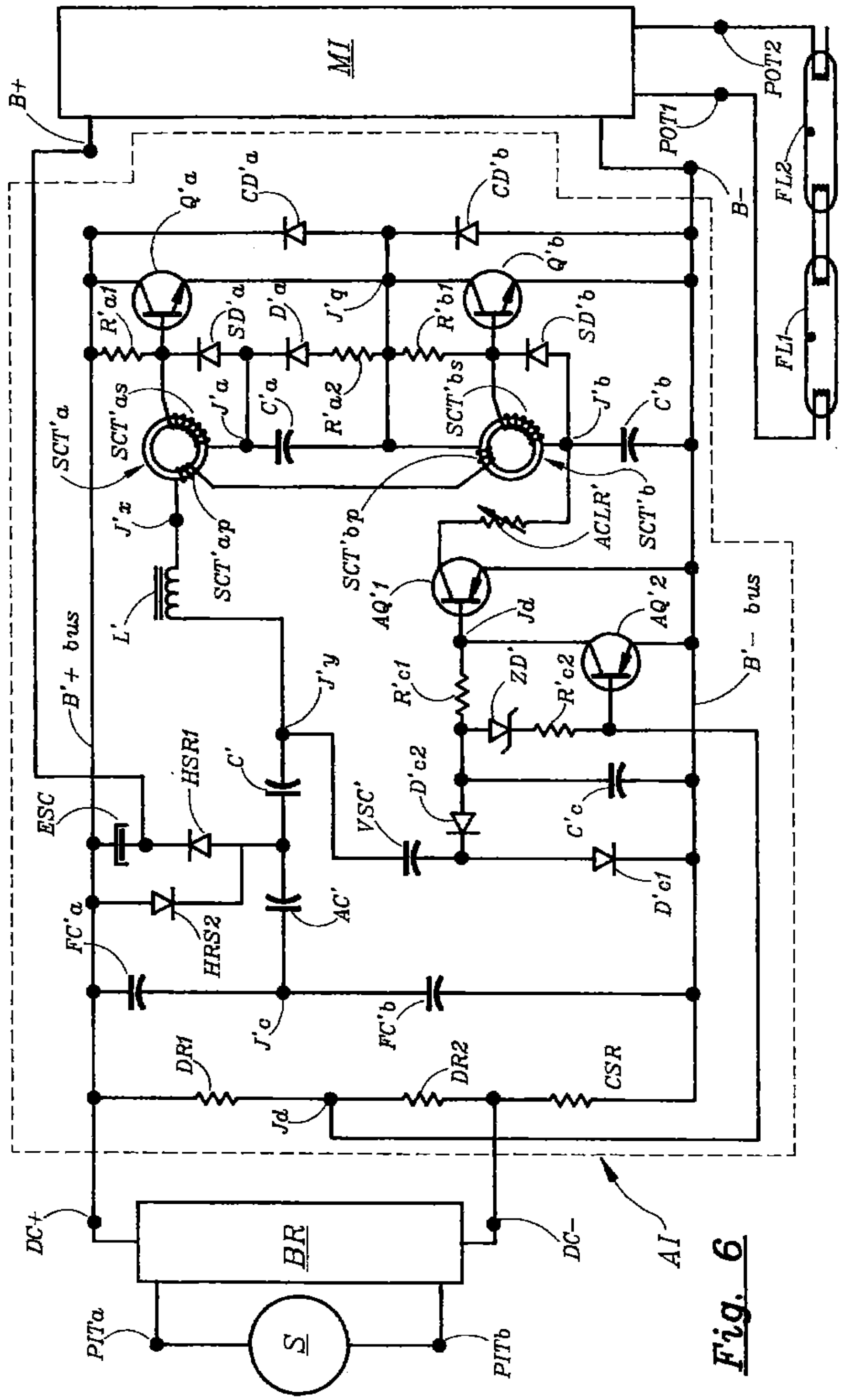

MI
B+
POT2
POT1
B−
FL2
FL1
CD'a
CD'b
Q'a
Q'b
R'a1
SD'a
D'a
J'q
R'b1
SD'b
SCT'a
SCT'as
J'a
C'a
R'a2
SCT'bs
J'b
C'b
J'x
SCT'ap
SCT'bp
ACLR'
SCT'b
AQ'1
Jd
AQ'2
L'
B'+ bus
B'− bus
J'y
R'c1
ZD'
R'c2
HSR1
C'
D'c2
ESC
VSC'
C'c
FC'a
HRS2
AC'
D'c1
J'c
FC'b
DR1
DR2
CSR
Jd
DC+
BR
DC−
AI
PITa
S
PITb

Fig. 6

CONTROLLED INVERTER-TYPE FLUORESCENT LAMP BALLAST

RELATED APPLICATIONS

Instant application is a Continuation-in-Part of Ser. No. 07/712,454 filed Jun. 10, 1991, now abandoned.

BACKGROUND OF THE INVENTION

1. Field of the Invention

The present invention relates to power-line-operated inverter-type power supplies for gas discharge lamps, especially of a type drawing power from the power line with a high power factor and with only a small amount of third harmonic current components.

2. Description of Prior Art

To provide a substantially constant-magnitude high frequency output current, an inverter-type ballast for a gas discharge lamp normally requires to be supplied with a nearly constant-magnitude DC voltage; and, when powered from a DC voltage obtained by straight-forward rectification and filtering of ordinary 60 Hz power line voltage, such a power supply draws current from the power line with a rather poor power factor.

Power factor correction is sometimes provided; but the associated power factor correction circuitry ordinarily involves the use of rather costly, bulky and energy-robbing 60 or 120 Hz inductors.

In particular, power supplies for gas discharge lamps must provide to such lamps an operating current of fairly low crest factor; yet, to be acceptable in most commercial applications, such power supplies must draw power from the power line with a high power factor as well as with a minimum amount of third harmonic current components.

SUMMARY OF THE INVENTION

Objects of the Invention

A basic object of the present invention is that of providing for a cost-effective inverter-type power supply means for gas discharge lamps.

A more specific object is a power-line-operated inverter-type power supply operable to provide to a gas discharge lamp a high frequency current of relatively low crest factor, yet drawing power from the power line with a relatively high power factor and low content of third harmonic current components.

Another object is an electronic fluorescent lamp ballast operable to power its lamp load with a current of low crest factor, yet to draw current from the power line at a high power factor and with minimal third harmonic content.

These as well as other objects, features and advantages of the present invention will become apparent from the following description and claims.

BRIEF DESCRIPTION

Unfiltered full-wave-rectified 60 Hz power line voltage is supplied to a first self-oscillating series-resonance-loaded inverter, the high frequency output current from which is magnitude-controlled, rectified and used for maintaining a substantially constant-magnitude DC voltage on an energy-storing capacitor. This constant-magnitude DC voltage is used in combination with the unfiltered full-wave-rectified 60 Hz power line voltage for powering a second self-oscillating series-resonance-loaded inverter, the high frequency current output from which is used for powering a fluorescent lamp load.

By frequency-modulating the first inverter at a 120 Hz rate, the current drawn from the power line is made to be of such waveshape as to result in a power factor well in excess of 90% while at the same time having a content of under 20% of odd triplets of third harmonic currents. The high frequency current provided to the fluorescent lamp load has a crest factor not in excess of 1.7.

Each of the two inverters is loaded by way of a series-tuned high-Q LC circuit connected across its output. In case of the first inverter, the energy-storing capacitor is charged with unidirectional current derived from across the tank-capacitor of its tuned LC circuit. In case of the second inverter, a pair of fluorescent lamps is series-connected across the tank-capacitor of its tuned LC circuit.

Each inverter has two bipolar transistors, each driven by an associated saturable current transformer that provides for a transistor ON-time dependent on the magnitude of an associated bias voltage.

One of the transistors of each inverter has a control arrangement connected in circuit with its associated saturable transformer and operative to control the magnitude of its associated bias voltage. As the magnitude of the bias voltage is controlled, so are the inverter's oscillating frequency, the magnitude of the voltage across the tank-capacitor, and the magnitude of the current available from the inverter's output.

In case of the first inverter, the magnitude of its bias voltage is automatically controlled such that the magnitude of the DC voltage developing across the energy-storing capacitor is maintained at a pre-determined substantially constant level, while at the same time providing for a desirable waveshape of the current drawn from the power line.

In case of the second inverter, before the fluorescent lamps ignite, the magnitude of the bias voltage is automatically controlled such that the magnitude of the high frequency voltage present across its associated tank-capacitor is maintained at a substantially constant level of magnitude somewhat higher than that of the operating voltage developing across the lamps after they have ignited. Then, after an initial one second period of cathode heating, to cause the lamps to ignite, the magnitude of the bias voltage is—for a period of about 10 milli-seconds—controlled such as to provide across the tank-capacitor a high frequency voltage of magnitude high enough to cause reliable lamp ignition. After the lamps have ignited, the magnitude of the bias voltage is controlled such as to maintain a lamp current of fairly constant magnitude.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1, in the form of an inverter-type power-line-operated fluorescent lamp ballast, diagrammatically illustrates an inverter of the type used for the first and second inverters.

FIG. 2 shows details of part of the control circuitry used in the ballast of FIG. 1.

FIG. 3 illustrates various voltage and current waveforms associated with the operation of the ballast of FIG. 1.

FIG. 4 shows various power line current waveforms having net zero current in the neutral wire of a three phase system.

FIG. 5 illustrates a preferred embodiment of the present invention, including the first and second inverters.

FIG. 6 illustrates an alternative embodiment of the present invention.

FIG. 7 illustrates some of the voltage and current waveforms associated with the embodiments of FIGS. 5 and 6.

DESCRIPTION OF THE PREFERRED EMBODIMENT

Details of Construction

In FIG. 1, a source S of ordinary 120 Volt/60 Hz power line voltage is applied to power input terminals PITa and PITb; which terminals, in turn, are connected with a bridge rectifier BR. The DC output from bridge rectifier BR is applied to a DC+ terminal and a DC− terminal, with the DC+ terminal being of positive polarity. A high-frequency filter capacitor HFFC is connected between the DC+ terminal and the DC− terminal. A first winding w1 of an EMI suppression inductor SI is connected between the DC+ terminal and a B+ terminal; and a second winding W2 of EMI suppression inductor SI is connected between the DC− terminal and a B− terminal. The B+ and the B− terminals are respectively connected with a B+ bus and a B− bus.

A filter capacitor FCa is connected between the B+ bus and a junction Jc; a filter capacitor FCb is connected between junction Jc and the B− bus. A switching transistor Qa is connected with its collector to the B+ bus and with its emitter to a junction Jq; a switching transistor Qb is connected with its collector to junction Jq and with its emitter to the B− bus. A commutating diode CDa is connected between the B+ bus and junction Jq, with its cathode connected with the B+ bus; a commutating diode CDb is similarly connected between junction Jq and the B− bus.

A saturable current transformer SCTa has a secondary winding SCTas connected between the base of transistor Qa and a junction Ja; a saturable current transformer SCTb has a secondary winding SCTbs connected between the base of transistor Qb and a junction Jb. Saturable current transformers SCTa and SCTb, respectively, have primary windings SCTap and SCTbp; which primary windings are series-connected between junction Jq and a junction Jx.

A resistor Ra1 is connected between the collector and the base of transistor Qa; a resistor Rb1 is connected between the collector and the base of transistor Qb. A capacitor Ca is connected between junction Ja and the emitter of transistor Qa; a capacitor Cb is connected between junction Jb and emitter of transistor Qb. A diode Da is connected with its cathode to junction Ja and, by way of a leakage resistor Ra2, with its anode to the emitter of transistor Qa. A shunt diode SDa is connected between the base of transistor Qa and junction Ja, with its anode connected with junction Ja. A shunt diode SDb is similarly connected between the base of transistor Qb and junction Jb.

An auxiliary transistor AQ1 is connected with its emitter to junction Jb and with its collector to the B− bus by way of an adjustable current-limiting resistor ACLR.

A tank-inductor L is connected between junction Jx and a junction Jy; and a tank-capacitor C is connected between junction Jy and a junction Jz. The primary winding of a cathode transformer CT is connected between junctions Jz and Jc, as is also an auxiliary capacitor AC. Cathode transformer CT has three secondary cathode heater windings collectively designated CHW.

A first power output terminal POT1 is connected with junction Jc; a second power output terminal POT2 is connected with junction Jy. First and second fluorescent lamps FL1 and FL2 are series-connected between power output terminals POT1 and POT2.

A voltage sampling capacitor VSC is connected between junction Jy and the cathode of a diode Dc1. The anode of diode Dc1 is connected with a bias bus BB, which is connected with junction Jb as well as with the emitter of auxiliary transistor AQ1. The anode of a diode Dc2 is connected with the cathode of diode Dc1; and the cathode of diode Dc2 is connected with the cathode of a Zener diode ZD.

A second auxiliary transistor AQ2 is connected with its emitter to bias bus BB and with its collector to the base of auxiliary transistor AQ1. A resistor Rc1 is connected between the base and the emitter of auxiliary transistor AQ2. A resistor Rc2 is connected between the base of auxiliary transistor AQ2 and the anode of Zener diode ZD.

A bistable resistor means BRM has a first terminal BT1 connected with the cathode of Zener diode ZD and a second terminal BT2 connected with the base of auxiliary transistor AQ1.

The elements connected together between the B+ and B− terminals on the one side and the POT1 and POT2 terminals on the other side is referred-to as main inverter MI.

FIG. 2 illustrates details of bistable resistor means BRM.

In FIG. 2, the emitters of first and second "bistable" transistors BQ1 and BQ2 are both connected with terminal BT2. The collector of transistor BQ1 is connected with terminal BT1 by way of a resistor BR1; the collector of transistor BQ2 is connected with terminal BT1 by way of a resistor BR2. A resistor BR3 is connected between terminal BT1 and the base of transistor BQ2. A resistor BR4 is series-connected with a timing capacitor BTC to form a series-combination; and this series-combination is connected between the collector of transistor BQ1 and the base of transistor BQ2. A resistor BR5 is connected between the collector of transistor BQ2 and the base of transistor BQ1.

FIG. 5 illustrates the preferred embodiment of the present invention.

In FIG. 5, a source S of ordinary 120 Volt/60 Hz power line voltage is applied to power input terminals PITa and PITb; which terminals, in turn, are connected with a bridge rectifier BR. The DC output from bridge rectifier BR is applied to a DC+ terminal and a DC− terminal, with the DC+ terminal being of positive polarity. The DC+ terminal is connected with a B'+ bus, and the DC− terminal is connected with a B'− bus by way of a current sensing resistor CSR.

A resistor DR1 is connected between the DC+ terminal and a junction Jd; and a resistor DR2 is connected between junction Jd and the DC− terminal.

An energy-storing capacitor ESC is connected between the B'− bus and the anode of a commutating rectifier CR1, the cathode of which is connected with the B+ terminal of main inverter MI. Another commutating rectifier CR2 is connected between the B'+ bus and the B+ terminal, its cathode being connected with the B+ terminal. The B− terminal of main inverter MI is connected with the B'− bus.

Fluorescent lamps FL1 and FL2 are series-connected across power output terminals POT1 and POT2 of main inverter MI.

A filter capacitor FC'a is connected between the B'+ bus and a junction J'c; a filter capacitor FC'b is connected between junction J'c and the B'− bus. A switching transistor Q'a is connected with its collector to the B'+ bus and with its emitter to a junction J'q; a switching transistor Q'b is connected with its collector to junction J'q and with its emitter to the B'+ bus. A commutating diode CD'a is connected between the B'+ bus and junction J'q, with its cathode connected with the B'+ bus; a commutating diode CD'b is similarly connected between junction J'q and the B'− bus.

A saturable current transformer SCT'a has a secondary winding SCT'as connected between the base of transistor

Q'a and a junction J'a; a saturable current transformer SCT'b has a secondary winding SCT'bs connected between the base of transistor Q'b and a junction J'b. Saturable current transformers SCT'a and SCT'b, respectively, have primary windings SCT'ap and SCT'bp; which primary windings are series-connected between junction J'q and a junction J'x.

A resistor R'a1 is connected between the collector and the base of transistor Q'a; a resistor R'b1 is connected between the collector and the base of transistor Q'b. A capacitor C'a is connected between junction J'a and the emitter of transistor Q'a; a capacitor C'b is connected between junction J'b and emitter of transistor Q'b. A diode D'a is connected with its cathode to junction J'a and, by way of a leakage resistor R'a2, with its anode to the emitter of transistor Q'a. A shunt diode SD'a is connected between the base of transistor Q'a and junction J'a, with its anode connected with junction J'a. A shunt diode SD'b is similarly connected between the base of transistor Q'b and junction J'b.

A tank-inductor L' is connected between junction J'x and a junction J'y; and a tank-capacitor C' is connected between junction J'y and a junction J'z. An auxiliary capacitor AC' is connected between junctions J'z and J'c.

A first high speed rectifier HSR1 is connected with its anode to junction J'z and with its cathode to the anode of rectifier CR1. A second high speed rectifier HSR2 is connected with its cathode to junction J'z and with its anode to the B'− bus.

A first auxiliary transistor AQ'1 is connected with its emitter to the B'− bus and with its collector to junction J'b by way of an adjustable current-limiting resistor ACLR'.

A voltage sampling capacitor VSC' is connected between junction J'y and the anode of a diode D'c1. The cathode of diode D'c1 is connected with the B'− bus. The cathode of a diode D'c2 is connected with the anode of diode D'c1; and the anode of diode D'c2 is connected with the anode of a Zener diode ZD'. A filter capacitor C'c is connected between the anode of diode D'c2 and the B'− bus. A resistor R'c1 is connected between the anode of diode D'c2 and the base of transistor AQ'1.

A second auxiliary transistor AQ'2 is connected with its emitter to the B'− bus and with its collector to the base of auxiliary transistor AQ'1; which base is connected with junction Jd. A resistor R'c2 is connected between the base of auxiliary transistor AQ'2 and the cathode of Zener diode ZD'.

The elements connected together between the DC+ and DC− terminals and the B+ and B− terminals are referred-to as auxiliary inverter AI.

FIG. 6 illustrates an alternative ambodiment of the present invention.

The arrangement of FIG. 6 is identical to that of FIG. 5 except as follows.

In FIG. 6: (i) rectifier CR1 has been replaced with a short circuit; (ii) rectifier CR2 has been replaced with an open circuit; (iii) the anode of rectifier HSR2 has been disconnected from the B'− bus and connected with the B'+ bus instead; and (iv) one of the terminals (the negative terminal) of energy storing capacitor ESC has been disconnected from the B'− bus and connected with the B'+ bus instead.

Details re Waveforms of FIG. 4

In a situation of an ordinary three-phase power distribution system, where each individual phase supplies a 60 Hz sinusoidal voltage, the waveforms of FIGS. *4a–4d* represent the currents resulting in each of the three power conductors under four different types of load conditions. In each of these four cases, the current flowing in the neutral conductor will be zero under conditions of symmetrical loading on each phase.

FIG. *4a* represents a situation where each of the loads draws a sinusoidal current; which, as is well known, results in zero current flowing in the three phase neutral conductor. That this is so may be ascertained by simply adding the instantaneous magnitudes of the three different waveforms: at each point along the time axis, the sum will be zero.

FIG. *4b* represents a situation where each of the loads draws a current of a constant magnitude for two thirds of the total duration of each half cycle of the power line voltage, and then a zero-magnitude current for the remainder of each half cycle. In this case the three currents also will add up to zero on an instantaneous basis all along the time axis.

FIGS. *4c* and *4e* represent situations equivalent to that of FIG. *4b*, except for having the current magnitudes increase and decrease more or less gradually as contrasted with the abrupt increases and decreases of the current magnitudes in FIG. *4b*. In each of these cases the three currents also will add up to zero on an instantaneous basis all along the time axis.

Details of Operation

The basic operation of the half-bridge inverter of FIG. 1 is conventional and is explained in conjunction with FIG. 3 of U.S. Pat. No. 4,307,353 to Nilssen.

For a given magnitude of the DC supply voltage, due to the effect of the high-Q LC circuit, the magnitude of the current provided to the fluorescent lamp load (or to any other load presented to the output) is a sensitive function of the frequency and the waveshape of the inverter's output voltage; which output voltage is a substantially squarewave voltage of controllable frequency and with peak-to-peak magnitude about equal to that of the instantaneous magnitude of the DC voltage present between the B− bus and the B+ bus.

The frequency of the inverter's squarewave output voltage is a sensitive function of the natural resonance frequency of the high-Q LC circuit as well as of the duration of the forward conduction period (i.e., the ON-period) of the two inverter switching transistors; which duration, in turn, is a sensitive function of the saturation characteristics of saturable current transformers SCTa and SCTb as combined with the magnitude of the bias voltages present on capacitors Ca and Cb. That is, the duration of the forward conduction period (the ON-time) of each switching transistor is determined by the volt-second product sustainable by its associated saturable current transformer as well as by the magnitude of the negative bias on capacitors Ca and Cb: the higher the volt-second product available before saturation, the longer the ON-time; the higher the negative bias on the Ca/Cb capacitors, the shorter the ON-time.

In the circuit arrangement of FIG. 1, the magnitude of the negative voltage on capacitors Ca and Cb is determined by the magnitude of the current provided to the bases of transistors Qa and Qb, less any current drained away through resistors Ra2 and AQ1/ACLR, all respectively. (Of course, a small amount of current is also drained away from bias capacitors Ca and Cb by resistors Ra1 and Rb1, respectively. However, this amount of charge leakage is in most situations negligible. Resistors Ra1 and Rb1 are principally used for getting the inverter to initiate oscillation.)

The magnitude of the base current provided to each transistor is directly proportional to the magnitude of the current flowing through the primary windings of saturable current transformers SCTa and SCTb. Thus, assuming transistor AQ1 to be conducting, for given values of resistors R2a and ACLR: the higher the magnitude of the inverter's output current, the higher the magnitude of the negative voltage on capacitors Ca and Cb.

Thus, for given values of resistors Ra2 and ACLR (assuming transistor AQ1 to be conducting), the circuit of FIG. 1 provides for a high degree of regulation of the magnitude of the inverter's output current.

By selecting a suitable resistance value for resistor Ra2, and assuming transistor AQ1 to be conducting, the magnitude of the inverter's output current may be adjusted by adjusting the resistance value of ACLR: a relatively low resistance value leads to an inverter output current of relatively high magnitude; a relatively high resistance value leads to an inverter output current of relatively low magnitude.

The higher the magnitude of the negative voltage on each bias capacitor, the higher the magnitude of the voltage that has to be provided from the secondary winding of each saturable current transformer; which, in turn, leads to a correspondingly shorter period before saturation is reached. Thus, as the magnitude of the negative bias on each bias capacitor is increased, the duration of each transistor's forward conduction period (ON-time) is decreased; which, in turn, leads to a reduction in the magnitude of the inverter's output current in comparison with what it otherwise would have been.

Whereas the base current provided to each transistor has to flow from its associated bias capacitor, the reverse or reset current provided from each of the saturable current transformer's secondary windings does not flow from the bias capacitor, but rather flows in a separate path through the reverse shunt diode (SDa or SDb) shunting the secondary winding of each saturable current transformer.

More particularly, the circuit and control arrangement of FIG. 1 operates as follows.

As power is applied at power input terminals PITa/PITb, the inverter starts to oscillate at a frequency near the natural self-resonance frequency of the LC circuit. The resulting inverter output current results in a positive feedback current provided to each base; and this feedback current, in turn, causes a negative bias to build up on each of bias capacitors Ca/Cb. As the magnitude of the negative bias voltage increases, the inverter's oscillation frequency increases as well. As a result, the magnitude of the inverter output current will stabilize at a some level determined by the effective resistance values of resistors Ra2 and AQ1/ACLR.

Transistor AQ1 conducts to a degree dependent on the magnitude of the base current it receives; which magnitude, in turn, depends on: (i) the magnitude of the high frequency voltage present at junction Jy; (ii) the reactance value of capacitor VSC; (iii) the magnitude of the Zener voltage of Zener diode ZD; and (iv) the effective resistance value of bistable resistance means BRM.

More particularly, the control circuit consisting of principal elements AQ1, ACLR, BRM, AQ2, ZD and VSC, operates as follows: (i) capacitor VSC is of small capacitance value relative to that of tank capacitor C, and—from the perspective of the magnitudes of the voltages present within the control circuit—may be considered as representing a current source; (ii) thus, the magnitude of the high frequency current supplied via capacitor VSC is a measure of the magnitude of the high frequency voltage present at junction Jy; (iii) the high frequency current provided via capacitor VSC is rectified and filtered, and results in a corresponding unidirectional current flowing through resistance gleans BRM and into the base of transistor AQ1, thereby causing this transistor to become fully conductive; (iv) as the magnitude of the high frequency current increases, so does the magnitude of the unidirectional current flowing through resistance means BRM as well as the magnitude of the DC voltage developing across resistance means BRM; (v) eventually a point is reached at which the magnitude of this DC voltage gets to be so high as to cause current to flow through Zener diode ZD and into the base of transistor QA2; (vi) as current flows into the base of transistor AQ2, it becomes conductive; (vii) as transistor AQ2 becomes conductive, it robs base current from transistor AQ1, thereby making transistor AQ1 less conductive; (viii) as transistor AQ1 becomes less conductive, the magnitude of the bias voltage across capacitor Cb will increase, thereby shortening the ON-time of transistor Qb; and (ix) the shortening ON-time of transistor Qc will cause the magnitude of the high frequency current supplied to tank capacitor C to decrease, thereby decreasing the magnitude of he high frequency voltage present at junction Jy.

Thus, a negative feedback loop exists: as the magnitude of the high frequency voltage provided at junction Jy increases beyond a pre-established level (which, for a given value of capacitor VSC, is determined by the Zener voltage of Zener diode ZD and the effective resistance value of resistance means BRM), the ON-time of transistor Qb becomes proportionally shortened. As a net result, the magnitude of the high frequency voltage at junction Jy will be regulated such as—in effect—not to exceed the above-indicated pre-established level. Thus, as long as the magnitude of the B+ supply voltage exceeds a certain minimum level, further increases in the magnitude of the B+ supply voltage will not cause further increases in the magnitude of the high frequency voltage.

By varying the effective resistance value of bistable resistance means BRM, the magnitude of the high frequency voltage at junction Jy will correspondingly vary. Bistable resistance means BRM is of such a nature as to self-oscillate between two distinct states: (i) a first state in which transistor BQ1 is ON and transistor BQ2 is OFF; and (ii) a second state in which transistor BQ1 is OFF and transistor BQ2 is ON. The components of bistable resistance means BRM were so chosen as to cause the first state to exist for about 10 milli-seconds and the second state to exist for about 1000 milli-seconds. As a consequence, resistors BR1 and BR2 are alternatingly switched in and out, thereby causing resistance means BRM to appear as a relatively low resistance value (equivalent to the effective resistance value of resistor BR2) for about 10 milli-seconds once each 1000 milli-seconds, while otherwise appearing as a relatively high resistance value (i.e., the effective resistance value of BR1).

As a result of causing bistable resistance means to alternate on the above-indicated manner, the magnitude of the high frequency voltage at junction Jy alternates correspondingly—being of a relatively large magnitude for 10 milli-seconds out of every 1000 milli-seconds.

The magnitude of the high frequency voltage during the indicated 10 milli-second periods is high enough to cause the fluorescent lamps to ignite, but only after the lamp cathodes have become thermionic. When provided with the proper amount of cathode heating power, the lamp cathodes will become thermionic within a period of less than 1000 milli-seconds.

With the fluorescent lamps connected and fully operating (i.e., fully loading the LC circuit), the magnitude of the high frequency voltage at junction Jy is kept below that magnitude required to cause current to flow into the base of transistor AQ2; which means that transistor AQ1 will be fully conductive. At this point, the control circuit is effectively disabled; and the magnitude of the lamp current is then determined by the resistance value of resistor ACLR: varying ACLR causes the magnitude of the lamp current to vary.

The resistance value of resistor Ra2 is selected such that the ON-time of transistor Qa corresponds to nearly a 50% duty-cycle. Similarly, the minimum resistance value of ACLR (which corresponds to maximum lamp current) is selected such that the ON-time of transistor Qb corresponds to nearly a 50% duty-cycle.

Heating of the lamp cathodes is accomplished by way of cathode transformer CT; which transformer is connected in parallel with auxiliary capacitor AC, the capacitance value of which is very large compared with that of tank capacitor C. Thus, the cathodes are heated with a substantially sinusoidal high frequency voltage.

When the circuit of FIG. 1 operates with no load (as before the lamp cathodes have become thermionic), the RMS magnitude of the high frequency voltage present at junction Jy is limited by the indicated negative feedback to be somewhat higher than the magnitude of the high frequency voltage present at junction Jy with the lamps connected and operating, but not so high as to cause instant-starting of the lamps. The magnitude of the cathode heating voltages provided during this no-load condition is chosen to be such as to cause cathode incandescence to occur within about one second. As a result, the lamps will normally ignite in a rapid-start manner on the first occasion where the magnitude of the high frequency output voltage is made extra high for the indicated 10 milli-second period.

During the 10 milli-second period, the magnitude of the cathode heating voltages also increases substantially, thereby aiding in lamp ignition. However, on an integrated RMS basis, this brief period of increased cathode heating voltage is of little consequence.

As soon as the lamps ignite, the magnitude of the high frequency voltage at junction Jy will decrease to a level so low that the amount of high frequency current provided via capacitor VSC will be insuffient to cause Zener diode ZD to conduct, thereby causing transistor AQ2 to enter its non-conducting state and to remain there for as long as the lamps remain ignited. As a result, transistor AQ1 will now be fully conducting on a continuous basis. This implies that the 10 milli-second pulses, which will continue to be provided every 1000 milli-seconds, will be of little consequence: they can't turn transistor AQ1 more than fully ON.

If the lamps were to be disconnected, however, the repeatedly provided 10 milli-second pulses will assure that affirmative lamp ignition will occur as soon as fully functional lamps are indeed connected.

As long as transistor AQ1 is conductive, the fluorescent lamps will be powered in a normal manner; and the magnitude of the lamp current flowing will depend on the particular setting of adjustable resistor ACLR.

FIG. 3 depicts various voltage and current waveforms associated with the circuit of FIG. 1.

For a situation with no loading presented to the high-Q LC circuit—that is, with the lamps disconnected, or before the lamps have ignitied—FIG. 3*a* shows the collector-to-emitter voltage Vb of transistor Qb and the corresponding inverter output current Ib. The part of Ib actually flowing through transistor Qb in the forward direction is shown in heavy solid line, the part of Ib flowing through commutating diode CDa is showed in light dashed line, and the part of Ib flowing through either Qa or CDb is shown in light solid line.

FIG. 3*b* shows the base-emitter voltage of transistor Qb as it corresponds to the waveforms of FIG. 3*a*.

For a situation where the LC circuit is substantially fully loaded by the two fluorescent lamps, FIG. 3*c* shows the collector-to-emitter voltage Vb' of transistor Qb and the corresponding inverter output current Ib'. The part of Ib' actually flowing through transistor Qb in the forward direction is shown in heavy solid line, the part of Ib' flowing through commutating diode CDa is showed in light dashed line, and the part of Ib' flowing through either Qa or CDb is shown in light solid line.

FIG. 3*d* shows the base-emitter voltage of transistor Qb as it corresponds to the waveforms of FIG. 3*c*.

FIG. 3*e* shows the waveshape of the high-frequency voltage present across the tank-capacitor under the condition of an unloaded LC circuit: a continuous substantially sinusoidal voltage of a relatively low magnitude, interrupted once each second (i.e., 1000 milli-seconds) with a 10 milli-second long burst of relatively high-magnitude substantially sinusoidal voltage.

In the arrangement of FIG. 5, there are two inverters: one is main inverter MI, whose operation is already described in connection with FIG. 1; the other is an auxiliary inverter, whose operation is essentially the same as that of main inverter MI. However, this auxiliary inverter is used differently.

More particularly, its high frequency output is rectified via high speed rectifiers HSR1 and HSR2, and the resulting unidirectional current is used for charging energy-storing capacitor ESC. In the absence of any loading on this energy-storing capacitor, the voltage-magnitude to which it charges is determined by the characteristics of the negative feedback control circuit: in the same manner as is the maximum magnitude of open circuit output voltage determined in the circuit arrangement of FIG. 1.

More particularly, the maximum voltage to which capacitor ESC charges is determined by the capacitance ratio between auxiliary capacitor AC' and tank capacitor C', as well as by the magnitude to which the high frequency voltage at junction J'y is regulated.

With commutating rectifier CR2 non-connected, or with the magnitude of the voltage on capacitor ESC larger than the peak magnitude of the rectified power line voltage, main inverter MI will be powered entirely from the voltage on capacitor ESC.

However, with commutating rectifier CR2 connected, and with the magnitude of the voltage on capacitor ESC lower than the peak magnitude of the rectified power line voltage, main inverter MI gets powered directly from the rectified power line voltage as long as the instantaneous magnitude of this rectified power line voltage is larger than the magnitude of the voltage on capacitor ESC. Then, when the instantaneous magnitude of the rectified power line voltage is lower than the magnitude of the voltage on capacitor ESC, main inverter MI gets powered by the voltage on capacitor ESC.

Otherwise, by virtue of the voltage established at junction Jd by action of the voltage divider consisting of resistors DR1 and DR2, as combined with the voltage established across current sensing resistor CSR (which is a resistor of relatively low resistance value, such as perhaps 1 to 10 Ohm), a situation is established whereby the instantaneous magnitude of the current drawn from the power line is prevented from exceeding a certain pre-established fraction of the instantaneous magnitude of the power line voltage; which pre-established fraction is determined by the resistance value of resistor DR2 versus that of resistor DR1.

More particularly, if the instantaneous magnitude of the current drawn from the power line through resistor CSR were to generate a larger voltage across CSR than the voltage established across resistor DR2, the net result would be that the voltage established at junction Jd would be negative with respect to the B'− bus. Thus, whenever the instantaneous magnitude of the voltage present across resistor CSR is larger than that of the voltage present across resistor DR2, junction Jd will present a voltage to the base of transistor AQ'2 that is negative with respect to its emitter, thereby causing this transistor to become more conductive, thereby causing transistor AQ'1 to become less conductive, thereby reducing the magnitude of the current delivered to capacitor ESC, thereby reducing the magnitude of the current drawn from the power line, etc.

The arrangement of FIG. 6 operates in the same manner as that of FIG. 5 except that energy-storing capacitor ESC has now been repositioned in such manner that its voltage is now added to the rectified power line voltage before being presented to main inverter MI.

FIG. 7 depicts various voltage and current waveforms associated with the arrangements of FIGS. 5 and 6.

FIG. 7*a* indicates the waveshape of the full-wave-rectified power line voltage, which is present between the DC+ and the DC− terminals.

FIG. 7*b* indicates the waveshape of the current drawn from the power line; which waveshape, under most conditions, is nearly sinusoidal.

FIG. 7*c* indicates the waveshape of the voltage presented to main inverter MI in the arrangement of FIG. 5 under a condition where the peak magnitude of the rectified power line voltage is about 25% higher that the magnitude of the voltage on capacitor ESC.

FIG. 7*d* indicates the waveshape of the voltage presented to main inverter MI in the arrangement of FIG. 6.

Additional Comments

a) Detailed information relative to a fluorescent lamp ballast wherein the fluorescent lamp is powered by way of a series-excited parallel-loaded L-C resonant circuit is provided in U.S. Pat. No. 4,554,487 to Nilssen.

b) The instantaneous peak-to-peak magnitude of the squarewave output voltage provided by each of the various half-bridge inverters between junctions Jq and Jc is substantially equal to the instantaneous magnitude of the DC voltage supplied to that inverter.

c) Saturable current transformers SCTa and SCTb require only a miniscule amount of voltage across their primary windings. Hence, the magnitude of the voltage-drop between junctions Jq & Jx is substantially negligible, and the inverter's full output voltage is therefore effectively provided across the LC circuit, which principally consists of tank-capacitor C and tank-inductor L. However, there is a small tuning effect associated with auxiliary capacitor AC of FIG. 1 and capacitor AC' of FIGS. 5 and 6.

d) in FIG. 3, the inverter frequency associated with the waveforms of FIGS. 3*a* and 3*b* is substantially higher than that associated with FIGS. 3*c* and 3*d*.

Also, current Ib is nearly 180 degrees out of phase with the fundamental frequency component of voltage Vb, while current Ib' is almost in phase with voltage Vb'.

e) In the situation associated with the waveform of FIG. 3*b*, the magnitude of the voltage "seen" by the secondary winding of saturable current transformer SCTb is about five times as high as that "seen" by the same secondary winding in the situation associated with FIG. 3*d*.

Correspondingly, the duration of the transistor ON-time in the situation associated with FIG. 3*d* is about five times longer than the transistor ON-time in the situation associated with the waveform of FIG. 3*b*.

f) As may be noticed in FIG. 3*a*, transistor Qb ceases to conduct in its forward direction while a substantial amount of current is still flowing from the inverter's output. After transistor Qb has ceased to conduct, the inverter's output current will continue to flow until the energy in the tank inductor has dissipated itself. However, the output current will continue its flow through commutating diode CDa, thereby discharging its energy into the DC power supply.

g) Forward conduction of a transistor is defined as current flowing, with the aid of forward base drive current, directly between the collector and the emitter; which, in case of transistor Qb for, instance, means that forward current is defined as positive current flowing from its collector to its emitter while drive current is being provided to its base.

A transistor's ON-time is defined as the period during which it conducts current in the forward direction.

h) In FIG. 3 it is noted that the fundamental frequency of the waveforms depicted in FIGS. 3*c* and 3*d* is lower by a certain factor as compared with the frequency associated with the waveforms of FIGS. 3*a* and 3*b*; yet the indicated duration of transistor ON-time associated with the waveforms of FIGS. 3*a* and 3*b* is shorter by a much larger factor as compared with the indicated duration of transistor ON-time associated with the waveforms of FIGS. 3*c* and 3*d*.

In fact, when the transistor ON-time is shortened by a given proportion, the fundamental frequency of the inverter's output voltage increases by a much smaller proportion. In instant case, with each transistor's ON-time shortened by a factor of about five, the inverter frequency increased only by a factor of about 1.3: from about 30 kHz to about 40 kHz.

i) The time constant associated with each bias capacitor and its associated charge leakage means (ex: Da and Ra2) is normally on the order of a single period of the high frequency inverter output voltage. For instance, for a situation where the power line input voltage is 120 Volt/60 Hz, the frequency of the inverter output voltage/current is on the order of 30 kHz, and the total inverter power output falls in the range between 10 and 100 Watt, the values of bias capacitors Ca and Cb might reasonably be in the range from 0.5 to 5 micro-Farad, the value for leakage resistor Ra2 might reasonably be in the range between 5 and 50 Ohm, and adjustable resistor ACLR might reasonably be adjustable over a range between 0 and 50 Ohm.

j) In the circuit arrangement of FIG. 1 there are two distinctly different kinds of current-magnitude-limiting provided. One is the ordinary kind associated with the natural characteristics of a series-excited parallel-loaded resonant LC circuit; another is due to the action of the control circuit associated with auxiliary transistors AQ1 and AQ2.

The former is the principal means for limiting the lamp current; the latter is the principal means for controlling the level at which the lamp current is limited as well as for limiting the inverter's output current in the absence of proper circuit loading.

k) The circuit arrangement of FIG. 1 may be described as an inverter that is loaded by way of a high-Q tuned LC circuit and arranged to self-oscillate by way of positive feedback derived from the inverter 's instantaneous output current (and/or voltage) while at the same time arranged to provide for controllable-magnitude output current (and/or voltage) by way of negative feedback derived from the average absolute magnitude of the inverter's output current (and/or voltage). That is, a larger magnitude output current provides for a larger magnitude bias voltage on capacitors Ca and Cb, thereby causing the inverter frequency to increase and the inverter's output current to decrease.

l) So as to fully reset the saturable cores each cycle, diodes SDa and SDb should each have a relatively high-magnitude forward voltage drop, such as might be obtained by using two ordinary diodes in series. However, instead of using special diodes with high-magnitude forward voltage drops, it is acceptable to use ordinary diodes with added series-resistors, thereby effectively to increase their forward voltage drops.

m) Some of the values associated with operating the ballast with the kind of waveform indicated by FIG. 3*e* are as follows: (i) substantially relaxed specifications for the tank-inductor; (ii) similarly relaxed specifications for the tank-capacitor; (iii reduced glow current prior to lamp ignition, thereby providing for increased lamp life; (iv) much improved lamp starting; (v) substantially reduced idling power; and (vi) more cost-effective compliance with U.L. specifications related to ground-fault current.

n) The RMS magnitude of the cathode heating voltage, which voltage is provided to each of the lamps' thermionic cathodes by way of cathode heating windings CHW, is such as to provide for proper cathode heating during the period before the lamps ignite, as well as on a continuous basis thereafter.

During the brief pulses provided by way of bistable resistor means BRM, the RMS magnitude of the cathode heating voltage is increased to about twice normal value. However, since the duration of each of these pulses is so very brief (about 10 milli-seconds) compared with the duration of each of the periods between such pulses (about 1000 milli-seconds), the net effect on the temperature of the cathodes is negligible. However, with respect to lamp ignition, the effect is substantial and beneficial. The briefly elevated RMS magnitude of the cathode voltage gives rise to ionization of the lamp gas along the cathodes' surfaces, thereby greatly facilitating the ignition of the main gas columns of the lamps.

o) By appropriately modifying the voltage divider means, which in the arrangements of FIGS. 5 and 6 is represented by resistors DR1 and DR2, the waveshape of the current drawn from the power line can be correspondingly modified. For instance, by placing a voltage-limiting means, such as a diode, in parallel with resistor DR2, the line current waveshape can be made to be like that illustrated in FIG. 4*d*.

p) When a resonant LC circuit is series-excited with a voltage source and parallel-loaded by a substantially fixed-magnitude voltage, such as is indeed the case with the arrangement of FIG. 5 (where the load is capacitor ESC; which capacitor, under normal conditions, will indeed be charged to a voltage of substantially fixed magnitude), the magnitude of the current drawn from the voltage source will be substantially constant regardless of the magnitude of the voltage provided from this voltage source.

This constant-voltage to constant-current conversion characteristic is basic with parallel-loaded series-excited resonant LC circuits; which have the characteristic of converting a constant-magnitude voltage series-input to a constant-magnitude current parallel-output; or, conversely, converting a constant-magnitude voltage parallel-output to a constant current series-input.

Thus, without any attempts at changing the frequency of the inverter's output voltage, the waveshape of the current drawn from the power line would have been substantially a continuous constant magnitude, regardless of the instantaneous magnitude of the power line voltage.

However, by dynamically changing the frequency of the inverter's output voltage in accordance with some desired pattern or algorithm (at a 120 Hz frequency), the magnitude of the current drawn by the inverter from the power line can be made to change correspondingly over a wide range. More particularly, the current drawn from the power line can readily be made to conform to various desired waveshapes, such as those of FIG. 4, for instance.

q) With reference to FIG. 7*c*, which illustrates the waveshape of the voltage presented to main inverter MI of FIG. 5, it is noted that the instantaneous magnitude of this unidirectional voltage is equal to the larger of: (i) a certain substantially constant level (as represented by the DC voltage present on capacitor ESC); and (ii) the full-wave-rectified power line voltage which is provided across the output terminals of rectifier BR and which is illustrated by FIG. 7*b*).

In this connection, it is noted that the voltage-drop across current-sensing resistor csr is assumed to be of negligible magnitude in comparison with the magnitude of the DC voltage supplied to main inverter MI.

r) With reference to FIGS. 5 and 6, for convenience, all the circuitry connected between power input terminals PITa and PITb and power output terminals POT1 and POT2 may hereinafter collectively be referred-to as a conditioner circuit.

What is claimed is:

1. An arrangement comprising:

an AC source operative to provide an AC power line voltage at a pair of power line terminals;

a gas discharge lamp having a pair of lamp terminals; and

conditioner circuit connected between the power line terminals and the lamp terminals; the conditioner circuit being characterized by:

(a) being operative to draw a low-frequency line current from the power line terminals;

(b) including an inverter sub-circuit powered from a unidirectional voltage whose instantaneous absolute magnitude is equal to the larger of: (i) the absolute instantaneous magnitude of a substantially constant DC voltage; and (ii) the absolute instantaneous magnitude of a sinusoidal voltage whose peak absolute magnitude is higher than that of the substantially constant DC voltage;

(c) being operative to draw current from the power line terminals even at times when the absolute instantaneous magnitude of the AC power line voltage is lower than that of the substantially constant DC voltage; and

(d) supplying a high-frequency lamp current to the lamp terminals and thereby to the gas discharge lamp; the frequency of the high-frequency lamp current being substantially higher than that of the AC power line voltage; the crest factor of the high-frequency lamp current being equal to or lower than 1.7.

2. The arrangement of claim 1 wherein the conditioner circuit is additionally characterized by being operative to cause the instantaneous magnitude of the low-frequency line current to be substantially proportional to the instantaneous magnitude of the AC power line voltage.

3. The arrangement of claim 1 wherein the conditioner circuit is additionally characterized by including at least two separate inverter circuits; each inverter circuit having a pair of series-connected switching transistors.

4. The arrangement of claim 1 wherein the conditioner circuit is additionally characterized by including at least two separate inverter circuits; each inverter circuit being operative to provide a non-sinusoidal output voltage across a pair of inverter output terminals; the one inverter output voltage being substantially different from the other inverter output voltage.

5. An arrangement comprising:

an AC source operative to provide an AC power line voltage at a pair of power line terminals;

a gas discharge lamp having a pair of lamp terminals; and

conditioner circuit connected between the power line terminals and the lamp terminals; the conditioner circuit being operative to draw a low-frequency line current from the power line terminals and to supply a high-frequency lamp current to the lamp terminals and thereby to the gas discharge lamp; the frequency of the high-frequency lamp current being substantially higher than that of the AC power line voltage; the conditioner circuit being further characterized by:

(a) being operative to cause: (i) the instantaneous magnitude of the low-frequency line current to be substantially proportional to the instantaneous magnitude of the AC power line voltage; and (ii) the crest factor of the high-frequency lamp current to be equal to or lower than 1.7; and

(b) including: (i) a first inverter circuit having a first pair of DC power supply terminals across which is supplied a first DC voltage; and (ii) a second inverter circuit having a second pair of DC power supply terminals across which is supplied a second DC voltage; at least part of the time, the instantaneous magnitude of the second DC voltage being distinctly different from that of the first DC voltage.

6. The arrangement of claim 5 wherein the absolute instantaneous magnitude of the first DC voltage is substantially equal to the absolute instantaneous magnitude of an alternating voltage of sinusoidal waveform.

7. The arrangement of claim 6 wherein the conditioner circuit is additionally characterized in that frequency of said alternating voltage is equal to that of the AC power line voltage.

8. The arrangement of claim 7 wherein the conditioner circuit is additionally characterized in that it draws current from the power line terminals even during times when the absolute instantaneous magnitude of the first DC voltage is lower than that of the second DC voltage.

9. The arrangement of claim 5 wherein the conditioner circuit is additionally characterized by causing the low-frequency line current to have an instantaneous magnitude that is substantially proportional to that of the AC power line voltage.

10. An arrangement comprising:

an AC source operative to provide an AC power line voltage at a pair of power line terminals;

a gas discharge lamp having a pair of lamp terminals; and

conditioner circuit connected between the power line terminals and the lamp terminals; the conditioner circuit being operative to draw a low-frequency line current from the power line terminals and to supply a high-frequency lamp current to the lamp terminals and thereby to the gas discharge lamp; the frequency of the

high-frequency lamp current being substantially higher than that of the AC power line voltage; the conditioner circuit being further operative to cause the crest factor of the high-frequency lamp current to be equal to or lower than 1.7; the conditioner circuit being further characterized by:

(a) having a pair of DC power supply terminals across which exists a unidirectional supply voltage whose absolute instantaneous magnitude is equal to the sum of: (i) the absolute instantaneous magnitude of a substantially constant DC voltage; and (ii) the absolute instantaneous magnitude of an alternating voltage having a substantially sinusoidal waveshape; and

(b) drawing current from the power line terminals even during periods when the absolute instantaneous magnitude of the alternating voltage is lower than that of the substantially constant DC voltage.

11. An arrangement comprising:

an AC source operative to provide an AC power line voltage at a pair of power line terminals;

a gas discharge lamp having a pair of lamp terminals; and

conditioner circuit connected between the power line terminals and the lamp terminals; the conditioner circuit being operative to draw a low-frequency line current from the power line terminals and to supply a high-frequency lamp current to the lamp terminals and thereby to the gas discharge lamp; the frequency of the high-frequency lamp current being substantially higher than that of the AC power line voltage; the conditioner circuit being further characterized by:

(a) being operative to cause the crest factor of the high-frequency lamp current to be equal to or lower than 1.7;

(b) having a pair of DC terminals across which exists a unidirectional supply voltage whose absolute instantaneous magnitude is equal to the larger of: (i) the absolute magnitude of a substantially constant DC voltage; and (ii) the absolute instantaneous magnitude of an alternating voltage having a substantially sinusoidal waveform and a peak absolute magnitude higher than the absolute magnitude of the DC voltage; and

(c) drawing current from the power line terminals even during times when the absolute instantaneous magnitude of the unidirectional supply voltage is equal to that of the substantially constant DC voltage.

12. An arrangement comprising:

an AC source operative to provide an AC power line voltage at a pair of power line terminals;

a gas discharge lamp having a pair of lamp terminals; and

conditioner circuit connected between the power line terminals and the lamp terminals; the conditioner circuit being operative to draw a low-frequency line current from the power line terminals and to supply a high-frequency lamp current to the lamp terminals and thereby to the gas discharge lamp; the frequency of the high-frequency lamp current being substantially higher than that of the AC power line voltage; the conditioner circuit being further characterized by:

(a) being operative to cause the crest factor of the high-frequency lamp current to be equal to or lower than 1.7;

(b) having a pair of DC terminals across which exists a unidirectional supply voltage whose absolute instantaneous magnitude is equal to the larger of: (i) the absolute magnitude of a substantially constant

DC voltage; and (ii) the absolute instantaneous magnitude of an alternating voltage having a substantially sinusoidal waveform and a peak absolute magnitude higher than the absolute magnitude of the substantially constant DC voltage; and

(c) including: (i) a first transistor conducting periodically and intermittently at a frequency equal to the frequency of the high-frequency lamp current, and (ii) a second transistor conducting periodically and intermittently at a different frequency.

13. An arrangement comprising:

an AC source operative to provide an AC power line voltage at a pair of power line terminals;

a gas discharge lamp having a pair of lamp terminals; and

conditioner circuit connected between the power line terminals and the lamp terminals; the conditioner circuit being operative to draw a low-frequency line current from the power line terminals and to supply a high-frequency lamp current to the lamp terminals and thereby to the gas discharge lamp; the frequency of the high-frequency lamp current being substantially higher than that of the AC power line voltage; the conditioner circuit being further characterized by having a pair of DC supply terminals across which exists a unidirectional supply voltage whose absolute instantaneous magnitude is equal to the sum of: (i) the absolute instantaneous magnitude of a substantially constant DC voltage; and (ii) the absolute instantaneous magnitude of an alternating voltage having a substantially sinusoidal waveform and a frequency equal to that of the AC power line voltage.

14. An arrangement comprising:

an AC source operative to provide an AC power line voltage at a pair of power line terminals;

a gas discharge lamp having a pair of lamp terminals; and

conditioner circuit connected between the power line terminals and the lamp terminals; the conditioner circuit being operative to draw a low-frequency line current from the power line terminals and to supply a high-frequency lamp current to the lamp terminals and thereby to the gas discharge lamp; the frequency of the high-frequency lamp current being substantially higher than that of the AC power line voltage; the conditioner circuit being further characterized by having a pair of DC supply terminals across which exists a unidirectional supply voltage whose absolute instantaneous magnitude is equal to the sum of: (i) the absolute magnitude of a substantially constant DC voltage; and (ii) the absolute instantaneous magnitude of an alternating voltage having a substantially sinusoidal waveform and an absolute peak magnitude equal to at least half of the absolute magnitude of the substantially constant DC voltage.

15. The arrangement of claim 14 wherein the conditioner circuit is additionally characterized by causing the low-frequency line current to flow during at least 90% percent of each complete cycle of the AC power line voltage.

16. The arrangement of claim 14 wherein the conditioner circuit is additionally characterized by drawing current from the power line terminals during at least 90% of the total duration of each complete cycle of the AC power line voltage.

17. An arrangement comprising:

an AC source operative to provide an AC power line voltage at a pair of power line terminals;

a gas discharge lamp having a pair of lamp terminals; and

conditioner circuit connected between the power line terminals and the lamp terminals; the conditioner circuit being operative to draw a low-frequency line current from the power line terminals and to supply a high-frequency lamp current to the lamp terminals and thereby to the gas discharge lamp; the frequency of the high-frequency lamp current being substantially higher than that of the AC power line voltage; the conditioner circuit being further characterized by: (a) drawing current from the power line terminals during more than 90% of the total duration of each complete cycle of the AC power line voltage; and (b) including a first inverter having a first transistor conducting periodically at a first frequency, and a second inverter having a second transistor also conducting periodically but at a frequency different from the first frequency.

18. The arrangement of claim 17 wherein the conditioner circuit is additionally characterized in that the low-frequency line current, when indeed being drawn, has an instantaneous magnitude that is substantially proportional to that of the AC power line voltage.

19. The arrangement of claim 17 wherein the conditioner circuit is additionally characterized by causing the high-frequency lamp current to have a crest factor no higher than about 1.7.

20. An arrangement comprising:

an AC source operative to provide an AC power line voltage at a pair of power line terminals;

a gas discharge lamp having a pair of lamp terminals; and

conditioner circuit connected between the power line terminals and the lamp terminals; the conditioner circuit being operative to draw a low-frequency line current from the power line terminals and to supply a high-frequency lamp current to the lamp terminals and thereby to the gas discharge lamp; the frequency of the high-frequency lamp current being substantially higher than that of the AC power line voltage; the conditioner circuit being further characterized by: (a) drawing current from the power line terminals during more than 90% of the total duration of each complete cycle of the AC power line voltage; and (b) having a pair of DC terminals across which exists a DC voltage whose instantaneous absolute magnitude is the larger of (i) a substantially constant magnitude, and (ii) the instantaneous absolute magnitude of an alternating voltage whose peak absolute magnitude is larger than said substantially constant magnitude.

* * * * *

EXHIBIT K

US006172464B1

(12) **United States Patent**
Nilssen

(10) **Patent No.: US 6,172,464 B1**
(45) **Date of Patent: Jan. 9, 2001**

(54) **COMPACT SCREW-IN FLUORESCENT LAMP**

(76) Inventor: **Ole K. Nilssen**, Caesar Dr., Barrington, IL (US) 60010

(*) Notice: Under 35 U.S.C. 154(b), the term of this patent shall be extended for 0 days.

(21) Appl. No.: **08/394,251**

(22) Filed: **Feb. 24, 1995**

Related U.S. Application Data

(63) Continuation-in-part of application No. 07/579,569, filed on Sep. 10, 1990, now abandoned, which is a continuation-in-part of application No. 06/787,692, filed on Oct. 15, 1985, now abandoned, which is a continuation of application No. 06/644,155, filed on Aug. 27, 1984, now abandoned, which is a continuation of application No. 06/555,426, filed on Nov. 23, 1983, now abandoned, which is a continuation of application No. 06/178,107, filed on Aug. 14, 1980, now abandoned.

(51) **Int. Cl.**[7] **H01J 13/46**; H01J 17/34; H01J 19/78

(52) **U.S. Cl.** **315/56**; 315/71; 315/DIG. 7; 315/DIG. 4; 315/DIG. 5; 315/219; 315/224; 315/205

(58) **Field of Search** .... 315/56, 71, 36, 315/DIG. 4, DIG. 5, DIG. 7, 307, 291, 219, 209 R, 224, 205

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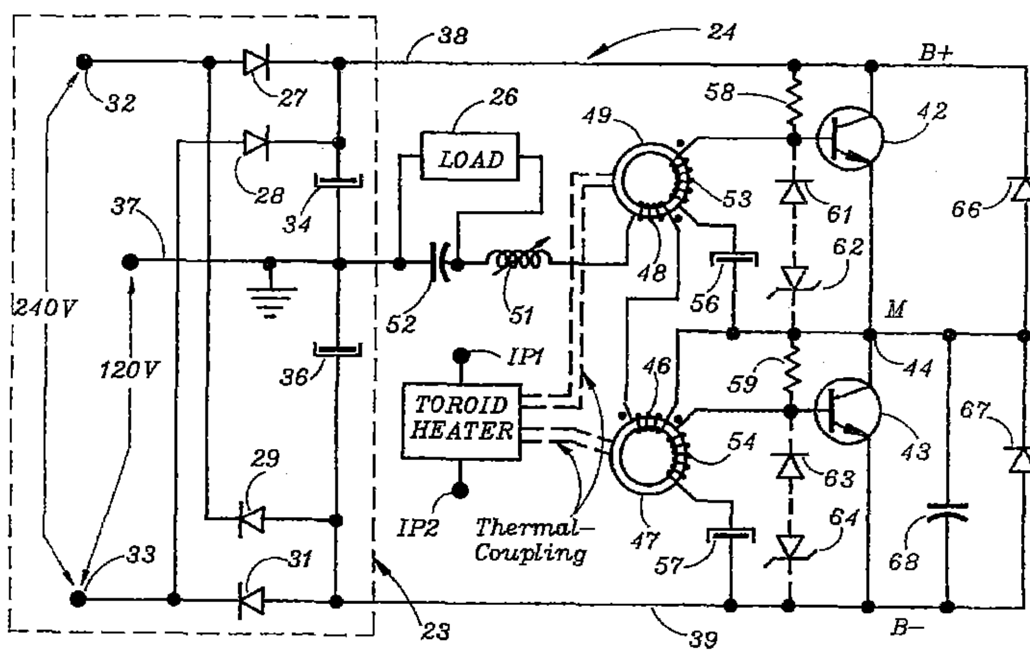
Primary Examiner—Michael B Shingleton

(57) **ABSTRACT**

In a high-frequency electronic ballast, a fluorescent lamp is connected with and powered by way of a series-resonant LC circuit. A resistive load is connected with the LC circuit, thereby to constitute a load therefor before ignition of the fluorescent lamp or in case the fluorescent lamp were to fail to ignite.

6 Claims, 3 Drawing Sheets



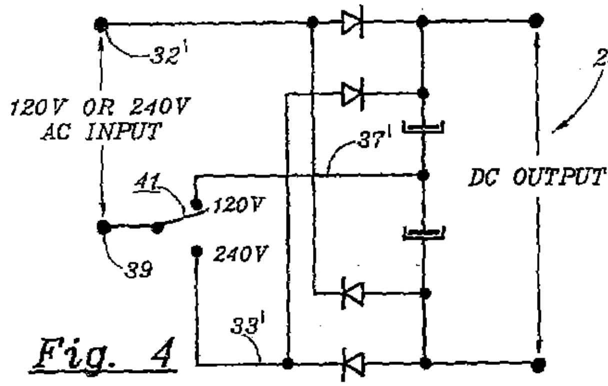


Fig. 4

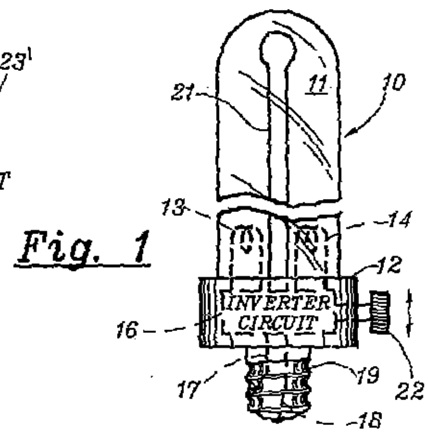


Fig. 1

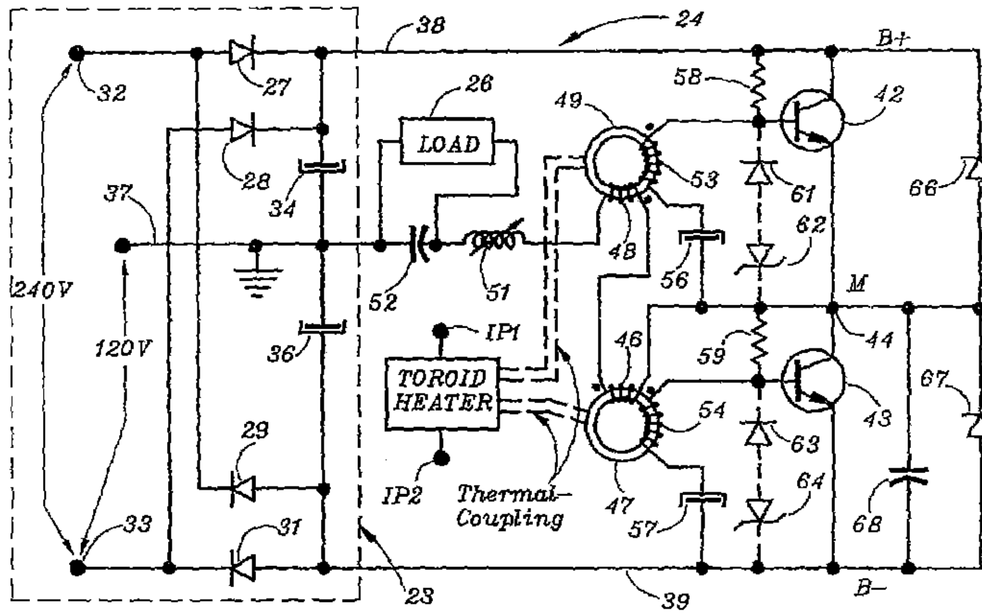


Fig. 2

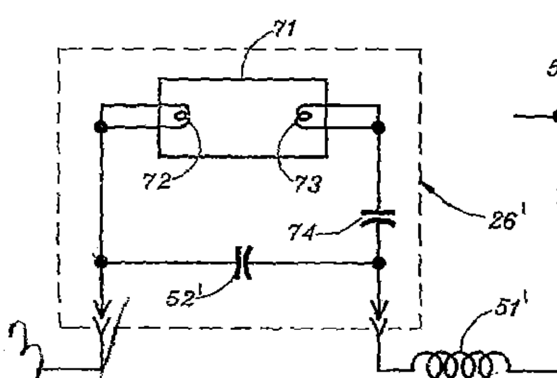


Fig. 5

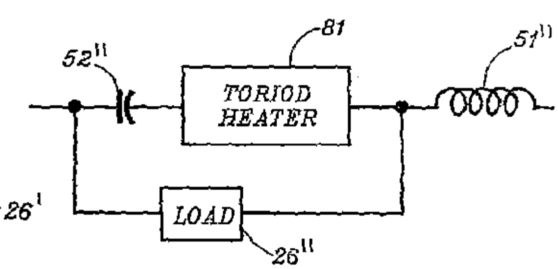


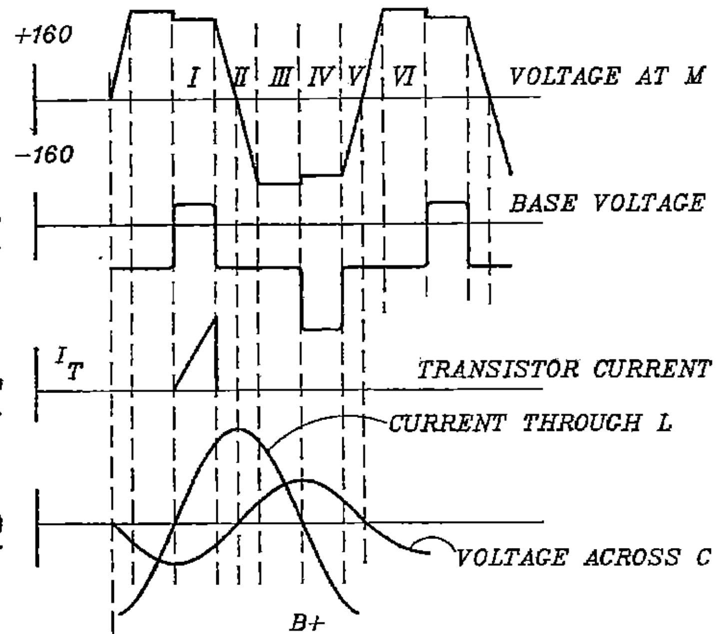
Fig. 6

Fig. 3A

Fig. 3B

Fig. 3C

Fig. 3D



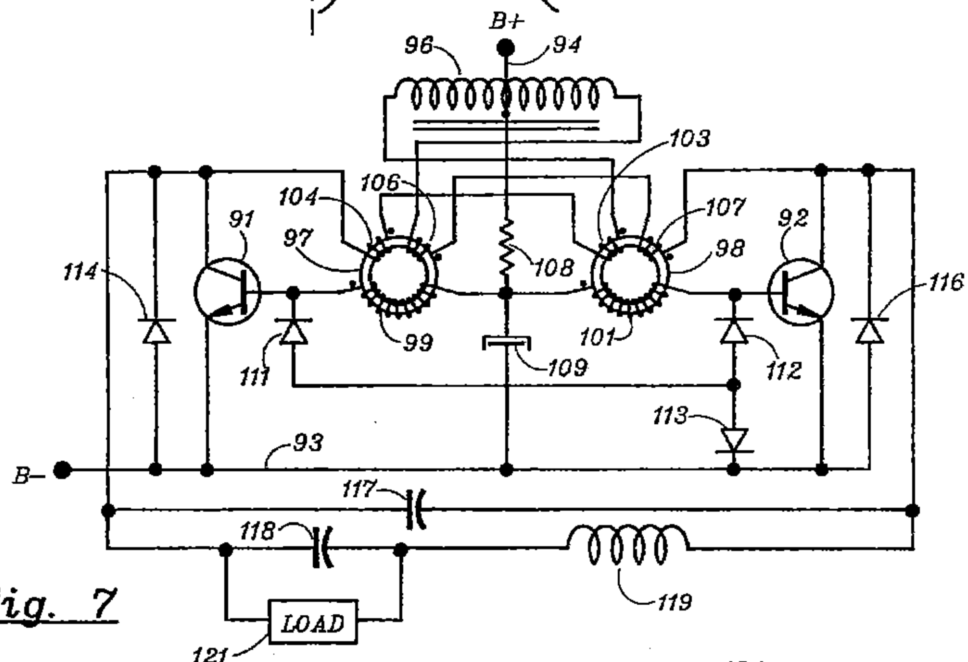


Fig. 7

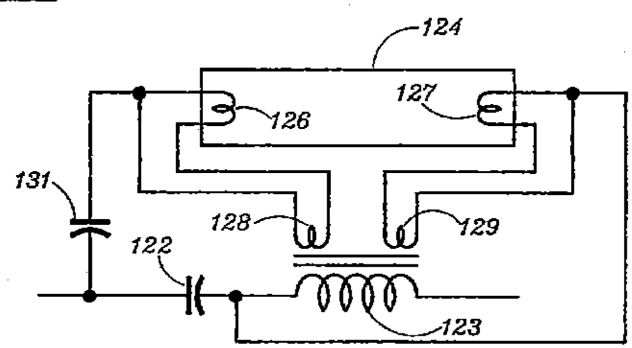


Fig. 8

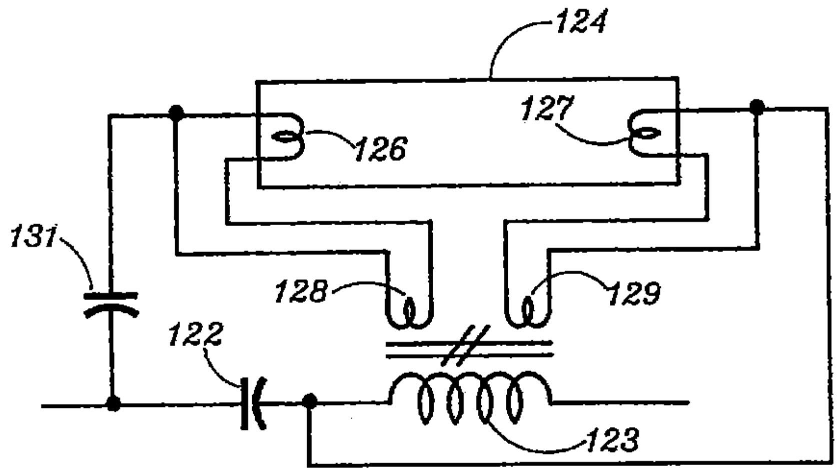
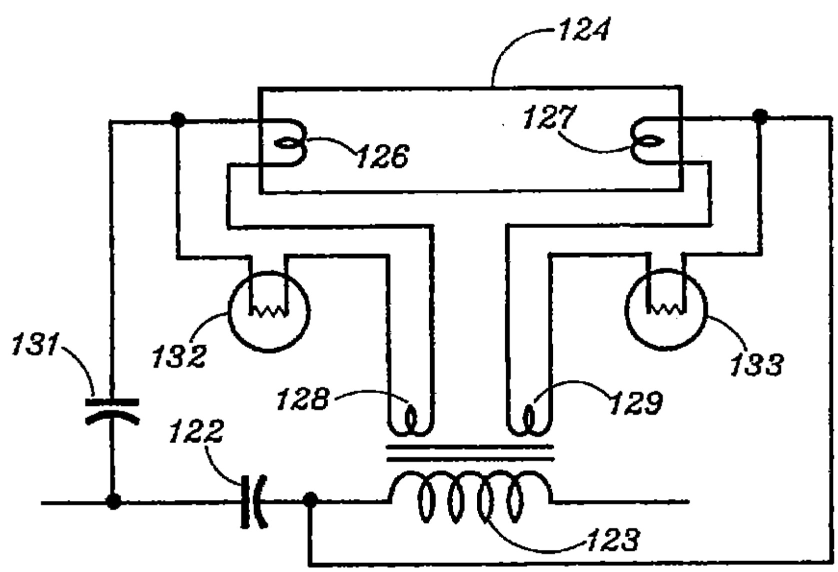

124
127
126
131
128
129
122
123

Fig. 9


124
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132
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128
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122
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Fig. 10

COMPACT SCREW-IN FLUORESCENT LAMP

RELATED APPLICATIONS

The present application is a Continuation-in-Part of Ser. No. 07/579,569 filed Sep. 10, 1990; which is a Continuation-in-Part of Ser. No. 06/787,692 filed Oct. 15, 1985; which is a Continuation of Ser. No. 06/644,155 filed Aug. 27, 1984, now abandoned; which was a Continuation of Ser. No. 06/555,426 filed Nov. 23, 1983, now abandoned; which was a Continuation of Ser. No. 06/178,107 filed Aug. 14, 1980, now abandoned.

BACKGROUND OF THE INVENTION

1. Field of Invention

This invention relates to electronic ballasts for rapid-start fluorescent lamps, particularly where the lamps are powered via a series-resonant LC circuit.

2. Description of Prior Art

For a description of pertinent prior art, reference is made to U.S. Pat. No. 4,677,345 to Nilssen; which patent issued from a Division of application Ser. No. 06/178,107 filed Aug. 14, 1980; which application is the original progenitor of instant application.

Otherwise, reference is made to the following U.S. Pat. No. 3,263,122 to Genuit; U.S. Pat. No. 3,320,510 to Locklair; U.S. Pat. No. 3,996,493 to Davenport et el.; U.S. Pat. No. 4,100,476 to Ghiringhelli; U.S. Pat. No. 4,262,327 to Kovacik et al.; U.S. Pat. No. 4,370,600 to Zansky; U.S. Pat. No. 4,634,932 to Nilssen; and U.S. Pat. No. 4,857,806 to Nilssen.

SUMMARY OF THE INVENTION

Objects of the Invention

Objects of the present invention are those of providing for cost-effective electronic ballasts as well as compact screw-in fluorescent lamps.

This as well as other objects, features and advantages of the present invention will become apparent from the following description and claims.

Brief Description

The present invention is directed to providing improved inverter circuits for powering and controlling gas discharge lamps. The inverter circuits according to the present invention are highly efficient, can be compactly constructed and are ideally suited for energizing gas discharge lamps, particularly "instant-start" and "self-ballasted" fluorescent lamps.

According to one form of the present invention, a series-connected combination of an inductor and a capacitor is provided in circuit with the inverter transistors to be energized upon periodic transistor conduction. Transistor drive current is preferably provided through the use of at least one saturable inductor to control the transistor inversion frequency to be equal to or greater than the nature resonant frequency of the inductor and capacitor combination. The high voltages efficiently developed by loading the inverter with the inductor and capacitor are ideally suited for energizing external loads such as gas discharge lamps. In such an application, the use of an adjustable inductor permits control of the inverter output as a means of adjusting the level of lamp illumination.

According to another important form of the present invention, reliable and highly efficient half-bridge inverters include a saturable inductor in a current feedback circuit to drive the transistors for alternate conduction. The inverters

also include a load having an inductance sufficient to effect periodic energy storage for self-sustained transistor inversion. Importantly, improved reliability is achieved because of the relatively low and transient-free voltages across the transistors in these half-bridge inverters.

Further, according to another feature of the present invention, novel and economical power supplies particularly useful with the disclosed inverter circuits convert conventional AC input voltages to DC for supplying to the inverters.

Yet further, according to still another feature of the invention, a rapid-start fluorescent lamp is powered by way of a series-resonant LC circuit; while heating power for the lamp's cathodes is provided via loosely-coupled auxiliary windings on the tank inductor of the LC circuit. Alternatively, cathode heating power is provided from tightly-coupled windings on the tank inductor; in which case output current-limiting is provided via a non-linear resistance means, such as an incandescent filament in a light bulb, connected in series with the output of each winding.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 is a front elevation of a folded fluorescent lamp unit adapted for screw-in insertion into a standard Edison incandescent socket;

FIG. 2 is a schematic diagram illustrating the essential features of a push-pull inverter circuit particularly suitable for energizing the lamp unit of FIG. 1;

FIGS. 3A–3D is a set of waveform diagrams of certain significant voltages and currents occurring in the circuit of FIG. 2;

FIG. 4 is a schematic diagram of a DC power supply connectable to both 120 and 240 volt AC inputs;

FIG. 5 is a schematic diagram which illustrates the connection of a non-self-ballasted gas discharge lamp unit to the FIG. 2 inverter circuit;

FIG. 6 is a schematic diagram which illustrates the use of a toroid heater for regulation of the inverter output;

FIG. 7 is an alternate form of push-pull inverter circuit accordind to the present invention;

FIG. 8 is a schematic diagram showing the connection of a gas discharge lamp of the "rapid-start" type to an inductor-capacitor-loaded inverter according to the present invention;

FIG. 9 is a modification of FIG. 8, showing loosely-coupled auxiliary windings on the tank inductor; and

FIG. 10 is another modification of FIG. 8, showing nonlinear current-limiting means connected with the output of tightly-coupled auxiliary windings on the tank inductor.

DESCRIPTION OF THE PREFERRED EMBODIMENTS

FIG. 1 illustrates a screw-in gas discharge lamp unit **10** comprising a folded fluorescent lamp **11** suitably secured to an integral base **12**. The lamp comprises two cathodes **13, 14** which are supplied with the requisite high operating voltage from a frequency-converting power supply and ballasting circuit **16**; which, because of its compact size, conveniently fits within the base **12**.

The inverter circuit **16** is connected by leads **17, 18** to a screw-type plug **19** adapted for screw-in insertion into a standard Edison-type incandescent lamp socket at which ordinary 120 Volt/60 Hz power line voltage is available. A ground plane comprising a wire or metallic strip **21** is disposed adjacent a portion of the fluorescent lamp **11** as a starting aid. Finally, a manually rotatable external knob 22 is connected to a shaft for mechanical adjustment of the air gap of a ferrite core inductor to vary the inductance value thereof in order to effect adjustment of the inverter voltage output connected to electrodes 13, 14 for controlled variation of the lamp illumination intensity.

With reference to FIG. 2, a power supply 23, connected to a conventional AC input, provides a DC output for supplying a high-efficiency inverter circuit 24. The inverter is operable to provide a high voltage to an external load 26, which may comprise a gas discharge device such as the fluorescent lamp 11 of FIG. 1.

The power supply 23 comprises bridge rectifier having four diodes 27, 28, 29 and 31 connectable to a 240 volt AC supply at terminals 32, 33. Capacitors 34, 36 are connected between a ground line 37 (in turn directly connected to the inverter 24) and to a B+ line 38 and a B− line 39, respectively. The power supply 23 also comprises a voltage doubler and rectifier optionally connectable to a 120 volt AC input taken between the ground line 37 and terminal 33 or 32. The voltage doubler and rectifier means provides a direct electrical connection by way of line 37 between one of the 120 volt AC power input lines and the inverter 24, as shown in FIG. 2. The bridge rectifier and the voltage doubler and rectifier provide substantially the same DC output voltage to the inverter 24 whether the AC input is 120 or 240 volts. Typical voltages are +160 volts on the B+ line 38 and −160 volts on the B− line 39.

With additional reference to FIG. 4, which shows an alternate power supply 23', the AC input, whether 120 or 240 volts, is provided at terminals 32' and 39. Terminal 39 is in turn connected through a single-pole double-throw selector switch 41 to terminal 37' (for 120 volt operation) or terminal 33' (for 240 volt operation). In all other respects, power supplies 23 and 23' are identical.

The inverter circuit 24 of FIG. 2 is a half-bridge inverter comprising transistors 42, 43 connected in series across the DC voltage output of the power supply 23 on B+ and B− lines 38 and 39, respectively. The collector of trasistor 42 is connected to the B+ line 38, the emitter of transistor 42 and the collector of transistor 43 are connected to a midpoint line 44 (designated "M") and the emitter of transistor 43 is connected to the B− line 39. The midpoint line 44 is in turn connected to the ground line 37 through primary winding 46 of a toroidal saturable core transformer 47, a primary winding 48 on an identical transformer 49, an inductor 51 and a series-connected capacitor 52. The inductor 51 and capacitor 52 are energized upon alternate transistor conduction in a manner to be described later.

An external load 26 is preferably taken off capacitor 52, as shown in FIG. 2. The inductor 51, preferably a known ferrite core inductor, has an inductance variable by mechanical adjustment of the air gap in order to effect variation in the level of the inductor and capacitor voltage and hence the power available to the load, as will be described. When the load is a gas discharge lamp such as lamp 11 in FIG. 1, variation in this inductance upon rotation of knob 22 accomplishes a lamp dimming effect.

Drive current to the base terminals of transistors 42 and 43 is provided by secondary windings 53, 54 of transformers 49, 47, respectively. Winding 53 is also connected to midpoint lead 44 through a bias capacitor 56, while winding 54 is connected to the B− lead 39 through an identical bias capacitor 57. The base terminals of transistors 42 and 43 are also connected to lines 38 and 44 through bias resistors 58 and 59, respectively. For a purpose to be described later, the base of transistor 42 can be optionally connected to a diode 61 and a series Zener diode 64 in turn connected to the midpoint line 44; similarly, a diode 63 and series Zener diode 64 in turn connected to the B− line 39 can be connected to the base of transistor 43. Shunt diodes 66 and 67 are connected across the collector-emitter terminals of transistors 42 and 43, respectively. Finally, a capacitor 68 is connected across the collector-emitter terminals of transistor 43 to restrain the rate of voltage rise across those terminals, as will be seen presently.

The operation of the circuit of FIG. 2 can best be understood with additional reference to FIG. 3, which illustrates significant portions of the waveforms of the voltage at midpoint M (FIG. 3A), the base-emitter voltage on transistor 42 (FIG. 3B), the current through transistor 42 (FIG. 3C), and the capacitor 52 voltage and the inductor 51 current (FIG. 3D).

Assuming that transistor 42 is first to be triggered into conduction, current flows from the B+ line 38 through windings 46 and 38 and the inductor 51 to charge capacitor 52 and returns through capacitor 34 (refer to the time period designated I in FIG. 3). When the saturable inductor 49 saturates at the end of period I, drive current to the base of transistor 42 will terminate, causing voltage on the base of the transistor to drop to the negative voltage stored on the bias capacitor 56 in a manner to be described, causing this transistor to become non-conductive. As shown in FIG. 3c, current-flow in transistor 43 terminates at the end of period I.

Because the current through inductor 51 cannot change instantaneously, current will flow from the B− bus 39 through capacitor 68, causing the voltage at midpoint line 44 to drop to −160 volts (period II in FIG. 3). The capacitor 68 restrains the rate of voltage change across the collector and emitter terminals of transistor 42. The current through the inductor 51 reaches its maximum value when the voltage at the midpoint line 44 is zero. During period III, the current will continue to flow through inductor 51 but will be supplied from the B− bus through the shunt diode 67. It will be appreciated that during the latter half of period II and all of period III, positive current is being drawn from a negative voltage; which, in reality, means that energy is being returned to the power supply through a path of relatively low impedance.

When the inductor current reaches zero at the start of period IV, the current through the primary winding 46 of the saturable inductor 47 will cause a current to flow out of its secondary winding 54 to cause transistor 43 to become conductive, thereby causing a reversal in the direction of current through inductor 51 and capacitor 52. When transformer 47 saturates at the end of period IV, the drive current to the base of transistor 43 terminates and the current through inductor 51 will be supplied through capacitor 68, causing the voltage at midpoint line 44 to rise (period V). When the voltage at the midpoint line M reaches 160 volts, the current will then flow through shunt diode 66 (period VI). The cycle is then repeated.

As seen in FIG. 3, saturable transformers 47, 49 provide transistor drive current only after the current through inductor 51 has diminished to zero. Further, the transistor drive current is terminated before the current through inductor 51 has reached its maximum amplitude. This coordination of base drive current and inductor current is achieved because of the series-connection between the inductor 51 and the primary windings 46, 48 of saturable transformers 47, 49, respectively.

The series-connected combination of the inductor **51** and the capacitor **52** is energized upon the alternate conduction of transistors **42** and **43**. With a large value of capacitance of capacitor **52**, very little voltage will be developed across its terminals. As the value of this capacitance is decreased, however, the voltage across this capacitor will increase. As the value of the capacitor **52** is reduced to achieve resonance with the inductor **51**, the voltage on the capacitor will rise and become infinite in a loss-free circuit operating under ideal conditions.

It has been found desirable to regulate the transistor inversion frequency, determined mainly by the saturation time of the saturable inductors **47**, **49**, to be equel to or higher than the natural resonance frequency of the inductor and capacitor combination in order to provide a high voltage output to external load **26**. A high voltage across capacitor **52** is efficiently developed as the transistor inversion frequency approaches the natural resonant frequency of the inductor **51** and capacitor **52** combination. Stated another way, the conduction period of each transistor is desirably shorter in duration than one quarter of the full period corresponding to the natural resonant frequency of the inductor and capacitor combination. When the inverter **24** is used with a self-ballasted gas discharge lamp unit, it has been found that the inversion frequency can be at least equal to the natural resonant frequency of the tank circuit. If the capacitance value of capacitor **52** is reduced still further beyond the resonance point, unacceptably high transistor currents will be experienced during transistor switching and transistor burn-out will occur.

It will be appreciated that the sizing of capacitor **52** is determined by the application of the inverter circuit **24**. Variation in the values of the capacitor **52** and the inductor **51** will determine the voltages developed in the inductor-capacitor tank circuit. The external load **26** may be connected in circuit with the inductor **51** (by a winding on the inductor, for example) and the capacitor may be omitted entirely. If the combined circuit loading of the inductor **51** and the external load **26** has an effective inductance of value sufficient to effect periodic energy storage for self-sustained transistor inversion, the current feedback provided by the saturable inductors **47**,**49** will effect alternate transistor conduction without the need for additional voltage feedback. When the capacitor **52** is omitted, the power supply **23** provides a direct electrical connection between one of the AC power input lines and the inverter load circuit.

Because the voltages across transistors **42**, **43** are relatively low (due to the effect of capacitors **34**, **36**), the half-bridge inverter **24** is very reliable. The absence of switching transients minimizes the possibility of transistor burn-out.

The inverter circuit **24** comprises means for supplying reverse bias to the conducting transistor upon saturation of its associated saturable inductor. For this purpose, the capacitors **56** and **57** are charged to negative voltages as a result of reset current flowing into secondary windings **53**, **54** from the bases of transistors **42**, **43**, respectively. This reverse current rapidly turns off a conducting transistor to increase its switching speed and to achieve inverter circuit efficiency in a manner described more fully in my co-pending U.S. patent application Ser. No. 103,624 filed Dec. 14, 1979 and entitled "Bias Control for High Efficiency Inverter Circuit" (now U.S. Pat. No. 4,307,353). The more negative the voltage on the bias capacitors **56** and **57**, the more rapidly charges are swept out of the bases of their associated transistors upon transistor turn-off.

When a transistor base-emitter junction is reversely biased, it exhibits the characteristics of a Zener diode having

a reverse breakdown voltage on the order of 8 to 14 Volt for transistors typically used in high-voltage inverters. As an alternative, to provide a negative voltage smaller in magnitude on the base lead of typical transistor **42** during reset operation, the optional diode **61** and Zener diode **62** combination can be used. For large values of the bias capacitor **56**, the base voltage will be substantially constant.

If the load **26** comprises a gas discharge lamp, the voltage across the capacitor **52** will be reduced once the lamp is ignited to prevent voltages on the inductor **51** and the capacitor **52** from reaching destructive levels. Such a lamp provides an initial time delay during which a high voltage, suitable for instant starting, is available.

FIG. **5** illustrates the use of an alternate load **26'** adapted for plug-in connection to an inverter circuit such as shown in FIG. **2**. The load **26'** consists of a gas discharge lamp **71** having electrodes **72**, **73** and connected in series with a capacitor **74**. The combination of lamp **71** and capacitor **74** is connected in parallel with a capacitor **52'** which serves the same purpose as capacitor **52** in the FIG. **2** circuit. However, when the load **26'** is unplugged from the circuit, the inverter stops oscillating and the development of high voltages in the inverter is prevented. The fact that no high voltages are generated by the circuit if the lamp is disconnected while the circuit is oscillating is important for safety reasons.

FIG. **6** illustrates a capacitor **52"** connected in series with an inductor **51"** through a heater **81** suitable for heating the toroidal inductors **47**, **49** in accordance with the level of output. The load **26"** is connected across the series combination of the capacitor **52"** and the toroid heater. The heater **81** is preferably designed to controllably heat the toroidal saturable inductors in order to decrease their saturation flux limit and hence their saturation time. The result is to decrease the periodic transistor conduction time and thereby increase the transistor inversion frequency. When a frequency-dependent impedance means, that is, an inductor or a capacitor, is connected in circuit with the AC voltage output of the inverter, change in the transistor inversion frequency will modify the impedance of the frequency-dependent impdance means and correspondingly modify the inverter output. Thus as the level of the output increases, the toroid heater **81** is correspondingly energized to effect feedback regulation of the output. Further, transistors **42**, **43** of the type used in high voltage inverters dissipate heat during periodic transistor conduction. As an alternative, the toroid heater **81** can use this heat for feedback regulation of the output or control of the temperature of transistors **42**, **43**.

The frequency dependent impedance means may also be used in a circuit to energize a gas discharge lamp at adjustable illumination levels. Adjustment in the inversion frequency of transistors **42**, **43** results in control of the magnitude of the AC current supplied to the lamp. This is preferably accomplished where saturable inductors **47**, **49** have adjustable flux densities for control of their saturation time.

FIG. **7** schematically illustrates an alternate form of inverter circuit, shown without the AC to DC power supply connections for simplification. In this Figure, the transistors are connected in parallel rather than in series but the operation is essentially the same as previously described.

In particular, this circuit comprises a pair of alternately conducting transistors **91**, **92**. The emitter terminals of the transistors are connected to a B− line **93**. A B+ lead **94** is connected to the center-tap of a transformer **96**. In order to provide drive current to the transistors **91**, **92** for control of their conduction frequency, saturable inductors **97**, **98** have secondary windings **99**, **101**, respectively, each secondary winding having one end connected to the base of its associated transistor; the other ends are connected to a common terminal **102**. One end of transformer **96** is connected to the collector of transistor **91** through a winding **103** on inductor **98** in turn connected in series with a winding **104** on inductor **97**. Likewise, the other end of transformer **96** is connected to the collector of transistor **92** through a winding **106** on inductor **97** in series with another winding **107** on inductor **98**.

The B+ terminal is connected to terminal **102** through a bias resistor **108**. A bias capacitor **109** connects terminal **102** to the B− lead **93**. This resistor and capacitor serve the same function as resistors **58**, **59** and capacitors **56**, **57** in the FIG. 2 circuit.

The bases of transistors **91**, **92** are connected by diodes **111**, **112**, respectively, to a common Zener diode **113** in turn connected to the B− lead **93**. The common Zener diode **113** serves the same function as individual Zener diodes **62**, **64** in FIG. 2.

Shunt diodes **114**, **116** are connected across the collector-emitter terminals of transistors **91**, **92**, respectively. A capacitor **117** connecting the collectors of transistors **91**, **92** restrains the rate of voltage rise on the collectors in a manner similar to the collector-emitter capacitor **68** in FIG. 2.

Inductive-capacitive loading of the FIG. 7 inverter is accomplished by a capacitor **118** connected in series with with an inductor **119**, the combination being connected across the collectors of the transistors **91**, **92**. A load **121** is connected across the capacitor **118**.

FIG. **8** illustrates how an inverter loaded with a series capacitor **122** and inductor **123** can be used to energize a "rapid-start" fluorescent lamp **124** (the details of the inverter circuit being omitted for simplication). The lamp **124** has a pair of cathodes **126**, **127** connected across the capacitor **122** for supply of operating voltage in a manner identical to that previously described. In addition, the inductor **123** comprises a pair of magnetically-coupled auxiliary windings **128**, **129** for electrically heating the cathodes **126**, **127**, respectively. A small capacitor **131** is connected in series with lamp **124**.

FIG. **9** illustrates the very same circuit arrangement as that of FIG. **8** except that the auxiliary windings **128**, **129** are only loosely coupled to the inductor **123**, thereby providing for a manifest limitation on the amount of current that can be drawn from each auxiliary winding in case it were to be accidentally short-circuited.

FIG. **10** also illustrates the very same circuit arrangement as that of FIG. **8** except that the cathodes **126**, **127** are connected with their respective auxiliary windings **128**, **129** by way of nonlinear current-limiting means **132** and **133**, respectively.

In FIG. **10**, the non-linear current-limiting means **132**, **133** are shown as being two (small) incandescent lamps. However, other types of non-linear resistance means could be used as well.

Both the FIG. **9** circuit and the FIG. **10** circuit serve the same basic purpose; which is that of preventing damage to the ballast circuit (such as that if FIG. 2) in case the leads used for connecting to one of the lamp cathodes **126**, **127** were to be accidentally shorted. This damage prevention is accomplished by providing for manifest limitation of the maximum amount of current that can be drawn from each one of the auxiliary windings **128**, **129**. In the circuit of FIG. **9**, this manifest limitation is accomplished by having the auxiliary windings **128**, **129** couple sufficiently loosely to the main inductor **123**—such as by providing a magnetic shunt between inductor **123** and the auxiliary windings—thereby correspondingly limiting the degree of impact resulting from an accidental short circuit. Such a short circuit would result in a net reduction in the effective inductance value of the tank inductor **123**; which net reduction in inductance may in turn cause a precipitous increase in the magnitude of the reactive current drawn from the inverter by the series-connected inductor **123** and capacitor **122**, thereby causing damage to the inverter.

Additional Explanations and Comments

(a) With reference to FIGS. 2 and **5**, adjustment of the amount of power supplied to load **26'**, and thereby the amount of light provided by lamp **71**, may be accomplished by applying a voltage of adjustable magnitude to input terminals IP1 and IP2 of the Toroid Heater; which is thermally coupled with the toroidal ferrite cores of saturable transformers **47**, **49**.

(b) With commonly available components, inverter circuit **24** of FIG. **2** can be made to operate efficiently at any frequency between a few kHz to perhaps as high as 50 kHz. However, for various well-known reasons (i.e., eliminating audible noise, minimizing physical size, and maximizing efficiency), the frequency actually chosen is in the range of 20 to 40 kHz.

(c) The fluorescent lighting unit of FIG. **1** could be made in such manner as to permit fluorescent lamp **11** to be disconnectable from its base **12** and ballasting means **16**. However, if powered with normal line voltage without its lamp load connected, frequency-converting power supply and ballasting circuit **16** is apt to self-destruct.

To avoid such self-destruction, arrangements can readily be made whereby the very act of removing the load automatically establishes a situation that prevents the possible destruction of the power supply and ballasting means. For instance, with the tank capacitor (**52**) being permanently connected with the lamp load (**11**)—thereby automatically being removed whenever the lamp is removed—the inverter circuit is protected from self-destruction.

(d) At frequencies above a few kHz, the load represented by a fluorescent lamp—once it is ignited—is substantially resistive. Thus, with the voltage across lamp **11** being of a substantially sinusoidal waveform (as indicated in FIG. 3*d*), the current through the lamp will also be substantially sinusoidal in waveshape.

(e) In the fluorescent lamp unit of FIG. **1**, fluorescent lamp **11** is connected with power supply and ballasting circuit **16** in the exact same manner as is load **26** connected with the circuit of FIG. **2**. That is, it is connected in parallel with the tank capacitor (**52**) of the L-C series-resonant circuit. As is conventional in instant-start fluorescent lamps—such as lamp **11** of FIG. **1**—the two terminals from each cathode are shorted together, thereby to constitute a situation where each cathode effectively is represented by only a single terminal. However, it is not necessary that the two terminals from each cathode be shorted together; in which case—for instant-start operation—connection from a lamp's power supply and ballasting means need only be made with one of the terminals of each cathode.

(f) with respect to the circuit arrangement of FIG. **9**, in situations where the tank inductor **123** includes a ferrite magnetic core having an air gap, one particularly cost-effective way of accomplishing the indicated loose coupling between the tank inductor **123** and the auxiliary windings **128**, **129** is that of arranging for the auxiliary windings to be placed in the air gap in such a manner that they each couple only with part of the magnetic flux crossing the air gap.

(g) in FIG. 1, the compact screw-in fluorescent lamp has a longitudinal central axis penetrating through the center of the bottom of base **19** (i.e., at the point where lead **18** is connected), passing up centrally between the two legs of lamp **11**, and emerging at the center of the very top of lamp **11**.

(h) In FIG. 1, as a skilled artisan would perceive by direct inspection, the visible parts are drawn to scale. Thus, for instance:

(i) the height and width (i.e., diameter) of screw-base **19** are in proper proportion to those of an actual screw-base on an ordinary household incandescent lamp;

(ii) the diameter of the individual straight legs of the folded fluorescent lamp **11** are shown in proportion to the diameter of the screw-base;

(iii) the diameter of the bent portion connecting the top parts of the two straight lamp legs is shown in proper proportion to the diameter of the lamp legs; and

(iv) the distance between the two straight lamp legs is shown in proper proportion to the diameter of those lamp legs.

Of course, for a screw-in fluorescent lamp to have maximum utility, it is imperative that it has dimensions sufficiently compact to permit it to be conveniently used in most places where an incandescent lamp would ordinarily be used. Thus, it is important that its maximum diameter not be any larger than those of an ordinary household incandescent lamp (whose maximum diameter is typically about twice that of its screw-base). The screw-in fluorescent lamp depicted in FIG. **1** clearly satisfies those requirements.

What is claimed is:

1. An arrangement comprising:

a screw-base inserted into and held by a lamp socket adapted to accept and hold an ordinary household incandescent light bulb; the screw-base having base terminals and being otherwise characterized by having a central screw-base axis around which the outer boundary of the screw-base forms an approximately cylindrical surface; the lamp socket having socket terminals at which is provided AC power line voltage such as that normally provided from an ordinary electric utility power line;

a fluorescent lamp having lamp terminals; the fluorescent lamp being characterized by including at least two straight cylindrical light-emitting glass-enclosed sections disposed parallel to each other as well as to a central lamp axis; a lamp terminal being disposed at one end of each of the two straight cylindrical light-emitting glass-enclosed sections; the other ends of the two straight cylindrical light-emitting glass-enclosed sections being connected together via a transversely disposed light-emitting glass-enclosed section; said transversely disposed section being of a maximum dimension substantially no longer than just sufficient to reach between said other ends;

electronic sub-assembly having power input terminals connected with the base terminals and power output terminals connected with the lamp terminals; the sub-assembly being operative, whenever supplied with AC power line voltage at its power input terminals, to supply an alternating lamp current to the lamp terminals; the alternating lamp current being of frequency distinctly higher than that of the AC power line voltage; the sub-assembly being additionally characterized by not including a transformer having a primary winding connected across the power input terminals; the electronic sub-assembly being further characterized by including two terminals across which exists a DC voltage having a substantially constant absolute magnitude that is distinctly higher than the peak absolute magnitude of the AC power line voltage; and

structure means operative to hold together the screw-base, the fluorescent lamp, and the electronic sub-assembly, thereby to form an integral screw-in lamp structure characterized by having an overall longitudinal axis parallel with the central screw-base axis as well as with the central lamp axis.

2. An arrangement comprising:

a screw-base inserted into and held by a lamp socket adapted to accept and hold an ordinary household incandescent light bulb; the screw-base having base terminals and being otherwise characterized by having a central screw-base axis around which the outer boundary of the screw-base forms an approximately cylindrical surface; the lamp socket having socket terminals at which is provided AC power line voltage such as that normally provided from an ordinary electric utility power line;

a fluorescent lamp having lamp terminals; the fluorescent lamp being characterized by including at least two straight cylindrical light-emitting glass-enclosed sections disposed parallel to each other as well as to a central lamp axis; a lamp terminal being disposed at one end of each of the two straight cylindrical light-emitting glass-enclosed sections; the other ends of the two straight cylindrical light-emitting glass-enclosed sections being connected together via a transversely disposed light-emitting glass-enclosed section; said transversely disposed section being of a maximum dimension substantially no longer than just sufficient to reach between said other ends;

electronic sub-assembly having power input terminals connected with the base terminals and power output terminals connected with the lamp terminals; the sub-assembly being operative, whenever supplied with AC power line voltage at its power input terminals, to supply an alternating lamp current to the lamp terminals; the alternating lamp current being of frequency distinctly higher than that of the AC power line voltage; the sub-assembly being additionally characterized by not including a transformer having a primary winding connected across the power input terminals; the electronic sub-assembly being further characterized by including a pair of terminals across which exists a DC voltage and between which are series-connected two transistors; the transistors being characterized by alternatingly conducting current; at least one of the transistors conducting current in the form of periodic unidirectional current pulses; the periodic unidirectional current pulses occurring at a frequency equal to that of the alternating lamp current; each individual unidirectional current pulse having a duration distinctly shorter than half of a complete period of the alternating lamp current; and

structure means operative to hold together the screw-base, the fluorescent lamp, and the electronic sub-assembly, thereby to form an integral screw-in lamp structure characterized by having an overall longitudinal axis parallel with the central screw-base axis as well as with the central lamp axis.

3. A structure characterized by having a central axis about which the following elements are assembled:

a screw-base operative to screw into a lamp socket of a type usually used for receiving and holding an ordinary household incandescent light bulb; the screw-base being otherwise characterized by having base terminals and by being disposed symmetrically about the central axis;

a fluorescent lamp having lamp terminals and plural cylindrical lamp segments disposed apart from, but parallel to, each other as well as to the central axis; each of the plural cylindrical lamp segments having a total length; the fluorescent lamp being further characterized in that a flat plane disposed perpendicular to the central axis and intersecting one of the cylindrical lamp segments anywhere along its total length creates a cross-sectional pattern that (i) is symmetrical with respect to a flat plane disposed parallel to the central axis, and (ii) includes nothing but cross-sections of substantially identical cylindrical lamp segments;

an electronic sub-assembly having input terminals and output terminals; the input terminals being connected with the base terminals; the output terminals being connected with the lamp terminals; the electronic sub-assembly being operative to supply an alternating voltage at its output terminals provided it be supplied with an AC power line voltage at its input terminals; the frequency of the alternating voltage being distinctly higher than that of the AC power line voltage; the electronic subassembly being additionally characterized by including a transistor through which flows unidirectional current pulses at a periodic rate equal to that of the alternating voltage; each current pulse having a duration distinctly shorter than half of the complete cycle of the alternating voltage; and

housing means mounted rigidly on the screw-base and operative to house the electronic sub-assembly as well as to hold and support the fluorescent lamp, thereby to form a fluorescent lamp entity adapted to be screwed into and powered from a lamp socket at which ordinary AC power line voltage is provided.

4. A structure characterized by having a central axis about which the following elements are assembled:

a screw-base operative to screw into a lamp socket of a type usually used for receiving and holding an ordinary household incandescent light bulb; the screw-base being otherwise characterized by having base terminals and by being disposed symmetrically about the central axis;

a fluorescent lamp having lamp terminals and plural cylindrical lamp segments disposed apart from, but parallel to, each other as well as to the central axis; each of the plural cylindrical lamp segments having a total length; the fluorescent lamp being further characterized in that a flat plane disposed perpendicular to the central axis and intersecting one of the cylindrical lamp segments anywhere along its total length creates a cross-sectional pattern that (i) is symmetrical with respect to a flat plane disposed parallel to the central axis, and (ii) includes nothing but cross-sections of substantially identical cylindrical lamp segments;

an electronic sub-assembly having input terminals and output terminals; the input terminals being connected with the base terminals; the output terminals being connected with the lamp terminals; the electronic sub-assembly being operative to supply an alternating voltage at its output terminals provided it be supplied with an AC power line voltage at its input terminals; the frequency of the alternating voltage being distinctly higher than that of the AC power line voltage; the electronic subassembly being additionally characterized by including a pair of terminals across which exists a DC voltage having a substantially constant absolute magnitude that is distinctly larger than the absolute peak magnitude of the AC power line voltage; and

housing means mounted rigidly on the screw-base and operative to house the electronic sub-assembly as well as to hold and support the fluorescent lamp, thereby to form a fluorescent lamp entity adapted to be screwed into and powered from a lamp socket at which ordinary AC power line voltage is provided.

5. A structure characterized by having a central axis about which the following elements are assembled:

a screw-base operative to screw into a lamp socket of a type usually used for receiving and holding an ordinary household incandescent light bulb; the screw-base being otherwise characterized by having base terminals and by being disposed symmetrically about the central axis;

a fluorescent lamp having lamp terminals and plural cylindrical lamp segments disposed apart from, but parallel to, each other as well as to the central axis; each of the plural cylindrical lamp segments having a total length; the fluorescent lamp being further characterized in that a flat plane disposed perpendicular to the central axis and intersecting one of the cylindrical lamp segments anywhere along its total length creates a cross-sectional pattern that (i) is symmetrical with respect to a flat plane disposed parallel to the central axis, and (ii) includes nothing but cross-sections of substantially identical cylindrical lamp segments;

an electronic sub-assembly having input terminals and output terminals; the input terminals being connected with the base terminals; the output terminals being connected with the lamp terminals; the electronic sub-assembly being operative to supply an alternating voltage at its output terminals provided it be supplied with an AC power line voltage at its input terminals; the frequency of the alternating voltage being distinctly higher than that of the AC power line voltage; the electronic subassembly being additionally characterized by including a pair of terminals across which (i) exists a DC voltage, and (ii) are series-connected two filter capacitors; and

housing means mounted rigidly on the screw-base and operative to house the electronic sub-assembly as well as to hold and support the fluorescent lamp, thereby to form a fluorescent lamp entity adapted to be screwed into and powered from a lamp socket at which ordinary AC power line voltage is provided.

6. A structure characterized by having a central axis about which the following elements are assembled:

a screw-base operative to screw into a lamp socket of a type usually used for receiving and holding an ordinary household incandescent light bulb; the screw-base being otherwise characterized by having base terminals and by being disposed symmetrically about the central axis;

a fluorescent lamp having lamp terminals and plural cylindrical lamp segments disposed apart from, but parallel to, each other as well as to the central axis; each of the plural cylindrical lamp segments having a total length; the fluorescent lamp being further characterized in that a flat plane disposed perpendicular to the central axis and intersecting one of the cylindrical lamp segments anywhere along its total length creates a cross-sectional pattern that (i) is symmetrical with respect to a flat plane disposed parallel to the central axis, and (ii) includes nothing but cross-sections of substantially identical cylindrical lamp segments;

an electronic sub-assembly having input terminals and output terminals; the input terminals being connected with the base terminals; the output terminals being connected with the lamp terminals; the electronic sub-assembly being operative to supply an alternating voltage at its output terminals provided it be supplied with an AC power line voltage at its input terminals; the frequency of the alternating voltage being distinctly higher than that of the AC power line voltage; the electronic subassembly being additionally characterized by including a voltage-doubling rectifier assembly connected with the base terminals as well as with a pair of DC terminals across which exists a DC voltage of substantially constant magnitude; which substantially constant magnitude is distinctly higher than the peak absoltr magnitude of the AC power line voltage; and

housing means mounted rigidly on the screw-base and operative to house the electronic sub-assembly as well as to hold and support the fluorescent lamp, thereby to form a fluorescent lamp entity adapted to be screwed into and powered from a lamp socket at which ordinary AC power line voltage is provided.

* * * * *

EXHIBIT L

US006211619B1

(12) **United States Patent**
Nilssen

(10) **Patent No.: US 6,211,619 B1**
(45) **Date of Patent: Apr. 3, 2001**

(54) **ELECTRONIC BALLAST CATHODE HEATING CIRCUIT**

(76) Inventor: **Ole K. Nilssen**, 408 Caesar Dr., Barrington, IL (US) 60010

(*) Notice: Subject to any disclaimer, the term of this patent is extended or adjusted under 35 U.S.C. 154(b) by 1252 days.

(21) Appl. No.: **08/502,817**

(22) Filed: **Jul. 14, 1995**

Related U.S. Application Data

(63) Continuation of application No. 08/394,251, filed on Feb. 24, 1995, now Pat. No. 6,172,464, which is a continuation-in-part of application No. 07/579,569, filed on Sep. 10, 1990, now abandoned, which is a continuation-in-part of application No. 06/787,692, filed on Oct. 15, 1985, now abandoned, which is a continuation of application No. 06/644,155, filed on Aug. 27, 1984, now abandoned, which is a continuation of application No. 06/555,426, filed on Nov. 23, 1983, now abandoned, which is a continuation of application No. 06/178,107, filed on Aug. 14, 1980, now abandoned.

(51) **Int. Cl.**[7] **H05B 37/02**

(52) **U.S. Cl.** **315/106**; 315/244; 315/225; 315/307; 315/DIG. 7

(58) **Field of Search** .... 315/244, 100, 315/101, 105, 99, 106, 107, DIG. 7, 307, 219, 209 R, 225

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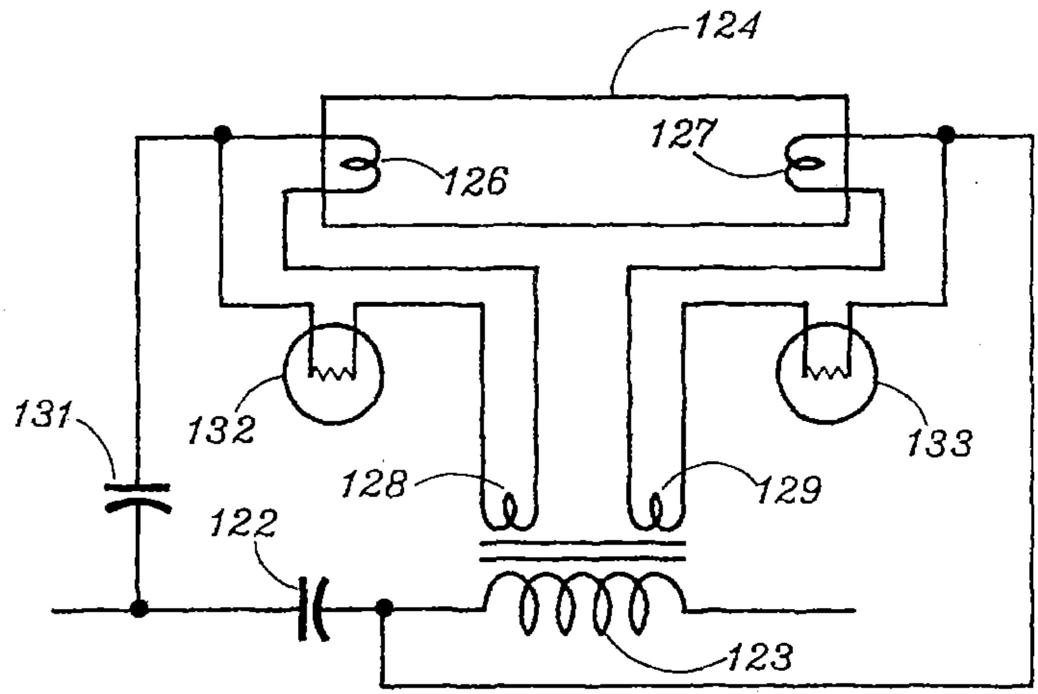
Primary Examiner—Michael B. Shingleton

(57) **ABSTRACT**

In a high-frequency electronic ballast, a screw-in fluorescent lamp is connected with and powered by way of a series-resonant LC circuit. A resistive load is connected with the LC circuit, thereby to constitute a load therefor before ignition of the fluorescent lamp or in case the fluorescent lamp were to fail to ignite.

8 Claims, 3 Drawing Sheets



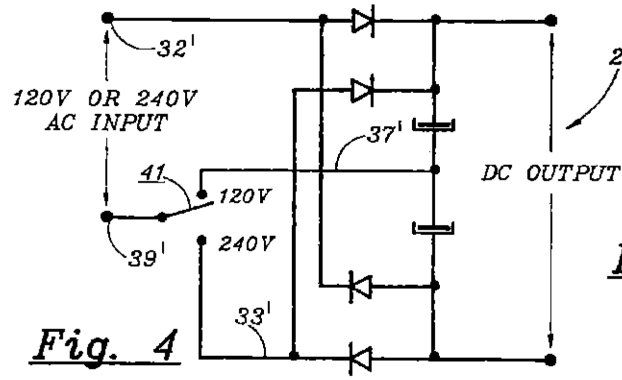


Fig. 4

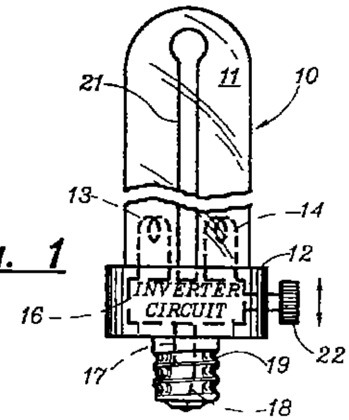


Fig. 1

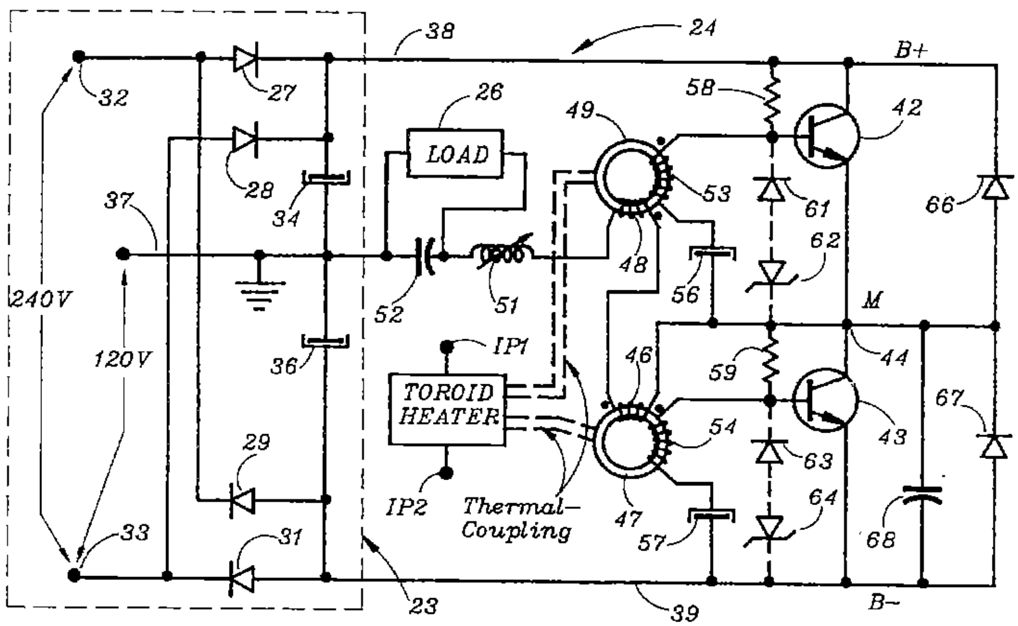


Fig. 2

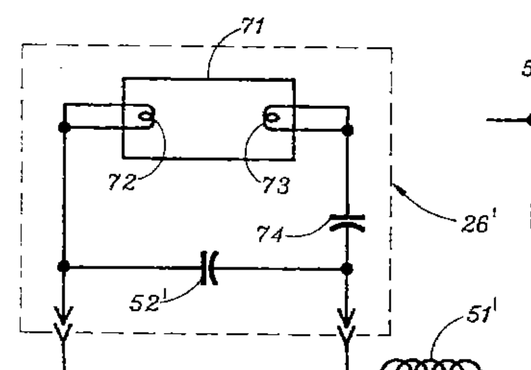


Fig. 5

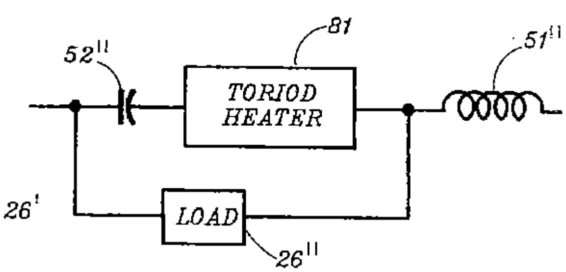


Fig. 6

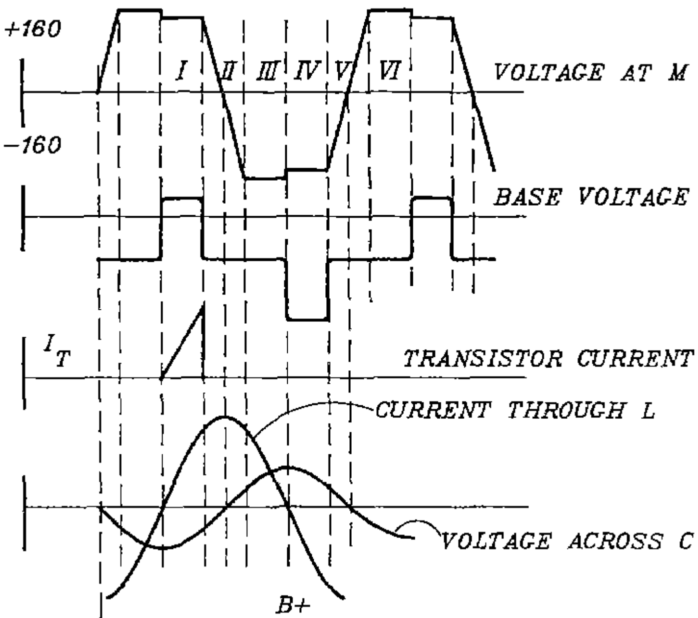


Fig. 3A

Fig. 3B

Fig. 3C

Fig. 3D

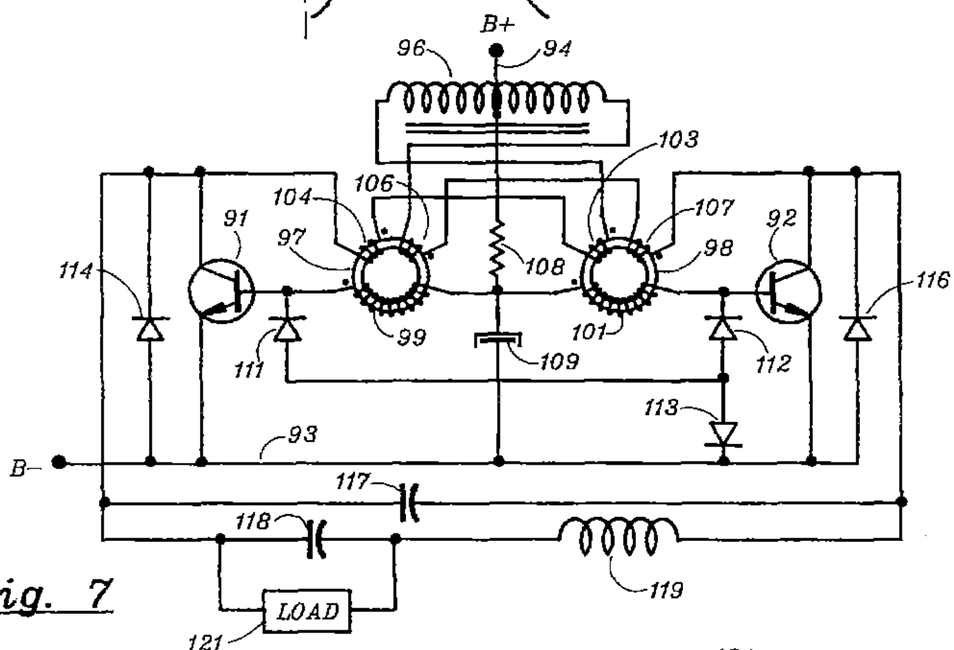


Fig. 7

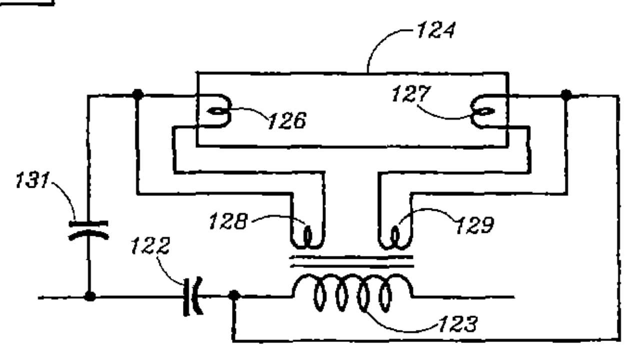


Fig. 8

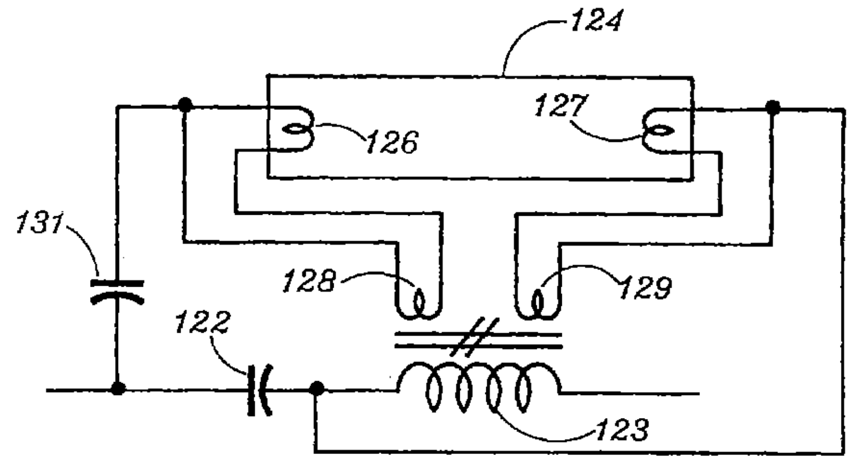

124
127
126
131
128
129
122
123

Fig. 9

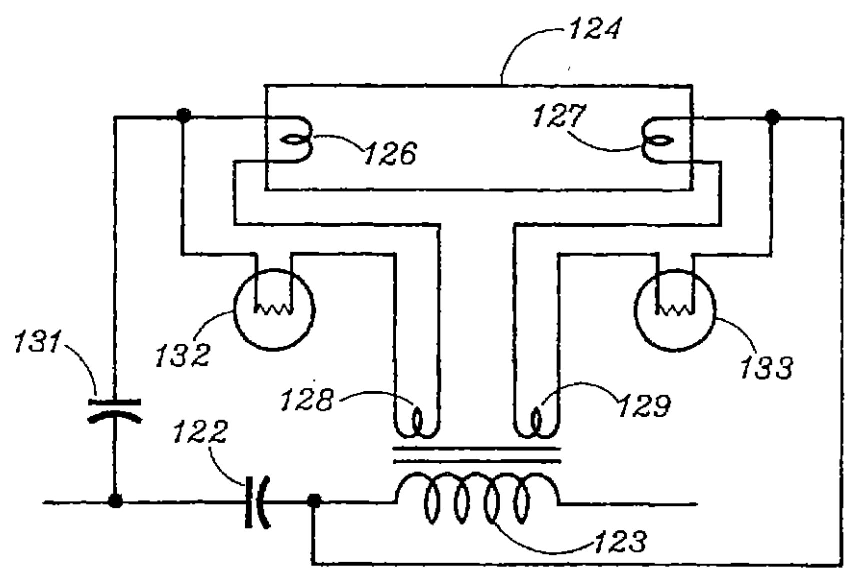

124
127
126
131
132
133
128
129
122
123

Fig. 10

ELECTRONIC BALLAST CATHODE HEATING CIRCUIT

RELATED APPLICATIONS

The present application is a Continuation of Ser. No. 08/394,251 filed Feb. 24, 1995 now U.S. Pat. No. 6,172,464; which is a Continuation-in-Part of Ser. No. 07/579,569 filed Sep. 10, 1990, now abandoned; which is a Continuation-in-Part of Ser. No. 06/787,692 filed Oct. 15, 1985 now abandoned; which is a Continuation of Ser. No. 06/644,155 filed Aug. 27, 1984, now abandoned; which is a Continuation of Ser. No. 06/555,426 filed Nov. 23, 1983, now abandoned; which is a Continuation of Ser. No. 06/178,107 filed Aug. 14, 1980, now abandoned.

BACKGROUND OF INVENTION

1. Field of Invention

This invention relates to electronic ballasts for rapid-start fluorescent lamps, particularly where the lamps are powered via a series-resonant LC circuit.

2. Description of Prior Art

For a description of pertinent prior art, reference is made to U.S. Pat. No. 4,677,345 to Nilssen; which patent issued from a Division of application Ser. No. 06/178,107 filed Aug. 14, 1980; which application is the original progenitor of instant application.

Otherwise, reference is made to the following U.S. Pat. Nos. 3,263,122 to Genuit; U.S. Pat. No. 3,320,510 to Locklair; U.S. Pat. No. 3,996,493 to Davenport et el.; U.S. Pat. No. 4,100,476 to Ghiringhelli; U.S. Pat. No. 4,262,327 to Kovacik et al.; U.S. Pat. No. 4,370,600 to Zansky; U.S. Pat. No. 4,634,932 to Nilssen; and U.S. Pat. No. 4,857,806 to Nilssen.

SUMMARY OF THE INVENTION

Objects of the Invention

Objects of the present invention are those of providing for cost-effective electronic ballasts as well as compact screw-in fluorescent lamps.

This as well as other objects, features and advantages of the present invention will become apparent from the following description and claims.

Brief Description

The present invention is directed to providing improved inverter circuits for powering and controlling gas discharge lamps. The inverter circuits according to the present invention are highly efficient, can be compactly constructed and are ideally suited for energizing gas discharge lamps, particularly "instant-start" and "self-ballasted" fluorescent lamps.

According to one form of the present invention, a series-connected combination of an inductor and a capacitor is provided in circuit with the inverter transistors to be energized upon periodic transistor conduction. Transistor drive current is preferably provided through the use of at least one saturable inductor to control the transistor inversion frequency to be equal to or greater than the nature resonant frequency of the inductor and capacitor combination. The high voltages efficiently developed by loading the inverter with the inductor and capacitor are ideally suited for energizing external loads such as gas discharge lamps. In such an application, the use of an adjustable inductor permits control

of the inverter output as a means of adjusting the level of lamp illumination.

According to another important form of the present invention, reliable and highly efficient half-bridge inverters include a saturable inductor in a current feedback circuit to drive the transistors for alternate conduction. The inverters also include a load having an inductance sufficient to effect periodic energy storage for self-sustained transistor inversion. Importantly, improved reliability is achieved because of the relatively low and transient-free voltages across the transistors in these half-bridge inverters.

Further, according to another feature of the present invention, novel and economical power supplies particularly useful with the disclosed inverter circuits convert conventional AC input voltages to DC for supplying to the inverters.

Yet further, according to still another feature of the invention, a rapid-start fluorescent lamp is powered by way of a series-resonant LC circuit; while heating power for the lamp's cathodes is provided via loosely-coupled auxiliary windings on the tank inductor of the LC circuit. Alternatively, cathode heating power is provided from tightly-coupled windings on the tank inductor; in which case output current-limiting is provided via a non-linear resistance means, such as an incandescent filament in a light bulb, connected in series with the output of each winding.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. **1** is a front elevation of a folded fluorescent lamp unit adapted for screw-in insertion into a standard Edison incandescent socket;

FIG. **2** is a schematic diagram illustrating the essential features of a push-pull inverter circuit particularly suitable for energizing the lamp unit of FIG. **1**;

FIG. **3A–3D** is a set of waveform diagrams of certain significant voltages and currents occurring in the circuit of FIG. **2**;

FIG. **4** is a schematic diagram of a DC power supply connectable to both 120 and 240 volt AC inputs;

FIG. **5** is a schematic diagram which illustrates the connection of a non-self-ballasted gas discharge lamp unit to the FIG. **2** inverter circuit;

FIG. **6** is a schematic diagram which illustrates the use of a toroid heater for regulation of the inverter output;

FIG. **7** is an alternate form of push-pull inverter circuit according to the present invention;

FIG. **8** is a schematic diagram showing the connection of a gas discharge lamp of the "rapid-start" type to an inductor-capacitor-loaded inverter according to the present invention;

FIG. **9** is a modification of FIG. **8**, showing loosely-coupled auxiliary windings on the tank inductor; and

FIG. **10** is another modification of FIG. **8**, showing non-linear current-limiting means connected with the output of tightly-coupled auxiliary windings on the tank inductor.

DESCRIPTION OF THE PREFERRED EMBODIMENTS

FIG. **1** illustrates a screw-in gas discharge lamp unit **10** comprising a folded fluorescent lamp **11** suitably secured to an integral base **12**. The lamp comprises two cathodes **13**, **14** which are supplied with the requisite high operating voltage from a frequency-converting power supply and ballasting circuit **16**; which, because of its compact size, conveniently fits within the base **12**.

The inverter circuit 16 is connected by leads 17, 18 to a screw-type plug 19 adapted for screw-in insertion into a standard Edison-type incandescent lamp socket at which ordinary 120 Volt/60 Hz power line voltage is available. A ground plane comprising a wire or metallic strip 21 is disposed adjacent a portion of the fluorescent lamp 11 as a starting aid. Finally, a manually rotatable external knob 22 is connected to a shaft for mechanical adjustment of the air gap of a ferrite core inductor to vary the inductance value thereof in order to effect adjustment of the inverter voltage output connected to electrodes 13, 14 for controlled variation of the lamp illumination intensity.

With reference to FIG. 2, a power supply 23, connected to a conventional AC input, provides a DC output for supplying a high-efficiency inverter circuit 24. The inverter is operable to provide a high voltage to an external load 26, which may comprise a gas discharge device such as the fluorescent lamp 11 of FIG. 1.

The power supply 23 comprises bridge rectifier having four diodes 27, 28, 29 and 31 connectable to a 240 volt AC supply at terminals 32, 33. Capacitors 34, 36 are connected between a ground line 37 (in turn directly connected to the inverter 24) and to a B+ line 38 and a B− line 39, respectively. The power supply 23 also comprises a voltage doubler and rectifier optionally connectable to a 120 volt AC input taken between the ground line 37 and terminal 33 0r 32. The voltage doubler and rectifier means provides a direct electrical connection by way of line 37 between one of the 120 volt AC power input lines and the inverter 24, as shown in FIG. 2. The bridge rectifier and the voltage doubler and rectifier provide substantially the same DC output voltage to the inverter 24 whether the AC input is 120 or 240 volts. Typical voltages are +160 volts on the B+ line 38 and −160 volts on the B− line 39.

With additional reference to FIG. 4, which shows an alternate power supply 23', the AC input, whether 120 or 240 volts, is provided at terminals 32' and 39. Terminal 39 is in turn connected through a single-pole double-throw selector switch 41 to terminal 37' (for 120 volt operation) or terminal 33' (for 240 volt operation). In all other respects, power supplies 23 and 23' are identical.

The inverter circuit 24 of FIG. 2 is a half-bridge inverter comprising transistors 42, 43 connected in series across the DC voltage output of the power supply 23 on B+ and B− lines 38 and 39, respectively. The collector of transistor 42 is connected to the B+ line 38, the emitter of transistor 42 and the collector of transistor 43 are connected to a midpoint line 44 (designated "M") and the emitter of transistor 43 is connected to the B− line 39. The midpoint line 44 is in turn connected to the ground line 37 through primary winding 46 of a toroidal saturable core transformer 47, a primary winding 48 on an identical transformer 49, an inductor 51 and a series-connected capacitor 52. The inductor 51 and capacitor 52 are energized upon alternate transistor conduction in a manner to be described later.

An external load 26 is preferably taken off capacitor 52, as shown in FIG. 2. The inductor 51, preferably a known ferrite core inductor, has an inductance variable by mechanical adjustment of the air gap in order to effect variation in the level of the inductor and capacitor voltage and hence the power available to the load, as will be described. When the load is a gas discharge lamp such as lamp 11 in FIG. 1, variation in this inductance upon rotation of knob 22 accomplishes a lamp dimming effect.

Drive current to the base terminals of transistors 42 and 43 is provided by secondary windings 53, 54 of transformers 49, 47, respectively. Winding 53 is also connected to midpoint lead 44 through a bias capacitor 56, while winding 54 is connected to the B− lead 39 through an identical bias capacitor 57. The base terminals of transistors 42 and 43 are also connected to lines 38 and 44 through bias resistors 58 and 59, respectively. For a purpose to be described later, the base of transistor 42 can be optionally connected to a diode 61 and a series Zener diode 64 in turn connected to the midpoint line 44; similarly, a diode 63 and series Zener diode 64 in turn connected to the B− line 39 can be connected to the base of transistor 43. Shunt diodes 66 and 67 are connected across the collector-emitter terminals of transistors 42 and 43, respectively. Finally, a capacitor 68 is connected across the collector-emitter terminals of transistor 43 to restrain the rate of voltage rise across those terminals, as will be seen presently.

The operation of the circuit of FIG. 2 can best be understood with additional reference to FIG. 3, which illustrates significant portions of the waveforms of the voltage at midpoint M (FIG. 3A), the base-emitter voltage on transistor 42 (FIG. 3B), the current through transistor 42 (FIG. 3C), and the capacitor 52 voltage and the inductor 51 current (FIG. 3D).

Assuming that transistor 42 is first to be triggered into conduction, current flows from the B+ line 38 through windings 46 and 38 and the inductor 51 to charge capacitor 52 and returns through capacitor 34 (refer to the time period designated I in FIG. 3). When the saturable inductor 49 saturates at the end of period I, drive current to the base of transistor 42 will terminate, causing voltage on the base of the transistor to drop to the negative voltage stored on the bias capacitor 56 in a manner to be described, causing this transistor to become non-conductive. As shown in FIG. 3C, current-flow in transistor 43 terminates at the end of period I.

Because the current through inductor 51 cannot change instantaneously, current will flow from the B− bus 39 through capacitor 68, causing the voltage at midpoint line 44 to drop to −160 volts (period II in FIG. 3). The capacitor 68 restrains the rate of voltage change across the collector and emitter terminals of transistor 42. The current through the inductor 51 reaches its maximum value when the voltage at the midpoint line 44 is zero. During period III, the current will continue to flow through inductor 51 but will be supplied from the B− bus through the shunt diode 67. It will be appreciated that during the latter half of period II and all of period III, positive current is being drawn from a negative voltage; which, in reality, means that energy is being returned to the power supply through a path of relatively low impedance.

When the inductor current reaches zero at the start of period IV, the current through the primary winding 46 of the saturable inductor 47 will cause a current to flow out of its secondary winding 54 to cause transistor 43 to become conductive, thereby causing a reversal in the direction of current through inductor 51 and capacitor 52. When transformer 47 saturates at the end of period IV, the drive current to the base of transistor 43 terminates and the current through inductor 51 will be supplied through capacitor 68, causing the voltage at midpoint line 44 to rise (period V). When the voltage at the midpoint line M reaches 160 volts, the current will then flow through shunt diode 66 (period VI). The cycle is then repeated.

As seen in FIG. 3, saturable transformers 47, 49 provide transistor drive current only after the current through inductor 51 has diminished to zero. Further, the transistor drive current is terminated before the current through inductor 51 has reached its maximum amplitude. This coordination of base drive current and inductor current is achieved because of the series-connection between the inductor 51 and the primary windings 46, 48 of saturable transformers 47, 49, respectively.

The series-connected combination of the inductor 51 and the capacitor 52 is energized upon the alternate conduction of transistors 42 and 43. With a large value of capacitance of capacitor 52, very little voltage will be developed across its terminals. As the value of this capacitance is decreased, however, the voltage across this capacitor will increase. As the value of the capacitor 52 is reduced to achieve resonance with the inductor 51, the voltage on the capacitor will rise and become infinite in a loss-free circuit operating under ideal conditions.

It has been found desirable to regulate the transistor inversion frequency, determined mainly by the saturation time of the saturable inductors 47, 49, to be equal to or higher than the natural resonance frequency of the inductor and capacitor combination in order to provide a high voltage output to external load 26. A high voltage across capacitor 52 is efficiently developed as the transistor inversion frequency approaches the natural resonant frequency of the inductor 51 and capacitor 52 combination. Stated another way, the conduction period of each transistor is desirably shorter in duration than one quarter of the full period corresponding to the natural resonant frequency of the inductor and capacitor combination. When the inverter 24 is used with a self-ballasted gas discharge lamp unit, it has been found that the inversion frequency can be at least equal to the natural resonant frequency of the tank circuit. If the capacitance value of capacitor 52 is reduced still further beyond the resonance point, unacceptably high transistor currents will be experienced during transistor switching and transistor burn-out will occur.

It will be appreciated that the sizing of capacitor 52 is determined by the application of the inverter circuit 24. Variation in the values of the capacitor 52 and the inductor 51 will determine the voltages developed in the inductor-capacitor tank circuit. The external load 26 may be connected in circuit with the inductor 51 (by a winding on the inductor, for example) and the capacitor may be omitted entirely. If the combined circuit loading of the inductor 51 and the external load 26 has an effective inductance of value sufficient to effect periodic energy storage for self-sustained transistor inversion, the current feedback provided by the saturable inductors 47,49 will effect alternate transistor conduction without the need for additional voltage feedback. When the capacitor 52 is omitted, the power supply 23 provides a direct electrical connection between one of the AC power input lines and the inverter load circuit.

Because the voltages across transistors 42, 43 are relatively low (due to the effect of capacitors 34, 36), the half-bridge inverter 24 is very reliable. The absence of switching transients minimizes the possibility of transistor burn-out.

The inverter circuit 24 comprises means for supplying reverse bias to the conducting transistor upon saturation of its associated saturable inductor. For this purpose, the capacitors 56 and 57 are charged to negative voltages as a result of reset current flowing into secondary windings 53, 54 from the bases of transistors 42, 43, respectively. This reverse current rapidly turns off a conducting transistor to increase its switching speed and to achieve inverter circuit efficiency in a manner described more fully in my co-pending U.S. patent application Ser. No. 103,624 filed Dec. 14, 1979 and entitled "Bias Control for High Efficiency Inverter Circuit" (now U.S. Pat. No. 4,307,353). The more negative the voltage on the bias capacitors 56 and 57, the more rapidly charges are swept out of the bases of their associated transistors upon transistor turn-off.

When a transistor base-emitter junction is reversely biased, it exhibits the characteristics of a Zener diode having a reverse breakdown voltage on the order of 8 to 14 Volt for transistors typically used in high-voltage inverters. As an alternative, to provide a negative voltage smaller in magnitude on the base lead of typical transistor 42 during reset operation, the optional diode 61 and Zener diode 62 combination can be used. For large values of the bias capacitor 56, the base voltage will be substantially constant.

If the load 26 comprises a gas discharge lamp, the voltage across the capacitor 52 will be reduced once the lamp is ignited to prevent voltages on the inductor 51 and the capacitor 52 from reaching destructive levels. Such a lamp provides an initial time delay during which a high voltage, suitable for instant starting, is available.

FIG. 5 illustrates the use of an alternate load 26' adapted for plug-in connection to an inverter circuit such as shown in FIG. 2. The load 26' consists of a gas discharge lamp 71 having electrodes 72, 73 and connected in series with a capacitor 74. The combination of lamp 71 and capacitor 74 is connected in parallel with a capacitor 52' which serves the same purpose as capacitor 52 in the FIG. 2 circuit. However, when the load 26' is unplugged from the circuit, the inverter stops oscillating and the development of high voltages in the inverter is prevented. The fact that no high voltages are generated by the circuit if the lamp is disconnected while the circuit is oscillating is important for safety reasons.

FIG. 6 illustrates a capacitor 52" connected in series with an inductor 51" through a heater 81 suitable for heating the toroidal inductors 47, 49 in accordance with the level of output. The load 26" is connected across the series combination of the capacitor 52" and the toroid heater. The heater 81 is preferably designed to controllably heat the toroidal saturable inductors in order to decrease their saturation flux limit and hence their saturation time. The result is to decrease the periodic transistor conduction time and thereby increase the transistor inversion frequency. When a frequency-dependent impedance means, that is, an inductor or a capacitor, is connected in circuit with the AC voltage output of the inverter, change in the transistor inversion frequency will modify the impedance of the frequency-dependent impdance means and correspondingly modify the inverter output. Thus as the level of the output increases, the toroid heater 81 is correspondingly energized to effect feedback regulation of the output. Further, transistors 42, 43 of the type used in high voltage inverters dissipate heat during periodic transistor conduction. As an alternative, the toroid heater 81 can use this heat for feedback regulation of the output or control of the temperature of transistors 42, 43.

The frequency dependent impedance means may also be used in a circuit to energize a gas discharge lamp at adjustable illumination levels. Adjustment in the inversion frequency of transistors 42, 43 results in control of the magnitude of the AC current supplied to the lamp. This is preferably accomplished where saturable inductors 47, 49 have adjustable flux densities for control of their saturation time.

FIG. 7 schematically illustrates an alternate form of inverter circuit, shown without the AC to DC power supply connections for simplification. In this Figure, the transistors are connected in parallel rather than in series but the operation is essentially the same as previously described.

In particular, this circuit comprises a pair of alternately conducting transistors **91**, **92**. The emitter terminals of the transistors are connected to a B− line **93**. A B+ lead **94** is connected to the center-tap of a transformer **96**. In order to provide drive current to the transistors **91**, **92** for control of their conduction frequency, saturable inductors **97**, **98** have secondary windings **99**, **10**, respectively, each secondary winding having one end connected to the base of its associated transistor; the other ends are connected to a common terminal **102**. One end of transformer **96** is connected to the collector of transistor **91** through a winding **103** on inductor **98** in turn connected in series with a winding **104** on inductor **97**. Likewise, the other end of transformer **96** is connected to the collector of transistor **92** through a winding **106** on inductor **97** in series with another winding **107** on inductor **98**.

The B+ terminal is connected to terminal **102** through a bias resistor **108**. A bias capacitor **109** connects terminal **102** to the B− lead **93**. This resistor and capacitor serve the same function as resistors **58**, **59** and capacitors **56**, **57** in the FIG. 2 circuit.

The bases of transistors **91**, **92** are connected by diodes **111**, **112**, respectively, to a common Zener diode **113** in turn connected to the B− lead **93**. The common Zener diode **113** serves the same function as individual Zener diodes **62**, **64** in FIG. 2.

Shunt diodes **114**, **116** are connected across the collector-emitter terminals of transistors **91**, **92**, respectively. A capacitor **117** connecting the collectors of transistors **91**, **92** restrains the rate of voltage rise on the collectors in a manner similar to the collector-emitter capacitor **68** in FIG. 2.

Inductive-capacitive loading of the FIG. 7 inverter is accomplished by a capacitor **118** connected in series with an inductor **119**, the combination being connected across the collectors of the transistors **91**, **92**. A load **121** is connected across the capacitor **118**.

FIG. **8** illustrates how an inverter loaded with a series capacitor **122** and inductor **123** can be used to energize a "rapid-start" fluorescent lamp **124** (the details of the inverter circuit being omitted for simplication). The lamp **124** has a pair of cathodes **126**, **127** connected across the capacitor **122** for supply of operating voltage in a manner identical to that previously described. In addition, the inductor **123** comprises a pair of magnetically-coupled auxiliary windings **128**, **129** for electrically heating the cathodes **126**, **127**, respectively. A small capacitor **131** is connected in series with lamp **124**.

FIG. **9** illustrates the very same circuit arrangement as that of FIG. **8** except that the auxiliary windings **128**, **129** are only loosely coupled to the inductor **123**, thereby providing for a manifest limitation on the amount of current that can be drawn from each auxiliary winding in case it were to be accidentally short-circuited.

FIG. **10** also illustrates the very same circuit arrangement as that of FIG. **8** except that the cathodes **126**, **127** are connected with their respective auxiliary windings **128**, **129** by way of non-linear current-limiting means **132** and **133**, respectively.

In FIG. **10**, the non-linear current-limiting means **132**, **133** are shown as being two (small) incandescent lamps. However, other types of non-linear resistance means could be used as well.

Both the FIG. **9** circuit and the FIG. **10** circuit serve the same basic purpose; which is that of preventing damage to the ballast circuit (such as that if FIG. **2**) in case the leads used for connecting to one of the lamp cathodes **126**, **127** were to be accidentally shorted. This damage prevention is accomplished by providing for manifest limitation of the maximum amount of current that can be drawn from each one of the auxiliary windings **128**, **129**. In the circuit of FIG. **9**, this manifest limitation is accomplished by having the auxiliary windings **128**, **129** couple sufficiently loosely to the main inductor **123**—such as by providing a magnetic shunt between inductor **123** and the auxiliary windings—thereby correspondingly limiting the degree of impact resulting from an accidental short circuit. Such a short circuit would result in a net reduction in the effective inductance value of the tank inductor **123**; which net reduction in inductance may, in turn cause a precipitous increase in the magnitude of the reactive current drawn from the inverter by the series-connected inductor **123** and capacitor **122**, thereby causing damage to the inverter.

Additional Explanations and Comments

(a) With reference to FIGS. 2 and **5**, adjustment of the amount of power supplied to load **26'**, and thereby the amount of light provided by lamp **71**, may be accomplished by applying a voltage of adjustable magnitude to input terminals IP1 and IP2 of the Toroid Heater; which is thermally coupled with the toroidal ferrite cores of saturable transformers **47**, **49**.

(b) With commonly available components, inverter circuit **24** of FIG. **2** can be made to operate efficiently at any frequency between a few kHz to perhaps as high as 50 kHz. However, for various well-known reasons (i.e., eliminating audible noise, minimizing physical size, and maximizing efficiency), the frequency actually chosen is in the range of 20 to 40 kHz.

(c) The fluorescent lighting unit of FIG. **1** could be made in such manner as to permit fluorescent lamp **11** to be disconnectable from its base **12** and ballasting means **16**. However, if powered with normal line voltage without its lamp load connected, frequency-converting power supply and ballasting circuit **16** is apt to self-destruct.

To avoid such self-destruction, arrangements can readily be made whereby the very act of removing the load automatically establishes a situation that prevents the possible destruction of the power supply and ballasting means. For instance, with the tank capacitor (**52**) being permanently connected with the lamp load (**11**)—thereby automatically being removed whenever the lamp is removed—the inverter circuit is protected from self-destruction.

(d) At frequencies above a few kHz, the load represented by a fluorescent lamp—once it is ignited—is substantially resistive. Thus, with the voltage across lamp **11** being of a substantially sinusoidal waveform (as indicated in FIG. 3D), the current through the lamp will also be substantially sinusoidal in waveshape.

(e) In the fluorescent lamp unit of FIG. **1**, fluorescent lamp **11** is connected with power supply and ballasting circuit **16** in the exact same manner as is load **26** connected with the circuit of FIG. 2. That is, it is connected in parallel with the tank capacitor (**52**) of the L-C series-resonant circuit. As is conventional in instant-start fluorescent lamps—such as lamp **11** of FIG. **1**—the two terminals from each cathode are shorted together, thereby to constitute a situation where each cathode effectively is represented by only a single terminal. However, it is not necessary that the two terminals from each cathode be shorted together; in which case—for instant-start operation—connection from a lamp's power supply and ballasting means need only be made with one of the terminals of each cathode.

(f) With respect to the circuit arrangement of FIG. 9, in situations where the tank inductor 123 includes a ferrite magnetic core having an air gap, one particularly cost-effective way of accomplishing the indicated loose coupling between the tank inductor 123 and the auxiliary windings 128, 129 is that of arranging for the auxiliary windings to be placed in the air gap in such a manner that they each couple only with part of the magnetic flux crossing the air gap.

(g) In FIG. 1, the compact screw-in fluorescent lamp has a longitudinal central axis penetrating through the center of the bottom of base 19 (i.e., at the point where lead 18 is connected), passing up centrally between the two legs of lamp 11, and emerging at the center of the very top of lamp 11.

What is claimed is:

1. An arrangement comprising:

a source providing an alternating voltage across a pair of source terminals; the alternating voltage having a fundamental frequency distinctly higher than that of the AC voltage on an ordinary electric utility power line;

a series-combination of an inductor and a capacitor; the series-combination being: (i) naturally resonant at a frequency lower than said fundamental frequency, (ii) effectively connected across the source terminals, thereby to draw a source current from the source terminals, and (iii) connected in circuit with a pair of output terminals across which is provided an approximately sinusoidal output voltage; the inductor being coupled with an auxiliary winding, thereby to cause an auxiliary voltage to be provided from this auxiliary winding; the coupling between the inductor and the auxiliary winding being sufficiently loose so that, in case an electrical short circuit were to be placed across the auxiliary winding, the magnitude of the source current would be prevented from increasing to a detrimentally high level; and

a gas discharge lamp means having a first thermionic cathode with a pair of cathode terminals connected with the auxiliary winding by way of a connect means; the lamp means also having a second thermionic cathode; the approximately sinusoidal output voltage being applied between the first and the second thermionic cathodes.

2. An arrangement comprising:

a source providing an alternating voltage across a pair of source terminals; the alternating voltage having a fundamental frequency distinctly higher than that of the AC voltage on an ordinary electric utility power line;

a series-combination of an inductor and a capacitor; the series-combination being: (i) naturally resonant at a frequency lower than said fundamental frequency, (ii) effectively connected across the source terminals, thereby to draw a source current from the source terminals, and (iii) connected in circuit with a pair of output terminals across which is provided an approximately sinusoidal output voltage; the inductor means being coupled with an auxiliary winding, thereby to cause an auxiliary voltage to be provided from this auxiliary winding; the coupling between the inductor and the auxiliary winding being sufficiently loose so that, in case an electrical short circuit were to be placed across the auxiliary winding, the magnitude of the source current would be prevented from increasing to a detrimentally high level; and

a gas discharge lamp having a first thermionic cathode with a pair of cathode terminals connected with the auxiliary winding by way of a connect means; the lamp also having a second thermionic cathode; the substantially sinusoidal output voltage being applied between the first and the second thermionic cathodes.

3. An arrangement comprising:

a source providing an alternating voltage across a pair of source terminals; the alternating voltage having a fundamental frequency distinctly higher than that of the AC voltage on an ordinary electric utility power line;

a series-combination of an inductor and a capacitor; the series-combination being: (i) naturally resonant at a frequency lower than said fundamental frequency, (ii) effectively connected across the source terminals, thereby to draw a source current from the source terminals, and (iii) connected in circuit with a pair of output terminals across which is provided an approximately sinusoidal output voltage; the inductor means being coupled with an auxiliary winding, thereby to cause an auxiliary voltage to be provided from this auxiliary winding; and

a gas discharge lamp means having a first thermionic cathode with a pair of cathode terminals connected with the auxiliary winding by way of a connect means; the lamp means also having a second thermionic cathode; the substantially sinusoidal output voltage being applied between the first and the second thermionic cathodes; the connect means being characterized by including a resistor means.

4. The arrangement of claim 3 wherein the resistor means is a non-linear resistor means.

5. The arrangement of claim 3 wherein the resistor means includes an incandescent filament means.

6. An arrangement comprising:

a source providing an alternating voltage across a pair of source terminals; the alternating voltage having a fundamental frequency distinctly higher than that of the AC voltage on an ordinary electric utility power line;

a series-combination of an inductor and a capacitor; the series-combination being: (i) naturally resonant at a frequency lower than said fundamental frequency, (ii) effectively connected across the source terminals, thereby to draw a source current from the source terminals, and (iii) connected in circuit with a pair of output terminals across which is provided an approximately sinusoidal output voltage; the inductor means being coupled with an auxiliary winding, thereby to cause an auxiliary voltage to be provided from this auxiliary winding; and

a gas discharge lamp means having a first thermionic cathode with a pair of cathode terminals connected with the auxiliary winding by way of a connect means; the lamp means also having a second thermionic cathode; the substantially sinusoidal output voltage being applied between the first and the second thermionic cathodes; the connect means including limiting means operative to manifestly limit to a pre-established level the magnitude of any current drawn from the auxiliary winding.

7. An arrangement comprising:

a source providing an alternating voltage across a pair of source terminals; the alternating voltage having a fundamental frequency distinctly higher than that of the AC voltage on an ordinary electric utility power line;

an inductive reactance combined with a capacitive reactance; the combination being: (i) naturally resonant at a frequency at or near said fundamental frequency, (ii) effectively connected across the source terminals, thereby to draw a source current from the source terminals, and (iii) connected in circuit with a pair of output terminals across which is provided an approximately sinusoidal output voltage; the inductive reactance being coupled with an auxiliary winding, thereby to cause an auxiliary voltage to be provided from this auxiliary winding; the auxiliary winding exhibiting a current-limiting feature by which, in case an electrical short circuit were to be placed across the auxiliary winding, the magnitude of any current thereby drawn therefrom would be manifestly limited to a level sufficiently low not to give rise to a damaging effect; and

a gas discharge lamp having a first thermionic cathode with a pair of cathode terminals connected with the auxiliary winding; the lamp means also having a second thermionic cathode;

the approximately sinusoidal output voltage being applied between the first and the second thermionic cathodes.

8. An arrangement comprising:

a source providing an alternating voltage across a pair of source terminals; the alternating voltage having a fundamental frequency distinctly higher than that of the AC voltage on an ordinary electric utility power line;

an inductive reactance combined with a capacitive reactance; the combination being: (i) naturally resonant at a frequency at or near said fundamental frequency, (ii) effectively connected across the source terminals, thereby to draw a source current from the source terminals, and (iii) connected in circuit with a pair of output terminals across which is provided a high-frequency output voltage; the inductive reactance being coupled with an auxiliary winding, thereby to cause an auxiliary voltage to be provided from this auxiliary winding; the auxiliary winding exhibiting a current-limiting feature by which, in case an electrical short circuit were to be placed across the auxiliary winding, the magnitude of any current drawn therefrom would be prevented from increasing to a detrimentally high level; and

a gas discharge lamp having a first thermionic cathode with a pair of cathode terminals connected with the auxiliary winding; the lamp means also having a second thermionic cathode; the high-frequency output voltage being applied between the first and the second thermionic cathodes.

* * * * *

EXHIBIT M

US006211625B1

(12) **United States Patent**
Nilssen

(10) **Patent No.:** **US 6,211,625 B1**
(45) **Date of Patent:** **Apr. 3, 2001**

(54) **ELECTRONIC BALLAST WITH OVER-VOLTAGE PROTECTION**

(76) Inventor: **Ole K. Nilssen**, Caesar Dr., Barrington, IL (US) 60010

(*) Notice: Subject to any disclaimer, the term of this patent is extended or adjusted under 35 U.S.C. 154(b) by 1388 days.

(21) Appl. No.: **08/571,634**

(22) Filed: **Dec. 13, 1995**

Related U.S. Application Data

(63) Continuation of application No. 08/236,125, filed on May 3, 1994, now abandoned, which is a continuation of application No. 07/579,569, filed on Sep. 10, 1990, now abandoned, which is a continuation-in-part of application No. 06/787,692, filed on Oct. 15, 1985, now abandoned, which is a continuation of application No. 06/644,155, filed on Aug. 27, 1984, now abandoned, which is a continuation of application No. 06/555,426, filed on Nov. 23, 1983, now abandoned, which is a continuation of application No. 06/178,107, filed on Aug. 14, 1980, now abandoned.

(51) **Int. Cl.**[7] **H05B 37/00**; H05B 37/02

(52) **U.S. Cl.** **315/225**; 315/311; 315/284; 315/205; 315/225

(58) **Field of Search** .... 315/311, 284, 315/205, 53, 225, 209 R

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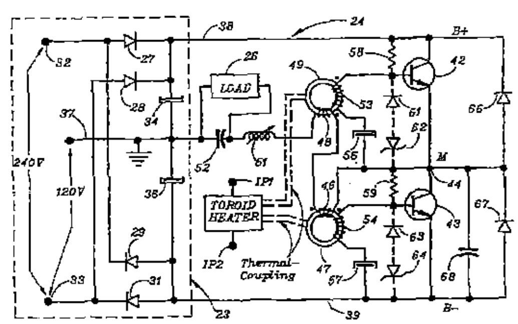
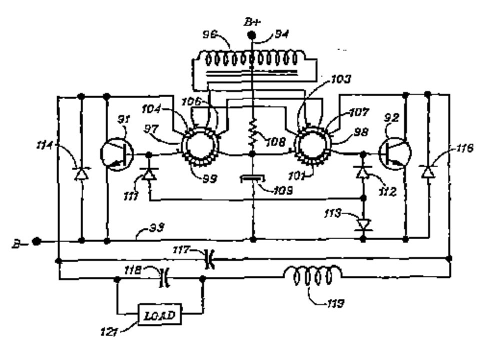
Primary Examiner—Arnold Kinkead

(57) **ABSTRACT**

In an electronic ballast, a rapid start fluorescent lamp is powered by being parallel-connected with the tank-capacitor of a series-resonant LC circuit. Beating power for the lamp's cathodes is obtained by way of loosely-coupled auxiliary windings on the tank inductor of the LC circuit. In case the fluorescent lamp were to be disconnected or otherwise were to fail to properly load the series-resonant LC circuit, due to so-called Q-multiplication, the magnitude of the voltage develped across the tank-capacitor would normally increase to a high and potentially destructive level. However, due to feedback operable to cause the inverter frequency to increase as a function of the magnitude of the voltage across the tank-capacitor, the magnitude of this tank-capacitor voltage is limited to a level substantially lower than what otherwise would be the case.

17 Claims, 3 Drawing Sheets





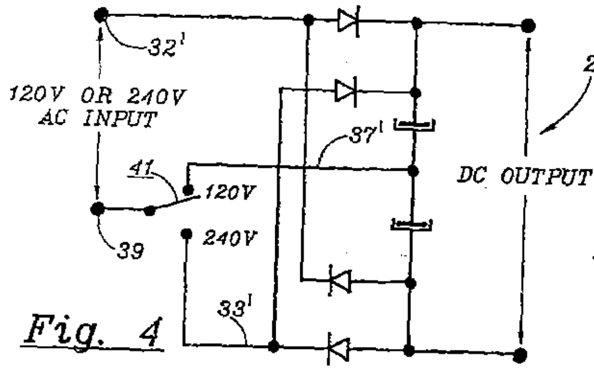


Fig. 4

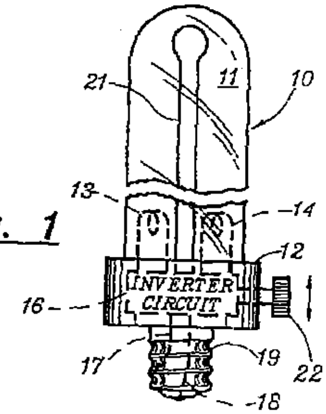


Fig. 1

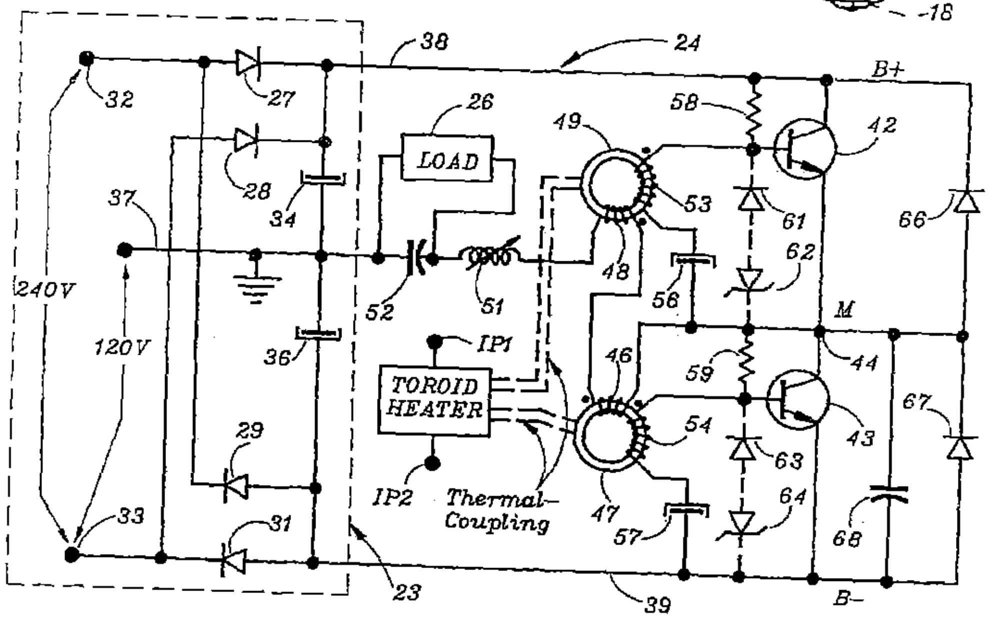


Fig. 2

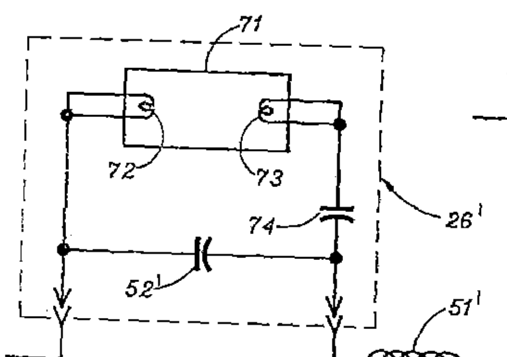


Fig. 5

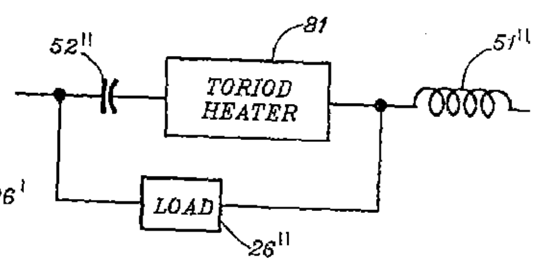


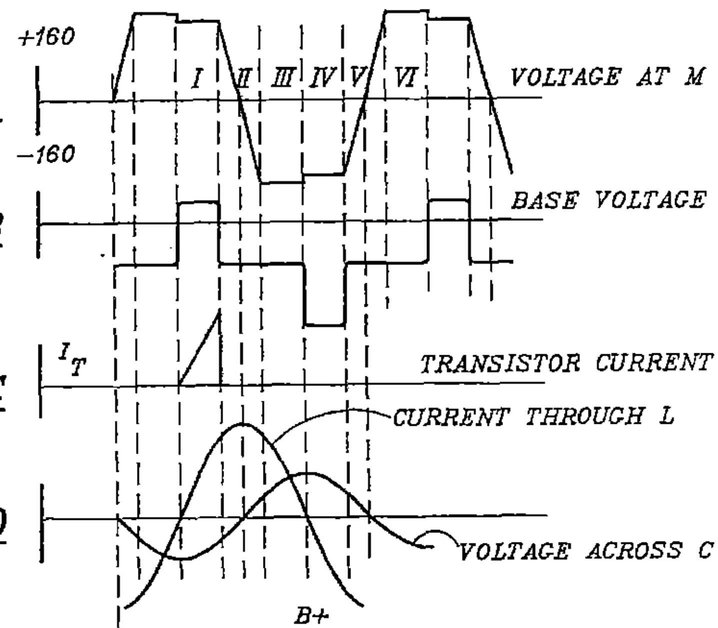
Fig. 6

Fig. 3A

Fig. 3B

Fig. 3C

Fig. 3D



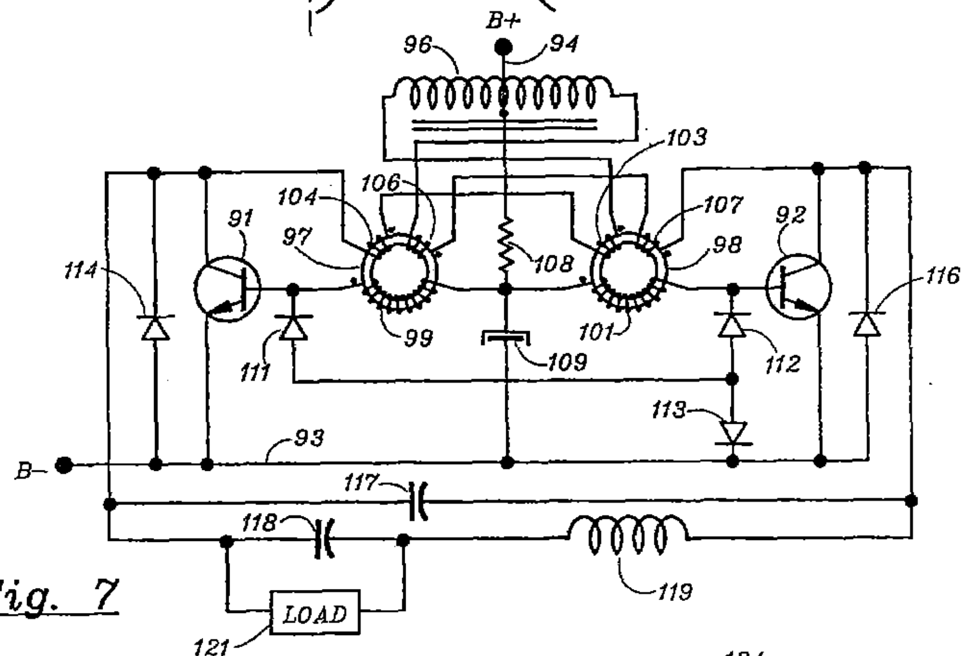


Fig. 7

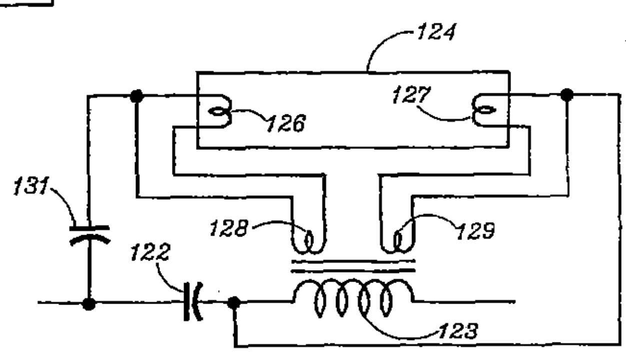


Fig. 8

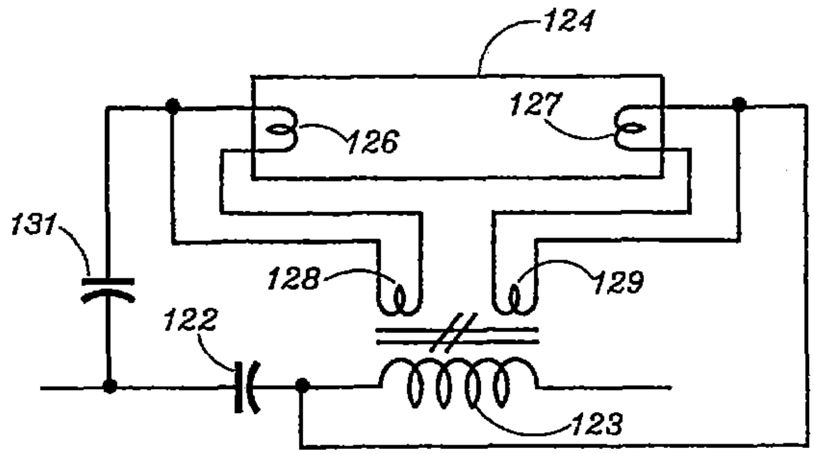

124
127
126
131
128
129
122
123

Fig. 9

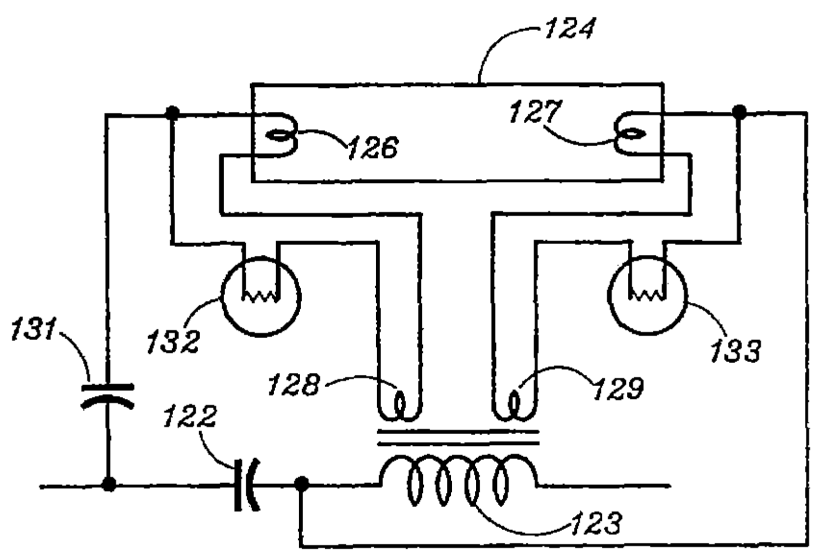

124
127
126
131
132
133
128
129
122
123

Fig. 10

ELECTRONIC BALLAST WITH OVER-VOLTAGE PROTECTION

RELATED APPLICATIONS

The present application is a Continuation of Ser. No. 08/236,125 filed May 3, 1994 which is a continuation of Ser. No. 07/579,569 filed Sep. 10, 1990, now abandoned; which is a Continuation-in-Part of Ser. No. 06/787,692 filed Oct. 15, 1985 now abandoned; which is a Continuation of Ser. No. 06/644,155 filed Aug. 27, 1984, now abandoned; which is a Continuation of Ser. No. 06/555,426 filed Nov. 23, 1983, now abandoned; which is a Continuation of Ser. No. 06/178,107 filed Aug. 14, 1980, now abandoned.

BACKGROUND OF THE INVENTION

1. Field of Invention

This invention relates to electronic ballasts for rapid-start fluorescent lamps, particularly where the lamps are powered via a series-resonant LC circuit.

2. Description of Prior Art

For a description of pertinent prior art, reference is made to U.S. Pat. No. 4,677,345 to Nilssen; which patent issued from a Division of application Ser. No. 06/178,107 filed Aug. 14, 1980; which application is the original progenitor of instant application.

Otherwise, reference is made to the following U.S. patents: No. 3,263,122 to Genuit; No. 3,320,510 to Locklair; No. 3,996,493 to Davenport et el.; No. 4,100,476 to Ghiringhelli; No. 4,262,327 to Kovacik et al.; No. 4,370,600 to Zansky; No. 4,634,932 to Nilssen; and No. 4,857,806 to Nilssen.

SUMMARY OF THE INVENTION

Objects of the Invention

An object of the present invention is that of providing for a cost-effective electronic ballast.

This as well as other objects, features and advantages of the present invention will become apparent from the following description and claims.

Brief Description

The present invention is directed to providing improved inverter circuits for powering and controlling gas discharge lamps. The inverter circuits according to the present invention are highly efficient, can be compactly constructed and are ideally suited for energizing gas discharge lamps, particularly "instant-start" and "self-ballasted" fluorescent lamps.

According to one form of the present invention, a series-connected combination of an inductor and a capacitor is provided in circuit with the inverter transistors to be energized upon periodic transistor conduction. Transistor drive current is preferably provided through the use of at least one saturable inductor to control the transistor inversion frequency to be equal to or greater than the natural resonant frequency of the inductor and capacitor combination. The high voltages efficiently developed by loading the inverter with the inductor and capacitor are ideally suited for energizing external loads such as gas discharge lamps. In such an application, the use of an adjustable inductor permits control of the inverter output as a means of adjusting the level of lamp illumination.

According to another important form of the present invention, reliable and highly efficient half-bridge inverters include a saturable inductor in a current feedback circuit to drive the transistors for alternate conduction. The inverters also include a load having an inductance sufficient to effect periodic energy storage for self-sustained transistor inversion. Importantly, improved reliability is achieved because of the relatively low and transient-free voltages across the transistors in these half-bridge inverters.

Further, according to another feature of the present invention, novel and economical power supplies particularly useful with the disclosed inverter circuits convert conventional AC input voltages to DC for supplying to the inverters.

Yet further, according to still another feature of the invention, a rapid-start fluorescent lamp is powered by way of a series-resonant LC circuit; while heating power for the lamp's cathodes is provided via loosely-coupled auxiliary windings on the tank inductor of the LC circuit. Alternatively, cathode heating power is provided from tightly-coupled windings on the tank inductor; in which case output current-limiting is provided via a non-linear resistance means, such as an incandescent filament in a light bulb, connected in series with the output of each winding.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. **1** is a front elevation of a folded fluorescent lamp unit adapted for screw-in insertion into a standard Edison incandescent socket;

FIG. **2** is a schematic diagram illustrating the essential features of a push-pull inverter circuit particularly suitable for energizing the lamp unit of FIG. **1**;

FIGS. **3A–3D** is a set of waveform diagrams of certain significant voltages and currents occurring in the circuit of FIG. **2**;

FIG. **4** is a schematic diagram of a DC power supply connectable to both 120 and 240 volt AC inputs;

FIG. **5** is a schematic diagram which illustrates the connection of a non-self-ballasted gas discharge lamp unit to the FIG. **2** inverter circuit;

FIG. **6** is a schematic diagram which illustrates the use of a toroid heater for regulation of the inverter output;

FIG. **7** is an alternate form of push-pull inverter circuit according to the present invention;

FIG. **8** is a schematic diagram showing the connection of a gas discharge lamp of the "rapid-start" type to an inductor-capacitor-loaded inverter according to the present invention;

FIG. **9** is a modification of FIG. **8**, showing loosely-coupled auxiliary windings on the tank inductor; and

FIG. **10** is another modification of FIG. **8**, showing non-linear current-limiting means connected with the output of tightly-coupled auxiliary windings on the tank inductor.

DESCRIPTION OF THE PREFERRED EMBODIMENTS

FIG. **1** illustrates a screw-in gas discharge lamp unit **10** comprising a folded fluorescent lamp **11** suitably secured to an integral base **12**. The lamp comprises two cathodes **13, 14** which are supplied with the requisite high operating voltage from a frequency-converting power supply and ballasting circuit **16**; which, because of its compact size, conveniently fits within the base **12**.

The inverter circuit **16** is connected by leads **17, 18** to a screw-type plug **19** adapted for screw-in insertion into a standard Edison-type incandescent lamp socket at which ordinary 120 Volt/60 Hz power line voltage is available. A ground plane comprising a wire or metallic strip **21** is disposed adjacent a portion of the fluorescent lamp 11 as a starting aid. Finally, a manually rotatable external knob 22 is connected to a shaft for mechanical adjustment of the air gap of a ferrite core inductor to vary the inductance value thereof in order to effect adjustment of the inverter voltage output connected to electrodes 13, 14 for controlled variation of the lamp illumination intensity.

With reference to FIG. 2, a power supply 23, connected to a conventional AC input, provides a DC output for supplying a high-efficiency inverter circuit 24. The inverter is operable to provide a high voltage to an external load 26, which may comprise a gas discharge device sich as the fluorescent lamp 11 of FIG. 1.

The power supply 23 comprises bridge rectifier having four diodes 27, 28, 29 and 31 connectable to a 240 volt AC supply at terminals 32, 33. Capacitors 34, 36 are connected between a ground line 37 (in turn directly connected to the inverter 24) and to a B+ line 38 and a B− line 39, respectively. The power supply 23 also comprises a voltage doubler and rectifier optionally connectable to a 120 volt AC input taken between the ground line 37 and terminal 33 or 32. The voltage doubler and rectifier means provides a direct electrical connection by way of line 37 between one of the 120 volt AC power input lines and the inverter 24, as shown in FIG. 2. The bridge rectifier and the voltage doubler and rectifier provide substantially the same DC output voltage to the inverter 24 whether the AC input is 120 or 240 volts. Typical voltages are +160 volts on the B+ line 38 and −160 volts on the B− line 39.

With additional reference to FIG. 4, which shows an alternate power supply 23', the AC input, whether 120 or 240 volts, is provided at terminals 32' and 39. Terminal 39 is in turn connected through a single-pole double-throw selector switch 41 to terminal 37' (for 120 volt operation) or terminal 33' (for 240 volt operation). In all other respects, power supplies 23 and 23' are identical.

The inverter circuit 24 of FIG. 2 is a half-bridge inverter comprising transistors 42, 43 connected in series across the DC voltage output of the power supply 23 on B+ and B− lines 38 and 39, respectively. The collector of trasistor 42 is connected to the B+ line 38, the emitter of transistor 42 and the collector of transistor 43 are connected to a midpoint line 44 (designated "M") and the emitter of transistor 43 is connected to the B− line 39. The midpoint line 44 is in turn connected to the ground line 37 through primary winding 46 of a toroidal saturable core transformer 47, a primary winding 48 on an identical transformer 49, an inductor 51 and a series-connected capacitor 52. The inductor 51 and capacitor 52 are energized upon alternate transistor conduction in a manner to be described later.

An external load 26 is preferably taken off capacitor 52, as shown in FIG. 2. The inductor 51, preferably a known ferrite core inductor, has an inductance variable by mechanical adjustment of the air gap in order to effect variation in the level of the inductor and capacitor voltage and hence the power available to the load, as will be described. When the load is a gas discharge lamp such as lamp 11 in FIG. 1, variation in this inductance upon rotation of knob 22 accomplishes a lamp dimming effect.

Drive current to the base terminals of transistors 42 and 43 is provided by secondary windings 53, 54 of transformers 49, 47, respectively. Winding 53 is also connected to midpoint lead 44 through a bias capacitor 56, while winding 54 is connected to the B− lead 39 through an identical bias capacitor 57. The base terminals of transistors 42 and 43 are also connected to lines 38 and 44 through bias resistors 58

and 59, respectively. For a purpose to be described later, the base of transistor 42 can be optionally connected to a diode 61 and a series Zener diode 64 in turn connected to the midpoint line 44; similarly, a diode 63 and series Zener diode 64 in turn connected to the B− line 39 can be connected to the base of transistor 43. Shunt diodes 66 and 67 are connected across the collector-emitter terminals of transistors 42 and 43, respectively. Finally, a capacitor 68 is connected across the collector-emitter terminals of transistor 43 to restrain the rate of voltage rise across those terminals, as will be seen presently.

The operation of the circuit of FIG. 2 can best be understood with additional reference to FIG. 3, which illustrates significant portions of the waveforms of the voltage at midpoint M (FIG. 3A), the base-emitter voltage on transistor 42 (FIG. 3B), the current through transistor 42 (FIG. 3C), and the capacitor 52 voltage and the inductor 51 current (FIG. 3D).

Assuming that transistor 42 is first to be triggered into conduction, current flows from the B+ line 38 through windings 46 and 38 and the inductor 51 to charge capacitor 52 and returns through capacitor 34 (refer to the time period designated I in FIG. 3). When the saturable inductor 49 saturates at the end of period I, drive current to the base of transistor 42 will terminate, causing voltage on the base of the transistor to drop to the negative voltage stored on the bias capacitor 56 in a manner to be described, causing this transistor to become non-conductive. As shown in FIG. 3c, current-flow in transistor 43 terminates at the end of period I.

Because the current through inductor 51 cannot change instantaneously, current will flow from the B− bus 39 through capacitor 68, causing the voltage at midpoint line 44 to drop to −160 volts (period II in FIG. 3). The capacitor 68 restrains the rate of voltage change across the collector and emitter terminals of transistor 42. The current through the inductor 51 reaches its maximum value when the voltage at the midpoint line 44 is zero. During period III, the current will continue to flow through inductor 51 but will be supplied from the B− bus through the shunt diode 67. It will be appreciated that during the latter half of period II and all of period III, positive current is being drawn from a negative voltage; which, in reality, means that energy is being returned to the power supply through a path of relatively low impedance.

When the inductor current reaches zero at the start of period IV, the current through the primary winding 46 of the saturable inductor 47 will cause a current to flow out of its secondary winding 54 to cause transistor 43 to become conductive, thereby causing a reversal in the direction of current through inductor 51 and capacitor 52. When transformer 47 saturates at the end of period IV, the drive current to the base of transistor 43 terminates and the current through inductor 51 will be supplied through capacitor 68, causing the voltage at midpoint line 44 to rise (period V). When the voltage at the midpoint line M reaches 160 volts, the current will then flow through shunt diode 66 (period VI). The cycle is then repeated.

As seen in FIG. 3, saturable transformers 47, 49 provide transistor drive current only after the current through inductor 51 has diminished to zero. Further, the transistor drive current is terminated before the current through inductor 51 has reached its maximum amplitude. This coordination of base drive current and inductor current is achieved because of the series-connection between the inductor 51 and the primary windings 46, 48 of saturable transformers 47, 49, respectively.

The series-connected combination of the inductor **51** and the capacitor **52** is energized upon the alternate conduction of transistors **42** and **43**. With a large value of capacitance of capacitor **52**, very little voltage will be developed across its terminals. As the value of this capacitance is decreased, however, the voltage across this capacitor will increase. As the value of the capacitor **52** is reduced to achieve resonance with the inductor **51**, the voltage on the capacitor will rise and become infinite in a loss-free circuit operating under ideal conditions.

It has been found desirable to regulate the transistor inversion frequency, determined mainly by the saturation time of the saturable inductors **47**, **49**, to be equel to or higher than the natural resonance frequency of the inductor and capacitor combination in order to provide a high voltage output to external load **26**. A high voltage across capacitor **52** is efficiently developed as the transistor inversion frequency approaches the natural resonant frequency of the inductor **51** and capacitor **52** combination. Stated another way, the conduction period of each transistor is desirably shorter in duration than one quarter of the full period corresponding to the natural resonant frequency of the inductor and capacitor combination. When the inverter **24** is used with a self-ballasted gas discharge lamp unit, it has been found that the inversion frequency can be at least equal to the natural resonant frequency of the tank circuit. If the capacitance value of capacitor **52** is reduced still further beyond the resonance point, unacceptably high transistor currents will be experienced during transistor switching and transistor burn-out will occur.

It will be appreciated that the sizing of capacitor **52** is determined by the application of the inverter circuit **24**. Variation in the values of the capacitor **52** and the inductor **51** will determine the voltages developed in the inductor-capacitor tank circuit. The external load **26** may be connected in circuit with the inductor **51** (by a winding on the inductor, for example) and the capacitor may be omitted entirely. If the combined circuit loading of the inductor **51** and the external load **26** has an effective inductance of value sufficient to effect periodic energy storage for self-sustained transistor inversion, the current feedback provided by the saturable inductors **47**,**49** will effect alternate transistor conduction without the need for additional voltage feedback. When the capacitor **52** is omitted, the power supply **23** provides a direct electrical connection between one of the AC power input lines and the inverter load circuit.

Because the voltages across transistors **42**, **43** are relatively low (due to the effect of capacitors **34**, **36**), the half-bridge inverter **24** is very reliable. The absence of switching transients minimizes the possibility of transistor burn-out.

The inverter circuit **24** comprises means for supplying reverse bias to the conducting transistor upon saturation of its associated saturable inductor. For this purpose, the capacitors **56** and **57** are charged to negative voltages as a result of reset current flowing into secondary windings **53**, **54** from the bases of transistors **42**, **43**, respectively. This reverse current rapidly turns off a conducting transistor to increase its switching speed and to achieve inverter circuit efficiency in a manner described more fully in my co-pending U.S. patent application Ser. No. 103,624 filed Dec. 14, 1979 and entitled "Bias Control for High Efficiency Inverter Circuit" (now U.S. Pat. No. 4,307,353). The more negative the voltage on the bias capacitors **56** and **57**, the more rapidly charges are swept out of the bases of their associated transistors upon transistor turn-off.

When a transistor base-emitter junction is reversely biased, it exhibits the characteristics of a Zener diode having

a reverse breakdown voltage on the order of 8 to 14 Volt for transistors typically used in high-voltage inverters. As an alternative, to provide a negative voltage smaller in magnitude on the base lead of typical transistor **42** during reset operation, the optional diode **61** and Zener diode **62** combination can be used. For large values of the bias capacitor **56**, the base voltage will be substantially constant.

If the load **26** comprises a gas discharge lamp, the voltage across the capacitor **52** will be reduced once the lamp is ignited to prevent voltages on the inductor **51** and the capacitor **52** from reaching destructive levels. Such a lamp provides an initial time delay during which a high voltage, suitable for instant starting, is available.

FIG. **5** illustrates the use of an alternate load **26'** adapted for plug-in connection to an inverter circuit such as shown in FIG. **2**. The load **26'** consists of a gas discharge lamp **71** having electrodes **72**, **73** and connected in series with a capacitor **74**. The combination of lamp **71** and capacitor **74** is connected in parallel with a capacitor **52'** which serves the same purpose as capacitor **52** in the FIG. 2 circuit. However, when the load **26'** is unplugged from the circuit, the inverter stops oscillating and the development of high voltages in the inverter is prevented. The fact that no high voltages are generated by the circuit if the lamp is disconnected while the circuit is oscillating is important for safety reasons.

FIG. **6** illustrates a capacitor **52"** connected in series with an inductor **51"** through a heater **81** suitable for heating the toroidal inductors **47**, **49** in accordance with the level of output. The load **26"** is connected across the series combination of the capacitor **52"** and the toroid heater. The heater **81** is preferably designed to controllably heat the toroidal saturable inductors in order to decrease their saturation flux limit and hence their saturation time. The result is to decrease the periodic transistor conduction time and thereby increase the transistor inversion frequency. When a frequency-dependent impedance means, that is, an inductor or a capacitor, is connected in circuit with the AC voltage output of the inverter, change in the transistor inversion frequency will modify the impedance of the frequency-dependent impdance means and correspondingly modify the inverter output. Thus as the level of the output increases, the toroid heater **81** is correspondingly energized to effect feedback regulation of the output. Further, transistors **42**, **43** of the type used in high voltage inverters dissipate heat during periodic transistor conduction. As an alternative, the toroid heater **81** can use this heat for feedback regulation of the output or control of the temperature of transistors **42**, **43**.

The frequency dependent impedance means may also be used in a circuit to energize a gas discharge lamp at adjustable illumination levels. Adjustment in the inversion frequency of transistors **42**, **43** results in control of the magnitude of the AC current supplied to the lamp. This is preferably accomplished where saturable inductors **47**, **49** have adjustable flux densities for control of their saturation time.

FIG. **7** schematically illustrates an alternate form of inverter circuit, shown without the AC to DC power supply connections for simplification. In this Figure, the transistors are connected in parallel rather than in series but the operation is essentially the same as previously described.

In particular, this circuit comprises a pair of alternately conducting transistors **91**, **92**. The emitter terminals of the transistors are connected to a B− line **93**. A B+ lead **94** is connected to the center-tap of a transformer **96**. In order to provide drive current to the transistors **91**, **92** for control of their conduction frequency, saturable inductors **97**, **98** have secondary windings **99**, **101**, respectively, each secondary winding having one end connected to the base of its associated transistor; the other ends are connected to a common terminal **102**. One end of transformer **96** is connected to the collector of transistor **91** through a winding **103** on inductor **98** in turn connected in series with a winding **104** on inductor **97**. Likewise, the other end of transformer **96** is connected to the collector of transistor **92** through a winding **106** on inductor **97** in series with another winding **107** on inductor **98**.

The B+ terminal is connected to terminal **102** through a bias resistor **108**. A bias capacitor **109** connects terminal **102** to the B− lead **93**. This resistor and capacitor serve the same function as resistors **58**, **59** and capacitors **56**, **57** in the FIG. **2** circuit.

The bases of transistors **91**, **92** are connected by diodes **111**, **112**, respectively, to a common Zener diode **113** in turn connected to the B− lead **93**. The common Zener diode **113** serves the same function as individual Zener diodes **62**, **64** in FIG. **2**.

Shunt diodes **114**, **116** are connected across the collector-emitter terminals of transistors **91**, **92**, respectively. A capacitor **117** connecting the collectors of transistors **91**, **92** restrains the rate of voltage rise on the collectors in a manner similar to the collector-emitter capacitor **68** in FIG. 2.

Inductive-capacitive loading of the FIG. **7** inverter is accomplished by a capacitor **118** connected in series with with an inductor **119**, the combination being connected across the collectors of the transistors **91**, **92**. A load **121** is connected across the capacitor **118**.

FIG. **8** illustrates how an inverter loaded with a series capacitor **122** and inductor **123** can be used to energize a "rapid-start" fluorescent lamp **124** (the details of the inverter circuit being omitted for simplication). The lamp **124** has a pair of cathodes **126**, **127** connected across the capacitor **122** for supply of operating voltage in a manner identical to that previously described. In addition, the inductor **123** comprises a pair of magnetically-coupled auxiliary windings **128**, **129** for electrically heating the cathodes **126**, **127**, respectively. A small capacitor **131** is connected in series with lamp **124**.

FIG. **9** illustrates the very same circuit arrangement as that of FIG. **8** except that the auxiliary windings **128**, **129** are only loosely coupled to the inductor **123**, thereby providing for a manifest limitation on the amount of current that can be drawn from each auxiliary winding in case it were to be accidentally short-circuited.

FIG. **10** also illustrates the very same circuit arrangement as that of FIG. **8** except that the cathodes **126**, **127** are connected with their respective auxiliary windings **128**, **129** by way of non-linear current-limiting means **132** and **133**, respectively.

In FIG. **10**, the non-linear current-limiting means **132**, **133** are shown as being two (small) incandescent lamps. However, other types of non-linear resistance means could be used as well.

Both the FIG. **9** circuit and the FIG. **10** circuit serve the same basic purpose; which is that of preventing damage to the ballast circuit (such as that if FIG. **2**) in case the leads used for connecting to one of the lamp cathodes **126**, **127** were to be accidentally shorted. This damage prevention is accomplished by providing for manifest limitation of the maximum amount of current that can be drawn from each one of the auxiliary windings **128**, **129**. In the circuit of FIG. **9**, this manifest limitation is accomplished by having the auxiliary windings **128**, **129** couple sufficiently loosely to the main inductor **123**—such as by providing a magnetic shunt between inductor **123** and the auxiliary windings—thereby correspondingly limiting the degree of impact resulting from an accidental short circuit. Such a short circuit would result in a net reduction in the effective inductance value of the tank inductor **123**; which net reduction in inductance may in turn cause a precipitous increase in the magnitude of the reactive current drawn from the inverter by the series-connected inductor **123** and capacitor **122**, thereby causing damage to the inverter.

ADDITIONAL EXPLANATIONS AND COMMENTS

(a) With reference to FIGS. 2 and 5, adjustment of the amount of power supplied to load **26**', and thereby the amount of light provided by lamp **71**, may be accomplished by applying a voltage of adjustable magnitude to input terminals IP1 and IP2 of the Toroid Heater; which is thermally coupled with the toroidal ferrite cores of saturable transformers **47**, **49**.

(b) With commonly available components, inverter circuit **24** of FIG. **2** can be made to operate efficiently at any frequency between a few kHz to perhaps as high as 50 kHz. However, for various well-known reasons (i.e., eliminating audible noise, minimizing physical size, and maximizing efficiency), the frequency actually chosen is in the range of 20 to 40 kHz.

(c) The fluorescent lighting unit of FIG. 1 could be made in such manner as to permit fluorescent lamp **11** to be disconnectable from its base **12** and ballasting means **16**. However, if powered with normal line voltage without its lamp load connected, frequency-converting power supply and ballasting circuit **16** is apt to self-destruct.

To avoid such self-destruction, arrangements can readily be made whereby the very act of removing the load automatically establishes a situation that prevents the possible destruction of the power supply and ballasting means. For instance, with the tank capacitor (**52**) being permanently connected with the lamp load (**11**)—thereby automatically being removed whenever the lamp is removed—the inverter circuit is protected from self-destruction.

(d) At frequencies above a few kHz, the load represented by a fluorescent lamp—once it is ignited—is substantially resistive. Thus, with the voltage across lamp **11** being of a substantially sinusoidal waveform (as indicated in FIG. 3*d*), the current through the lamp will also be substantially sinusoidal in waveshape.

(e) In the fluorescent lamp unit of FIG. **1**, fluorescent lamp **11** is connected with power supply and ballasting circuit **16** in the exact same manner as is load **26** connected with the circuit of FIG. 2. That is, it is connected in parallel with the tank capacitor (**52**) of the L-C series-resonant circuit. As is conventional in instant-start fluorescent lamps—such as lamp **11** of FIG. **1**—the two terminals from each cathode are shorted together, thereby to constitute a situation where each cathode effectively is represented by only a single terminal. However, it is not necessary that the two terminals from each cathode be shorted together; in which case—for instant-start operation—connection from a lamp's power supply and ballasting means need only be made with one of the terminals of each cathode.

(f) With respect to the circuit arrangement of FIG. 9, in situations where the tank inductor **123** includes a ferrite magnetic core having an air gap, one particularly cost-effective way of accomplishing the indicated loose coupling between the tank inductor **123** and the auxiliary windings

128, 129 is that of arranging for the auxiliary windings to be placed in the air gap in such a manner that they each couple only with part of the magnetic flux crossing the air gap.

(i) In FIG. 1, as would be readily perceived by a person having ordinary skill in the art pertinent hereto, folded fluorescent lamp 11 includes two straight parallel-disposed tubular sections rising perpendicularly up from the top surface of base 12; which two tubular sections are, a short distance distally from base 12, joined together by way of a transverse relatively narrow tubular section.

What is claimed is:

1. An arrangement comprising:

a power source operative to supply a power line voltage at a pair of power line terminals;

a first circuit assembly connected with the power line terminals and operative to provide a DC voltage between a pair of DC output terminals;

a gas discharge lamp having a pair of lamp terminals and being functional to provide luminous output when supplied with a usual amount of lamp current; and

a second circuit assembly having a pair of DC input terminals connected with the DC output terminals and a pair of AC output terminals connected with the lamp terminals; the second circuit assembly being further characterized by:

(a) having a tuned L-C circuit connected with the AC output terminals; the L-C circuit having a tank-inductor and a tank-capacitor; the tank-capacitor being effectively connected across the AC output terminals; and

(b) producing an AC output voltage across the AC output terminals; the magnitude of the AC output voltage being determined by: (i) a Q-multiplying effect associated with the L-C circuit; (ii) the amount of power drawn by the lamp from the AC output terminals; and (iii) an internal feedback effect responsive to the magnitude of the AC output voltage and operative to diminish the Q-multiplying effect by causing the frequency of the AC output voltage to change away from the natural resonance frequency of the L-C circuit; such that the internal feedback effect is functional, under a condition when no power is being drawn from the AC output terminals, to cause the magnitude of the AC output voltage to be lower than it would have been if determined solely by said Q-multiplying effect;

such that the magnitude of the AC output voltage is: (i) at a minimum level whenever the lamp is drawing its usual lamp current; and (ii) at a maximum level whenever the lamp fails to draw power, the maximum level being distinctly lower than a level which would have prevailed in the absence of said internal feedback effect.

2. The arrangement of claim 1 wherein the second circuit assembly is additionally characterized in that the fundamental frequency of the AC output voltage is lower whenever the lamp draws the usual amount of lamp current than when the lamp draws substantially less than the usual amount of lamp current.

3. An arrangement comprising:

a power source operative to supply a power line voltage at a pair of power line terminals;

a gas discharge lamp having a pair of lamp terminals and being functional to provide luminous output when supplied with an amount of lamp current; and

a main assembly having (i) a pair of power input terminals connected with the power line terminals, and (ii) a pair of AC output terminals connected with the lamp terminals; the main assembly including electronic circuitry and being otherwise characterized by:

(a) having: (i) a tuned L-C circuit connected with the AC output terminals, the L-C circuit having a tank-inductor and a tank-capacitor, the tank-capacitor being effectively connected across the AC output terminals; and (ii) a magnitude-limiting sub-assembly connected within the electronic circuitry as well as with the AC output terminals; and

(b) being functional to cause an AC output voltage to exist across the AC output terminals; the magnitude of the AC output voltage being determined by a combination of: (i) a Q-multiplying effect associated with the L-C circuit; (ii) the frequency of the AC output voltage; (iii) the amount of power drawn by the lamp from the AC output terminals; and (iv) a magnitude-limiting effect provided by the magnitude-limiting sub-assembly in response to the magnitude of the AC output voltage; the magnitude-limiting effect, by which an increasing magnitude of the AC output voltage causes a change in the frequency of the AC output voltage away from the natural resonance frequency of the L-C circuit, resulting in a diminishment of the Q-multiplying effect in such a manner as to cause the magnitude of the AC output voltage, whenever no power is being drawn by the lamp, to be distinctly lower than it would have been in the absence of said magnitude-limiting effect.

4. The arrangement of claim 3 wherein the main assembly is additionally characterized in that the tuned L-C circuit has a natural resonance frequency that is not higher than the fundamental frequency of the AC output voltage.

5. The arrangement of claim 3 wherein the main assembly is additionally characterized by including a pair of terminals between which a DC voltage exists, the absolute magnitude of which DC voltage is substantially constant as well as distinctly higher than the peak absolute magnitude of the power line voltage.

6. The arrangement of claim 3 wherein the main assembly is additionally characterized by including a pair of terminals between which two transistors are series-connected.

7. The arrangement of claim 3 wherein the gas discharge lamp includes two parallel-disposed straight sections of glass tubing, and wherein the main assembly is additionally characterized by including a housing having a screw-base operable to be screwed into and held by an ordinary Edison-type lamp socket.

8. The arrangement of claim 3 further combined with a mechanical structure operative to rigidly hold together the gas discharge lamp and the main assembly, thereby to form a single lamp-ballast entity operable to be powered from a power line voltage; the gas discharge lamp being characterized by including: (i) two parallel-disposed relatively long straight sections of tubing, each one having a first cross-sectional area and protruding perpendicularly from a substantially flat surface of said mechanical structure; and (ii) a relatively short section of tubing having a second cross-sectional area that is substantially smaller than said first cross-sectional area.

9. The arrangement of claim 8 wherein the lamp-ballast entity includes a screw-base operable to be screwed into and held by an ordinary Edison-type lamp socket.

10. An arrangement comprising:

a power source operative to supply a power line voltage at a pair of power line terminals;

a gas discharge lamp having a pair of lamp terminals and being functional to provide luminous output when supplied with lamp current; and

a main assembly having (i) a pair of power input terminals connected with the power line terminals, and (ii) a pair of AC output terminals connected with the lamp terminals; the main assembly including electronic circuitry and being otherwise characterized by:

(a) having: (i) a tuned L-C circuit connected with the AC output terminals; and (ii) a magnitude-limiting sub-assembly connected within the electronic circuitry as well as with the AC output terminals; and

(b) being functional to cause an AC output voltage to exist across the AC output terminals; the magnitude of the AC output voltage being determined by a combination of: (i) a Q-multiplying effect associated with the L-C circuit; (ii) the amount of power drawn by the lamp from the AC output terminals; and (iii) a magnitude-limiting effect provided by the magnitude-limiting sub-assembly in response to the magnitude of the AC output voltage; the magnitude-limiting effect, by which an increasing magnitude of the AC output voltage causes a change in the frequency of the AC output voltage away from the natural resonance frequency of the L-C circuit, resulting in a diminishment of the Q-multiplying effect in such a manner as to cause the magnitude of the AC output voltage, whenever no power is being drawn by the lamp, to be distinctly lower than it would have been in the absence of said magnitude-limiting effect.

11. An arrangement comprising:

a power source operative to supply a power line voltage at a pair of power line terminals;

a lamp means having lamp terminals and being operative to provide light when drawing a lamp current; and

a main assembly characterized by: (i) having power input terminals connected with the power line terminals; (ii) having AC output terminals connected with the lamp terminals; (iii) being operative to provide an AC output voltage at the AC output terminals, the frequency of the AC output voltage being higher than that of the power line voltage; (iv) having a sub-assembly connected within the main assembly as well as with the AC output terminals, the sub-assembly being operative under a certain condition, by causing a change in the frequency of the AC output voltage, to limit the magnitude of the AC output voltage to a level lower than what it would have been if the sub-assembly had not been so connected, said certain condition occurring if the lamp means were to fail to draw said lamp current; such that if the sub-assembly were to be disconnected, and if the lamp means were to fail to draw said lamp current, the magnitude of the AC output voltage would attain a level substantially higher than the level attained with the sub-assembly connected; and (v) having a pair of terminals between which two transistors are series-connected and across which exists a DC supply voltage.

12. An arrangement comprising:

a power source operative to supply a power line voltage at a pair of power line terminals;

a lamp having lamp terminals and being operative to provide light when being supplied with a lamp current; and

a main assembly characterized by: (i) having power input terminals connected with the power line terminals; (ii) having AC output terminals connected with the lamp terminals, thereby to cause said lamp current to be supplied to the lamp; (iii) being operative to provide an AC output voltage at the AC output terminals, the frequency of the AC output voltage, which is much higher than that of the power line voltage, increasing with increasing magnitude of the AC output voltage; (iv) having an L-C circuit connected with the AC output terminals, the L-C circuit having a natural resonance frequency not higher than the fundamental frequency of the AC output voltage; (v) having a pair of transistors series-connected between two terminals across which is provided a DC voltage whose absolute magnitude always is equal to or higher than that of the power line voltage; and (vi) having a sub-assembly connected within the main assembly as well as with the AC output terminals, the sub-assembly being operative under a certain condition to limit the magnitude of the AC output voltage to a level lower than what it would have been if the sub-assembly had not been so connected, said certain condition occurring if the lamp were to fail to draw said lamp current; such that, if the sub-assembly were to be disconnected, the magnitude of the AC output voltage would reach an undesirably high level in case the lamp were to be disconnected or otherwise fail to draw said lamp current.

13. An arrangement comprising:

a power source operative to supply a power line voltage at a pair of power line terminals;

a lamp having lamp terminals and being operative to provide light when drawing a lamp current; and

a main assembly characterized by: (i) having power input terminals connected with the power line terminals; (ii) having AC output terminals connected with the lamp terminals; (iii) providing an AC output voltage at the AC output terminals, the AC output voltage having a frequency that changes in response to the magnitude of the AC output voltage; (iv) having two terminals between which a pair of transistors are series-connected and across which exists a DC voltage whose absolute magnitude is about equal to or higher than that of the power line voltage; and (v) having a sub-assembly connected within the main assembly as well as with the AC output terminals, the sub-assembly being operative under a certain condition to limit the magnitude of the AC output voltage to a level lower than what it would have been if the sub-assembly had not been so connected, said certain condition occurring if the lamp were to fail to draw said lamp current; such that if the sub-assembly were to be disconnected, and if the lamp were to fail to draw said lamp current, the magnitude of the AC output voltage would attain a level substantially higher than the level attained with the sub-assembly connected.

14. The arrangement of claim 13 wherein the sub-assembly is further characterized in that it is operative under said certain condition to cause the frequency to increase.

15. An arrangement comprising:

a power source operative to supply a power line voltage at a pair of power line terminals;

a lamp having lamp terminals and being operative to provide light when drawing a lamp current; and

a main assembly characterized by: (i) having power input terminals connected with the power line terminals; (ii) having AC output terminals connected with the lamp terminals; (iii) providing an AC output voltage at the AC output terminals; (iv) having two terminals between which a pair of alternatingly switching transistors are series-connected and across which is provided a DC voltage whose absolute magnitude is always distinctly higher than that of the power line voltage; and (v) having a sub-assembly connected within the main assembly as well as with the AC output terminals, the sub-assembly being operative under a certain condition to cause the frequency of the AC output voltage to change, thereby to limit the magnitude of the AC output voltage to a level lower than what it would have been if the sub-assembly had not been so connected, said certain condition occurring if the lamp means were to fail to draw said lamp current; such that if the sub-assembly were to be disconnected, and if the lamp means were to fail to draw said lamp current, the magnitude of the AC output voltage would attain a level substantially higher than the level attained with the sub-assembly connected.

16. The arrangement of claim **15** wherein the main assembly is additionally characterized in that said DC voltage is of substantially constant magnitude.

17. An arrangement comprising:

a power source operative to supply an AC power line voltage at a pair of power line terminals; and

a combination of:

(A) a gas discharge lamp having a pair of lamp terminals and being functional to provide luminous output when supplied with an amount of lamp current; the gas discharge lamp being further characterized by including: (i) two parallel-disposed relatively long straight sections of tubing, each one having a first cross-sectional area; and (ii) a relatively short section of tubing having a second cross-sectional area that is substantially smaller than said first cross-sectional area;

(B) a main assembly having: (i) a pair of power input terminals adapted to connect with a pair of terminals across which exists said AC power line voltage; and (ii) a pair of AC output terminals connected with the lamp terminals; the main assembly including electronic circuitry and being otherwise characterized by:

(1) having a tuned L-C circuit connected with the AC output terminals, the L-C circuit having a tank-inductor and a tank-capacitor, the tank-capacitor being effectively connected across the AC output terminals; and

(2) whenever an ordinary AC power line voltage is present across the power input terminals, being operative to cause an AC output voltage to exist across the AC output terminals; and

(C) a mechanical structure having a screw-base operable to be screwed into and held by an ordinary Edison-type lamp socket; the mechanical structure also being functional to rigidly hold together the gas discharge lamp and the main assembly, thereby to form a single lamp-ballast entity operable to be screwed into and powered from the AC power line voltage usually provided at an ordinary Edison-type lamp socket.

* * * * *

EXHIBIT N

US006472827B1

(12) **United States Patent**
Nilssen

(10) **Patent No.: US 6,472,827 B1**
(45) **Date of Patent: *Oct. 29, 2002**

(54) **PARALLEL-RESONANT INVERTER-TYPE FLUORESCENT LAMP BALLAST**

(76) Inventor: **Ole K. Nilssen**, 602 Caesar Dr., Rte. 5, Barrington, IL (US) 60010

(*) Notice: This patent issued on a continued prosecution application filed under 37 CFR 1.53(d), and is subject to the twenty year patent term provisions of 35 U.S.C. 154(a)(2).

Subject to any disclaimer, the term of this patent is extended or adjusted under 35 U.S.C. 154(b) by 0 days.

(21) Appl. No.: **08/032,569**

(22) Filed: **Mar. 18, 1993**

Related U.S. Application Data

(63) Continuation of application No. 07/719,671, filed on Jun. 24, 1991, now abandoned, which is a continuation of application No. 07/346,292, filed on May 1, 1989, now abandoned, which is a continuation of application No. 06/697,949, filed on Feb. 4, 1985, now abandoned, which is a continuation-in-part of application No. 06/658,423, filed on Oct. 5, 1984, now abandoned.

(51) **Int. Cl.**[7] **H05B 37/01**

(52) **U.S. Cl.** .... **315/209 R**; 315/219; 315/222; 315/224; 315/279; 315/307; 315/DIG. 7; 363/19; 363/23; 363/133; 323/249; 323/302

(58) **Field of Search** .... 315/209 R, 219, 315/222, 224, 279, 307, DIG. 7; 363/19, 23, 133; 323/249, 302

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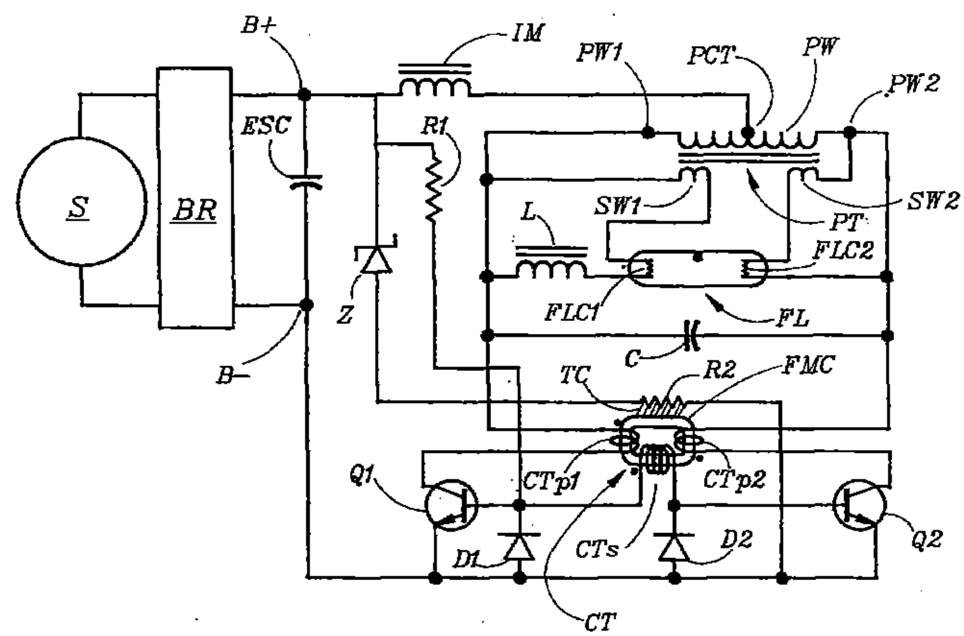
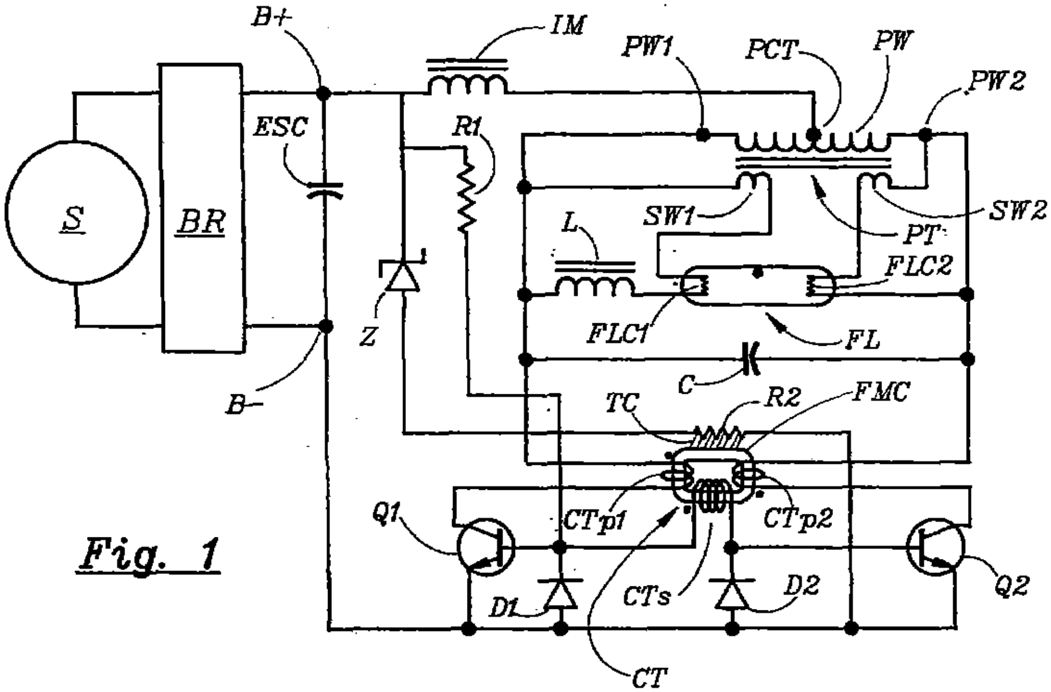
Primary Examiner—Michael B Shingleton

(57) **ABSTRACT**

A push-pull inverter is supplied from an inductively current-limited DC voltage source by way of a center-tap on a transformer having significant inductance. This transformer inductance is parallel-coupled with a capacitance means. The inverter is made to self-oscillate through positive feedback provided by way of a saturable current transformer. The inverter frequency is determined by the saturation time of this current transformer, which saturation time is designed to be somewhat longer than the half-cycle period of the natural resonance frequency of the transformer inductance combined with the capacitance means. By controlling the length of this saturation time, the magnitude of the current provided to the fluorescent lamp is controlled, thereby permitting control of the light output in response to changes in the magnitude of the power line voltage.

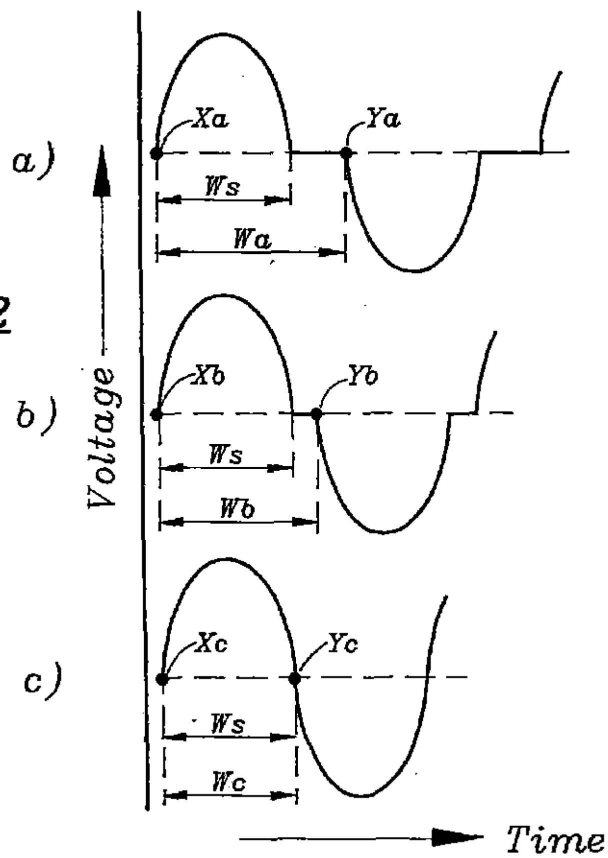
26 Claims, 1 Drawing Sheet





B+
IM
PW1
PCT
PW
PW2
ESC
R1
S
BR
SW1
SW2
PT
L
FLC2
FLC1
FL
Z
B−
C
TC
R2
FMC
Q1
CTp1
CTp2
CTs
D1
D2
Q2
CT

Fig. 1



a)
Xa
Ya
Ws
Wa
b)
Xb
Yb
Ws
Wb
c)
Xc
Yc
Ws
Wc
Voltage
Time

Fig. 2

PARALLEL-RESONANT INVERTER-TYPE FLUORESCENT LAMP BALLAST

RELATED APPLICATIONS

The present application is a continuation of application Ser. No. 07/719,671 filed Jun. 24, 1991, now abandoned which is a continuation of application Ser. No. 07/346,292 filed May 1, 1989 now abandoned, which a continuation of application Ser. No. 06/697,949 filed Feb. 2, 1985, now abandoned; which was a Continuation-in-Part of Ser. No. 06/658,423 filed Oct. 4, 1984 now abandoned.

BACKGROUND OF THE INVENTION

1. Field of Invention

The present invention relates to power-line-operated inverter-type fluorescent lamp ballasts, particularly of the kind using a current-excited parallel-loaded resonant L-C circuit for matching the inverter output to the lamp load.

2. Prior Art

Power-line-operated inverter-type fluorescent lamp ballasts using a current-excited parallel-loaded resonant (or near resonant) L-C circuit for matching the inverter output to the lamp load are well known and widely used. An example of such a ballast is described in U.S. Pat. No. 4,277,726 to Burke.

One significant problem generally associated with these ballasts relates to their relatively poor ability to regulate lamp light output in response to variations in the magnitude of the power line voltage.

SUMMARY OF THE INVENTION

Objects of the Invention

An object of the present invention is that of providing cost-effective means for controlling and/or regulating the light output associated with inverter-type fluorescent lamp ballasts.

This as well as other important objects and advantages of the present invention will become apparent from the following description.

Brief Description

A push-pull inverter is supplied from an inductively current-limited DC voltage source by way of a center-tap on a transformer having significant inductance. This transformer inductance is parallel-coupled to a capacitance means. A fluorescent lamp is series-connected with a current-limiting inductor, and this lamp-inductor series-combination is connected in parallel with the capacitance means.

The inverter is made to self-oscillate through positive feedback provided by way of a saturable current transformer. The inverter frequency is determined by the saturation time of this current transformer, which saturation time is designed to be somewhat longer than the half-cycle period of the natural resonance frequency of the transformer inductance means combined with the parallel-coupled capacitance means as well as the parallel-coupled lamp-inductor series-combination.

The saturable current transformer comprises a ferrite magnetic core, and the length of the saturation time of this transformer is determined by the magnitude of the magnetic saturation flux of this ferrite core.

The magnitude of the magnetic saturation flux is determined by the temperature of the ferrite magnetic core: the higher the temperature, the lower the magnitude of the magnetic saturation flux. By heating the ferrite core as a function of the magnitude of the DC voltage supplying power to the inverter, the inverter frequency is made to vary in inverse relationship with the magnitude of this DC voltage. As an overall result, lamp light output is kept substantially constant in spite of substantial variations in the magnitude of the DC supply voltage.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 provides a schematic circuit diagram of the preferred embodiment of the invention.

FIG. 2 provides illustration of various voltage waveforms associated with the embodiment of FIG. 1.

DESCRIPTION OF THE PREFERRED EMBODIMENT

Details of Construction

FIG. 1 shows an AC voltage source S, which in reality is an ordinary 120 Volt/60 Hz electric utility power line. Connected directly across S is a bridge rectifier BR, the DC output from which is applied to a B+ terminal and a B− terminal—with the B+ terminal being of positive polarity in respect to the B− terminal. Connected directly between the B+ and B− terminals is an energy-storing capacitor ESC.

Connected between the B+ terminal and a center-tap PCT of the primary winding PW of a power transformer PT is an inductor means IM. Primary winding PW has two terminals PW1 and PW2 in addition to its center-tap PCT. Across terminals PW1 and PW2 is connected a capacitor C, as well as a series-combination of a fluorescent lamp FL and an inductor L.

Power transformer PT has two secondary windings SW1 and SW2 connected respectively with cathodes FLC1 and FLC2 of fluorescent lamp FL.

The collector of a first transistor Q1 is connected to terminal PW1 by way of a first primary winding CTp1 of a saturable feedback current transformer CT; and the collector of a second transistor Q2 is connected to terminal PW2 by way of a second primary winding CTp2 of current transformer CT.

Secondary winding CTs of current transformer CT is connected directly between the bases of transistors Q1 and Q2. Current transformer CT has a ferrite magnetic core FMC.

A diode D1 is connected between the base and the emitter of transistor Q1, with the diode's cathode being connected to the base. Similarly, a diode D2 is connected between the base and the emitter of transistor Q2, with the diode's cathode being connected with the base. The emitters of transistors Q1 and Q2 are both connected with the B− terminal. Connected between the B+ terminal and the base of transistor Q1 is a resistor R1.

A Zener diode Z is connected with its cathode to the B+ terminal and with its anode to one terminal of a resistor R2. The other terminal of resistor R2 is connected with the B− terminal.

Resistor R2 is placed in close proximity with the ferrite magnetic core FMC of current transformer CT; and a thermal conduction means TC is placed between resistor R2 and this ferrite magnetic core FMC.

Details of Operation

In FIG. 1, the 120 Volt/60 Hz power line voltage is rectified by rectifier BR and, due to the filtering effect of capacitor ESC, provides a DC voltage of substantially constant magnitude between the B+ and the B− terminals. At low line voltage, the magnitude of this DC voltage is about 144 Volt; at normal line voltage it is 160 Volt; at high line voltage, it is 176 Volt.

The Zenering voltage of Zener diode Z is about 144 Volt; which implies that substantially no current flows through resistor R2 at low line voltage. At high line voltage, on the other hand, current-flow through the Zener diode is only limited by the resistance of resistor R2, which at this point sees a voltage of about 32 Volt magnitude. The value of resistor R2 is chosen such as to provide a certain amount of heating of the ferrite magnetic core of current transformer CT.

The purpose of providing heat to the ferrite magnetic core is that of controlling the oscillating frequency of the inverter, which oscillating frequency is principally determined by the saturation time associated with the saturable feedback current transformer CT. This saturation time is principally determined by the saturation magnetic flux density of the ferrite magnetic core FMC; and the saturation flux density, in turn, is a function of the temperature of the ferrite magnetic core: the higher the core temperature, the lower the magnitude of the saturation flux density.

Otherwise, the operation of the inverter circuit itself, which consists of inductor means IM, power transformer PT, saturable feedback current transformer CT, transistors Q1 and Q2, diodes D1 and D2, and biasing resistor R1, may be understood as a very special combination of the basic inverter circuits described in U.S. Pat. No. 4,277,726 to Burke and in U.S. Pat. No. 4,279,011 to Nilssen. By combining features of these two types of inverter circuits, useful effects may be obtained.

In the basic inverter circuit described by Burke, inverter oscillating frequency is determined by the inductance of power transformer PT as naturally interacting with the net capacitive reactance parallel-connected therewith (i.e., predominantly C).

On the other hand, in the basic inverter circuit described by Nilssen, inverter oscillating frequency is determined by the saturation time of the saturable feedback current transformer.

Thus, in the circuit of FIG. 1, inverter oscillating frequency is principally determined by the saturation time of saturable feedback current transformer PCT. However, for this to be the case, it is necessary that this saturation time be longer than the half-period of the natural resonance frequency associated with the inductance of the power transformer as interacting with the net capacitive reactance represented by capacitor C and the effect of the load circuit (which consists of inductor L in series-connection with the fluorescent lamp).

By way of various voltage waveforms, FIG. 2 effectively illustrates the operation of the inverter circuit of FIG. 1.

FIG. 2*a* shows the intermittently sinusoidal waveform of the voltage present across power transformer PT as observed between its center-tap PCT and terminal PW1 for the situation of low line voltage; in which situation the magnitude of the DC supply voltage is 144 Volt. Thus, in this situation, the average magnitude of the voltage existing during a complete half-cycle (i.e., between points Xa and Ya in FIG. 2*a*) must by basic necessity be 144 Volt.

By choice of saturation time of current transformer CT, and with no heating provided by resistor R2, the width Wa of this complete half-cycle is made to be about 22% wider than the width Ws of the base of the sinusoidal half-cycle existing between points Xa and Ya. As a necessary consequence of this fact, the amplitude of the sinusoidal half-cycle per se must be about 22% larger than would have been the case if the width Ws of this sinusoidal half-cycle had been equal to that of the complete inverter half-cycle Wa—the reason being that the average magnitude of the complete inverter half cycle must in this case by basic necessity be equal to 144 Volt.

In FIG. 2*a*, it is noted that the sinusoidal half-cycle located between points Xa and Ya is a half-cycle of the natural resonance action between the inductance of power transformer PT and the combination of capacitor C and the load circuit consisting of inductor L connected in series with the fluorescent lamp FL. Thus, the degree that the distance between points Xa and Ya is longer than the base of the sinusoidal half-cycle corresponds to the degree that the inverter's oscillating frequency is lower than the natural resonance frequency of the inductance of power transformer PT as interacting with the combination of capacitor C and the load circuit.

FIG. 2*b* shows the corresponding voltage waveforms existing under the condition of normal line voltage, in which case the magnitude of the DC supply voltage is 160 Volt. In this case, due to heating by resistor R2 of the ferrite magnetic core of transformer CT, the saturation time of this saturable transformer has been shortened to the point where the width Wb of the complete inverter half-cycle is only about 11% wider than the width Ws of the sinusoidal half-cycle. In this case, the average magnitude of the voltage existing between points Xb and Yb must by necessity be equal to 160 Volt; which implies that the magnitude of the sinusoidal half-cycle per se must be about 11% larger than it would have been if Ws had been equal to Wb.

FIG. 2*c* shows the corresponding voltage waveforms existing under the condition of maximum line voltage, in which case the magnitude of the DC supply voltage is 176 Volt. In this case, due to further heating by resistor R2 of the ferrite magnetic core of transformer CT, the saturation time has been been shortened to the point where the width of Wc of the complete inverter half-cycle is about equal to the width Ws of the sinusoidal half-cycle. In this case, the average magnitude of the voltage existing between points Xc and Yc must by basic necessity be equal to 176 Volt. Also, at this point, the inverter oscillating frequency equals the natural resonance frequency of the inductance of power transformer PT as interacting with the combination of capacitor C and the load circuit.

Comparing FIGS. 2*a*, 2*b* and 2*c*, it is seen that the magnitude of the sinusoidal half-cycles remains roughly constant even though the magnitude of the DC supply voltage varies over a range of +/−10%. The inverter frequency, on the other hand, changed substantially in inverse proportion to the magnitude of the DC supply voltage. With an inductor as the lamp current-limiting means, the overall implication is such that the power provided to the lamp remains substantially constant as the magnitude of the DC supply voltage changes by +/−10%.

The degree of controlling effect achieved depends on the degree of heating provided by resistor R2 to the magnetic ferrite core of transformer CT: the lower the resistance of R2, the better the thermal coupling between R2 and the magnetic ferrite core MFC, the smaller the physical size of transformer CT, the lower the heat losses from MFC, etc., the larger the degree of control. It is readily possible to provide for a change in the saturation time of some +/−20% as a result of the +/−10% change in the magnitude of the DC supply voltage, in which case the result would be over-regulation.

It is readily possible to provide for an additional and/or separate means for controlling the amount of heating provided to the ferrite core. For instance, a variable resistor means could be used to provide a selectable initial temperature bias; which, inter alia, could be used for control of light output irrespective of the magnitude of the DC supply voltage.

Or, the input from a light control means, such as a photo-cell, could be used for automatically affecting the core temperature and thereby to correspondingly control the amount of light provided by the lamp.

If, instead of an inductor, a capacitor were to be used as a lamp current-limiting means, it would still be possible to achieve light output regulation by way of controlling the saturation flux limits of the magnetic ferrite core of the saturable feedback transformer. However, in this case it would be necessary to heat the magnetic ferrite material in inverse relationship with the magnitude of the DC supply voltage.

By making R2 non-linear (such as, for instance, combining it with a Zener diode and/or a thermistor), it is possible to provide an improved regulation profile, thereby attaining more accurate regulation of light output as function of line voltage magnitude variations.

The voltage waveforms illustrated by FIG. 2, such as the one shown by FIG. *2a*, correspond to the voltage provided to the series-combination of inductor L and fluorescent lamp FL. These waveforms may be described as sinusoidal half-cycles interconnected with periods of substantially zero voltage.

Also, by selectively modifying the functional relationship between the magnitude of the DC supply voltage and the amount of heat provided to the ferrite magnetic core, it is possible to arrange for the net inverter output voltage (as illustrated by FIG. 2) to behave in a wide variety of different ways: it would readily be possible to make the RMS magnitude remain constant as the magnitude of the DC supply voltage changes; or, to have the power provided to the load circuit (i.e., to the lamp) remain constant as the magnitude of the DC supply voltage changes; or, to have the magnitude of the fundamental frequency-component of the waveforms in FIG. 2 decrease in direct proportion with the basic repetition frequency of the waveforms (i.e., the basic inversion frequency); or, conversely, to make this repetition frequency decrease in direct proportion to the magnitude of this fundamental frequency component; etc.

In respect to the effect of temperature on the magnitude of the magnetic saturation flux of ferrite magnetic cores, reference is made to various handbooks and product catalogs relating to magnetic ferrites for inverter applications.

It is believed that the present invention and its several attendant advantages and features will be understood from the preceeding description. However, without departing from the spirit of the invention, changes may be made in its form and in the construction and interrelationships of its component parts, the form herein presented merely representing the presently preferred embodiment.

What is claimed is:

1. An arrangement comprising:

an ordinary electric utility AC power line operable to provide an AC power line voltage at a pair of AC power line terminals;

a voltage conditioning circuit having a set of AC input terminals drawing an AC input current and being operative to provide a DC supply voltage at a pair of DC supply terminals; the magnitude of the DC supply voltage being substantially constant;

an inverter circuit having a pair of DC input terminals and a pair of AC output terminals; the inverter circuit being characterized in that its pair of DC input terminals draws a DC current and its AC output terminals supplies an AC output voltage; the AC output voltage being of a frequency substantially higher than that of the AC power line voltage and having a waveform consisting of sinusoidally-shaped half-cycles of voltage of alternating polarity; the waveform having a repetition period and being substantially sinusoidal except for a distinct discontinuity occurring over a brief period during each repetition period at or near each point-in-time where the instantaneous magnitude of the AC output voltage changes polarity; said brief period being substantially shorter than half of the duration of one of said half-cycles; the inverter circuit being further characterized by including two periodically conducting transistors, one of which is always conducting;

a gas discharge lamp;

the equivalent of an inductive impedance; and

the equivalent of a reactive impedance;

the arrangement being inter-connected such that:

(1) the AC input terminals are connected with the AC power line terminals;

(2) the DC input terminals are connected with the DC output terminals by way of said equivalent of an inductive impedance; and

(3) the lamp and said equivalent of a reactive impedance are series-connected across the AC output terminals.

2. The arrangement of claim **1** further characterized in that the magnitude of the AC output voltage is distinctly lower during said brief period, as averaged over said brief period, than it would have been had said waveform been smoothly sinusoidal.

3. The arrangement of claim **1** wherein the instantaneous magnitude of the AC output voltage is at or near zero during said brief part.

4. The arrangement of claim **1** wherein the inverter circuit is further characterized by including the equivalent of an inductive reactance parallel-connected with the equivalent of a capacitive reactance, thereby to form a parallel-resonant circuit having a natural resonance frequency distinctly different from that of the AC output voltage.

5. The arrangement of claim **1** further characterized by having the equivalent of a control circuit operative to control the duration of said brief part.

6. The arrangement of claim **1** further characterized by not including a capacitor series-connected with an inductor so as to form a near resonant circuit.

7. An arrangement comprising:

an ordinary electric utility AC power line operable to provide an AC power line voltage at a pair of AC power line terminals;

a voltage conditioning circuit having a set of AC input terminals drawing an AC input current and being operative to provide a DC supply voltage at a pair of DC supply terminals; the magnitude of the DC voltage being substantially constant;

an inverter circuit having a pair of DC input terminals and a pair of AC output terminals; the inverter circuit being characterized in that its pair of DC input terminals draws a DC current and its AC output terminals supplies an AC output voltage; the AC output voltage having a waveform and being of a frequency substantially higher than that of the AC power line voltage; the waveform having a repetition period and being substantially sinusoidal except for a distinct discontinuity occurring over a brief period during each repetition period at or near each point-in-time where the instantaneous magnitude of the AC output voltage changes polarity; said brief period being substantially shorter than one quarter of said repetition period; the inverter circuit being further characterized by including two periodically conducting transistors, neither one of which is non-conducting at the same time;

a gas discharge lamp having a pair of lamp terminals;

the equivalent of an inductive impedance having inductor terminals; and

the equivalent of a reactive impedance having reactor terminals;

the arrangement being structured such that:

(1) the AC input terminals are connected with the AC power line terminals,

(2) the DC input terminals are connected with the DC output terminals by way of the inductor terminals; and

(3) the lamp terminals are connected with the AC output terminals by way of the reactor terminals.

8. The arrangement of claim **7** further characterized by not including a capacitor series-connected with an inductor so as to form a resonant circuit that is naturally resonant near said frequency.

9. The arrangement of claim **7** wherein the arrangement is still further characterized by not having a capacitor parallel-connected with the fluorescent lamp.

10. An arrangement comprising:

an inverter-type power supply having a pair of AC input terminals connected to an ordinary AC power line voltage and a pair of AC output terminals at which is provided an AC output current of frequency much higher than that of the AC power line voltage; the power supply being further characterized by including: (i) a first inductor parallel-connected with a first capacitor so as to form a parallel-resonant circuit with a natural resonance frequency higher than the frequency of the AC output current; (ii) two alternatingly conducting transistors arranged so that at least one of the two transistors is always conducting; and (iii) a control sub-circuit operative to maintain the magnitude of the AC output current at an approximately constant level irrespective of variations in the magnitude of the AC power line voltage by as much as plus-minus ten percent; the power supply being further characterized by not including a second inductor and a second capacitor series-connected so as to form a near reasonant circuit; the control sub-circuit being further characterized by not including the equivalent of two transistors arranged in a Darlington circuit; and

a gas discharge lamp connected with the AC output terminals so as to receive said AC output current.

11. The arrangement of claim **10** wherein the two alternatingly conducting transistors periodically exhibit simultaneous conduction.

12. The arrangement of claim **10** further characterized by including a pair of terminals across which exists a DC voltage of substantially constant magnitude higher than 140 Volt.

13. An arrangement comprising:

an ordinary electric utility AC power line operable to provide an AC power line voltage at a pair of AC power line terminals;

an inverter-type power supply having a pair of AC input terminals and a pair of AC output terminals; the AC input terminals being connected with the AC power line terminals; the power supply being characterized by: (i) being operative to provide an AC output voltage across the AC output terminals; (ii) including a first inductor parallel-connected with a first capacitor so as to form a parallel-resonant circuit with a natural resonance frequency distinctly different from the frequency of the AC output voltage; and (iii) by not including a capacitor series-connected with an inductor so at to form a resonant circuit; the AC output voltage being of a frequency substantially higher than that of the AC power line voltage and having a waveform consisting of sinusoidally-shaped half-cycles of voltage of alternating polarity; the waveform having a repetition period equal to that of the AC output voltage and being sinusoidal except for a distinct discontinuity occurring over a brief period during each repetition period at or near each point-in-time where the instantaneous magnitude of the AC output voltage changes polarity; said brief period being shorter than half the duration of one of said half-cycles; the power supply being further characterized by including two periodically conducting transistors, at least one of which is conducting at any given time; and

a gas discharge lamp connected with the AC output terminals through the equivalent of a reaction impedance, thereby to draw a manifestly magnitude-limited current from the AC output terminals.

14. The arrangement of claim **13** further characterized in that the two transistors conduct simultaneously for a short time at least once during each repetition period.

15. The arrangement of claim **13** wherein the absolute magnitude of the AC output voltage at or near zero during said brief period.

16. An arrangement comprising:

a voltage conditioning circuit connected with an ordinary AC power line voltage and operative to provide a DC supply voltage at a pair of DC supply terminals; the voltage conditioning circuit being operative to supply a substantially constant-magnitude voltage at a pair of DC supply terminals;

an inverter circuit connected with the DC supply terminals by way of an inductive reactor means and operative to supply an AC output voltage at a pair of AC output terminals; the AC output voltage being of a frequency substantially higher than that of the AC power line voltage and having a waveform consisting of sinusoidally-shaped half-cycles of voltage of alternating polarity; the waveform having a repetition period and being substantially sinusoidal except for a distinct discontinuity occurring over a short time-period during each repetition period at or near each point-in-time where the magnitude of the AC output voltage changes polarity; said short time-period having a duration substantially shorter than half of that of one of said half-cycles; the inverter circuit being further characterized by including two periodically conducting transistors, at least one of which is conducting at any given time;

the equivalent of a reactive current-limiting means having two reactor terminals;

a gas discharge lamp having two lamp terminals; one of the lamp terminals being effectively connected directly with one of the AC output terminals; the other one of the lamp terminals being connected with the other AC output terminal through said equivalent of a reactive impedance means.

17. The arrangement of claim **16** further characterized in that the magnitude of the AC output voltage is distinctly lower during said short time-period, as averaged over said short time-period, than it would have been had said waveform been smoothly sinusoidal.

18. The arrangement of claim **17** wherein the instantaneous magnitude of the AC output voltage is at or near zero during said brief time-period.

19. The arrangement of claim **16** wherein the waveform of the AC output voltage is substantially sinusoidal except for a brief time-span during which the absolute magnitude of the slope of the waveform is distinctly lower than it would have been at said point-in-time if the waveform had indeed been smoothly sinusoidal.

20. An arrangement comprising:

a voltage conditioning circuit connected with ordinary AC power line voltage and operative to provide a DC supply voltage at a pair of DC supply terminals; the magnitude of the DC voltage being substantially constant;

an inverter circuit being connected with the DC supply terminals and having a pair of AC output terminals at which is provided an AC output voltage; the AC output voltage having a waveform and being of a frequency substantially higher than that of the AC power line voltage; the waveform having a repetition period and being substantially sinusoidal except for a distinct discontinuity occurring during each repetition period at or near each point-in-time where the instantaneous magnitude of the AC output voltage changes polarity and occupying a time-span that is substantially shorter than one quarter of said repetition period; the inverter circuit being further characterized by including two periodically conducting transistors, at least one of which is conducting at any given time;

a current-limiting reactive impedance having a pair of reactor terminals;

a gas discharge lamp having a pair of lamp terminals; one of the lamp terminals being connected directly with one of the AC output terminals; the other one of the lamp terminals being connected with the other one of the AC output terminals through said reactive impedance;

the arrangement being further characterized by: (i) not including a high frequency leakage transformer; and (ii) not including an inductor series-connected with a lamp-capacitor parallel-combination.

21. An arrangement comprising:

a voltage conditioning circuit connected with ordinary AC power line voltage and operative to provide a DC supply voltage at a pair of DC supply terminals; the magnitude of the DC voltage being substantially constant;

an inverter circuit being connected with the DC supply terminals and having a pair of AC output terminals at which is provided an AC output voltage; the AC output voltage having a waveform consisting of sinusoidally-shaped half-cycles of voltage of alternating polarity, such that the waveform is intermittently sinusoidal in that each half-cycle is separated by a brief time-period during which the waveform temporarily but distinctly deviates from its otherwise sinusoidal nature; the waveform having a repetition period; the brief time-period being substantially shorter than one quarter of the repetition period; the inverter circuit being further characterized by including two periodically conducting transistors, at least one of which is conducting at any given time;

a current-limiting reactive impedance having a pair of reactor terminals;

a gas discharge lamp having a pair of lamp terminals; one of the lamp terminals being connected directly with one of the AC output terminals; the other one of the lamp terminals being connected with the other one of the AC output terminals through said reactive impedance.

22. An arrangement comprising:

an inverter circuit powered from a source of DC voltage of substantially constant magnitude and operative to provide an AC output voltage at a pair of AC output terminals; the AC output voltage having a waveform consisting of sinusoidally-shaped half-cycles of voltage of alternating polarity; the waveform being intermittently sinusoidal in that each sinusoidally-shaped half-cycle is separated by a brief time-period during which the waveform temporarily but distinctly deviates from its otherwise sinusoidal nature; the waveform having a repetition period; the brief time-period being shorter than one quarter of the repetition period; the inverter circuit being further characterized by including two periodically conducting transistors, at least one of which is conducting at any given time; the transistors each having a base-emitter junction; the base-emitter junctions jointly receiving a unidirectional bias current of substantially constant magnitude;

a current-limiting reactive impedance;

a gas discharge lamp;

the lamp and the current-limiting reactive impedance being series-connected across the AC output terminals.

23. The arrangement of claim **22** wherein the inverter circuit is further characterized by including a capacitor parallel-connected with an inductor so as to form a parallel-resonant circuit having a natural resonance frequency that is distinctly different from the frequency of the AC output voltage.

24. The arrangement of claim **22** further characterized by not including: (i) an inductor series-connected with a parallel-combination of a capacitor and a gas discharge lamp; and (ii) a leakage reactance transformer.

25. An arrangement comprising:

an inverter circuit powered from a source of constant-magnitude DC voltage and operative to provide an AC voltage at a pair of AC terminals; the AC voltage being of a frequency substantially higher than that of ordinary AC power line voltage and having a waveform consisting of sinusoidally-shaped half-cycles of voltage of alternating polarity; the waveform having a repetition period and being substantially sinusoidal except for a distinct discontinuity occurring during a brief time-span at least once during each repetition period at or near a point-in-time where the instantaneous magnitude of the AC output voltage changes polarity; said brief time-span being substantially shorter than half of the duration of one of said half-cycles; the inverter circuit being further characterized by including two periodically conducting transistors, one of which is always conducting; each transistor having a base-emitter junction; the two base-emitter junctions jointly receiving a forward bias current of substantially constant magnitude;

a current-limiting reactive impedance;

a gas discharge lamp;

the lamp and the current-limiting reactive impedance being series-connected across the AC output terminals.

26. The arrangement of claim **25** including a parallel-connected L-C circuit having a natural resonance frequency distinctly different from the frequency of the AC voltage.

* * * * *

EXHIBIT O

US006479074B2

(12) **United States Patent**
Murdock et al.

(10) **Patent No.: US 6,479,074 B2**
(45) **Date of Patent: *Nov. 12, 2002**

(54) **METHODS AND TRANSDERMAL COMPOSITIONS FOR PAIN RELIEF**

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(73) Assignee: **Pharmaceutical Applications Associates LLC**, Yakima, WA (US)

(*) Notice: Subject to any disclaimer, the term of this patent is extended or adjusted under 35 U.S.C. 154(b) by 0 days.

This patent is subject to a terminal disclaimer.

(21) Appl. No.: **09/825,375**

(22) Filed: **Apr. 2, 2001**

(65) **Prior Publication Data**

US 2002/0009487 A1 Jan. 24, 2002

Related U.S. Application Data

(60) Division of application No. 09/652,662, filed on Aug. 31, 2000, and a division of application No. 09/342,679, filed on Jun. 29, 1999, now abandoned, and a continuation-in-part of application No. 09/106,684, filed on Jun. 29, 1998, now Pat. No. 6,290,986, which is a continuation-in-part of application No. PCT/US97/19651, filed on Oct. 24, 1997, and a continuation-in-part of application No. 08/957,485, filed on Oct. 24, 1997, now abandoned.

(60) Provisional application No. 60/122,903, filed on Mar. 5, 1999, and provisional application No. 60/029,120, filed on Oct. 24, 1996.

(51) **Int. Cl.**^{7} **A61F 13/00**

(52) **U.S. Cl.** **424/449**; 424/447; 424/448; 424/484; 514/78; 514/906

(58) **Field of Search** 424/447, 448, 424/449, 484; 514/906, 78

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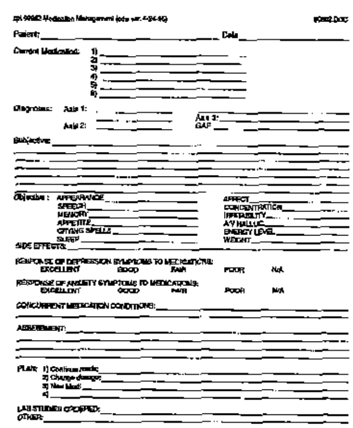
(57) **ABSTRACT**

The present invention features methods and compositions for transdermal administration. In one embodiment, the invention features methods and compositions for transdermal administration of an amine containing compound having biphasic solubility and/or an agent which enhances the activity of the amine containing compound having biphasic solubility, e.g., a muscle relaxant, to relieve pain.

17 Claims, 11 Drawing Sheets



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cpt 90862 Medication Management (cdw ver. 4-24-95) 90862.DOC

Patient: ______________________ Date ______________

Current Medication: 1) ______________
2) ______________
3) ______________
4) ______________
5) ______________
6) ______________

Diagnoses: Axis 1: ______________ ______________
______________ Axis 3: ______________
Axis 2: ______________ GAF ______________

Subjective: ______________________________

Objective : APPEARANCE______________ AFFECT______________
SPEECH ______________ CONCENTRATION______________
MEMORY______________ IRRITABILITY______________
APPETITE______________ A/V HALLUC.______________
CRYING SPELLS ______________ ENERGY LEVEL______________
SLEEP ______________ WEIGHT______________

SIDE EFFECTS:______________________________

RESPONSE OF DEPRESSION SYMPTOMS TO MEDICATIONS:
EXCELLENT GOOD FAIR POOR N/A

RESPONSE OF ANXIETY SYMPTOMS TO MEDICATIONS:
EXCELLENT GOOD FAIR POOR N/A

CONCURRENT MEDICATION CONDITIONS:______________________________

ASSESSMENT:______________________________

PLAN: 1) Continue meds:______________
2) Change dosage:______________
3) New Med:______________
4) ______________

LAB STUDIES ORDERED:______________
OTHER:______________

Fig. 1

Case Processing Summary(a)

	Cases					
	Included		Excluded		Total	
	N	Percent	N	Percent	N	Percent
ankle * MEDS * Composition	4	3.1%	127	96.9%	131	100.0%
arm * MEDS * Composition	10	7.6%	121	92.4%	131	100.0%
Back * MEDS * Composition	69	52.7%	62	47.3%	131	100.0%
elbow * MEDS * Composition	11	8.4%	120	91.6%	131	100.0%
headache * MEDS * Composition	131	100.0%	0	.0%	131	100.0%
Knee * MEDS * Composition	19	14.5%	112	85.5%	131	100.0%
hip * MEDS * Composition	15	11.5%	116	88.5%	131	100.0%
Neck * MEDS * Composition	28	21.4%	103	78.6%	131	100.0%
leg * MEDS * Composition	13	9.9%	118	90.1%	131	100.0%
shoulder * MEDS * Composition	25	19.1%	106	80.9%	131	100.0%
wrist * MEDS * Composition	26	19.8%	105	80.2%	131	100.0%
a Limited to first 150 cases						

P

						Case Number	ankle	arm
	c-dox-gu	Composition	5/5/10	1		26	.	.
				2		33	.	.
				3		41	.	.
				4		59	.	.
				5		73	.	.
				6		80	.	.
				Total	N			
					Mean			
			4/5/10	1		98	.	.
				2		112	.	.
				Total	N			
					Mean			

Fig. 2A

			Total	N				
				Mean				
	c-gab-do	Composition	5/5/5	1		34	.	.
				Total	N			
					Mean			
			Total	N				
				Mean				
	carb	Composition	4	1		5	.	.
				Total	N			
					Mean			
			6	1		81	.	.
				Total	N			
					Mean			
			Total	N				
				Mean				
	carb-ami	Composition	4/5	1		12	.	.
				2		40	.	.
				3		49	.	.
				4		64	.	.
				Total	N			
					Mean			
			Total	N				
				Mean				
	carb-gab	Composition	4/4	1		27	.	mild-moderate
				2		35	.	.
				3		126	.	moderate
				Total	N			2
					Mean			1.750
			Total	N				2
				Mean				1.750
				1		4	.	moderate
				2		13	.	.
				3		15	.	.

Fig. 2B

	dox	Composition	5	4		42	.	.
				5		46	.	none
				6		74	.	.
				7		95	moderate	.
				8		116	.	.
				9		121	.	.
				10		128	.	moderate
				Total	N		1	3
					Mean		2.000	1.333
			Total	N			1	3
				Mean			2.000	1.333
	dox-chl	Composition	7/13	1		10	.	.
				Total	N			
					Mean			
			5/10	1		21	.	.
				2		29	.	.
				3		30	.	.
				Total	N			
					Mean			
			7/10	1		83	.	.
				Total	N			
					Mean			
			Total	N				
				Mean				
				1		7	.	.
				2		9	.	.
				3		14	.	.
				4		18	.	.
				5		20	.	.
				6		25	.	.
				7		36	.	.
				8		50	.	.

Fig. 2C

			5/10	9	58	.	.
				10	71	.	.
				11	76	.	.
				12	77	.	.
				13	90	.	.
				14	97	.	.
				15	101	.	.
				16	103	.	.
				17	108	.	.
	dox-guai	Compsition		18	123	.	.
				19	131	.	.
				Total N			
				Total Mean			
			7/10	1	22	.	.
				2	47	.	.
				3	111	.	.
				Total N			
				Total Mean			
			10/10	1	23	.	.
				2	48	.	.
				3	53	.	.
				4	57	.	.
				5	67	.	.
				Total N			
				Total Mean			
			Total	N			
				Mean			
			4/5/10	1	11	.	.
				Total N			
				Total Mean			
				1	1	.	.
				2	32	.	.
				3	39	.	.

Fig. 2D

	g-dox-gu	Composition	5/5/10	4		44	.	.
				5		51	.	.
				6		54	.	.
				7		62	.	.
				8		72	.	.
				9		85	.	.
				10		87	.	.
				11		93	.	.
				12		119	.	.
				13		129	.	.
				Total	N			
					Mean			
			Total	N				
				Mean				
	gab-dox	Composition	5/5	1		37	.	.
				2		65	.	.
				3		68	.	.
				Total	N			
					Mean			
			Total	N				
				Mean				
	k-ca-dox	Composition	10/4/5	1		86	.	.
				Total	N			
					Mean			
			Total	N				
MEDS				Mean				
			10/6/3	1		43	.	.
				Total	N			
					Mean			
				1		102	.	.

Fig. 2E

k-car-pi	Composition	10/4/3	2		104	.	.
			Total	N			
				Mean			
		Total	N				
			Mean				
k-dox-ch	Composition	20/10/5	1		100	.	.
			Total	N			
				Mean			
		Total	N				
			Mean				
k-dox-gu	Composition	3/5/5	1		6	.	.
			Total	N			
		20/5/10	1		63	.	.
			Total	N			
				Mean			
		Total	N				
			Mean				
k-dox-pi	Composition	10/4/3/5	1		122	.	.
			Total	N			
				Mean			
		Total	N				
			Mean				
k-g-do-g	Composition	10/4/5/10	1		17	.	.
			Total	N			
				Mean			
		Total	N				
			Mean				
k-gab	Composition	20/4	1		115		none
			Total	N		.	1
				Mean			.000
		Total	N				1
			Mean				.000
			1		117	.	.

Fig. 2F

	k-gab-am	Composition	20/5/5	Total	N			
					Mean		.	.
			Total	N				
				Mean				
	k-gab-do	Composition	20/4/5	1		55	.	.
				Total	N			
					Mean			
			10/5/4	1		99	major	.
				Total	N		1	
					Mean		3.000	
			10/5/5	1		113	.	.
				Total	N			
					Mean			
			20/5/5	1		118	.	.
				Total	N			
					Mean			
			Total	N			1	
				Mean			3.000	
	k-gab-gu	Composition	20/4/4/1	1		94	.	.
				Total	N			
					Mean			
			20/5/5	1		105	.	.
				Total	N			
					Mean			
			Total	N				
				Mean				
				1		2	.	.
				2		8	.	major
				3		19	.	.
				4		31	.	.
				5		38	.	.
				6		45	.	none

Fig. 2G

			10/4/3	7		56	.	.
				8		78	.	.
				9		89	.	.
				10		109	.	.
				11		120	.	.
				12		124	.	.
				13		130	.	.
				Total	N			2
					Mean			1.500
	k-gab-pi	Composition	10/4/1	1		16	.	.
				2		28	.	mild
				3		52	.	.
				4		66	.	.
				5		69	.	.
				6		75	moderate	.
				7		82	.	.
				8		84	.	.
				9		88	.	.
				10		91	.	.
				11		96	major	.
				12		125	.	.
				Total	N		2	1
					Mean		2.500	1.000
			10/1/3	1		114	.	.
				Total	N			
			Total	N			2	3
				Mean			2.500	1.333
	k-pi	Composition	10/3	1		127	.	.
				Total	N			
					Mean			
			Total	N				
				Mean				
				1		110	.	.

Fig. 2H

la-li-gu	Composition	5/5/10	Total	N			
				Mean			
		Total	N				
			Mean				
lam-chl	Composition	7/10	1		3	.	moderate-major
			Total	N			1
				Mean			2.500
		10/10	1		24	.	.
			2		70	.	.
			3		106	.	.
			Total	N			
				Mean			
		Total	N				1
			Mean				2.500
n-dox-ch	Composition	30/5/5	1		79	.	.
			Total	N			
				Mean			
		Total	N				
			Mean				
naproxen	Composition	30	1		60	.	.
			Total	N			
		Total	N				
tri-chl	Composition	7/10	1		61	.	.
			Total	N			
				Mean			
		7/13	1		92	.	.
			2		107	.	.
			Total	N			
				Mean			
		Total	N				
			Mean				

Fig. 2I

	Total	N		4	10
		Mean		2.500	1.400
June 2 1999 N=131					
a Limited to first 150 cases					

Fig. 2J

METHODS AND TRANSDERMAL COMPOSITIONS FOR PAIN RELIEF

CROSS-REFERENCE TO RELATED APPLICATIONS

This application claims priority to divisional application Ser. No. 09/652,662 filed Aug. 31, 2000, U.S. Pat. application No. 09/342,679 filed on Jun. 29, 1999, abandoned U.S. Provisional Patent Application No. 60/122,903 filed on Mar. 5, 1999, and is also a continuation-in-part of U.S. patent application Ser. No. 09/106,684 filed on Jun. 29, 1998, now U.S. Pat. No. 6,290,986 which is a continuation-in-part of PCT Application Ser. No. PCT/US97/19651 filed on Oct. 24, 1997, and a continuation-in-part of U.S. patent application Ser. No. 08/957,485 filed on Oct. 24, 1997, now abandoned, and U.S. Provisional Patent Application Ser. No. 60/029,120 filed on Oct. 24, 1996. The contents of all of the aforementioned application(s) are hereby incorporated by reference.

FIELD OF THE INVENTION

The present invention is directed to methods and compositions for transdermal administration. In particular, the present invention is directed to methods and compositions for the transdermal administration of an amine containing compound having biphasic solubility and/or an agent which enhances the activity of the amine containing compound having biphasic solubility, e.g., a muscle relaxant, to relieve pain.

BACKGROUND OF THE INVENTION

It is believed that damage to somatic sensory nerves causes a somatic sensory loss. Such damage can be caused by a variety of means including trauma, diseases such as diabetes, herpes zoster and late-stage cancer, chemotherapy, or by a chemical injury. It is believed that neural pain circuits rewire themselves, both anatomically and biochemically, after nerve injury. In many patients suffering from damage to somatic sensory nerves, negative symptoms such as numbness are joined by positive sensations, involving a sort of false sensation of pain. The experience can range from mild dysesthesia to excruciating pain, rendering some patients unable to work, walk or do other daily activities.

In the past, patients were generally treated by administration of analgesics to relieve pain. A vast majority of such patients receive doses of these agents orally. Unfortunately, in some situations, oral administration of such agents has been associated with a variety of side effects, such as liver damage, kidney damage, gastrointestinal side effects, addiction, sedation, and/or weight gain which cannot be tolerated well by the patient. In other cases, malabsorption of oral preparations have resulted in subtherapeutic plasma levels. In other cases, the agents have relatively short plasma half-lives, necessitating inconveniently frequent dosing. In general, oral delivery involves a time delay as the analgesic is absorbed via the digestive system before entering the bloodstream. A number of agents which have traditionally been administered orally or by injection have been inappropriate or suboptimal for some patients when so-administered. There are a number of medications which, in at least some patients, are not tolerated well when orally administered (e.g. which cause undesirable gastrointestinal or other side effects) and/or which provide undesirably high or low concentrations or delayed concentrations in a target tissue. In some cases, dosages which are appropriate for oral administration, upon being distributed more or less uniformly throughout the body, are undesirably low in a particular area, e.g., tissue, to achieve desired results. Oral or injection administration may result in too slow or too rapid increase in blood plasma levels, e.g., may involve an undesirably long time delay as the analgesic is absorbed by the digestive system before entering the bloodstream, or may result in a "spike" in blood plasma levels followed by an undesirably low level, where a more constant level would be preferable. Some analgesics are particularly prone to cause or contribute to kidney or liver damage when administered orally.

Although other forms of delivery of pharmaceuticals agents are known, each has its drawbacks. Parenteral (i.e., intravenously or intramuscularly injected) administration is inconvenient and expensive, and is rarely used outside the hospital. Inhalation is believed to be not feasible with many analgesic agents currently in use. Therefore, there is a need for an analgesic delivery system which provides effective and acceptable levels, while preferably avoiding or reducing undesired effects such as liver damage or gastrointestinal side effects.

SUMMARY OF THE INVENTION

The present invention provides a transdermal composition for the treatment of pain in a subject, particularly a human subject. The transdermal composition for the treatment of pain in a subject includes an amine containing compound having biphasic solubility in an amount effective to treat pain in a subject and a pharmaceutically acceptable carrier suitable for transdermal delivery of the amine containing compound, e.g., a lecithin organogel carrier. In a preferred embodiment, the transdermal composition further includes an agent which enhances the activity of the amine containing compound having biphasic solubility, e.g., a muscle relaxant, such as guaifenesin, chlorzoxazone, dantrolene sodium, metaxalone, carisoprodol, and combinations thereof. Preferably, the agent which enhances the activity of the amine containing compound having biphasic solubility, e.g., the muscle relaxant, also has a biphasic solubility.

In one embodiment of the present invention, the amine containing compound having biphasic solubility is an antidepressant compound, such as a tricyclic antidepressant compound, e.g., doxepin or trimipramine.

In another embodiment of the present invention, the amine containing compound having biphasic solubility is a sodium channel blocker, an anti-epileptic compound, or an anti-convulsant compound.

Another embodiment of the invention features a transdermal composition which includes an amine-containing compound as described herein and an anti-inflammatory compound, such as a nonsteroidal anti-inflammatory compound, e.g., celecoxib, etodolac, mefanamic acid, nabumetone, salsalate, naproxen, vioxx®, and combinations thereof. Such a composition can further include an agent which enhances the activity of the amine containing compound, e.g., a muscle relaxant such as guaifenesin.

In another aspect, the invention features a transdermal composition for the treatment of pain in a subject including an amine containing compound having biphasic solubility in an amount effective to treat pain in a subject; a muscle relaxant in an amount effective to enhance the activity of the amine containing compound having biphasic solubility; and a pharmaceutically acceptable carrier suitable for transdermal delivery of the amine containing compound having biphasic solubility and the muscle relaxant.

In yet another aspect, the invention features a transdermal composition for the treatment of pain in a subject including doxepin in an amount effective to treat pain in a subject-;.guaifenesin in an amount effective to enhance the activity of doxepin; and a pharmaceutically acceptable carrier suitable for transdermal delivery of the doxepin and the guaifenesin.

Other aspects of the invention feature methods for treating pain in a subject in which the subject is contacted with a transdermal composition including an amine containing compound having biphasic solubility in an amount effective to treat pain in the subject; and a pharmaceutically acceptable carrier suitable for transdermal delivery of the amine containing compound to thereby treat pain in the subject. In a preferred embodiment, the transdermal composition is applied to the skin of the subject.

Another aspect of the invention features a method for selecting a compound suitable for treating pain in a subject. The method includes transdermally administering an amine containing compound having biphasic solubility to a subject; and determining whether pain is treated in the subject to thereby select a compound suitable for treating pain in a subject. In a preferred embodiment, the method can further include modeling the compound using a computer equipped with a three-dimensional chemical structure modeling program; and determining whether the three dimensional chemical structure of the compound possesses sufficient characteristics to be useful as a sodium channel blocker, thereby selecting a compound suitable for treating pain in a subject.

In another aspect, the invention features a transdermal composition suitable for transdermal delivery, which includes a therapeutically effective amount of a pharmaceutical compound (e.g., a serotonin specific reuptake inhibitor, a mood stabilizing compound, a dopamine compound, a compound suitable for treating attention deficit hyperactivity disorder, a compound suitable for treating hypertension and akathisia, an analgesic compound, or a compound used in the treatment of impotence) and a pharmaceutically acceptable carrier suitable for transdermal delivery of the pharmaceutical compound, e.g., a lecithin organogel carrier.

In yet another aspect, the invention features a transdermal composition for treatment of pain in a subject which includes a compound capable of blocking afferent neuron transmission in an amount effective to block afferent neuron transmission in a subject; and a pharmaceutically acceptable carrier suitable for transdermal delivery of the compound.

Other features and advantages of the invention will be apparent from the following detailed description and claims.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 is an evaluation form used in evaluating an embodiment of the present invention.

FIG. 2 is a table depicting the results from clinical experiments using compositions of the invention.

DETAILED DESCRIPTION OF THE INVENTION

The present invention provides a transdermal composition suitable for treatment of pain in a subject. The transdermal composition includes an amine containing compound having biphasic solubility in an amount effective to treat pain in a subject; and a pharmaceutically acceptable carrier suitable for transdermal delivery of the amine containing compound having biphasic solubility.

As used herein, the term "subject" includes a mammal, such as a human, a horse, a pig, a cow, a mouse, a rat, a rabbit, or a goat. In preferred embodiment, the subject is a human.

As used herein, the term "pain" is art recognized and includes a bodily sensation elicited by noxious chemical, mechanical, or thermal stimuli, in a subject, e.g., a mammal such as a human. The term "pain" includes chronic pain, such as lower back pain; pain due to arthritis, e.g., osteoarthritis; joint pain, e.g., knee pain or carpal tunnel syndrome; myofascial pain, and neuropathic pain. The term "pain" further includes acute pain, such as pain associated with muscle strains and sprains; tooth pain; headaches; pain associated with surgery; or pain associated with various forms of tissue injury, e.g., inflammation, infection, and ischemia.

As used herein, the term "amine containing compound having biphasic solubility" includes compounds having at least one amine moiety and having sufficient lipid solubility (e.g., solubility in polar solvents such as ethanol, ethoxydiglycerol, ethoxydiglycol, chloroform, benzene, and the like) such that the compound passes through the stratum corneum, and has sufficient aqueous solubility to be active in the aqueous environment of the dermis and the underlying tissue.

Transdermal compositions of the present invention include an amine containing compound having biphasic solubility in an amount effective to treat pain in a subject. As used herein, the terms "amount effective to treat pain in a subject" and "effective amount" are used interchangeably herein and include an amount effective, at dosages and for periods of time necessary, to achieve the desired result, e.g., sufficient to treat pain in a subject. An effective amount of an amine containing compound or a pharmaceutical compound as defined herein may vary according to factors such as the disease state, age, and weight of the subject, and the ability of the amine containing compound or pharmaceutical compound to elicit a desired response in the subject. Dosage regimens may be adjusted to provide the optimum therapeutic response. An effective amount is also one in which any toxic or detrimental effects of the amine containing compound having biphasic solubility or pharmaceutical compound are outweighed by the therapeutically beneficial effects.

The transdermal compositions of the invention can further include an agent which enhances the activity of the amine containing compound having biphasic solubility. As used herein, an "agent which enhances the activity of the amine containing compound having biphasic solubility" includes an agent which enhances the pharmacological activity of the amine containing compound hiving biphasic solubility (e.g., the ability of the amine containing compound to treat pain), or enhances the transdermal delivery of the amine containing compound having biphasic solubility (e.g., the ability of the amine containing compound to cross the stratum corneum), or enhances both the pharmacological activity and the transdermal delivery of the amine containing compound. Examples of agents which enhance the activity of the amine containing compound having biphasic solubility, include muscle relaxants, described in further detail below.

As used herein, the term "transdermal" composition includes compositions capable of passing through the stratum corneum of a subject. The term transdermal further includes compositions capable of passing through the epidermis of a subject, compositions capable of passing through the dermis of a subject, and compositions capable of passing through the hypodermis of a subject. In preferred embodiments, the term transdermal includes compositions capable of passing through the skin of a subject and reaching the underlying tissues and organs.

As used herein, the term "transdermal delivery" includes delivery of, for example, a compound through the stratum corneum of a subject. The term transdermal delivery further includes delivery of, for example, a compound through the epidermis of a subject, delivery of, for example, a compound through the dermis of a subject, and delivery of, for example, a compound through the hypodermis of a subject. In preferred embodiments, the term transdermal delivery includes delivery of, for example, a compound through the skin of a subject to the underlying tissues and organs.

The present invention further features a transdermal composition for treatment of pain in a subject which includes a compound capable of blocking afferent neuron transmission in an amount effective to block afferent neuron transmission in a subject; and a pharmaceutically acceptable carrier suitable for transdermal delivery of the compound.

As used herein, the term "compound capable of blocking afferent neuron transmission" includes a compound which is capable of blocking the ability of an afferent neuron, i.e., a sensory neuron, to carry an impulse toward the central nervous system.

Various aspects of the invention are described in further detail in the following subsections:

Amine Containing Compounds Having Biphasic Solubility

Amine containing compounds having biphasic solubility for use in the transdermal compositions of the invention include antidepressant compounds, antiepileptic compounds, anticonvulsant compounds, and sodium channel blockers. As used herein, the term "antidepressant compounds" includes compounds capable of alleviating the symptoms of depression. Examples of antidepressant compounds include all tricyclic antidepressants (e.g., amitriptyline, dothiepin, or lofepramine), bupropion (sold under the trade name Wellbutrin), reboxetine (sold under the trade name Edronax), nefazodone (sold under the trade name Serzone) and trazone (sold under the trade name Desyrel). Antidepressant compounds are described in, for example, the 1998 SIGMA catalogue and the "The Merck Index", 12th Ed., Budavari et al., eds., Merck & Co., Inc., Rahway, N.J., 1996, the contents of which are incorporated herein by reference.

In one embodiment of the present invention, the antidepressant compounds of the present invention contain a tricyclic moiety. Therefore, in a preferred embodiment, a transdermal composition of the present invention includes a tricyclic antidepressant compounds. Exemplary tricyclic antidepressants include adinazolam, amitriptylinoxide, amoxapine, clomipramine, demexiptiline, dimetacrine, dothiepin, doxepin, imipramine N-oxide, iprindole, lofepramine, melitracen, metapramine, noxiptilin, pizotyline, propizepine, quinupramine, tianeptine, and trimipramine. A particularly preferred tricyclic antidepressant for use in the compositions of the invention is doxepin.

Tricyclic antidepressant compounds are described in, for example, "Guide to Clinical Neurology" by J. P. Mohr et al. (Churchill Livingstone, 1995), the contents of which are incorporated herein by reference.

Preferably, the tricyclic antidepressant compound is selected from the group consisting of doxepin, trimipramine, other tricyclics having biphasic solubility, and combinations thereof. When combined with other compounds, such as an agent which enhances the activity of the amine containing compound, e.g., a muscle relaxant, and/or an anti-inflammatory compound, e.g., a nonsteroidal anti-inflammatory compound, as discussed below, the tricyclic antidepressant preferably constitutes from about 1% by weight (% by wt.) to about 30% by wt. of the total amount of the pharmaceutical, more preferably from about 3% by wt. to about 15% by wt., and most preferably from about 5% by wt. to about 13% by wt.

The amine containing compounds having biphasic solubility used in the transdermal compositions of the invention further include antiepileptic compounds. As used herein, the term "antiepileptic compound" includes compounds capable of alleviating the symptoms of epilepsy. Exemplary antiepileptic compounds for use in the compounds of the invention include lamotrigine, felbamate, and carbamazepine. Preferably, the antiepileptic compound is selected from the group consisting of lamotrigine, felbamate, carbamazepine, and combinations thereof. When combined with other compounds, such as an agent which enhances the activity of the amine containing compound, e.g., a muscle relaxant, and/or an anti-inflammatory compound, e.g., a nonsteroidal anti-inflammatory compound as discussed below, the antiepileptic compound constitutes from about 1% by wt. to about 30% by wt. of the total amount of the pharmaceutical, more preferably from about 3% by wt. to about 20% by wt., and most preferably from about 5% by wt. to about 15% by wt. Antiepileptic compounds are described in, for example, the 1998 SIGMA catalogue, the "The Merck Index", 12th Ed., Budavari et al., eds., Merck & Co., Inc., Rahway, N.J., 1996, and the "Guide to Clinical Neurology" by J. P. Mohr et al. (Churchill Livingstone, 1995) the contents of which are incorporated herein by reference.

In another aspect of the present invention, the amine containing compounds having biphasic solubility of the present invention include anticonvulsant compounds. As used herein, the term "anticonvulsant compound" includes compounds capable of alleviating the symptoms of convulsion, i.e., the violent involuntary tetanic contractions of an entire group of muscles. Exemplary anticonvulsant compounds which for use in the compositions of the invention include felbamate, lamotrigine and carbamazepine. Preferably, the anticonvulsant compound is selected from the group consisting of felbamate, lamotrigine, and combinations thereof. When combined with other compounds, such as an agent which enhances the activity of the amine containing compound, e.g., a muscle relaxant, and/or an anti-inflammatory compound, e.g., a nonsteroidal anti-inflammatory compound as discussed below, the anticonvulsant compound constitutes from about 1% by wt. to about 30% by wt. of the total amount of the pharmaceutical, more preferably from about 3% by wt. to about 20% by wt., and most preferably from about 5% by wt. to about 15% by wt. Anticonvulsant compounds are described in, for example, the 1998 SIGMA catalogue, the "The Merck Index", 12th Ed., Budavari et al., eds., Merck & Co., Inc., Rahway, N.J., 1996, and the "Guide to Clinical Neurology" by J. P. Mohr et al. (Churchill Livingstone, 1995) the contents of which are incorporated herein by reference.

In yet another aspect of the present invention, the amine containing compounds having biphasic solubility of the present invention include adrenergic agonist compounds. Preferably, the adrenergic agonist compound is tizanidine. When combined with other compounds, such as a muscle relaxant and/or nonsteroidal anti-inflammatory compound as discussed below, the adrenergic agonist compound constitutes from about 1% by wt. to about 30% by wt. of the total amount of the pharmaceutical, more preferably from about 3% by wt. to about 20% by wt., and most preferably from about 5% by wt. to about 15% by wt. Adrenergic agonist compounds are described in, for example, the 1998 SIGMA catalogue, the "The Merck Index", 12th Ed., Budavari et al., eds., Merck & Co., Inc., Rahway, N.J., 1996, and the "Guide to Clinical Neurology" by J. P. Mohr et al.(Churchill Livingstone, 1995) the contents of which are incorporated herein by reference.

The amine containing compounds having biphasic solubility used in the transdermal compositions of the invention further include sodium channel blockers. As used herein, the term "sodium channel blockers" includes compounds which are capable of blocking the activity of a sodium channel. Examples of sodium channel blockers include tetrodoxin, flecainide, disopyramide, and terfenadine. Sodium channel blockers are described in, for example, the 1998 SIGMA catalogue, the "The Merck Index", 12th Ed., Budavari et al., eds., Merck & Co., Inc., Rahway, N.J., 1996, and the "Guide to Clinical Neurology" by J. P. Mohr et al.(Churchill Livingstone, 1995) the contents of which are incorporated herein by reference.

Whenever nerves are damaged, for example, by trauma, by diseases such as diabetes, herpes zoster, or late-stage cancer, or by chemical injury (e.g., as an untoward consequence of agents including the false-nucleoside anti-HIV pharmaceuticals), neural pain circuits rewire themselves, anatomically and/or biochemically. Thus, following an injury, new sodium channels are formed which is believed to constitute the basis for chronic pain development. Through a similar action in the dorsal root ganglia, chronic regional pain syndromes may develop. Each time one of these sodium channels depolarizes, a nerve impulse originates. Because there are so many sodium channels, there may be a constant cascade of nerve impulses, causing allodynia, burning sensations, and/or dysesthesias. It is believed that some chronic pains may be mediated through sodium channels in nerve cells. Thus, it is believed that amine containing compounds having biphasic solubility which can block sodium channels may also be used in the transdermal compositions of the invention.

In one embodiment of the invention, the amine moiety of the amine containing compounds having biphasic solubility of the present invention may function similar to a sodium ion upon entry into the sodium channel of a nerve cell membrane. A non-polar moiety, which is preferably present in the amine containing compound having biphasic solubility of the present invention may interact with the nerve cell membrane, perhaps through Van der Waals forces. In such cases, it is believed that the presence of the non-polar moiety prevents or inhibits a complete uptake of the amine containing compound having biphasic solubility through the nerve cell membrane. It is believed that one or more these interactions prevent or reduce the amount and/or the rate of depolarization and ion exchange involved in stimulus conduction, thereby decreasing pain sensation.

The amount of an amine containing compound having biphasic solubility useful in relieving pain transdermally may be determined by methods known in the art, and typically ranges from about 1 mg to about 300 mg per subject per dose, preferably from about 5 mg to about 100 mg per subject per dose, and more preferably from about 10 mg to about 50 mg per subject per dose, depending on a variety of factors including the particular amine containing compound having biphasic solubility used, whether the area of transdermal application is the site of action, and the intended size of the site of action. In a preferred embodiment, the amount of an amine containing compound having biphasic solubility useful in relieving pain transdermally, is 5, 10, 15, 20, 25, 30, 35, 40, 45, 50, 55, 60, 65, 70, 75, 80, 85, 90, 95, 100 mg, 150 mg, 200 mg, 250 mg, or 300 mg per subject per dose.

Muscle Relaxants

Transdermal compositions of the present invention may also include a muscle relaxant. As used herein, the term "muscle relaxant" includes compounds which facilitate or enhance the relaxation of muscles (e.g., provide relief from muscle spasm) and, thus, facilitate or enhance the transdermal delivery of the transdermal compositions of the invention. Exemplary muscle relaxants include both skeletal muscle relaxants and smooth muscle relaxants such as anticholinergics, antispasmodics, bronchodilators, and vasodilators. Muscle relaxants are described in, for example, the 1998 SIGMA catalogue, the "The Merck Index", 12th Ed., Budavari et al., eds., Merck & Co., Inc., Rahway, N.J., 1996, pp. THER-1 to THER-28, and the "Guide to Clinical Neurology" by J. P. Mohr et al. (Churchill Livingstone, 1995) the contents of which are incorporated herein by reference. Preferably, the muscle relaxant is selected from the group consisting of guaifenesin, benzodiazepines (e.g., clozapine or diazopam), chlorzoxazone, dantrolene sodium, metaxalone, carisoprodol, other muscle relaxants having biphasic solubility, and combinations thereof More preferably, the muscle relaxant is selected from the group consisting of guaifenesin, chlorzoxazone, and combinations thereof. A preferred muscle relaxant for use in the compositions of the invention is guaifenesin.

Preferably the muscle relaxant has biphasic solubility. Preferably the muscle relaxant, when present in the pharmaceutical composition, constitutes from about 1% by wt. to about 30% by wt. of the total amount of the pharmaceutical, more preferably from about 3% by wt. to about 20% by wt., and most preferably from about 5% by wt to about 15% by wt.

Anti-Inflammatory Compounds

The transdermal compositions of the present invention may also include an anti-inflammatory compound. As used herein, the term "anti-inflammatory compound" includes a compound which is capable of reducing cell migration, caused by ischemic and trauma associated events, and therefore reduces edema formation to thereby provide pain relief. Preferably, the anti-inflammatory compound is a nonsteroidal anti-inflammatory compound (i.e., NTHE) including ketoprofen. Anti-inflammatory compounds, e.g., NTHEs, are described in, for example, the 1998 SIGMA catalogue, the "The Merck Index", 12th Ed., Budavari et al., eds., Merck & Co., Inc., Rahway, N.J., 1996, pp. THER-1 to THER-28, and the "Guide to Clinical Neurology" by J. P. Mohr et al. (Churchill Livingstone, 1995) the contents of which are incorporated herein by reference. Preferably, the NTHE is selected from the group consisting of celecoxib, etodolac, mefanamic acid, nabumetone, salsalate, naproxen, Vioxx®, COX-2 NTHEs having biphasic solubility, and combinations thereof.

More preferably, the NTHE is selected from the group consisting of celecoxib, etodolac, naproxen, COX-2 NTHEs having biphasic solubility, and combinations thereof. Preferably, the NTHE has biphasic solubility. The NTHE, when present in the transdermal composition, preferably, constitutes from about 1% by wt. to about 30% by wt. of the total amount of the pharmaceutical, more preferably from about 3% by wt. to about 30% by wt., and most preferably from about 5% by wt. to about 30% by wt.

Dosages

The concentration as well as the quantity of the amine containing compounds having biphasic solubility, the agents which enhance the activity of the amine containing compounds, e.g., the muscle relaxants, and the anti-inflammatory compounds can be varied independently in order to achieve the desired effect. For example, higher concentrations of the amine containing compounds having biphasic solubility, the muscle relaxants, and the anti-inflammatory compounds contained in a dosage form of decreased viscosity may result in an analgesic with fast onset and short duration. High concentrations of the amine containing compounds having biphasic solubility, the muscle relaxants, and the anti-inflammatory compounds contained in a dosage form of increased viscosity may result in potent analgesic with fast onset and long duration. Low concentrations of the amine containing compounds having biphasic solubility, the muscle relaxants, and the anti-inflammatory compounds in a dosage form of decreased viscosity may result in mild analgesic with longer onset and short duration. Low concentrations of the amine containing compounds having biphasic solubility, the muscle relaxants, and the anti-inflammatory compounds contained in a dosage form of increased viscosity may have mild analgesic properties with longer onset and longer duration. The ability to vary the concentration of the amine containing compounds having biphasic solubility, the muscle relaxants, and the anti-inflammatory compounds from very low to high of the total composition, combined with the ability to coat thin (about 0.1 mm) or thick (about 0.5 mm) enables the practitioner of the invention to vary the dosage of the system as needed for particular level of pain and anatomical sites of interest. It should be appreciated, however, that, onset time as yell as duration of analgesic effect of the transdermal composition of the present invention will vary from subject to subject as well as on the basis of the site of application, and properties of the amine containing compounds having biphasic solubility, the muscle relaxants, and the anti-inflammatory compounds.

Generally, the concentration of the amine containing compounds having biphasic solubility, the muscle relaxants, and the anti-inflammatory compounds can range, on a weight basis, from about 1% to about 30% of the total composition, preferably from about 3% to about 20%, and more preferably from about 5% to about 15%.

Pharmaceutically Acceptable Carriers

The transdermal compositions of the present invention also includes a pharmaceutically acceptable carrier which is capable of transdermal delivery of the amine containing compound having biphasic solubility. As used herein, the term "pharmaceutically acceptable carrier suitable for transdermal delivery" includes a carrier capable of delivering the amine containing compound transdermally as defined above. Suitable carriers for transdermal delivery of pharmaceuticals are described in U.S. Pat. No. 5,446,070, the contents of which are incorporated herein by reference. Briefly, pharmaceutically acceptable carriers of the present invention include any suitable finite (i.e, solid) or non-finite (i.e, non-solid, such as liquid or semi-liquid) carrier including liquids, semi-liquids or solid carriers, such as a bioadhesive. Thus, the amine containing compounds having biphasic solubility may be admixed with a pharmaceutically acceptable carrier such as a cream, gel, emulsion, lotion, salve, paste, plaster, ointment, spray solution, or any other "non-finite" carrier known in the art of pharmaceutical delivery. For example, the base of a non-finite carrier may be lipid including phospholipids such as lecithins; fatty oils; lanolin; vasoline; paraffins; glycols; higher fatty acids; and higher alcohols.

The term "bioadhesive" as used herein includes an adhesive which attaches to a biological surface such as skin or mucosal tissue. Preferably, the bioadhesive of the present invention is self-adhesive in that it attaches to the site of interest without the need to reinforce its attachment by way of another adhesive. Suitable bioadhesive include natural or synthetic polysaccharides such as cellulose derivatives including methylcellulose, cellulose acetate, carboxymethylcellulose, hydroxyethylcellulose and the like; pectin; a mixture of sulfated sucrose and aluminum hydroxide; hydrophilic polysaccharide gums including natural plant exudates, such as karaya gum, ghatti gum, tragacanth gum, xanthan gum, jaraya gum and the like; seed gums including guar gum, locust bean gum, psillium seed gum and the like; and lecithins such as soya lecithin. In addition to the above ingredients, compositions of the present invention may also include other ingredients such as various pharmaceutically acceptable additives available to those skilled in the art. These additives include binders, stabilizers, preservatives, flavorings, fiances, and pigments.

In another embodiment, the pharmaceutically acceptable carrier of the present invention includes van pen cream (cetyl alcohol, stearyl alcohol, steric acid, gllycerol monosterate, isopropyl myristate, soya lecithin, BHT alcohol 95%, simethicone, sodium hydroxide 30% solution, polyoxyl stearate, edetate disodium 5%, purified water, urea).

Other Pharmaceutical Compounds

In another aspect, the invention features a transdermal composition suitable for transdermal delivery, which includes a therapeutically effective amount of a pharmaceutical compound (e.g., a serotonin specific reuptake inhibitor, a mood stabilizing compound, a dopamine compound, a compound suitable for treating attention deficit hyperactivity disorder, a compound suitable for treating hypertension and akathisia, an analgesic compound, or a compound used in the treatment of impotence) and a pharmaceutically acceptable carrier suitable for transdermal delivery of the pharmaceutical compound.

As used herein, the term "pharmaceutical compound" includes compounds suitable for treating a targeted condition and capable of being delivered in active form, in vivo. Examples of pharmaceuticals include drugs, enzymes, chemical compounds, combinations of chemical compounds, biological macromolecules and analogs thereof. Examples of pharmaceutical compounds are described in detail below.

In one embodiment of the invention, the pharmaceutical compound is a serotonin specific reuptake inhibitor (SSRI). SSRIs are commonly prescribed for patients with diagnoses of mood disorders, some forms of anxiety disorder (particularly panic disorder), obsessive compulsive disorders, some forms of menopausal disorders, and eating disorders (especially bulimia nervosa). Examples of such SSRIs include sertraline (sold under the trade name Zoloft), paroxetine (sold under the trade name Paxil), fluoxetine (sold under the trade name Prozac), venlafaxine (sold under the trade name Effexor), and fluvoxamine (sold under the trade name Luvox).

In another embodiment of the invention, the pharmaceutical compound is a mood stabilizing medication, such as carbamazepine (sold under the trade name Tegretol) and valproic acid (sold under the trade name Depakote). These agents are used frequently in psychiatric practice as either augmentation medications (to render antidepressants more effective) or as anti-manic medications in the treatment of bipolar mood disorder. Mood stabilizing medications are also used in neurologic practice for the treatment of seizure disorders and for the treatment of certain pain disorders.

In yet another embodiment of the invention, the pharmaceutical compound is a compound used for treating Attention Deficit Hyperactivity Disorder (ADHD), one example of which is permoline, sold under the trade name Cylert. Permoline is a medication that is used in the treatment of Attention Deficit Hyperactivity Disorder in children and adults. It is practically insoluble in water, but soluble in ethylene glycol and lipids, making it a good candidate for transdermal administration

In a further embodiment of the invention, the pharmaceutical compound is a dopamine compound, used for treating Parkinson's disease, examples of which are pergolide, sold under the trade name Permax and bromocriptine mesylate, sold under the trade name Parlodel.

In yet another embodiment of the invention, the pharmaceutical compound is a compound used for treating hypertension and akathisia, one example of which is propranalol, sold under the trade name Inderal.

In yet a further embodiment of the invention, the pharmaceutical compound is a compound used in the treatment of impotence such as sildenafil, sold under the tradename Viagra. It is believed that transdermal administration of sildenafil may be useful, for at least some subjects, as compared to oral administration which has been found, in at least some situations, to be associated with gastrointestinal side effects.

Methods For Preparing The Transdermal compositions

Another embodiment of the present invention provides a method for preparing the above described transdermal compositions, by admixing a therapeutically effective amount of the amine containing compound having biphasic solubility, optimally an agent which enhances the activity of the amine containing compound, e.g., a muscle relaxant, optimally an anti-inflammatory compound with the carrier suitable for transdermal delivery of the amine containing compound.

In one embodiment of the present invention, a transdermal composition is prepared by dispersing or dissolving crushed tablets, capsules or other preparation(s) of the amine containing compound having biphasic solubility, the muscle relaxants, and the anti-inflammatory compounds, which were intended for oral delivery, in a gel formed of soya lecithin and isopropyl palmitate or isopropyl myristate, alcohol, or ethoxy diglycol. In another embodiment of the present invention, Pluronic gel, formed of Pluronic such as Pluronic F127, potassium sorbate and water is used.

In a particular embodiment of the present invention, a transdermal composition including a combination of doxepin with guaifenesin is useful for treating pain. It is believed that transdermal administration of such combination can be advantageous, for at least some patients, as compared to oral administration, because higher local pharmaceutical concentrations at the site(s), e.g., of injury, can be achieved yielding an improved therapeutic response without systemic side effects such as weight gain, drowsiness, gastrointestinal upset and/or other known side effects of these pharmaceuticals.

Methods For Use

In one embodiment, the invention feature methods for treating pain in a subject in which the subject is contacted with a transdermal composition including an amine containing compound having biphasic sole in an amount effective to treat pain in the subject; and a pharmaceutically acceptable carrier suitable for transdermal delivery of the amine containing compound to thereby treat pain in the subject In a preferred embodiment, the transdermal composition is applied to the skin of the subject as often as needed for the alleviation of pain. For example, the transdermal composition may be applied daily, weekly, monthly, yearly, for a length of time sufficient to alleviate pain.

Detailed examples of the preparation are provided below, along with examples of results obtained from transdermal administration to human patients. Preferably, a gel preparation is applied to the skin at the site or sites of pain. Patients can be evaluated by means of a structured evaluation form, e.g., completed at a frequency of at least one time per week. Evaluation of patients are for the present symptoms as well as any side effects from currently administered medications. This makes it possible to note changes on an ongoing basis.

Compositions of the invention can be self-administered doses in the form of a gel applied to the skin by the patient, or be implemented by providing a transdermal preparation in premeasured doses preferably in connection with an adhesive or other covering or patch so that the dosage may be administered e.g., by placing the adhesive patch on the skin of the patient. Although some embodiments of the invention have been described in connection with positioning the pharmaceutical gel on the arm of a patient, other positioning on the skin of a patient can also be used. Because, depending on the formulation, speed or duration of transdermal delivery may vary as function of skin location, in one embodiment the location of the skin to which the pharmaceutical is applied is selected so as to relatively increase or decrease the delay, speed, duration, or rate of delivery of the pharmaceutical, either with respect to a particular tissue or systemically.

For example, when a rapid rise in blood serum levels is desired, a placement which enhances delivery rate, such as behind the ear, can be used. When it is desired to enhance dose or delivery rate locally, the transdermal formulation may be positioned adjacent the desired treatment area. Membranes or matrices, such as a polymer matrix, may be used to limit or control delivery rates. In addition to transdermal gel or patch delivery, delivery of the transdermal or aerosol formulation can be achieved, e.g. by administration as nose drops, eardrops, eyedrops and/or suppositories.

In one embodiment, medications dispensed in transdermal gel form will be dispensed in unit doses, such as blister packs. The gel will be extruded from the blister pack, and rubbed on the administration site. The dosage will be adjusted by varying the number of unit dose applied. This will ensure accurate dosimetry and will avoid contamination of the gel.

Methods For Selecting A Compound Suitable For Treating Pain

In a further aspect, the invention features a method for selecting a compound suitable for treating pain in a subject. The method includes transdermally administering an amine containing compound having biphasic solubility to a subject; and determining whether pain is treated in the subject to thereby select a compound suitable for treating pain in a subject. In a preferred embodiment, the method can further include modeling the compound using a computer equipped with a three-dimensional chemical structure modeling program (e.g., Molecules-3D Professional Edition, version 2.60, copyright 1991–1998, Molecular Arts Corp., © 1994–1998 WCB/McGraw Hill); and determining whether the three-dimensional chemical structure of the compound possesses sufficient characteristics to be useful as a sodium channel blocker, thereby selecting a compound suitable for treating pain in a subject.

The effectiveness of the amine containing compound having biphasic solubility to treat pain can be tested in vitro or in vivo. An animal model for pain, e.g., such as the one described in Kral M. G. et al. (1999) *Pain* 81(1–2):15–24 can, for example, be used for testing such compounds.

This invention is further illustrated by the following examples which should not be construed as limiting. The contents of all references, patents and published patent applications cited throughout this application, as well as the Figures are incorporated herein by reference.

EXAMPLES

Example 1

One hundred grams of lecithin soya (granular) and 0.66 grams sorbic acid NF-FCC powder) were dispersed in 100 grams (117 milliliters (mL)) of isopropyl palmitate NF and allowed to stand overnight. Approximately 220 milliliters of lecitbin-isopropyl palmitate in a form of a liquid of a syrup consistency was formed.

Example 2

One hundred grams of lecithin soya (granular) and 0.66 grams sorbic acid (NF-FCC powder) is dispersed in 100 grams (117 milliliters) of isopropyl myristate NF and allowed to stand overnight. Approximately 220 milliliters of lecithin-isopropyl myristate in a form of a liquid of a syrup consistency was formed.

Example 3

A beaker was prepared by treasuring to a volume of 100 milliliters. It was considered important to measure the volume accurately rather than using beaker markings. An amount of Pluronic F127 NF (20 grams for a 20 percent gel, 30 grams for a 30 percent gel, 40 grams for a 40 percent gel) was mixed with 0.3 grams potassium sorbate NF. Refrigerated purified water was added in an amount sufficient to bring the volume to 100 milliliters. When all of the granules had been wet the gel was refrigerated. Solution took place upon cooling, taking 12 to 24 hours. The resulting 100 milliliters of Pluronic gel was kept refrigerated, since the gel will solidify at room temperature.

Example 4

Nine grams of carbamazepine in tablet form was ground in mortar and pestle. 4.3 milliliters of ethoxy diglycol was added and mixed to form a creamy paste. 13.2 milliliters of soya lecithin was added and mixed until smooth. The resulting 24 cc of solution was put into a 60 cc syringe. About 36 cc Pluronic F127 gel 20 percent (made according to Example 3) was placed in another syringe. The material was mixed well between syringes to yield 60 cc of carbamazepine organogel having a strength of 150 milligrams (mg) per milliliter. In some cases, the mixture was run through an ointment mill to reduce particle size.

Example 5

Sixty 100 milligram tablets of buproprion were ground and strained to form a fine powder. The buproprion powder was dissolved in 30 cc purified water, placed in a filter and washed with 10 to 20 cc purified water. The filtrate was used to make a 20 percent Pluronic gel using the procedures from Example 3, substituting filtrate for an equivalent volume of water, and stored in a refrigerator. Thirteen milliliters of soya lecithin was mixed with one-half the buproprion Pluronic gel and mixed between syringes to form a first batch. Thirteen milliliters of soya lecithin was mixed with the second half of the buproprion Pluronic gel and mixed between syringes to form a second batch. To each batch was added sufficient Pluronic gel F127 (made according to example 3) to yield a total of two 60 cc batches of buproprion HCl organogel having a strength of 15 milligrams per milliliter.

Example 6

600 milligrams of fluoxetine HCl (in the form of thirty 20 milligram capsules) was placed in a beaker and dissolved in approximately 18 cc of 95 percent ethyl alcohol. The solution was filtered through a filter funnel using fine filter paper. The residue was washed with 95 percent alcohol. The filtrate was heated, maintaining a temperature less than 85° C., to evaporate the alcohol to concentrate to 1 to 2 milliliters. 600 milligrams of isopropyl palmitate was combined with 600 milligrams of soya lecithin (granular), set aside and allowed to liquefy. Upon liquefaction, a thick syrupy consistency was obtained. 1.2 grams of the mixture was drawn into a 10 milliliter syringe and the alcoholic solution of fluoxetine HCl was drawn into another syringe. The two syringes were attached together with a Luer-Luer adapter and the gel was thoroughly mixed. All of the organogel was then transferred into one syringe and the empty syringe was disconnected. Sufficient quantity of 20 percent Pluronic F127 gel (formed as described in Example 3) was drawn into the empty syringe to make a total of 6 milliliters when added to the volume in the other syringe. A Luer-Luer adapter was attached and the contents of the two syringes was remixed until a smooth creamy mixture was obtained. All the mixture was transferred into one syringe, the empty syringe was removed and the Luer-Luer adapter was removed.

A Luer-oral adapter was attached to the mixture and transferred to six 1 milliliter oral syringes, was filled with 1 milliliter of the gel. In this way, each syringe contained five 20 milligram doses, or ten 10 milligram doses to yield a total of 60 doses of fluoxetine in lecithin organogel having a strength of 10 milligrams per 0.1 milliliters.

Example 7

Twelve 250 milligram tablets of nefazadone were crushed in a mortar and pestle and put through a strainer. 4.8 milliliters of ethoxy diglycol (8 percent) was added and mixed. In cases in which all particles were not dissolved, 2 milliliters of Pluronic were added and mixed. 13.6 milliliters of soya lecithin were added and mixed. The resulting mixture was put into syringes with a Luer adapter and mixed well. Sufficient Pluronic F127 gel, prepared according to Example 3, was added to achieve a volume of 60 cc and mixed well to yield 60 cc of nefazadone organogel having a strength of 50 milligrams per milliliter.

Example 8

Thirty 40 milligram tablets of paroxetine were crushed and run through a strainer, discarding green coating material. 4.8 milliliters of ethoxy diglycol was added to the powder and mixed in a mortar and pestle. Forty milliliters of Pluronic F127 gel 20 percent, formed according to Example 3, was added in graduated amounts to the powder and mixed until smooth using a spatula. 13.2 milliliters of soya lecithin was added and mixed well and the resulting material placed into syringes and sufficient quantity of Pluronic gel was added to bring the volume to 60 milliliters. In those such cases where particle size of the resulting material was too large, the cream was run through an ointment mill to yield 60 milliliters of paroxetine organogel having a strength of 20 milligrams per milliliter.

Example 9

Thirty 100 milligram tablets of sertraline were crushed into a fine powder and strained, discarding the yellow coating. Sufficient amount of Pluronic F127 gel 20 percent (formed according to Example 3) was added to achieve a volume of 38 milliliters and mixed well in a mortar and pestle until a smooth cream was achieved. This material was placed into syringes and mixed between the syringes to obtain a compact cream. 13.2 milliliters of soya lecithin was added and mixed well between the syringes using about 20 pumps. Sufficient quantity of Pluronic F127 gel 20 percent was added to yield 60 milliliters of sertraline gel having a strength of 15 milligrams per milliliter.

Example 10

Venlafaxine hydrochloride has a solubility in water of 572 mg/mL (adjusted to ionic strength of 0.2 M with sodium chloride). Forty-five 100 milligram tablets of venlafaxine were crushed and put through a strainer. The powder was dissolved in 15 cc purified water, the solution placed into a filter and washed with 10 cc purified water. The filtrate was used to make a 20 percent Pluronic gel using the procedures of Example 3 (substituting the filtrate for an equivalent amount of water) and placed into a refrigerator overnight 13.2 milliliters of soya lecithin were drawn into a syringe with a Luer loc. The venlafaxine Pluronic gel was drawn into another syringe coupled to the first syringe and mixed well. Sufficient Pluronic F127 gel was added to achieve a volume of 60 cc with a strength of 75 mg. per cc.

Example 11

15 grams of sodium valproate (Depakote) was ground in mortar and pestle. 4 mL of ethoxy diglycol was added and mixed well to form a creamy paste. 19.8 mL of soya lecithin was added and mixed until smooth. The resulting 24 cc of solution was put into 2 syringes with a Luer Loc and mixed well. The mixture was divided so that half is in each syringe. Using another 60 cc syringe, Pluronic 30% gel was added to each to bring each syringe to a volume of 45 mL.

Example 12

Paroxetine hydrochloride has a solubility in water of 5.4 mg/mL. Paroxetine (Paxil) gel was prepared, according to the procedures of example 8. A dosage of 40 mg per day was self-administered by a 59 year old male patient by application to the skin, for a period of at least 1 hour. No skin irritation was reported. After 210 days, blood was drawn and blood serum level of Paxil was determined to be 0 nanograms (ng) per mL, while typical reference levels are 49±26 ng/mL, indicating possible poor absorption or lab error. Clinical evaluation of the patient over a 210 day period of such transdermal administration indicated benefit to patient without GI side effects similar to that noted with oral preparation.

Example 13

Sertraline hydrochloride is slightly soluble in water and isopropyl alcohol and sparingly soluble in ethanol. Sertraline (Zoloft) gel was prepared, according to the procedures of example 9. A dosage of 100 mg per day was self-administered by a 54 year old female patient by application to the skin, for a period of at least 1 hour. No skin irritation was reported. After 19 days, blood was drawn and blood serum level of Zoloft was determined to be 5 ng/mL, while typical reference levels are 30–200 mg/mL indicating possible limited absorption or lab error.

Example 14

Fluoxetine hydrochloride has a solubility in water of 14 mg/mL. Fluoxetine (Prozac) gel was prepared, according to the procedures of example 6. A dosage of 20 mg per day was self-administered by a 54 year old female patient by application to the skin, for a period of at least 1 hour. No skin irritation was reported. After 7 days, blood was drawn and blood serum level of fluoxetine was determined to be 45 ng/ml, while the plasma level of the primary active metabolite norfluoxetin was also 45 ng/ml. There was evidence of patient benefit from the clinical evaluation.

Example 15

Carbamazepine is practically insoluble in water and soluble in alcohol and in acetone. Carbamazepine (Tegretol) gel was prepared, according to the procedures of example 4. A dosage of 400 mg per day was self-administered by a 55 year old male patient by application to the skin, for a period of at least 1 hour. No skin irritation was reported. After 120 days, blood was drawn and blood serum level of Tegretol was determined to be 4.6 micrograms (μg) per mL, while typical therapeutic levels are 4–10 11 μg/ml indicating good absorption. There were no GI side effects and the patient demonstrated clinical improvement.

Example 16

Carbamazepine (Tegretol) gel was prepared, according to the procedures of example 4. A dosage of 200 mg per day was self-administered by a 53 year old male patient by application to the skin, for a period of at least 1 hour. No skin irritation was reported. After 60 days, blood was drawn and blood serum level of Tegretol was determined to be 10.8 μg/mL, while typical therapeutic levels are 4–10 11 μg/mL indicating excellent absorption. There were no GI side effects and the patient demonstrated clinical improvement.

Example 17

Sertraline (Zoloft) gel was prepared, according to the procedures of example 9. A dosage of 50 mg per day was self-administered by a 53 year old male patient by application to the skin, for a period of at least 1 hour. No skin irritation was reported. After 63 days, blood was drawn and blood serum level of Zoloft was determined to be 23 ng/mL, while typical reference levels are 30–200 mg/mL. The patient demonstrated a good clinical response without GI side effects.

Example 18

Carbamazepine (Tegretol) gel was prepared, according to the procedures of example 4. A dosage of 200 mg per day was self-administered by a 47 year old male patient by application to the skin, for a period of at least 1 hour. No skin irritation was reported. After 91 days, blood was drawn and blood serum level of Tegretol was determined to be less than 0.5 μg/mL, while typical therapeutic levels are 4–10 μg/mL, indicating poor absorption, lab error, or patient non-compliance.

Example 19

Buproprion is highly soluble in water. Buproprion (Wellbutrin) gel was prepared, according to the procedures of example 5. A dosage of 100 mg per day was selfadministered by a 47 year old male patient by application to the skin, for a period of at least 1 hour. No skin irritation was reported. After 44 days, blood was drawn and blood serum level of Wellbutrin was determined to be less than 0.5 ng/mL, while typical therapeutic levels are 10–30 indicating poor absorption, lab error, or patient non-compliance.

Example 20

Fluoxetine gel was prepared, according to the procedures of example 6. Typically, a total daily adult dosage of fluoxetine as applied to the skin according to the present invention is between about 20 mg and 200 mg, more preferably between about 120 mg and about 200 mg. Dosages for non-adults and/or non-human mammals may need to be adjusted, e.g. proportionally to body weight A dosage of 20–60 mg per day was self-administered by 5 patients, including that of example 13 and also including a 44 year old male patient, a 53 year old female patient, a 47 year old male patient and a 36 year old female patient by application to the skin, for a period of at least 1 hour. No skin irritation or gastrointestinal side effects were reported. Clinical evaluation of the patients over a 30–180 day period of such transdermal administration indicated a clinical response ranging from complete remission of symptoms to moderate improvement.

Example 21

Fluoxetine gel was prepared, according to the procedures of example 6. A dosage of 80–160 mg per day was self administered by a 50 year old female by application to the skin, for a period of at least 1 hour. No skin irritation was reported. After 7 days at the 80 mg dosage level blood was drawn and the blood serum of fluoxetine was determined to be 34 ng/mL fluoxetine and 25 ng/mL norfluoxetine, while typical reference levels are 50–480 ng/mL, indicating good absorption. There was evidence of patient benefit from the clinical evaluation. The dosage was then increased to 160 mg per day and administered by the same method. After 7 days at the 160 mg dosage level blood was drawn and the blood serum level of fluoxetine was determined to be 90 ng/mL fluoxetine and 25 ng/mL norfluoxetine, indicating good absorption. There was evidence of increased patient benefit at this higher dosage level which correlated positively with the higher plasma level. The patient has been receiving the medication continuously for a period of 5 months.

Example 22

Fluoxetine gel was prepared, according to the procedures of example 6. A dosage of 80–160 mg/day was self administered by a 38 year old female by application to the skin, for a period of at least 1 hour. No skin irritation was reported. After 7 days at the 80 mg dosage level, blood was drawn and the blood serum level of fluoxetine was determined to be 25 ng/mL of fluoxetine and 25 ng/mL norfluoxetine. There was evidence of patient benefit from the clinical evaluation. The dosage was then increased to 160 mg per day and administered by the same method.

Example 23

Sertraline (Zoloft) gel was prepared, according to the procedures of example 9. A dosage of 50–200 mg per day was self-administered by 6 patients, including those of examples 12 and 16 and also including a 60 year old male patient, a 53 year old male patient, a 48 year old male patient, a 38 year old male patient and a 47 year old male patient, by application to the skin, for a period of at least 1 hour. No skin irritation or gastrointestinal side effects were reported. Clinical evaluation of the patients over a 7–90 day period of such transdermal administration indicated responses ranging from complete resolution of depression to no noticeable response.

Example 24

Carbamazepine (Tegretol) gel was prepared, according to the procedures of example 4. A dosage of 200–400 mg per day was self-administered by 6 patients, including those of examples 14, 15 and 17, and also including a 48 year old female patient, a 48 year old male patient and a 54 year old female patient, by application to the skin, for a period of at least 1 hour. No skin irritation or gastrointestinal side effects were reported. The clinical evaluation of the patients over a 30–300 day period of such transdermal administration indicated responses ranging from moderate improvement to no positive clinical response.

Example 25

Paroxetine (Paxil) gel was prepared, according to the procedures of example 8. A dosage of 20 mg per day was self-administered by the patient of example 12 as well as by a 15 year old female patient by application to the skin, for a period of at least 1 hour. No skin irritation was reported. Clinical evaluation of the patients over a 30–210 day period of such transdermal administration indicated equivocal clinical improvement of the depression which may (or may not) have been related to the transdermally administered Paxil.

Example 26

Five 150 mg tablets of amitriptyline were crushed and run through a strainer. The powder was put into syringes with a Luer Loc and mixed well with 2 mL ethoxy diglycol. About 6 mL Pluronic Gel 20% was added and mixed well. 6.6 mL Soya Lecithin was added and mixed well. This mixture was thinned to 30-mL total volume with Pluronic Gel 20% and mixed well. The resulting mixture having a strength of 25 mg/mL was placed in appropriate dispensing device.

Example 27

Amitriptyline (Elavil) gel was prepared, according to the procedure of example 26. A dosage of 25 mg per day was self-administered by a 47 year old male patient. Administration was by application to the skin, for a period of at least 1 hour. No skin irritation or gastrointestinal side effects were reported. Clinical evaluation of the patients over a 100 day period of such transdermal administration indicated an apparently good clinical response, comparable to that achieved with oral medication.

Example 28

Trazodone (Desyrel) gel was prepared, according to a procedure similar to that of example 7. A dosage of 50–150 mg per day was self-administered by 2 patients, including a 36 year old female patient and a 47 year old male patient. Administration was by application to the skin, for a period of at least 1 hour. No skin irritation or gastrointestinal side effects were reported. Clinical evaluation of the patients over a 42–90 day period of such transdermal administration indicated a good to excellent clinical response.

Example 29

Venlafaxine (Effexor) gel was prepared, according to a procedure similar to that of example 9. A dosage of 150–225 mg per day was self-administered by 2 patients, including a 54 year old female patient and a 55 year old male patient. Administration was by application to the skin, for a period of at least 1 hour. No skin irritation or gastrointestinal side effects were reported. Clinical evaluation of the patients over a 15–165 day period of such transdermal administration indicated a response ranging from no clinical improvement to mild clinical improvement.

Example 30

Propranolol (Inderal) gel was prepared, according to a procedure similar to that of example 8 to produce a gel having a strength of 40 mg of propranalol per mL of gel. A dosage of 80 mg per day was self-administered by 2 patients, including a 36 year old female patient and a 47 year old male patient. Administration was by application to the skin, for a period of at least 1 hour. No skin irritation or gastrointestinal side effects were reported. Clinical evaluation of the patients over a 100 day period of such transdermal administration indicated results comparable to those achieved with oral medication.

Example 31

Buproprion (Wellbutrin) gel was prepared, according to a procedure described in example 5. A dosage of 150–200 mg per day was self-administered by 3 patients, including that of example 18, and also including a 38 year old male patient and a 53 year old female patient. Administration was by application to the skin, for a period of at least 1 hour. No skin irritation or gastrointestinal side effects were reported. Clinical evaluation of the patients over a 5–45 day period of such transdermal administration indicated equivocal results.

Example 32

Valproic acid (Depakote) gel was prepared, according to a procedure similar to that of example 4. A dosage of 1000 mg per day was self-administered by a 38 year old male patient. Administration was by application to the skin, for a period of at least 1 hour. No skin irritation or gastrointestinal side effects were reported. Clinical evaluation of the patients over a 30 day period of such transdermal administration indicated results comparable to those achieved with oral medication.

Example 33

Valproic acid (Depakote) gel was prepared according to the procedure of example 11. A dosage of 500–1000 mg was self administered by two male patients, ages 41 and 49. Administration was by application to the skin, for a period of at least one hour. Significant skin irritation occurred with one patient, but no gastrointestinal side effects were reported. Clinical evaluation of the patients over a period of two months revealed improvement, but upon longer term follow-up it appeared that other factors may have been responsible. After 28 days, blood was drawn and a serum valproic acid level of 26 μg/mL was obtained for the 49 year old patient (while taking 250 mg twice daily), with a therapeutic reference range of 50–150 μg/mL. This indicated poor to fair absorption, and the dosage was raised to 500 mg twice daily, with a further improvement in clinical response. The 41 year old patient reported a good clinical response to an initial dosage of250 mg adored twice daily, but a serum valproic acid level of only 1 μg/mL was obtained. The dosage was increased to 500 mg twice daily, and a similar serum valproic acid level was obtained. The disparity between the clinical response and the plasma level might be explained either by laboratory error or placebo effect.

Example 34

A gel containing reboxetine (sold under the trade name Edronax) is prepared according to a procedure similar to that described in example 5 but using reboxetine in place of boproprion. The resulting mixture will be self administered by patients by application to the skin for a period of at least 1 hour. No skin irritation or gastrointestinal side effects are expected. Clinical evaluation of patients over a 5–45 day period of such transdermal administration is expected to indicate a good response to treatment.

Example 35

Nefazodone (Serzone) gel was prepared, according to a procedure described in example 7. A dosage of 100 mg per day was self-administered by a 61 year old (male, female) patient. Administration was by application to the skin, for a period of at least 1 hour. No skin irritation or gastrointestinal side effects were reported. Clinical evaluation of the patients over a 21 day period of such transdermal administration indicated a good response to treatment.

Example 36

1 gram of permoline tablets are crushed in a mortar and then dissolved in propylene glycol, just sufficient to effect dissolution. 3 mL of propylene glycol or 95% ethyl alcohol is added to form a paste. 6.6 mL soya lecithin is added to the mixture in the mortar. The mixture is placed in two syringes with a Luer Loc and mixed thoroughly. Each syringe is filled to 30 mL Pluronic F127 20% gel and mixed between syringes to produce a mixture having a strength of 33 mg/mL. The mixture is put in an appropriate dispensing device.

Example 37

A 16-year-old female with an established diagnosis of Attention Deficit Disorder had been treated successfully with oral pemoline (Cylert) for about 6 months. To potentially decrease the risk of liver damage associated with long-term use, permoline prepared according to the procedure of example 36 will be administered transdermally, by application to the skin in the post auricular region for a period of at least one hour, at two sites, twice daily. No skin irritation is expected. The clinical results are expected to be comparable to those obtained with the oral medication, although the dosage may have to be adjusted upwards to achieve adequate plasma levels, and more time may be required to achieve satisfactory plasma levels.

For psychiatric patients, some have received two or more psychopharmaceuticals, and in some cases, two or more of the above examples describe different evaluations for the same period of administration of a psychopharmaceutical agent.

Of the patients who have received prescriptions for one or more of the medications as described in the examples above, each had previously demonstrated a significant intolerance to oral administration of one or more medications, prior to instituting transdermal administration. The laboratory measures of plasma blood levels described above for transdermally administered fluoxetine and carbamazepine are believed to demonstrate good absorption transdermally using lecithin organogel matrix as the vehicle. Valproic acid and sertraline do not appear to be absorbed well or reliably. Valproic acid appears to cause skin irritation in some patients necessitating discontinuation. Both the laboratory measure of Buproprion and the patient clinical responses indicated poor or equivocal absorptions and results. Patient tolerance of transdermal administration has been good to excellent. Patients in the example above who suffered very severe GI side effects using oral preparations were more tolerant of the inconvenience of rubbing on the gel than were patients who had experienced only mild to moderate side effects. In general, more highly motivated and treatment-compliant patients also had a higher rate of sustained compliance.

Patients in the examples above were evaluated by means of a structured evaluation form depicted in FIG. 1, which was completed at a frequency of at least one time per week for each patient receiving transdermal medication according to the present invention. The patients were evaluated both for all present psychiatric symptoms as well as any side effects from currently-administered medications. In general, it is believed that patients with the most clear cut and uncomplicated diagnosis of major depression experienced the best results. In general, patients with severe personality disorders or with concealed substance abuse disorders did less well.

Example 38

1800 mg of gabapentin in powder form is dissolved with 1 mL propylene glycol in syringes with a Luer Loc. 6.6 mL of Soya lecithin is added and mixed thoroughly between syringes. The resulting material is placed in a device for dispensing measured amounts.

Example 39

Gabapentin mixtures of 2% and 4% will be prepared by substituting 1200 mg gabapentin or 600 mg gabapentin in place of 1800 mg gabapentin, in example 38.

Example 40

Gabapentin, prepared according to Example 38 or 39, will be combined with either 3% or 5% Lidocaine in varying ratios.

Example 41

4% gabapentin, prepared according to Example 38 or 39, will be combined with 7% carbamazepine and 7% amitriptyline.

Example 42

2% gabapentin, prepared according to Example 38 or 39, will be combined with 2% carbamazepine and 1% Piroxicam, which is expected to yield better penetration into muscle tissue.

Example 43

Gabapentin, prepared according to Example 38 or 39, in concentrations ranging from 2%–6% will be combined with clonidine in concentrations between 0.2% and 0.3%.

Example 44

A 56-year-old woman had pill upper and lower extremity spasms as a result of spastic quadriparesis resulting from an injury. Oral gabapentin, an anticonvulsant, had been administered previously, but had caused a "drugged" feeling, one of the commonly reported side effects with this agent It was believed that use of transdermal gabapentin might provide local relief by achieving high local tissue concentrations near the site of administration without correspondingly elevated blood plasma levels. It is known that other anticonvulsants, such as carbamazepine, are useful in reducing neurogenic pain. Gabapentin's solubility in water exceeds 10%, making systemic absorption less likely. Gabapentin prepared according to the procedure of example 38 was self-administered by application to the skin in the area of pain. The patient reported moderate relief of spasms over a period of one week, with no systemic side effects and no report of skin irritation.

Example 45

Six grams of amitriptyline powder was placed in 40 milliliters of Pluronic F127 33% gel and placed under refrigeration to dissolve. Two milliliters of ethoxy diglycol was added to 4.8 grams of carbamazepine and mixed to form a smooth paste. 16.4 grams of soya lecithin was added to the resulting paste and mixed well. The dissolved amitriptyline composition was added to the carbamazepine composition and sufficient Pluronic F127 20% was added to make 120 milliliters and the resulting composition was mixed well to yield a composition having 5% amitriptyline and 4% carbamazepine.

Example 46

6 grams of doxepin was added to 20 milliliters Pluronic 33% F127 and put into a refrigerator to dissolve. 24 grams of ketoprofen and 12 grams of guaifenesin was added to 10 milliliters of 95% alcohol and mixed well. 26.4 milliliters of soya lecithin was added and mixed well and the doxepin composition was mixed with the ketoprofen/ guaifenesin composition. The resulting mixture was added to sufficient Pluronic 33% to yield 120 milliliters. The resulting composition was mixed well to yield a composition having about 20% ketoprofen, 5% doxepin and 10% guaifenesin

Example 47

6 grams of doxepin was added to 26 milliliters Pluronic 33% and refrigerated to dissolve. 2 milliliters ethoxy diglycol was added 4.8 grams carbamazepine and mixed. The resultant mixture was added to 24 grams ketoprofen and six milliliters alcohol and the result was mixed well. 26.4 milliliters soya lecithin was added to the ketoprofen composition and mixed well. The doxepin composition was mixed with the carbamazepine/ketoprofen composition and sufficient Pluronic 33% was added to yield 120 milliliters. The resultant composition was mixed well to yield a composition having about 20% ketoprofen, 4% carbamazepine and 5% doxepin.

Example 48

0.15 grams sildenafil was crushed and strained and dissolved in 5 milliliters Pluronic 20% F127 and mixed between syringes. 2.2 milliliters of soya lecithin was added and mixed. Sufficient Pluronic 20% was added to yield 10 milliliters and the resultant composition was mixed well to yield a composition having the strength of about 15 milligrams per milliliter.

Example 49

A mixture of Sildenafil 15 mg/ml was applied to the penis and scrotum of a 51 year old male. An immediate and strong erection resulted with sexual stimulation, without any irritation or burning. It is believed the composition will possess the therapeutic results claimed for orally administered Sildenafil, without any time delay, without any systemic GI side effects, and possibly without the degree of drug interaction with nitrates used in cardiac disease. It is believed that this will contribute both to the convenience of use of the pharmaceutical and to its safety.

Example 50

Compositions according the examples 45 through 47, 53, 55 were transdermally applied to numerous patients, for the purpose of treating pain including as described in other examples herein, with the results summarized in Table I below. The meaning of certain entries in Table I is indicated in Table II below. Blank results indicate no treatment at the pertinent site for this patient. Where a given line of Table I shows more than one site, one "best" (greatest pain relief) result if shown in bold.

TABLE I

Patient	Age	Gender	Surgery	Pain	Medication Wt % in lecithin organogel Ketoprofen	Gabapentinm	Piroxicam	doxepin	carbamazepine	amitriptyline	guifenesin	Duration	Result (Best result in Bold) shoulder	back	neck	elbow	Knee	Wrist	Arm	Ankle Ankle	Hip Hip	Leg Leg
A	50	2	2	3	10	3	4					2		0								
B	61	1	1	3				5				4				2.0						
B	61	1	1	3					4			12	2.0	2.0	2.0	2.0						
B	61	1	1	3	10	4	3					6							3.0			
C	41	2	1	2					4	5		2		1.0								
D	53	1	2	1	10	4	1					1			.0							
E	57	2	2	3	10	4		5			1	1		2.0			1.5					1.0
E	57	2	2	3	10	4	3					2	1.0	2.0				1.0				
F	38	2	2	3				10	5		5	2	2.0			3.0						
F	38	2	2	3		4			4			8	2.0						1.5			
F	38	2	2	3	10	4	1		4			4	2.0						1.0			
G	39	1	1	2	20	4		5	4			6					3.0					
H	61	1	1	3	10	4	3					4		2.0								
I	49	1	1	3	10	4	3					12						2.0				
I	49	1	1	3				5	5		10	1						1.0				
I	49	1	1	3			4		4			2						3.0				
J	54	1	1	3					5	5		2		1.5								
K	40	1	2	3				5				6	4.0									
K	40	1	2	3	10		3	6				4	1.0									
L	55	2	2	2	10	4	3					8	1.0						.0			
L	55	2	2	2				5				6		3.0					.0	1.5	2.0	
M	38	1	2	1				4	5			2	1.5					.0				
N	47	2	1	2	20	2				5		3	3.0	3.0	4.0		1.0					
N	47	2	1	2	10	4	1					2.0	.0		2.0			2.0				
O	57	2	1	2	20	4		5				24	2.0		3.0							
O	57	2	2	2	10	4	3					24	1.0	.0								
P	51	2	2	2	15	5		5				2		4.0								
Q	51	2	1	2	20			5			10	1		2.0								
R	35	1	1	2					4	5		0					1.5					
R	35	1	1	2	10	4	1					1					.0					
S	55	1	1	2	10	4	1					16		1.0								
T	50	2	2	1	10	4	1					16					2.0	1.0		2.0		
U	45	1	2	2	10	4	3					2		.0								
V	57	2	1	3					6			8					3.0					
V	57	2	1	3	10	4	1					3					1.0					
W	35	1	2	1	10	4	1					8		1.0								
X	46	1	1	3	10			5	4			8		2.0	2.0	2.0						
Y	48	1	1	3				5				4	2.0	2.0								
Y	48	2	1	3	10	4	1					4				1.5		1.5			.0	
AA	53	2	2	1	10	4	1					4		1.0								
BB	58	2	1	3	20	4			4	Hand	1	8		2.0							2.0	

TABLE I-continued

					Medication Wt % in lecithin organogel								Result (Best result in Bold)									
Pa-tient	Age	Gen-der	Sur-gery	Pain	Keto-profen	Gabapentinm	Piroxicam	doxepin	carbamazepine	amitriptyline	guifenesin	Dur-ation	shoul-der	back	neck	elbow	Knee	Wrist	Arm	Ankle Ankle	Hip Hip	Leg Leg
CC	59	1	1	2				5				2			2.0		2.0			**2.0**		
CC	59	1	1	2	10	4	1					20		1.0	2.0		**3.0**	2.0				
CC	59	1	1	2	10	4	5					1					**3.0**			3.0		
DD	58	1	1	2	10	4	3					12			1.0			**2.0**				
EE	45	2	2	2	10	4	3					24	**1.5**		1.0							
FF	44	2	1	3	10	4	3					20	**2.0**									
GG	35	1	1	3	20	4						4				1.0		**1.0**				
GG	35	1	1	3				5				8				1.0		**1.0**				
GG	35	1	1	3	20				5	5		2						**.0**				
GG	35	1	1	3	20			5	5			2						**2.0**				
GG	35	1	1	3		5		5			10	2				1.0		**2.5**				
HH	40	1	2	2	10	4	3					4		**1.0**			1.0					
II	40	1	2	3				5				8		**1.5**								1.5
II	40	1	1	3	10	4	3	5				8		**2.0**								
JJ	45	1	2	2	10	4	3					2		**1.0**								
KK	37	2	2	2	10	4	1					8		**1.0**								
LL	54	1	1	3	10	4	3					6					**1.0**					
LL	54	1	1	3					4	5		2					**.0**					
MM	42	2	1	3					4			8		.0	**4.0**				2.0			1.0
MM	42	2	1	3	10		3		4			12		**.0**								
MM	42	2	1	3				5				4							2.0			1.0
NN	41	1	2	2	10	4	3					2		**.0**								

TABLE II

Gender: 1 = male	2 = female	
Surgery: 1 = one or more surgeries	2 = no surgeries	
Pain: 1 = mild	2 = moderate	3 = severe-sufficient to produce observed tears

Duration: length of treatment trial in weeks
Result: 0 = no benefit
1 = mild benefit
2 = moderate benefit (greater than 25% pain reduction)
3 = major benefit (greater than 40–45% pain reduction)
4 = almost complete relief (greater than 80% pain reduction)

Certain results drawn from the information of Table I are summarized in Table III and IV.

TABLE III

		Percent reported pain relief					
Site	N (Number of data points)	None	Mild	Mild-moderate	moderate	Major	Total
Wrist	13	16.7	33.3	8.3	41.7		
Shoulder	14	7.1	21.4	14.3	42.9	7.1	7.1
Elbow	5		40	20	20	20	
Back	25	24	32	8	28	8	
Arm	7	28.6	14.3	14.3	28.6	14.3	
Neck	11	9.1	18.2		45.5	9.1	18.2
Knee	13	15.4	46.2	15.4	7.7	15.4	

TABLE IV

		(percent reported pain relief)					
	N	None	Mild	Mild-moderate	moderate	Major	Total
Best result without tricyclic	36	16.7	36.1	8.3	27.8	8.3	2.8
Best result with any tricyclic	20	10	10	20	35	15	10
Either tricyclic -sole agent	7		14.3	14.3	42.9	14.3	14.3
Best result with ketoprofen gabapentin piroxicam	25	16	44	4	28	8	
Best result without doxepin	43	18.6	32.6	14	23.3	7	4.7
Best result with doxepin	13		7.7	7.7	53.8	23.1	7.7

Example 51

A 51 year old female administered a composition prepared according to example 46, containing 20% ketoprofen, 5% doxepin, and 10% guaifenesin to her back for a period of 2 weeks. She reported moderate pain relief, lasting several hours, after each application. She reported no skin irritation nor any other side effects. Oral medications had produced no relief, and had caused significant GI side effects.

Example 52

A 34 year old man administered a composition containing 20% ketoprofen, 4% carbamazepine, and 5% doxepin to a very severely scarred wrist that had undergone 4 surgeries for carpel tunnel syndrome. He reported moderate pain relief, lasting for several hours after each application. No other treatment, including opiate oral pain medication, had been effective in providing even minor pain relief.

Example 53

24 grams ketoprofen and sufficient guaifenesin to result in a 10% final guaifenesin concentration, was mixed well with 10 milliliters 95% alcohol. 1200 mg gabapentin was dissolved in one ml propylene glycol in a syringe with a luer loc. 26.4 ml of soya lecithin was added to the ketoprofen-guaifenesin-alcohol mixture and mixed well. The resulting mixture was added to the gabapentin-propylene glycol mixture and mixed well. 4.8 gm of carbamazepine was combined with the resultant combination and mixed well to form a smooth paste. The resulting paste was combined with the ketoprofen-guaifenesin-alcoholgabapentin mixture and mixed well with sufficient pluronic to yield 120 ml of a composition containing ketoprofen 20%, carbamazepine 4%, gabapentin 4%, guaifenesin 10%.

Example 54

A 58 year old female with damage to her cervical spinal cord with a resultant spastic quadreparesis reported moderate relief of both pain and muscle spasms when she applied a mixture prepared generally according to example 53, containing ketoprofen 20%, carbamazepine 4%, gabapentin 4%, guaifenesin 10% for a period of 8 weeks to her back and hip. She had been unable to tolerate both oral carbamazepine and oral gabapentin because of systemic side effects, including skin rash with the carbamazepine and dizziness and sedation with the gabapentin. She experienced no skin irritation nor other side effects with the transdermal formulation.

Example 55

Six grams of doxepin powder combined with 26 milliliters pluronic and placed in the refrigerator until dissolved. 1200 mg gabapentin was mixed with 1 ml propylene glycol and placed in a syringe with luer lock. 6.6 ml of soya lecithin was added and mixed well between syringes. 24 gm of ketoprofen and 8 milliliters alcohol was mixed well between two syringes with luer loc. The doxepin mixture was mixed well with the gabapentin mixture and subsequently the ketoprofen mixture was added and mixed well. Sufficient pluronic 20% (about 54 ml) was added to yield 60 ml of a composition having about 20% ketoprofen, 4% weight percent gabapentin and 5% weight percent doxepin.

Example 56

A 57 year old female applied a mixture, prepared generally according to example 55, containing ketoprofen 20%, gabapentin 4%, and doxepin 5% for a period of 8 weeks to her neck and reported major relief. She applied the same mixture to her shoulder and reported moderate relief. A mixture that substituted piroxicam for the doxepin produced only mild shoulder relief.

Example 57

A 35 year old man with a history of knee injury with vascular compromise and 3 surgeries applied a mixture, prepared generally according to example 45, containing 4% carbamazepine and 5% amitriptyline to his knee, and reported mild to moderate pain relief, without skin irritation nor other side effects.

Example 57A

A 41 year old woman with history of back surgery applied a mixture, prepared generally according to example 45, containing 4% carbamazepine and 5% gabapentin to her back for a period of 2 weeks. She reported mild pain relief.

Example 58

A 53 year old man with a history of two total bilateral knee replacements applied a mixture, prepared generally according to example 45, containing 4% carbamazepine and 5% amitriptyline to both knees for a period of 4 weeks. He reported no pain relief.

Example 58A

A 54 year old man with a history of 7 back surgeries applied a mixture, prepared generally according to example 45, containing 4% carbamazepine and 5% amitriptyline to his back for a period of 2 weeks. He reported mild to moderate pain relief, over and above that he was receiving from a transdermal opiate medication (Duragesic). He reported no side effects, and specifically no skin irritation.

Example 59

A 38 year old man with a history of shoulder strain applied a mixture, prepared generally according to example 45, containing 4% carbamazepine and 5% amitriptyline to his shoulder for a period of 2 weeks. He reported mild to moderate pain relief, and reported no skin irritation nor other side effects.

Example 61

Sufficient carbamazepine and gabapentin was added to a combination of soya lecithin and pluronic to yield a lecithin organogel out 4%/a carbamazeprine 5% gabapentin.

Example 62

A 42 year old woman with a history of 3 back surgeries and cervical degenerative disc disease applied a mixture, prepared according to example 61, containing 4% carbamazepine and 5% gabapentin to her neck and reported total relief of pain. She reported no side effects, and no skin irritation. She noted the complete and rapid resolution of a migraine like headache at the same time. Administration of the same mixture to her arm and her wrist, affected by a diagnosed condition of reflex sympathetic dystrophy, yielded moderate pain relief.

Example 63

3.6 grams gabapentin was dissolved with 5.4 ml ethoxy diglycol using a mortar and pestle. 9.6 grams ketoprofen and 2.7 grams piroxicam were added and the resultant composition mixed well. 19.8 milliliters soya lecithin was added and resultant mixture mixed well and added to a sufficient quantity of 20% pluronic gel to yield 90 milliliters of a composition having about 10 percent ketoprofen, 4% gabapentin and 3% piroxicam.

Example 64

3.6 grams gabapentin was dissolved with 5.4 ml ethoxy diglycol using a mortar and pestle. 9 grams ketoprofen and 0.9 grams piroxicam were added and mixed well. 19.8 milliliters soya lecithin was added to the resultant mixture and mixed well. Sufficient amount of pluronic gel 20% was added to yield 90 milliliters of a composition having approximately 10% ketoprofen, 4% gabapentin and 1% prioxicam.

Example 65

12 g doxepin was mixed with 50 ml Pluronic F 127 33% and placed in a refrigerator to dissolve. 12 g gabapentin was dissolved in 9 ml ethoxy diglycol and mixed to form a smooth paste. 52.8 ml of soya lecithin was added and mixed well. The doxepin/Pluronic mixture was added and mixed well. Sufficient quantity of Pluronic F 127 20% was added to produce 240 ml of a composition having about 5 wt % gabapentin and 5 wt % doxepin.

Example 66

A 36 year old man with a knee injury involving joint surface damage and vascular comprise applied a mixture, prepared generally according to Example 65 to his knee several times per day. He reported moderate to major (40%) relief of pain that persisted for 4 to 6 hours. An earlier trial of carbamazepine-amitriptyline gel produced no relief when applied to his knee.

Example 67

6 gm doxepin was mixed with 18 ml of Pluronic 33% to and placed in a refrigerator to dissolve. 6 gm gabapentin was ground in a mortar and pestle to a fine powder, added to 6 ml ethoxy diglycol and mixed to form a smooth paste. 12 gm guaifenesin was added and mixed well. 26.4 ml soya lecithin was added and mixed well. The doxepin/Pluronic mixture was added and mixed well. Sufficient quantity of Pluronic gel (25.2 ml of 33% Pluronic, although 30% or 20% Pluronic can be used), was added to produce 120 ml of a composition having about 5 wt % gabapentin, about 5 wt % doxepin and about 10 wt % guaifenesin.

Example 68

A 55 year old woman with a back and shoulder injury sustained as a nursing care provider applied a mixture, prepared generally according to Example 67, to her back three times per day for a period of two weeks and achieved major relief She applied the same mixture to her hip and leg and reported moderate to major relief A mixture containing only doxepin provided only moderate relief to her back, and mild to moderate relief to her hip and leg. A mixture that contained only ketoprofen, gabapentin and piroxicam provided only mild relief to her back.

Example 69

A 59 year old woman with cervical and back strain applied a mixture, prepared generally according to example 51, but without steps involving ketoprofen) containing about 5 wt % doxepin and about 10 wt % guaifenesin, to her neck for a period of two weeks, two to four times per day, and achieved total relief. She applied the same mixture to her back and achieved major to total relief.

Example 70

4.5 gm of doxepin HCl was dissolved using 2.5 ml 95% alcohol and mixed well between syringes. It is also possible to mix the doxepin with 5 ml Pluronic 20% and place in a refrigerator to dissolve. Sufficient quantity of 20% Pluronic F127 was added to produce 90 ml of a composition having about 5 wt % doxepin. Preferably this and other disclosed compositions are protected from light.

Example 71

A 61 year old man with injuries to his back, neck and arm applied a mixture (prepared generally according to Example 70) to his neck four times per day and achieved major relief. He applied the same mixture to his elbow and achieved moderate relief.

Example 72

A formulation of 7% antidepressant and about 10% muscle relaxant was prepared by dissolving 3.15 g of trimipramine and 4.5 g of guaifenesin in a mixer jar using 2.7 mL of ethoxy diglycol. About 9.9 mL of soya lecithin was added and the mixture was mixed well. Sufficient quantity of Pluronic F127 NF (20%) to make total volume of about 45 mL was added and mixed well.

Example 73

A gel formulation of 30% NTHE was prepared from 36 g of celecoxib, 7.2 mL of ethoxy diglycol, 26.4 mL of soya lecithin and sufficient quantity of Pluronic F127 NF (20%) to make total volume of 120 mL.

Example 74

A gel formulation containing about 7% antidepressant and about 13% muscle relaxant was prepared from 14.4 g of doxepin, 31.2 g of guaifenesin, 12 mL of ethoxy diglycol, 52.8 mL of soya lecithin and sufficient quantity of Pluronic F127 NF (33%) to make total volume of 240 mL.

Example 75

A gel formulation containing 5% antiepileptic was prepared from 6 g of lamotrigine, 6 mL of ethoxy diglycol, 26.4 mL of soya lecithin and sufficient quantity of Pluronic F127 NF (33%) to make total volume of 120 mL.

Example 76

A gel formulation containing 10% adrenergic agonist was prepared from 12 g of crushed tizanidine, 6 mL of ethoxy diglycol, 26.4 mL of soya lecithin and sufficient quantity of Pluronic F127 NF (33%) to make total volume of 120 mL.

Example 77

A gel formulation containing 10% muscle relaxant was prepared from 12 g of crushed metaxalone, 6 mL of ethoxy diglycol, 26.4 mL of soya lecithin and sufficient quantity of Pluronic F127 NF (33%) to make total volume of 120 mL.

Example 78

A gel formulation containing 10% muscle relaxant was prepared from 12 g of crushed carisoprodol, 6 mL of ethoxy diglycol, 26.4 mL of soya lecithin and sufficient quantity of Pluronic F 127 NF (33%) to make total volume of 120 mL.

Example 79

A gel formulation containing 10% methocarbamol was prepared from 12 g of crushed methocarbamol, 6 mL of ethoxy diglycol, 26.4 mL of soya lecithin and sufficient quantity of Pluronic F127 NF (33%) to make total volume of 120 mL.

Example 80

A gel formulation containing 10% muscle relaxant was prepared from 12 g of crushed dantrolene sodium, 6 mL of ethoxy diglycol, 26.4 mL of soya lecithin and sufficient quantity of Pluronic F127 NF (33%) to make total volume of 120 mL.

Example 81A

A gel formulation containing 7% antidepressant, 10% muscle relaxant was prepared from 8.4 g of crushed doxepin, 12 g of chlorzoxazone, 6 mL of ethoxy diglycol, 26.4 mL of soya lecithin and sufficient quantity of Pluronic F127 NF (33%) to make total volume of 120 mL.

Example 82

A series of experiments in human subjects were performed using various combinations of pharmaceuticals. The results are indicated in FIG. 2.

Values of pain relief as rated by the patients are provided for each body part for which the medication was administered. The scale used in FIG. 2, is as follows:

0 = None	no benefit or equivalent benefit
1 = Mild	less than 15% pain reduction
1.5 = Mild-moderate	15–33% pain reduction
2.0 = Moderate	25–25% pain reduction
2.5 = Moderate-major	33–45% pain reduction
3.0 = Major	45–60% pain reduction
3.5 = Major-total	60–80% pain reduction
4.0 = Total	greater than 80% pain reduction

For each body part and for each percentage composition of each compounded medication, the individual ratings as well as a mean, which is the statistical mean of the values given according to the scale listed above, are provided. For example, 3 patients were administered doxepin 5% to their back, and the mean level of relief was 2.333. By contrast, 13 patients received the 5%/10% doxepin-guaifenesin combination, and their mean level of pain relief was 2.885. Results for 7/10 and 10/10 compositions of doxepin guaifenesin are also given, and the mean for the entire sample of dox-guai in all combinations is provided at the end of the section, namely 2.722.

The abbreviations used in FIG. 2 are as follows:

Abbreviations	Generic Pharmaceutical names
c-dox-gu	carbamazepine doxepin guaifenesin
c-gab-do	carbamazepine gabapentin doxepin
carb	carbamazepine
carb-ami	carbamazepine amitriptyline
carb-gab	carbamazepine gabapentin
dox	doxepin
dox-chl	doxepin chlorzoxazone
dox-guai	doxepin guaifenesin
g-dox-gu	gabapentin doxepin guaifenesin
gab-dox	gabapentin doxepin
k-ca-dox	ketoprofen carbamazepine doxepin
k-car-pi	ketoprofen carbamazepine piroxicam
k-dox-ch	ketoprofen doxepin chlorozoxazone
k-dox-gu	ketoprofen doxepin guaifenesin
k-dox-pi	ketoprofen doxepin piroxicam
k-g-do-g	ketoprofen gabapentin doxepin guaifenesin
k-gab	ketoprofen gabapentin
k-gab-ami	ketoprofen gabapentin amitriptyline
k-gab-do	ketoprofen gabapentin doxepin
k-gab-gu	ketoprofen gabapentin guaifenesin
k-gab-pi	ketoprofen gabapentin piroxicam
k-pi	ketoprofen piroxicam
la-li-gu	lamotrigine lidocaine guaifenesin
lam-chl	lamotrigine chlorzoxazone
n-dox-ch	naproxen doxepin chlorzoxazone
naproxen	naproxen
tri-chl	trimipramine chlorzoxazone

Based on the results described herein, doxepin appears to be an effective pain relief medication when administered transdermally and appears to be substantially free of side effects when administered transdermally as described herein.

Doxepin appears to provide about three times the positive response rate compared to at least some other pharmaceutical agents described herein, regardless of whether such other pharmaceutical agents are administered singly or in combination. Doxepin appears to be substantially more effective than amitriptyline as a pain, e.g., neuropathic pain agent when administered transdermally. This appears to be true regardless of whether doxepin is administered as a single agent or is administered in combination with other pharmaceuticals as described herein.

Carbamazepine appears to provide positive effects as a pain, e.g., neuropathic pain agent, at least in properly selected patients. Carbamazepine appears to cause a rash in at least some patients, requiring its discontinuation.

These side effects appear similar to those that are noted for oral administration of carbamazepine. Gabapentin appears to be free of side effects when administered transdermally. Although some patients appear to derive some benefit from a combination of transdermally administered ketoprofen, gabapentin, and prioxicam, the effect appears to be relatively weak compared to the effect provided by doxepin.

Guaifenesin appears to provide benefit as an adjunctive treatment, of painful spasticity. For the patient population described herein, amitriptyline appeared to offer limited pain relief when administered transdermally. It appears that combining gabapentin with doxepin may offer some additional benefit. The addition of guaifenesin to doxepin may be of particular value when painful spasticity is present.

In view of the above, the invention provides treatment to patients for whom oral delivery is suboptimal, such as patients who experience gastrointestinal or other side effects, patients who experience poor absorption for orally delivered pharmaceuticals and/or patients who benefit from delivery over an extended period or a relatively rapid delivery or higher rate of increase of plasma levels. The present invention achieves delivery of therapeutic amounts of pharmaceuticals, for at least some patient populations, substantially without skin irritation, gastrointestinal or other side effects associated with orally-delivered pharmaceuticals, especially psychopharmaceuticals, and yields clinical benefits comparable to or greater than those received by patients to whom corresponding pharmaceuticals were administered orally. In view of the above reasons, particularly effective pain medications are those described in examples 65, 67, 69 and 70.

A number of variations and modifications of the invention can also be used. It is believed that blood plasma levels may be increased by providing for two or more transdermal applications per day and/or applying a transdermal composition to two or more sites.

In at least one case, application of a Prozac gel formulation twice daily appeared to approximately double the plasma level. It is believed that an approach such as applying a Prozac gel formulation twice daily to two sites will yield middle range therapeutic levels of about 140–250 ng/ml. At least partially on the basis of the results described herein for fluoxetine, it is believed olanzapine (sold under the trade name Zyprexa) or a fluoxetine/olanzapine mixture in a lecithin organogel will prove useful.

Other types of psychotropic or psychopharmaceutical medications for which the described transdermal delivery may be used including psychostimulant medications. One example of a psychostimulant medication is Methylphenidate (sold under the trade name Ritalin) used in the treatment of attention deficit hyperactivity disorder (ADHD). Methylphenidate typically has a 2–4 hour duration of action necessitating frequent dosing of a patient which is particularly difficult to accomplish with children in school. It is believed that by using transdermal administration, it will be possible to achieve an extension of effective dosing throughout the day, eliminating the need for frequent oral medication administration. It is believed that transdermal administration will also eliminate peaks and valleys of blood plasma levels which, it is believed, will be more clinically effective. It is believed similar results will be obtained with other pharmaceuticals, for example, Dextroamphetamine (under the trade name Dexedrine) although it is believed the need is less acute since a time release "spansule" form of the medication is available which typically has a 5–6 hour duration of action. Another group of psychotropic medications which, it is believed, will benefit from transdermal delivery includes antipsychotic medication such as those used in the treatment in schizophrenia.

Embodiments of the invention include, but are not necessarily limited to, use by patients with enteric absorption deficits.

Equivalents

Those skilled in the art will recognize, or be able to ascertain using no more than routine experimentation, many equivalents to the specific embodiments of the invention described herein. Such equivalents are intended to be encompassed by the following claims.

What is claimed is:

1. A method for treating pain in a subject comprising contacting said subject with a transdermal composition comprising:

a psychopharmaceutical and guaifenesin in an amount effective to treat pain, and lecithin organogel,

thereby treating pain in said subject.

2. The method of claim 1, wherein said psychopharmaceutical is selected from the group consisting of sertraline,

fluoxetine, carbamazepine, amitriptyline, trazodone, fluvoxamine, pemoline, pergolide, bromocriptine mesylate, propranolol, buproprion, reboxetine, valproic acid, nefazodone and doxepin.

3. The method of claim 1, wherein said transdermal composition further comprises Pluronic F127.

4. The method of claim 1, wherein said psychopharmaceutical is doxepin.

5. The method of claim 1, wherein said transdermal composition comprises about 5 wt % doxepin.

6. The method of claim 1, wherein said transdermal composition comprises about 10 wt % guaifenesin.

7. The method of claim 1, wherein said transdermal composition comprises about 5 wt % doxepin and about 10 wt % guaifenesin.

8. A method for treating pain in a subject comprising contacting said subject with a transdermal composition comprising:

doxepin and guaifenesin in an amount effective to treat pain, and lecithin organogel,

thereby treating pain in said subject.

9. The method of claim 8, wherein said transdermal composition comprises about 5 wt % doxepin.

10. The method of claim 8, wherein said transdermal composition comprises about 10 wt % guaifenesin.

11. The method of claim 8, wherein said transdermal composition comprises about 5 wt % doxepin and about 10 wt % guaifenesin.

12. A method for treating pain in a subject comprising contacting said subject with a transdermal composition comprising:

doxepin and guaifenesin in an amount effective to treat pain, Pluronic F127, and lecithin organogel,

thereby treating pain in said subject.

13. The method of claim 12, wherein said transdermal composition comprises about 5 wt % guaifenesin.

14. The method of claim 12, wherein said transdermal composition comprises about 10 wt % guaifenesin.

15. The method of claim 12, wherein said transdermal composition comprises about 5 wt % doxepin and about 10 wt % guaifenesin.

16. A method for treating pain in a subject comprising contacting said subject with a transdermal composition comprising:

about 5 wt % doxepin, about 10 wt % guaifenesin, and lecithin organogel,

thereby treating pain in said subject.

17. A method for treating pain in a subject comprising contacting said subject with a transdermal composition comprising:

about 5 wt % doxepin, about 10 wt % guaifenesin, Pluronic F127, and lecithin organogel,

thereby treating pain in said subject.

* * * * *

EXHIBIT P

US006495969B1

(12) **United States Patent**
Nilssen

(10) **Patent No.: US 6,495,969 B1**
(45) **Date of Patent: Dec. 17, 2002**

(54) **SERIES-RESONANT BALLAST HAVING OVERLOAD CONTROL**

(76) Inventor: **Ole K. Nilssen**, Caesar Dr., Route #5, Barrington, IL (US) 60010

(*) Notice: Subject to any disclaimer, the term of this patent is extended or adjusted under 35 U.S.C. 154(b) by 0 days.

(21) Appl. No.: **08/425,256**

(22) Filed: **Apr. 17, 1995**

Related U.S. Application Data

(63) Continuation of application No. 07/280,943, filed on Dec. 7, 1988, which is a continuation-in-part of application No. 07/080,865, filed on Aug. 3, 1987, now Pat. No. 4,819,146.

(51) **Int. Cl.**[7] **H05B 37/02**

(52) **U.S. Cl.** .... **315/209 R**; 315/208; 315/246; 315/219; 315/DIG. 4; 315/DIG. 7

(58) **Field of Search** .... 315/DIG. 4, DIG. 7, 315/244, 219, 297, 307, 207, 208, 246, 209 R, 217; 363/98, 132; 331/113 A

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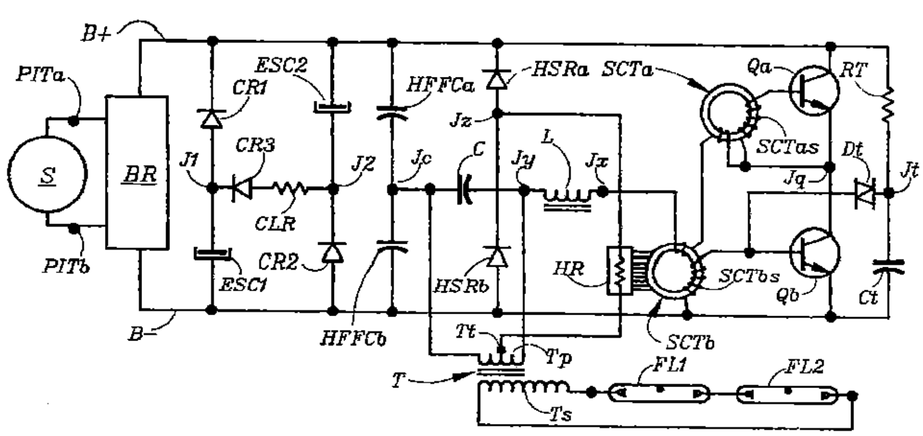
Primary Examiner—Michael B Shingleton

(57) **ABSTRACT**

An inverter is powered from a DC voltage whose instantaneous absolute magnitude is equal to that of the AC power line voltage except for being prevented from falling below a level equal to half the peak of the power line voltage. The inverter's output is loaded via a series-tuned high-Q LC circuit. Two fluorescent lamps are loading the secondary winding of a transformer whose primary winding is connected across the capacitor of the LC circuit. The magnitude of the voltage present across the capacitor is normally limited by the loading of the fluorescent lamps. However, with the lamps removed, if not expressly prevented from doing so, the magnitude of the voltage across the capacitor will increase to a destructive level due to Q-multiplication. To prevent this, an auxiliary winding on the transformer is used for limiting the magnitude for the voltage across the capacitor by rectifying the output from the auxiliary winding and feeding the resulting DC to the inverter's input; whereby the magnitude of the voltage across the capacitor will be limited to a level determined by the instantaneous magnitude of the DC voltage; which level is set such as to result in acceptable lamp starting while at the same time preventing any voltage-limiting from taking place at any time as long as the lamps are operating.

13 Claims, 1 Drawing Sheet



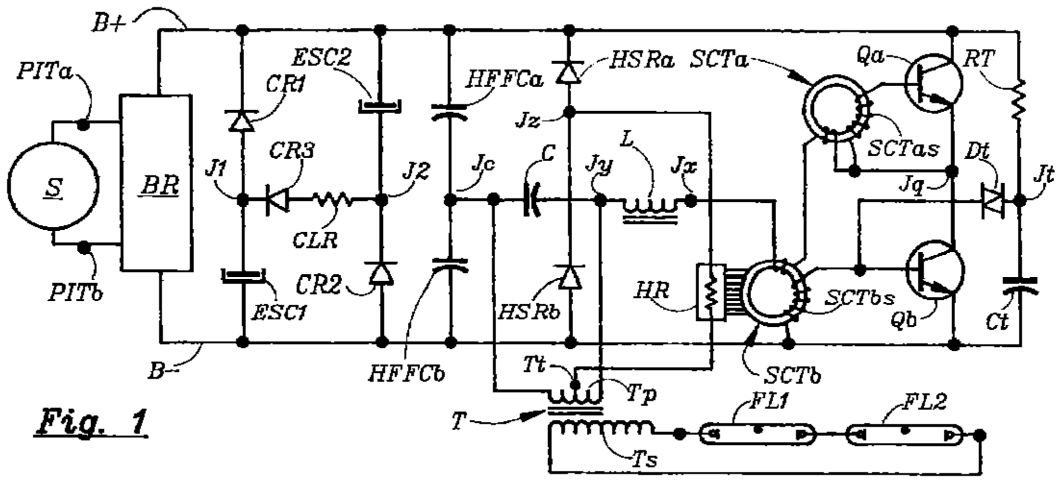


Fig. 1

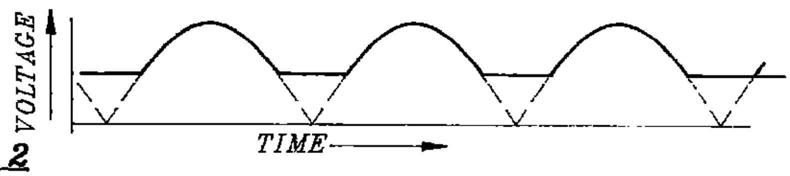


Fig. 2

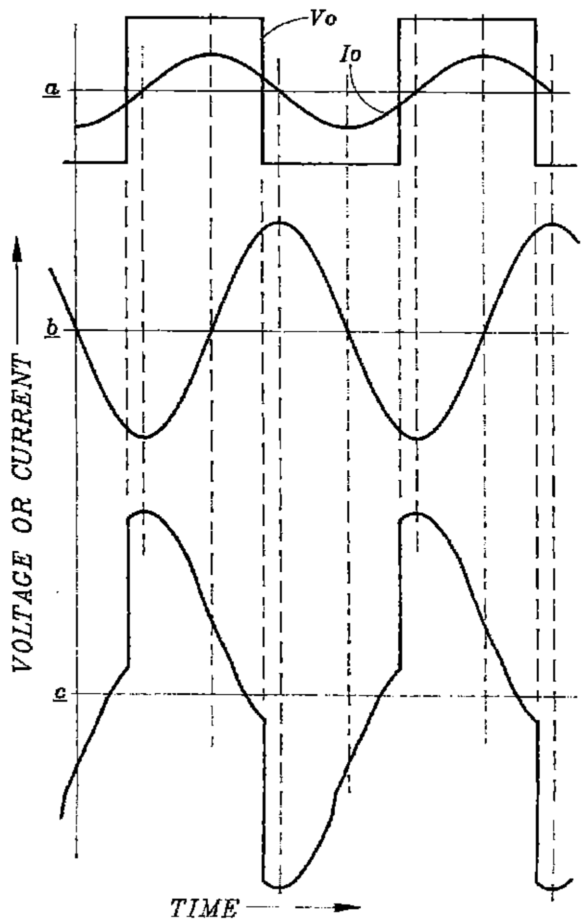


Fig. 3

SERIES-RESONANT BALLAST HAVING OVERLOAD CONTROL

RELATED APLICATION

This is a continuation of Ser. No. 07/280,943 filed Dec. 7, 1988 which is Continuation-in-Part of Ser. No. 07/080,865 filed Aug. 3, 1987, now U.S. Pat. No. 4,819,146.

BACKGROUND OF THE INVENTION

1. Field of the Invention

The present invention relates to series-resonant-loaded inverters, particularly as used for powering gas discharge lamps.

2. Description of Prior Art

In an inverter where a gas discharge lamp load is parallel-connected across the tank capacitor of a high-Q LC circuit that is resonantly series-excited by a high-frequency voltage output of the inverter, it is necessary to provide some means to protect against the high currents and voltages resulting due to so-called Q-multiplication whenever the lamp load is removed or otherwise fails to constitute a proper load for the LC circuit.

In U.S. Pat. No. 4,370,600 to Zansky, circuit protection is provided by way of providing to the LC circuit an alternative load in the form of a voltage-clamping means; which voltage-clamping means acts to load the LC circuit during any period when the lamp does not constitute a proper load therefor.

The voltage-clamping is accomplished by rectifying the Q-multiplied voltage output of the LC circuit and by applying the resulting DC output to the inverter's DC power source.

However, during any period when voltage-clamping does occur, a relatively large amount of power circulates within the electronic ballast means: from the inverter's output, through the LC circuit, and back into the inverter's DC power source by way of the voltage-clamping means.

SUMMARY OF THE INVENTION

Objects of the Invention

An object of the present invention is that of providing overload-protection means in a tuned inverter.

This as well as other objects, features and advantages of the present invention will become apparent from the following description and claims.

Brief Description

AC power line voltage is rectified and filtered such as to result in a DC voltage whose instantaneous absolute magnitude is equal to that of the power line voltage except for being prevented from ever falling below about half of the peak magnitude of the power line voltage. A half-bridge inverter is powered from this highly rippled DC voltage. The inverter's output voltage is loaded by way of a series-tuned high-Q LC circuit. Two fluorescent lamps are series-connected across the secondary winding of a transformer whose primary winding is connected across the tank capacitor of the LC circuit.

The magnitude of the voltage present across the tank capacitor is normally limited by the loading represented by the fluorescent lamps. However, with the lamps removed, if not expressly prevented from doing so, the magnitude of the voltage across the tank capacitor will increase to a destructive level due to Q-multiplication. To prevent this from happening, an auxiliary winding on the transformer is used for limiting the magnitude of the voltage across the tank capacitor by rectifying the output from the auxiliary winding and feeding the resulting DC to the inverter's DC input. As a result, the magnitude of the voltage across the tank capacitor will be limited to a level determined by the instantaneous magnitude of the DC voltage; which level is set such as to result in acceptable lamp starting while at the same time preventing any voltage-limiting from taking place at any time when the lamps are indeed operating.

To limit internal power dissipation during periods when the lamps are not connected, the effective magnitude of the inverter's output voltage is made to decrease in a time-delayed manner as a function of the magnitude of the current flowing from the auxiliary secondary winding.

The inverter is of the self-oscillating type and control of the effective magnitude of the inverter's output voltage is effected by controllably heating the ferrite material of one of two saturable current transformers used in the feedback loop.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 provides a basic electrical circuit diagram of the preferred embodiment of the invention.

FIG. 2 indicates the waveshape of the DC voltage used for powering the inverter.

FIG. 3 indicates the waveshape of various high frequency voltages present within the circuit.

DESCRIPTION OF THE PREFERRED EMBODIMENT

Details of Construction

FIG. 1 schematically illustrates the electrical circuit arrangement of the preferred embodiment of the present invention.

In FIG. 1, a source S of ordinary 120 Volt/60 Hz power line voltage is applied to power input terminals PITa and PITb; which terminals, in turn, are connected with a bridge rectifier BR. The DC output from bridge rectifier BR is applied to a B+ bus and a B− bus, with the B+ bus being of positive polarity.

A first communitating rectifier CR1 is connected with its cathode to the B+ bus and with its anode to a junction J1. A first energy-storing capacitor ESC1 is connected between junction J1 and the B− bus. A second communitating rectifier CR2 is connected with its cathode to a junction J2 and with its anode to the B− bus. A second energy-storing capacitor ESC2 is connected between junction J2 and the B+ bus. A third commutating rectifier CR3 is series-connected with a current-limiting resistor CLR to form a series-combination; which series-combination is connected between junctions J1 and J2 in such manner that the cathode of rectifier CR3 is connected with junction J1.

A first high-frequency filter capacitor HFFCa is connected between the B+ bus and a junction Jc; and a second high-frequency filter capacitor HFFCb is connected between junction Jc and the B− bus.

A first switching transistor Qa is connected with its collector to the B+ bus and with its emitter to a junction Jq; a second switching transistor Qb is connected with its collector to junction Jq and with its emitter to the B− bus.

A first saturable current transformer SCTa has a secondary winding SCTas connected across the base-emitter junction of transistor Qa; a second saturable current transformer SCTb has a secondary winding SCTbs connected across the base-emitter junction of transistor Qb. The saturable current transformers each has a primary winding; which primary windings are series-connected between a junction Jx and junction Jq.

A resistor Rt is connected between the B+ bus and a junction Jt; a capacitor Ct is connected between junction Jt and the B− bus; and a Diac Dt is connected between junction Jt and the base of transistor Qb.

A tank inductor L is connected between a junction Jy and junction Jx; and a tank capacitor C is connected between junctions Jy and Jc.

The primary winding Tp of a transformer T is connected between junctions Jc and Jy; while two fluorescent lamps FL1 and FL2 are series-connected across its secondary winding Ts. A tap Tt on primary winding Tp is connected, by way of a heating resistor HR, to a junction Jz; which heating resistor HR is thermally coupled with the magnetic core of saturable current transformer SCTb. A first high-speed rectifier HSRa is connected with its anode to junction Jz and with its cathode to the B+ bus. A second high-speed rectifier HSRb is connected with its cathode to junction Jz and with its anode to the B− bus.

Details of Operation

Except for effects associated with heating the ferrite core of saturable current transformer SCTb, the operation of the half-bridge inverter of FIG. 1 is conventional and is explained in conjunction with FIG. 8 of U.S. Pat. No. Re. 31,758 to Nilssen.

For a given magnitude of the DC supply voltage, due to the effect of the high-Q LC circuit, the magnitude of the current provided to the fluorescent lamp load (or to any other load presented to the output) is a sensitive function of the waveshape of the inverter's output voltage; which output voltage is a squarewave voltage of controllable symmetry and having peak-to-peak magnitude about equal to that of the instantaneous magnitude of the DC voltage present between the B− bus and the B+ bus.

The symmetry of the inverter's squarewave output voltage is a sensitive function of the magnetic flux saturation characteristics of the magnetic core of saturable current transformer SCTb; which flux saturation characteristics determine the duration of the ON-time of transistor Qb.

For a situation where the ferrite cores of both saturable current transformers are heated, the result is increasing inverter frequency with increasing core temperature; which situation is explained in detail in U.S. Pat. No. 4,513,364 to Nilssen. Specifically, as the saturation flux density of the two saturable current transformers is reduced, the inverter's oscillation frequency increases.

One way of reducing the transformers' saturation flux density is that of increasing the temperature of the ferrite magnetic cores used in those transformers; which effect is further explained in U.S. Pat. No. 4,513,364 to Nilssen. However, in instant situation, only one of the two magnetic cores is heated; and the net result is that the duration of the ON-time of transistor Qa stays constant while the duration of the ON-time of transistor Qb decreases.

On the other hand, the fundamental frequency of the inverter's output voltage is determined by the duration of the ON-time of the transistor that has the longest ON-time. Thus, for the circuit arrangement of FIG. 1, as long as the saturation flux of the ferrite core of saturable current transformer SCTa remains unaffected, the inverter's oscillation frequency stays approximately constant even as the saturation flux of the ferrite core of saturable current transformer is reduced. As an overall result, as the duration of the ON-time of transistor Qb is reduced (such as by heating the ferrite core of saturable current transformer SCTb), the fundamental frequency as well as the peak-to-peak magnitude of the inverter's squarewave output voltage remains approximately constant, but the symmetry of this output voltage is modified such as to reduce the magnitude of the fundamental frequency component thereof.

In fact, by sufficiently heating the ferrite core of saturable current transformer SCTb, the duration of the ON-time of transistor Qb may be reduced to near zero, thereby resulting in a dramatic reduction of the magnitude of the fundamental frequency component of the inverter's squarewave output voltage.

Thus, in view of FIG. 1, it is clear that: i) the more heating power provided to heating resistor HR, ii) the higher be the resulting temperature of the magnetic core of saturable current transformer SCTb, iii) the more reduction there-be in the saturation flux density of this current transformers' ferrite magnetic core, iv) the shorter be the duration of the ON-time of transistor Qb, v) the lower be the magnitude of the fundamental frequency component of the inverter's output voltage, and vi) due to the frequency-dependence of the output circuit, the lower be the magnitude of the current provided to the load.

In other words, since the heating power for heating resistor HR is provided by the clamping-current flowing out of tap Tt—as long as the heating resistor is an ordinary linear resistor means—the amount of heating power provided will be determined by the square of the magnitude of the clamping-current; which implies that the magnitude of the fundamental frequency component of the inverter's squarewave output voltage will decrease sensitively as a function of increasing magnitude of the clamping-current. Thus, a negative feedback condition exists: in the end, the magnitude of the clamping-current is manifestly prevented from exceeding the level at which the temperature of the magnetic cores of the saturable current transformers reaches the Curie-point; at which point the inverter ceases to oscillate.

In particular, the various component values were so chosen that the magnitude of the clamping-current stabilized at a level at which the amount of power associated with the clamping function is quite modest; which level is very much lower than the level of power associated with the clamping function prior to heating the magnetic core.

In the circuit of FIG. 1, the LC circuit is normally loaded with the two fluorescent lamps (via transformer T); and when the LC circuit is so loaded, the magnitude of the voltage present across the tank capacitor will be too low to cause clamping-current to flow. However, with the lamp loading removed, due to Q-multiplication, the magnitude of the voltage across the tank capacitor will rise until clamping-current flows.

With 120 Volt/60 Hz provided to rectifier BR, as indicated by FIG. 2, the magnitude of the DC voltage provided to the inverter varies periodically—at a frequency of 120 Hz—by a factor of about two: from a minimum of about 84 Volt to a maximum of about 168 Volt.

It is important that no clamping occur during normal operation of the fluorescent lamps. During such normal operation, the magnitude of the voltage present across the lamps will be essentially constant: determined by basic lamp characteristics. However, as indicated, the magnitude of the

DC voltage is highly variable; which implies that the voltage magnitude at which clamping starts must be at least twice as high as the normal lamp operating voltage.

In fact, taking into account variability in the magnitudes of the power line voltage and the lamps' operating voltage, it is prudent to arrange for a situation where clamping does not start to occur until the magnitude of the voltage across the tank capacitor reaches a level that is about three times as high as the level associated with normal lamp operation.

With no heating of the ferrite core of saturable current transformer SCTb, the inverter's output voltage will be as indicated by Vo of FIG. 3. The corresponding inverter output current will be as indicated by IO of FIG. 3. The voltage across tank capacitor C will be as indicated in FIG. 3; and the voltage across tank inductor L will be as indicated in FIG. 3.

With heating provided to the ferrite core of saturable current transformer SCTb, the inverter's output voltage Vo will become non-symmetrical; although the inverter frequency will remain substantially constant. As a result, with other things being approximately equal, the magnitude of the inverter's output current will decrease.

When operating normally, the fluorescent lamp load will draw about 60 Watts from the inverter and limit the magnitude of the output voltage across the secondary winding Ts to about 200 Volt. If the lamp load were to be disconnected, the magnitude of the output voltage would increase from about 200 Volt to about 600 Volt, at which point it would be prevented from increasing further due to the clamping effect resulting from the clamping-current flowing from tap Tt of primary winding Tp of transformer T. With other things being approximately equal, with the magnitude of the load voltage effectively increasing by a factor of three, the power drawn from the inverter will likewise increase by a factor of three: from 60 Watts to about 180 Watts; which 180 Watts would circulate: from the inverter's DC source, through the inverter, through the tuned LC circuit, through primary winding Tp, through heating resistor HR, and finally back to the inverter's DC source.

Of course, this circulating power must not damage the inverter nor any other other parts of the circuit. Moreover, since it is indeed possible that the lamp load be left disconnected for extended periods of time, it is important that an excessive amount of power not be dissipated as a result of this continuous circulation of a relatively large amount of power. Thus, absent means for reducing the amount of circulating power, it becomes necessary to use component parts capable continuously and efficiently of handling about three times more power than required under normal conditions where the lamp is indeed connected and operative to constitute a proper loading for the tuned LC circuit.

An important feature of instant invention relates to means for reducing the amount of power circulating when the tuned LC circuit is left externally unloaded. This feature is attained by feeding the clamping-current through heating resistor HR, thereby increasing the temperature of the ferrite core of saturable current transformer SCTb. In turn, this increased temperature significantly reduces the amount of circulating power, thereby reducing dissipation and component stresses.

The control sensitivity is arranged so as to cause the ferrite core of saturable current transformer SCTb to reach its Curie-point at a level of clamping-current that is just slightly higher than the maximum desired level of clamping-current. As a result, if the tuned LC circuit were in fact to be left externally unloaded for some extended period of time, the magnitude of the clamping-current would gradually reduce to but a small fraction of the level it would assume in the absence of controlling the temperature of the ferrite core.

In particular, with the gain in the negative feedback loop set such as to limit the magnitude of the clamping current to 50 milli-Ampere or so, the circulating power is limited to about 30 Watts. That is, the amount of power fed back to the DC source from tap Tt of transformer T is only about 30 Watts.

Of course, as soon as a functioning fluorescent lamp load is connected across secondary winding Ts, the magnitude of the voltage thereacross will fall to about 200 Volt, and clamping current will cease to flow.

Additional Comments

a) Detailed information relative to a fluorescent lamp ballast wherein the fluorescent lamp is powered by way of a series-excited parallel-loaded L-C resonant circuit is provided in U.S. Pat. No. 4,554,487 to Nilssen.

One effect of such a ballasting arrangement is that of making the waveshape of the voltage provided across the output to the fluorescent lamps very nearly sinusoidal, even though the output from the inverter itself, at the input to the series-resonant LC circuit, is more-or-less a squarewave.

b) The instantaneous peak-to-peak magnitude of the more-or-less squarewave output voltage provided by the half-bridge inverter between junctions Jq and Jc is substantially equal to the instantaneous magnitude of the DC supply voltage. Thus, as the magnitude of the DC supply voltage varies, so does the peak-to-peak magnitude of the more-or-less squarewave output voltage.

c) During a substantial part of each half-cycle of the 120 Volt/60 Hz power line input voltage, the instantaneous absolute magnitude of the DC supply voltage is substantially equal to that of the power line input voltage.

d) Saturable current transformers SCTa and SCTb require only a miniscule amount of voltage across their primary windings. Hence, the magnitude of the voltage-drop between junctions Jq and Jx is substantially negligible, and the inverter's full output voltage is therefore effectively provided between junctions Jx and Jc.

e) In the arrangement of FIG. 1, voltage-clamping is effectively accomplished by parallel-loading the tank capacitor, as opposed to parallel-loading the tank inductor.

That is, the voltage-clamping is accomplished at a point where the available current is manifestly magnitude-limited—at the fundamental inverter frequency as well as at all harmonics thereof. As a result, no problems exist with respect to component stresses due to peak currents of excessive magnitudes resulting from the clamping process.

The voltage across the tank inductor is substantially non-magnitude-limited with respect to currents at the various harmonics of the inverter frequency. That is, there is no manifest magnitude-limitation on the currents that will flow in the voltage-clamping means when that voltage-clamping means is effectively connected across the tank inductor; and the resulting component stresses are apt to be excessive.

While the voltage across the tank capacitor is apt to be essentially void of harmonics of the inverter frequency (see FIG. 3), the voltage across the tank inductor (see FIG. 3) is particularly rich in harmonics of the inverter frequency. In fact, the voltage across the tank inductor is effectively equal to the (vector) sum of the voltage across the capacitor and the (more-or-less squarewave) voltage provided from the inverter's output.

In the circuit arrangement of FIG. 4 in U.S. Pat. No. 4,370,600 to Zansky, voltage-clamping is applied to the voltage present across the tank inductor; and, as a result, severe component stresses occur as a result of clamping a non-current-limited voltage.

f) The fundamental oscillating frequency of the inverter of FIG. 1 is approximately 30 kHz. During normal operation—that is, when fully loaded with its lamp load—the inverter's output voltage is a symmetrical squarewave, as indicated by Vo of FIG. 3*a*; and the ON-time of each of the two switching transistors has a duration of about 16 micro-seconds.

As the temperature of the ferrite core of saturable current transformer SCTb is gradually increased, the duration of the ON-time of transistor Qb gradually decreases—while the duration of the ON-time of transistor Qa remains substantially constant. As a result, the inverter's output voltage becomes non-symmetrical. Ultimately, the inverter's output voltage will become brief negative (square) voltage pulses alternating with relatively long positive (square) voltage pulses. As a result, the magnitude of the fundamental frequency component of the inverter's output voltage decreases when increasing the temperature of the ferrite core of saturable current transformer SCTb.

g) As may be observed by comparing waveforms Io and Vo of FIG. 3, the current flowing from the inverter's output is somewhat delayed compared with the squarewave output voltage; which is to say that the inverter's loading is slightly inductive at the squarewave fundamental frequency.

That is, forward conduction of each switching transistor stops before the forward current has reached zero magnitude. As a result, at the point of switching, the current that was forward-flowing in one transistor will—as soon as that transistor ceases to conduct—continue to flow as reverse current in the other transistor. In this other transistor, this reverse current will flow from the emitter, through the (saturated) secondary winding of the associated current transformer, and through the base-collector junction.

Thus, in a situation where saturable current transformer SCTb has a relatively high temperature and the ON-time of transistor Qb is relatively short, the inverter's more-or-less sinusoidal output current will most of the time flow through transistor Qa: either in the forward direction, or in the reverse direction via secondary winding SCTas and the base-collector junction.

h) Controlling the inverter's output by way of controlling the symmetry of its squarewave output voltage has an advantage compared with controlling its output by way of controlling the inverter's frequency.

The inverter's frequency can be controlled by heating the ferrite cores of both saturable current transformers. However, as frequency increases, the resulting output current will become more-and-more out of phase with the inverter's output voltage; which implies that each transistor will switch at a point where the magnitude of the forward-flowing current is relatively large; which, in turn, leads to high switching losses.

On the other hand, by heating only one of the ferrite cores, transistor switching occurs at a more favorable point—particularly in the situation of minimum power output.

i) It is believed that the present invention and its several attendant advantages and features will be understood from the preceding description. However, without departing from the spirit of the invention, changes may be made in its form and in the construction and interrelationships of its component parts, the form herein presented merely representing the presently preferred embodiment.

What is claimed is:

1. A combination comprising:

a source of AC power line voltage;

a rectifier circuit connected with the AC power line voltage and operative to provide a DC voltage at a DC output; the instantaneous absolute magnitude of the DC voltage being: (i) substantially equal to that of the AC power line voltage whenever the instantaneous absolute magnitude of the AC power line voltage is higher than about half its peak absolute magnitude; and (ii) approximately constant whenever the instantaneous absolute magnitude of the AC power line voltage is lower than about half its peak absolute magnitude;

an inverter circuit connected with the DC output and operative to provide an inverter output voltage at a set of inverter terminals; and

a load circuit connected with the inverter terminals; the load circuit including an LC circuit effectively series-connected across the inverter terminals; the LC circuit having a tank capacitor and a tank inductor; a pair of output terminals being effectively parallel-connected with the tank capacitor; the LC circuit being resonant at or near the fundamental frequency of the inverter output voltage; an output voltage developing across the output terminals; the magnitude of the output voltage being determined by the nature of a final load connected in circuit therewith; the final load comprising an effective parallel-combination of a gas discharge lamp means and a voltage-clamping sub-circuit; the voltage-clamping sub-circuit comprising a rectifier sub-assembly connected in circuit with the DC terminals; the final load being operative to absorb output power from the output terminals; the output power being absorbed by: (i) the gas discharge lamp means whenever it is in fact connected and functional, or (ii) the voltage-clamping subcircuit whenever the gas discharge lamp means is effectively non-connected.

2. The combination of claim 1 wherein the inverter circuit comprises control circuitry operative to effect control of the amount of power delivered to the final load from the inverter output irrespective of the magnitude of the DC voltage.

3. The combination of claim 1 wherein the inverter circuit comprises control circuitry operative to effect control of the effective magnitude of the inverter output voltage irrespective of the magnitude of the DC voltage.

4. An arrangement comprising:

a source operative to provide a DC voltage across a pair of DC terminals;

an inverter circuit connected with the DC terminals and operative to provide an alternating inverter output voltage at a pair of inverter output terminals; a tank-inductor and a tank-capacitor being series-connected across the inverter output terminals, thereby to form a series-tuned LC circuit; the series-tuned LC circuit having a natural resonance at or near the fundamental frequency of the inverter output voltage; a ballast output voltage being present across the tank-capacitor's terminals; and

a load circuit connected with the tank-capacitor's terminals; the load circuit being characterized by:

(a) including a gas discharge lamp sub-assembly effectively parallel-connected across the tank-capacitor;

(b) including a transformer having a first winding effectively parallel-connected across the tank-capacitor and a second winding connected with a voltage-clamping sub-circuit; and

(c) being operative to absorb output power from the tank-capacitor's terminals; this output power being absorbed by (i) the gas discharge lamp sub-assembly whenever it is in fact connected and functional, or (ii) the voltage-clamping sub-circuit whenever the gas discharge lamp sub-assembly is effectively non-connected.

5. The arrangement of claim **4** wherein the load circuit is further characterized in that the second winding is connected with the DC terminals by way of a rectifier sub-assembly.

6. The arrangement of claim **4** wherein the load circuit is further characterized in that the voltage-clamping sub-circuit includes a rectifier sub-assembly connected with the DC terminals.

7. The arrangement of claim **4** wherein the transformer is further characterized in that the first winding and the second winding have one terminal in common; the common terminal being connected with one of the DC terminals.

8. The arrangement of claim **4** wherein the load circuit is further characterized in that, whenever the gas discharge lamp sub-assembly is effectively non-connected, output power from the tank-capacitor's terminals is, via the voltage-clamping sub-circuit, rectified and supplied to the DC terminals.

9. The arrangement of claim **4** wherein the inverter circuit is further characterized in that: (i) the alternating inverter output voltage has a waveshape approximating that of a squarewave voltage; and (ii) whenever the gas discharge lamp sub-assembly is in fact connected and functional, the ballast output voltage has a waveshape approximating that of sinusoidal voltage.

10. An arrangement comprising:

a source operative to provide a DC voltage across a pair of DC terminals;

an inverter circuit connected with the DC terminals and operative to provide an alternating inverter output voltage at a pair of inverter output terminals; a tank-inductor and a tank-capacitor being series-connected across the inverter output terminals, thereby to form a series-tuned LC circuit; the series-tuned LC circuit having a natural resonance at or near the fundamental frequency of the inverter output voltage; a ballast output voltage being present across the tank-capacitor's terminals; and

a load circuit connected with the tank-capacitor's terminals; the load circuit being characterized by:

(a) including a gas discharge lamp sub-assembly effectively parallel-connected across the tank-capacitor;

(b) including a transformer having a first winding effectively parallel-connected across the tank-

capacitor and a second winding connected with the DC terminals by way of a rectifier sub-assembly;

(c) being operative to absorb output power from the tank-capacitor's terminals; this output power being absorbed by (i) the gas discharge lamp sub-assembly whenever it is in fact connected and functional, or (ii) the DC terminals whenever the gas discharge lamp sub-assembly is effectively non-connected.

11. The arrangement of claim **10** wherein the transformer is additionally characterized in that the first winding and the second winding have one terminal in common; which common terminal is connected with one of the DC terminals.

12. The arrangement of claim **10** wherein the transformer is additionally characterized in that the first winding and the second winding in combination are represented by a single winding with a tap; the tap being connected with one of the DC terminals by way of the rectifier sub-assembly.

13. A combination comprising:

a source of AC power line voltage;

a rectifier circuit assembly connected with the AC power line voltage and operative to provide a DC voltage between a B− bus and a B+ bus; the instantaneous absolute magnitude of the DC voltage being: (i) substantially equal to that of the AC power line voltage whenever the instantaneous absolute magnitude of the AC power line voltage is higher than about half its peak absolute magnitude; and (ii) approximately constant whenever the instantaneous absolute magnitude of the AC power line voltage is lower than about half its peak absolute magnitude; the rectifier circuit being further characterized by including a first and a second capacitor, each having a pair of terminals; a substantially constant DC voltage existing between each pair of terminals; the absolute magnitude of this constant DC voltage being no larger than half the peak absolute magnitude of the AC power line voltage; one of the terminals of the first capacitor being directly connected with the B− bus; one of the terminals of the second capacitor being directly connected with the B+ bus; and

an inverter circuit assembly connected between the B− bus and the B+ bus; the inverter circuit assembly being operative to provide an inverter output voltage at a set of inverter output terminals; connected with the inverter output terminals is an LC circuit with a tank capacitor and a tank inductor; the LC circuit being resonant at or near the fundamental frequency of the inverter output voltage; a gas discharge lamp being connected in circuit with the inverter output terminals.

* * * * *

IN THE UNITED STATES DISTRICT COURT
FOR THE NORTHERN DISTRICT OF ILLINOIS

OLE K. NILSSEN and GEO FOUNDATION, LTD., Plaintiffs,	)	Civil Action No. 01-CV-3585
v.	)	Judge John W. Darrah
OSRAM SYLVANIA, INC. and OSRAM SYLVANIA PRODUCTS, INC., Defendants.	)	JURY TRIAL DEMANDED

03C 2962

JUDGE AMY ST. EVE

MAGISTRATE JUDGE ASHMAN

JURY DEMAND

The undersigned demands a jury trial.

OLE K. NILSSEN and GEO FOUNDATION, LTD.

Dated: May 2, 2003

Respectfully Submitted,

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DOCKETED
MAY 0 6 2003

UNITED STATES DISTRICT COURT
NORTHERN DISTRICT OF ILLINOIS

Cat

Civil Cover Sheet

This automated JS-44 conforms generally to the manual JS-44 approved by the Judicial Conference of the United States in September 1974. The data is required for the use of the Clerk of Court for the purpose of initiating the civil docket sheet. The information contained herein neither replaces nor supplements the filing and service of pleadings or other papers as required by law. This form is authorized for use only in the Northern District of Illinois.

03C 2962

Plaintiff(s): OLE K. NILSSEN and GEO FOUNDATION, LTD.

County of Residence: McHenry

Plaintiff's Atty: Raymond N. Nimrod, George S. Bosy, John E. Titus, Gregory D. Bonifield
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Defendant(s):OSRAM SYLVANNIA, INC. and OSRAM SYLVANIA PRODUCTS, INC.

County of Residence:

JUDGE AMY ST. EVE

Defendant's Atty:

FILED
MAY 0 2 2003
MICHAEL W. DOBBINS
CLERK, U.S. DISTRICT COURT

II. Basis of Jurisdiction: **3. Federal Question (U.S. not a party)**

III. Citizenship of Principal Parties **(Diversity Cases Only)**
Plaintiff:- **N/A**
Defendant:- **N/A**

IV. Origin : **1. Original Proceeding**

V. Nature of Suit: **830 Patent**

VI.Cause of Action: **35 U.S.C. 271 Patent Infringement**

VII. Requested in Complaint
Class Action:
Dollar Demand:
Jury Demand: **Yes**

VIII. This case **IS NOT** a refiling of a previously dismissed case.

Signature: [signature]

Date: 5/2/03

1-2

DOCKETED

MAY 06 2003

UNITED STATES DISTRICT COURT
NORTHERN DISTRICT OF ILLINOIS

EASTERN DIVISION

In the Matter of

OLE K. NILSSEN and GEO FOUNDATION, LTD.
v.
OSRAM SYLVANIA, INC. and OSRAM SYLVANIA PRODUCTS, INC.

Case Number: 03C 2962

APPEARANCES ARE HEREBY FILED BY THE UNDERSIGNED AS ATTORNEY(S) FOR:

FILED

MAY 02 2003

MICHAEL W. DOBBINS
CLERK, U.S. DISTRICT COURT

MAGISTRATE JUDGE ASHMAN

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TRIAL ATTORNEY?	YES ☑ NO ☐	TRIAL ATTORNEY?	YES ☑ NO ☐
DESIGNATED AS LOCAL COUNSEL?	YES ☐ NO ☐	DESIGNATED AS LOCAL COUNSEL?	YES ☐ NO ☐

1-3