Battery Management 'Cook Book'

AN+DS/RSC/IS

Battery Management 'Cook Book'

Inhoudsopgave

AN+DS/RSC/IS

INTEGRATED CIRCUITS

DATA SHEET

TEA1100; TEA1100T
Battery monitor for NiCd and NiMH chargers

Preliminary specification
File under Integrated Circuits, IC03

May 1992

Philips Semiconductors

PHILIPS

FEATURES

- Accurate regulation of charge current settings in co-operation with a switched mode power supply
- Accurate detection of fully charged batteries by currentless battery voltage sensing
- Switch over from fast to normal charging when batteries are fully charged
- Adjustable fast charging level (1 C to 5 C)
- Adjustable normal charging level (0.05 C to 0.25 C)
- Temperature guarding by means of an NTC resistor
- Tracking of maximum fast charging time with fast charging current level
- Protections against short-circuited and open batteries
- Large battery voltage range
- Both DC and PWM outputs with polarity switch

APPLICATIONS

- Charge systems for NiCd and NiMH batteries

GENERAL DESCRIPTION

The TEA1100 is manufactured in a BICMOS process intended to be used as a battery monitor circuit in charge systems for NiCd and NiMH batteries.

The circuit has to be situated on the secondary side in mains-isolated systems where it monitors the battery voltage and the charge current. The circuit drives, by means of an opto-coupler or a pulse transformer interface, an SMPS circuit, situated on the primary side of the system, thus controlling the charge current of the batteries.
The circuit can drive the external power transistor in switched mode systems, which have a DC power source, via a driver stage.

ORDERING INFORMATION

EXTENDED TYPE NUMBER	PACKAGE			
	PINS	PIN POSITION	MATERIAL	CODE
TEA1100	16	DIL	plastic	SOT38G
TEA1100T	16	SO16L	plastic	SOT162A

QUICK REFERENCE DATA

SYMBOL	PARAMETER	CONDITIONS	MIN.	TYP.	MAX.	UNIT
V_P	positive supply voltage range		5.65	–	11.5	V
I_P	supply current	outputs off	–	–	4.1	mA
V_{VAC}	voltage range of battery-full detection		0.385	–	3.85	V
dV_{VAC}/V_{VAC}	–dV detection level w.r.t. top value	note 1	–	1	–	%
I_{VAC}	input current battery monitor		–	–	1	nA
V_{VAC}	voltage protection					
	battery low		–	0.3	–	V
	battery high		–	4.25	–	V
	charging level	$I_{charge} = R1/R_s \times I$; see Fig. 3				
I_{ref}	fast	$I = I_{ref}$	20	–	100	µA
I_n	normal	$I = 1/p \times 0.1 \times I_n$ (p = prescale factor)	10	–	50	µA
f_{OSC}	oscillator frequency		10	–	100	kHz

Note to the quick reference data

1. The –dV detection level can be adjusted by use of an external voltage regulator diode to increase the sensitivity.

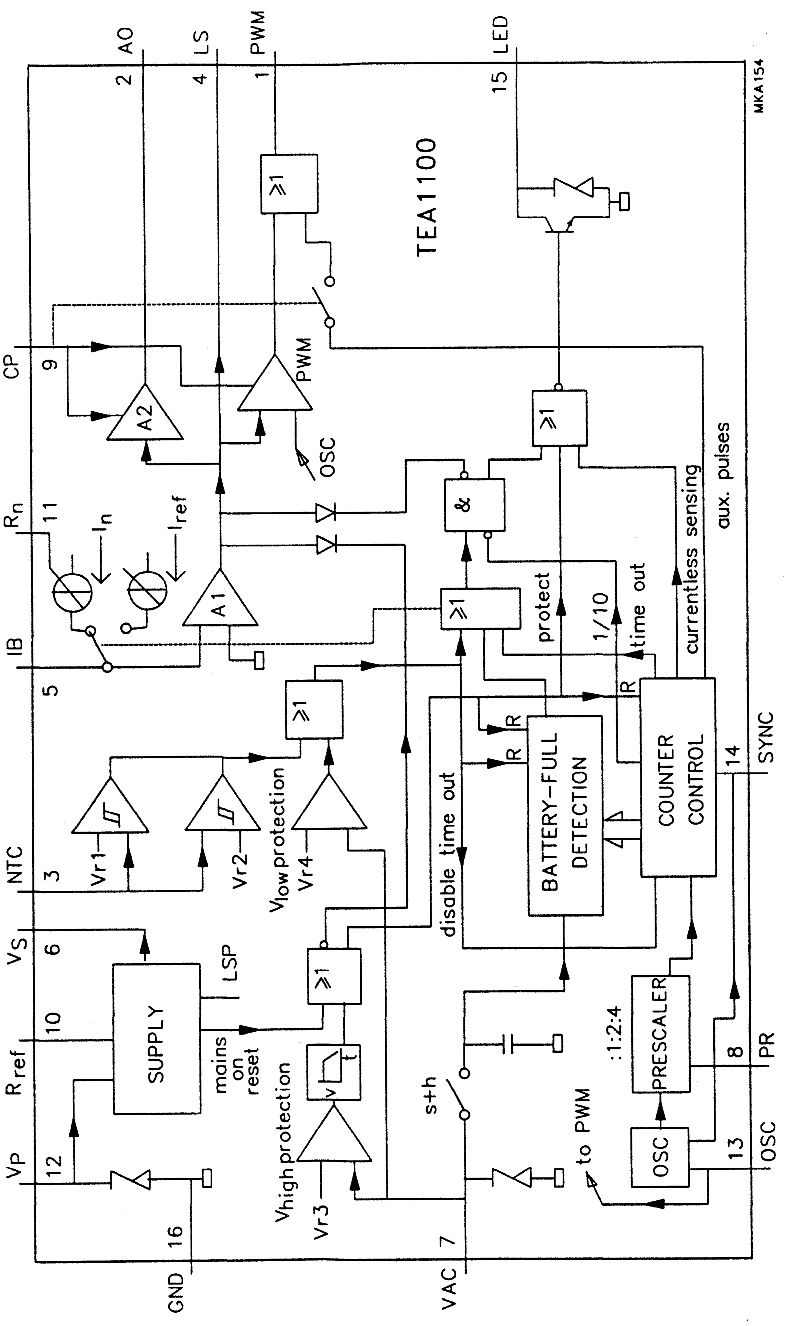


Fig.1 Block diagram.

PINNING

SYMBOL	PIN	DESCRIPTION
PWM	1	pulse width modulator
AO	2	analog output
NTC	3	temperature sensor input
LS	4	loop stability
IB	5	charge current
V_S	6	stabilized supply voltage
VAC	7	battery voltage
PR	8	prescaler
CP	9	change polarity
R_{ref}	10	reference resistor
R_n	11	normal charge reference resistor
V_P	12	positive supply voltage
OSC	13	oscillator input
SYNC	14	synchronization input
LED	15	LED output
GND	16	ground

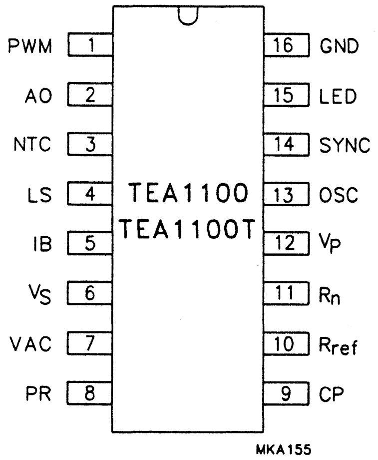


Fig.2 Pin configuration.

FUNCTIONAL DESCRIPTION

The operation of the circuit will be explained with the aid of Fig.1 (block diagram) and Fig.3 (application diagram). The circuit is divided into several blocks which are described separately.

Supply block

The circuit needs a supply voltage on pin V_P with a value between 5.65 and 11.5 V. Above 6.4 V typ., the circuit starts up assuming that mains is connected to the system and the charge session begins. This supply can be generated by a separate winding on the transformer, as shown in Fig.3 (application diagram), in either the flyback or the forward stroke. Another possibility is rectification from the mains secondary winding (at the connection D1 and L2). Considerations for choosing the way of supplying the IC are:

- supply voltage range of 5.65 to 11.5 V under all circumstances (also during the 90% pause at normal charging, the standby current then is 1 mA typ.)
- maximum battery voltage (flyback stroke)
- minimum power delivered by the primary SMPS (normal charging)

The supply block delivers the following outputs:

- By using an external resistor R_{ref} at pin 10 (R_{ref}) a reference current is obtained which defines all external related currents (charge reference currents, oscillator).
- Externally available 4.25 V stabilized voltage source (V_S). This source is used internally for a large part of the circuit and can be used to set the NTC biasing and to supply other external circuitry. V_S is cut off in the 90% pause during normal charging.
- Low Supply voltage Protection signal (LSP). When the supply voltage is lower than 5.25 V typ., there is supply voltage enough left to switch off the power regulation and hereafter the IC current is limited to the start level of 35 µA typ.
- Mains on reset pulse resets all digital circuitry after a start or restart due to an interrupted supply (V_P).

Charge current regulation

The charge current has to be sensed by means of a low-ohmic resistor in series with diode D1. The waveform on resistor R_s (see Fig.4 for a flyback converter) has the form of a negative-going ramp and after filtering a negative DC voltage is obtained. Across resistor R1 a positive voltage is created by means of the current sources set by the pins R_{ref} and R_n. The error amplifier A1 references the result to ground and via the regulation loop of the SMPS, the secondary current will be regulated to a value which is defined by:

$I_{ch} \times R_s = R1 \times I_{ref}$ (fast charge) or,
$I_{ch} \times R_s = R1 \times I_n$ (normal charge)

The I_{ref} current is the fast charging reference current, the I_n current is used for regulation after a full battery is detected. The I_{ref} current is the reference current set by R_{ref} while I_n is dependent on the resistor at pin R_n. With no resistor on pin R_n, the I_n current has a default value which is half the I_{ref} current.
By choosing the correct resistor values R_s, R1, R_{ref} and R_n, a wide range of charge currents can be set as well as a wide range of the ratio fast charge current as a function of normal charge current. For determination of the normal charge current the 1:10 duty cycle and the programmable prescale factor (p) should be taken into account (see Logic block); $I_n = 1/p \times 0.1 \times I_n$. The output of amplifier A1 is available at the loop stability pin (LS), so the time constant of the SMPS loop can be set at the secondary side of the system.

NTC block

The voltage at the NTC pin is compared with two reference voltages. When the NTC voltage is between V_{r1} and V_{r2}, the charge current regulation is unaffected.
When the NTC voltage is outside this window, the power of the SMPS is reduced to the normal charge level.
The NTC input can be used for temperature protection as shown in Fig.3 (application diagram) by using a suitable NTC resistor. To avoid switching on and off with temperature, a hysteresis is built in for both levels.

Output drivers

The SMPS regulation signal is available at different pins:

- Analog voltage output (push or pull) at AO (pin 2) to drive an opto-coupler in mains separated applications when an external resistor is connected between AO and the opto-coupler. The maximum current through the opto-coupler diode is 2 mA. The voltage gain of amplifier A2 is: $A = (V_{LS} - 1.4) \times 4$ and is typ. 12 dB. The voltage at AO can also be used to drive a PWM input of an SMPS circuit directly. During 'inhibit SMPS' the AO output is fixed to zero charge current for currentless sensing.
- The LS voltage is compared internally with the oscillator voltage to deliver a pulse width modulated output at PWM (pin 1) to drive an output device in a DC/DC converter application via a driver stage. The PWM output is latched to prevent multi-pulsing. Moreover with the latch a kind of current mode control is possible. The maximum duty cycle is internally fixed to 78% (typ.). The 'PWM' output can be used for synchronization and duty cycle control of a primary SMPS via a pulse transformer (the SMPS inhibit and auxiliary pulses are also available at pin PWM).
- The AO and PWM outputs can be changed in polarity by programming the change polarity pin CP. The PWM output in the on-state pushes current (CP = 0) or pulls current (CP = 1). The appearance of the auxiliary pulses at pin PWM can also be programmed with CP.

The 'LED' output pin offers the following output signals:

- 10/90% signal for driving a LED when the duty cycle is too small during the 10% time. This occurs when there is a large difference between fast and normal charge currents. The LED frequency is $f_{LED} = 2^{-12} \times 1/p \times f_{OSC}$
- An SMPS inhibit period (duration 10 OSC pulses) for currentless VAC sensing.
- VAC high voltage protection signals.

Battery monitor

At higher battery voltages it is advised to divide the battery voltage with a factor 5 before offering this to pin VAC (Voltage ACcumulator). It is also possible to take a tap on the chain of batteries.
The VAC voltage range has to be between 0.385 V and 3.85 V.
The VAC voltage is sampled at a low cycle frequency ($f_{cycle} = 2^{-16} \times f_{OSC}$) and the analog value of VAC is digitized and stored in a register. One cycle later, the digitized value is converted back to the analog value and compared with the actual value of VAC. If the actual value is higher, then the new VAC voltage is stored in the register, otherwise no conversion is done. So the VAC top value is stored and it is possible to detect an increasing VAC indicating 'not yet full batteries' or decreasing VAC indicating that the batteries are probably fully charged.

The circuit waits until the battery voltage has dropped 1% below the top value before indicating 'full batteries'. However, by applying a voltage regulator diode in the battery voltage sense-line (see Fig.7) an increased sensitivity of the –dV detection level can be obtained, e.g. 0.5% or even a lower value.
In Fig.5 the battery voltage as function of the charging time is shown. The negative slope depends on the charge current and is approximately 3 mV/cell/K.

The switching of the SMPS can cause interference on the battery voltage and therefore it has been necessary to stop the SMPS during the inhibit time (see Fig.8). This can be done automatically via the regulation pins AO and PWM or by using the SYNC output of the logic block. The SMPS is stopped for 10 periods at the end of which sampling is done. The VAC voltage will now be sensed currentless. To avoid false decisions concerning a falling VAC voltage, VAC is digitally filtered and analog stored in a sample-and-hold circuit. This approach ensures, even at very high –dV sensitivity (<1%) accurate detection of the battery full condition. Immediately hereafter decisions and VAC digitizing takes place. The benefit of a sample-and-hold circuit is that at high frequencies the noise on the VAC voltage is filtered and the VAC manipulations like decisions and digitizing are done on the same VAC voltage available in the sample-and-hold circuit.

When a –dV is detected, the reference current I_{ref} is switched off, the normal current I_n is switched on during 10% of regulation and the outputs are high-ohmic during 90%. This 1:10 ratio in active regulation, together with the ratio in reference currents (I_n as a function of I_{ref}), ensures that the resulting charge current is low enough to be allowed to flow through the batteries for a long time to overcome the self-discharge of the batteries without causing memory effects. In case the prescale factor p is programmed, the I_n current has to be lowered with the p factor, so $I_n = 1/p \times 0.1 \times I_n$.

Protections

- The circuit goes into standby (not active, low current consumption) when the supply voltage is less than 5.25 V (LSP).
- When the divided battery voltage exceeds the V_{r3} level (nominal 4.25 V) this is recognized as open or removed batteries and the output control signals terminate to stop the SMPS operation. This over-voltage sensing is digitally filtered. In above cases the 'battery full detector' and the 'counter/control' will be reset.
- When the divided battery voltage is less than V_{r4} (0.3 V), the circuit assumes short-circuited batteries, the charge current is reduced to the normal charge level. As soon as the voltage exceeds V_{r4}, the fast charging starts.
- The temperature protections are already mentioned in section NTC. In the case of short-circuited batteries or active temperature protections the 'battery full detector' is reset and the 'counter/control' is stopped.

Oscillator and control logic

The whole timing of the circuit is controlled by the oscillator.
The period time is defined by:
$T_{OSC} = 0.93 \times R_{ref} \times C_{OSC}$.

The counter block defines a maximum fast charge time called 'Time Out' (TO). As the charge current and the oscillator frequency (and so the TO) are both set by R_{ref}, changing one effects the other. Initially the oscillator capacitor can be chosen such that the fast charge time is half the TO time. This means that in case of a one hour (1C) charger, the TO signal occurs at 2 hours, in case of a quarter of an hour (4C) charger, the TO signal is active after half an hour. After that the circuit switches over to normal charging.

To adapt the SMPS switching frequency in the synchronized mode to the required oscillator frequency of the timing logic, the timer logic is preceded by a programmable divider. By means of the PR pin the divider ratio can be set to 1, 2 or 4 (p factor). Doing so the oscillator frequency can be increased with the factor p without changing TO.

Fast charging current:
$I_{ch} = R1/R_s \times V_{ref}/R_{ref}$
Time out:
$TO = 2^{26} \times 0.93 \times R_{ref} \times C_{OSC} \times p$
Normal charging current:
$I_{ch} = R1/R_s \times 1/p \times 0.1 \times V_{ref}/R_n$

The control block determines the following timing sequences:

- VAC sampling; this takes 1 clock pulse every interval cycle.
  The power converter is switched off during VAC sampling. As there are several types of converters, there also are several control signals available at :
  - pin 'SYNC' for synchronization in analog voltage controlled primary SMPS circuits
  - pin 'PWM' for digital controlled primary SMPS and DC/DC converters
  - pin 'LED' in special applications

- Disabling –dV during 2^{-5} x TO (3% of TO) for proper start with flat or inverse polarized batteries. Disabling is active at each fast charge cycle.
- Maximum fast charging time (TO): the maximum timer is stopped during VAC low voltage protection and outside temperature range.
- The normal charge duty cycle is 1/p x 0.1
- Auxiliary pulses to support the supply voltage of the primary SMPS circuit via pin PWM: the pulses can be programmed on and off at an appearance rate of $f_{OSC}/8$ with a duty cycle of 14%; programming is achieved by activating CP.

The timing logic and the –dV recognition circuitry are reset after each supply voltage failure and after a battery over-voltage recognition. The –dV circuit is also reset during normal charging.

The SYNC output delivers negative-going synchronization pulses which are suppressed during the sampling of the battery voltage. With these sync pulses the SMPS can be synchronized.The polarity of the sync pulses is chosen so that in case of an open SYNC pin in the synchronization mode, the power is regulated to a minimum. During the VAC sampling the absense of sync pulses causes the SMPS to stop thus minimizing interference (see Fig.8, synchronization waveforms).

During the 90% pause, only the oscillator and the control logic are operative to save current. In the pause V_P is never allowed to become less than V_{LSP}. This would cause a 'mains-on-reset' and so fast charging.

Programming

With pins 'CP' (change polarity) and 'PR' (prescaler) several functions can be programmed.

By defining the current (V_{ref}/R_{CP}) at pin CP, the following functions can be activated :

1	change polarity	CP = 0, normal polarity CP = 1, changed polarity
2	no auxilliary pulses at PWM	aux = 0
3	auxilliary pulses at f/8	aux = 8

CP PIN	**FUNCTIONS**	
	CP	**aux**
open pin	0	0
10 µA	0	8
22 µA	1	0
57 µA	1	8

By defining the voltage at pin PR, the following functions can be activated :

PR PIN	**FUNCTIONS**
V_S	prescaler divide by 1
open pin	prescaler divide by 2
ground	prescaler divide by 4

Formulas

DESCRIPTION	**SYMBOL**	**FORMULA**	**FUNCTION**
timing	T_{OSC}	$0.93 \times R_{ref} \times C_{OSC}$	repetition
	TO	$2^{-26} \times p \times T_{OSC}$	duration
	$T_{disable}$	$2^{-5} \times TO$	duration
	$T_{LED} = T_{trickle}$	$2^{12} \times p \times T_{OSC}$	repetition
	$T_{LED} = T_{trickel}$	$3/4 \times 2^{9} \times T_{OSC}$	duration
	$T_{inhibit}$	$2^{16} \times T_{OSC}$	repetition
	$T_{inhibit}$	$10 \times T_{OSC}$	duration
charge currents	I_{fast}	$R1/R_s \times V_{ref}/R_{ref}$	
	I_{normal}	$R1/R_s \times 1/p \times 0.1V_{ref}/R_n$	

LIMITING VALUES

In accordance with the Absolute Maximum Rating System (IEC 134)

All voltages with respect to ground; positive currents flow into the IC; all pins not mentioned in the voltage list are not allowed to be voltage driven. The voltage ratings are valid provided other ratings are not violated; current ratings are valid provided the power rating is not violated.

SYMBOL	PARAMETER	CONDITIONS	MIN.	MAX.	UNIT
Voltages					
V_P	positive supply voltage (pin 12)		–0.5	13.2	V
V_{LED}	LED voltage (pin 15)		–0.5	13.2	V
$V_{1,8,4,3}$	voltage at PWM (pin 1), PR (pin 8), LS (pin 4), NTC (pin 3)		–0.5	V_P	V
V_{IB}	voltage at IB (pin 5)		–0.5	–1	V
Currents					
I_{VS}	current at V_S (pin 6)		–3	+0.01	mA
I_{LED}	current at LED (pin 15)		–	25	mA
I_{AO}	current at AO (pin 2)		–5	+5	mA
I_{PWM}	current at PWM (pin 1)		–15	+15	mA
I_{SYNC}	current at SYNC (pin 14)		–2	+2	mA
$I_{11,10,9}$	current at R_n (pin 11), R_{ref} (pin 10), CP (pin 9)		–1	+0.01	mA
$I_{4,5,7}$	current at LS (pin 4), IB (pin 5), VAC (pin 7)		–1	+1	mA
I_P	current at V_P (pin 6)		–	15	mA
Dissipation					
P_{tot}	total power dissipation	T_{amb} = 85 °C SOT38G SOT162A	 – –	 0.6 0.3	 W W
Temperatures					
T_{amb}	operating ambient temperature		–20	+85	°C
T_j	junction temperature		–	+150	°C
T_{stg}	storage temperature		–55	+150	°C

CHARACTERISTICS

T_{amb} = 25 °C; V_P = 10 V; R_{ref} = 33 kΩ; R_n = 68 kΩ; C_{OSC} = 1 nF; CP is open; PR connected to V_S; unless otherwise specified

SYMBOL	PARAMETER	CONDITIONS	MIN.	TYP.	MAX.	UNIT
Supply (V_P, V_S, R_{ref})						
V_P	supply voltage range		5.65	–	11.5	V
V_{PC}	clamp voltage	I_{PC} = 10 mA	11.5	–	12.8	V
V_{PS}	start voltage		6.1	6.4	6.7	V
V_{PLSP}	low supply protection level		4.85	5.25	5.65	V
V_{PLSPH}	hystereris of V_{PLSP}		0.3	0.95	–	V
I_P	supply current	outputs off	–	–	4.1	mA
I_{PP}	supply pause current	V_P = 6 V	–	–	1.71	mA
I_{PSB}	standby current	V_P = 4 V	–	35	45	µA
V_S	source voltage (stabilized)	I_S = 1 mA	4.03	4.25	4.46	V
V_{ref}	reference voltage	I_{ref} = 20 µA	1.18	1.25	1.31	V
TC_{Vref}	temperature coefficient of V_{ref}	T_{amb} = 0 to 45 °C	–	±100	±200	ppm/K
dV_{ref}/dV_P	power supply rejection ratio (PSRR) of V_{ref}	f = 100 Hz; dV_P = 2 V (peak-to-peak value); V_P = 8 V	–46	–	–	dB
ΔV_{ref}	voltage difference	dI_S = 1 mA	–	–	5	mV
I_{Rref}	current range of R_{ref}		10	–	100	µA
Charge current regulation (IB, R_n, R_{ref})						
V_n	voltage at pin R_n	I_n = 10 µA; I_{ref} = 20 µA	1.17	1.25	1.32	V
I_n	current range at R_n		5	–	50	µA
I_{IB}/I_{ref}	input current ratio	R_n not connected				
	normal charging	V_{IB} = 0	0.475	0.5	0.525	
	fast charging	V_{IB} = 0	0.95	1	1.05	
I_{IB}/I_n	input current ratio normal charging	R_n connected	0.90	0.97	1.04	
V_{thIB}	threshold voltage at IB	T_{amb} = 25 °C	–5	–	+5	mV
		T_{amb} = 0 to 45 °C	–7	–	+7	mV
NTC input						
V_{NTCSPH}	switching protection voltage on high temperatures		0.75	0.81	0.87	V
V_{NTCHH}	hysteresis of V_{NTCSPH}		60	90	120	mV
V_{NTCSPL}	switching protection voltage on low temperatures		2.78	3.00	3.20	V
V_{NTCHL}	hysteresis of V_{NTCSPL}		65	100	135	mV
I_{NTC}	input current	V_{NTC} = 2 V	–5	–	+5	µA

SYMBOL	PARAMETER	CONDITIONS	MIN.	TYP.	MAX.	UNIT
Output drivers (AO, LS, PWM, LED)						
$I_{AOsource}$	source current	V_{AO} = 3 V; CP = 0	–	–	–2	mA
I_{AOsink}	sink current	V_{AO} = 0.5 V; CP = 1	2	–	–	mA
g_m	transconductance A1	V_{IB} = 50 mV	–	300	–	µS
G_{v1}	voltage gain A1 x A2	V_{AO} = 2 V (peak-to-peak value)	–	72	–	dB
G_{v2}	voltage gain A2	V_{AO} = 2 V (peak-to-peak value)	–	12	–	dB
$I_{LSsource}$	maximum source current	V_{LS} = 2.25 V	–25	–21	–16	µA
I_{LSsink}	maximum sink current	V_{LS} = 2.25 V	16	21	25	µA
I_{PWMH}	HIGH level output current	V_{PWM} = 3 V	–18	–14	–10	mA
I_{PWML}	LOW level output current	V_{PWM} = 0.5 V	7	12	17	mA
$I_{PWMleak}$	leakage current	V_{PWM} = 4.25 V	–	0.2	10	µA
δ_{PWM}	maximum duty cycle		70	78	86	%
δ_{PWMaux}	auxiliary pulse duty cycle		12.6	14	15.4	%
V_{LEDsat}	saturation voltage	I_{LED} = 15 mA	–	–	600	mV
$I_{LEDleak}$	leakage current	V_{LED} = 10 V	–	–	5	µA
Battery monitor (VAC)						
I_{VAC}	input current	V_{VAC} = 4.25 V	–	1	–	nA
V_{VAC}	voltage range of –dV detection		0.385	–	3.85	V
dV_{VAC}/V_{VAC}	–dV detection level w.r.t. top level	V_{VAC} = 2 V	0.85	1	1.15	%
ΔV_{VAC}	resolution –dV		0.42	0.6	0.78	mV
T_{-dV}	temperature range of –dV detection		0	–	50	°C
Protections (VAC)						
V_{VACLBP}	low battery voltage protection		–	0.3	0.33	V
V_{VACHBP}	high battery voltage protection	with respect to V_P	–	0	150	mV
Oscillator, logic (OSC, SYNCH)						
V_{OSCH}	oscillator switching level HIGH		–	2.5	–	V
V_{OSCL}	oscillator switching level LOW		–	1.5	–	V
K	period time $T_{OSC} = K \times R_{ref} \times C_{OSC}$		0.84	0.93	1.02	
f_{OSC}	oscillator frequency range		10	–	100	kHz
V_{SYNCH}	SYNC output level HIGH	I_{SYNCH} = –0.4 mA	3.4	–	–	V
V_{SYNCL}	SYNC output level LOW	I_{SYNCL} = 0.4 mA	–	–	0.85	V

SYMBOL	PARAMETER	CONDITIONS	MIN.	TYP.	MAX.	UNIT
Programming (CP)						
I_{CP}	programming currents					
	CP = 0; aux = 0	R_{CP} = 330 kΩ	–	–	4.2	µA
	CP = 0; aux = 8	R_{CP} = 120 kΩ	9.4	10.4	11.4	µA
	CP = 1; aux = 0	R_{CP} = 56 kΩ	20.0	22.3	24.5	µA
	CP = 1; aux = 8	R_{CP} = 22 kΩ	51.1	56.8	62.5	µA

QUALITY SPECIFICATION

General quality specification for integrated circuits: UZW-B0/FQ-0601.

Remark: for the synchronization pin (14), the ESD positive zap voltage is restricted to a maximum of 1000 V.

TEST AND APPLICATION INFORMATION

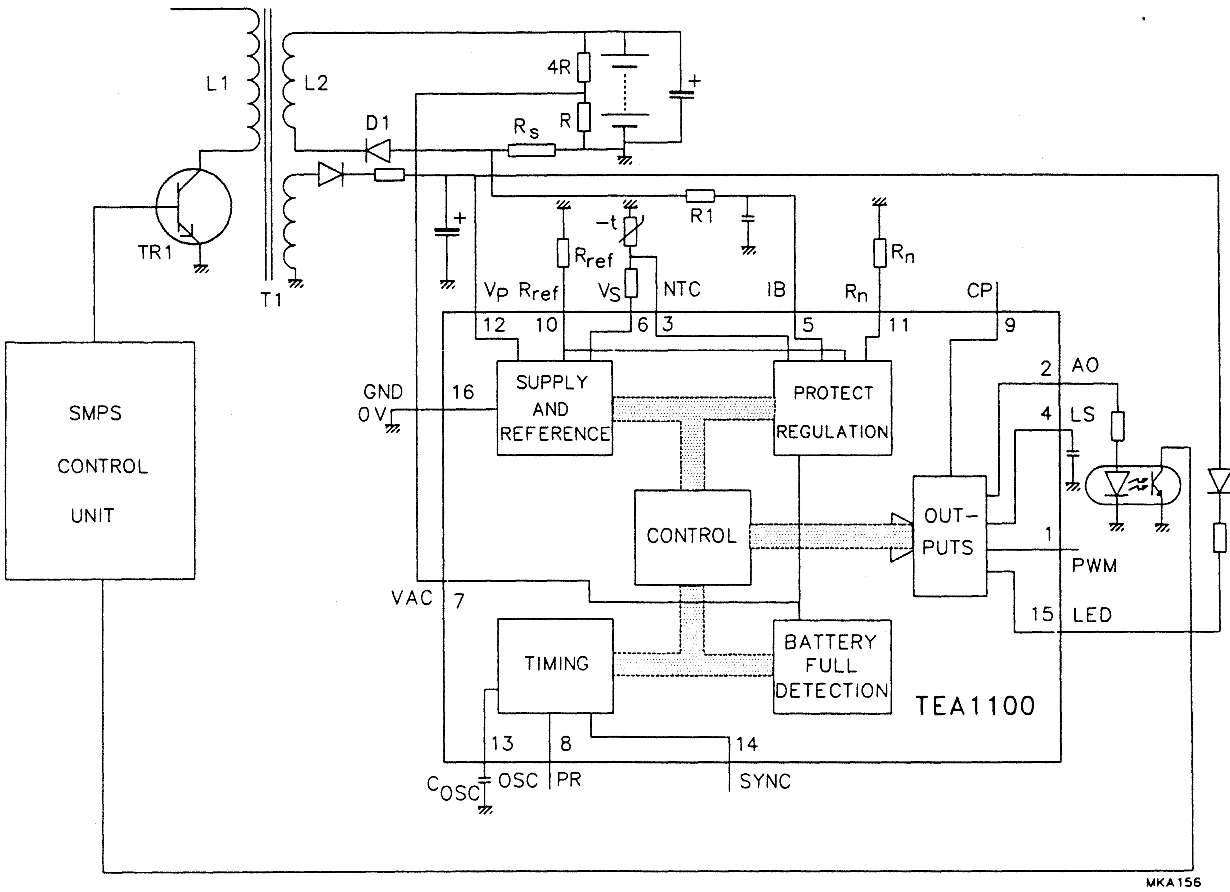


Fig.3 Application diagram.

Notes to Fig.3

1. Signaling the status of the charging session can be done by a LED-diode-resistor combination parallel to L2 (transformer T1). During the fast charging period the LED will burn continuously. During normal charging the LED will switch with the 10/90% rhythm. With mains-off the LED is off, thus not discharging the batteries. If at normal charging the duty cycle is too low during the 10% because of a very large difference between the fast charge and the normal charge levels, the LED can be driven by the LED pin.
2. With R_s = 50 mΩ and a required fast charging current level of 6 A (5C for 1.2 Ah batteries), the average current sense level is 300 mV. Power dissipation in R_s = 1.8 W.
   With a 3 kΩ resistor for R1, the required I_{ref} current is 300 mV/3 kΩ = 100 µA. For a normal charge level of 0.25C (300 mA) the voltage drop over R_s is 15 mV. Taken into account the duty cycle of 10%, the voltage drop over R1 = 150 mV. So the I_{Rn} current has to be: 150 mV/3 kΩ = 50 µA (p = 1).

Method to increase –dV sensitivity

The basic, direct battery sensing via a resistive divider, which adapts the battery voltage within the V_{AC} range, is shown in Fig.6. Detection occurs at $-dV = 1\%$ of $V_{B(max)}$.

The position of the Zener diode is shown in Fig.7. The TEA1100 now senses the voltage V_R, which is the battery voltage minus the Zener-diode voltage ($V_R = V_B - V_Z$).

Detection occurs at $-dV_R = 1\%$ of V_{Rmax}. This detection corresponds with a –dV in the battery in accordance with:

$dV_{bat} = (V_Z/V_{B(max)}) - 1$ (in %).

If the Zener voltage is half the maximum battery voltage, the dV_B detection will be at –0.5%.

Design example for six-cells' battery and 5% –dV cut-off

Conditions:

- maximum battery voltage (1.7 V/cell) = 10.2 V
- sense network current ≈ 300 µA
- maximum monitor sense voltage $V_{AC} = 3.6$ V (<3.85 V)

For $-dV \approx 0.5\%$, a Zener voltage of about half the battery voltage is required; choose $V_Z = 5$ V. Now V_R at top level is ≈5.2 V and the required divider factor (V_{AC}/V_R) is 0.69. R1 and R2 become 5.6 kΩ and 12 kΩ respectively (see Figs 6 and 7).

During charge the battery voltage rises with a minimum rate of ≈8 mV per minute for 6 cells under 1 C charge. The temperature coefficient of the Zener diode and its ambient temperature change should not cause a premature full detection.

With the following equation the allowance of the Zener-diode temperature change can be derived:

$\text{K/minute} < \{0.5\%(V_B - V_Z) + (8\ \text{mV/minute})\}/S_Z$

At the early state of charging, $V_B \approx 8.4$ V for 6 cells; 0.5% is a safe value for 1% detection in the TEA1100 and $V_Z = 5$ V.

Assuming a S_Z of +0.5 mV/K (typ. 0.2 mV/K for e.g. Zener diode PLVA450A), this results in maximum 50 K/minute.

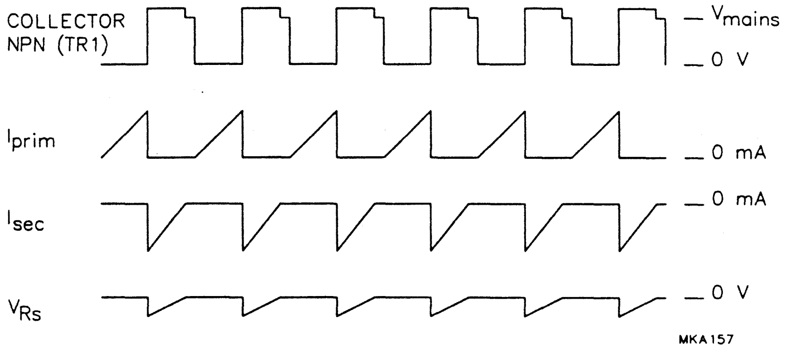


Fig.4 SMPS voltage and current waveforms.

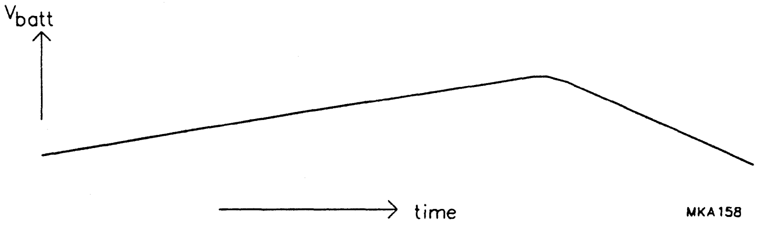


Fig.5 Battery voltage during charging.

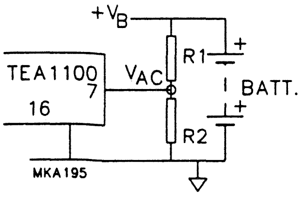


Fig.6 Basic circuit for direct battery sensing.

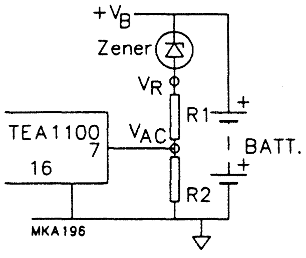


Fig.7 Basic circuit showing the use of a Zener diode.

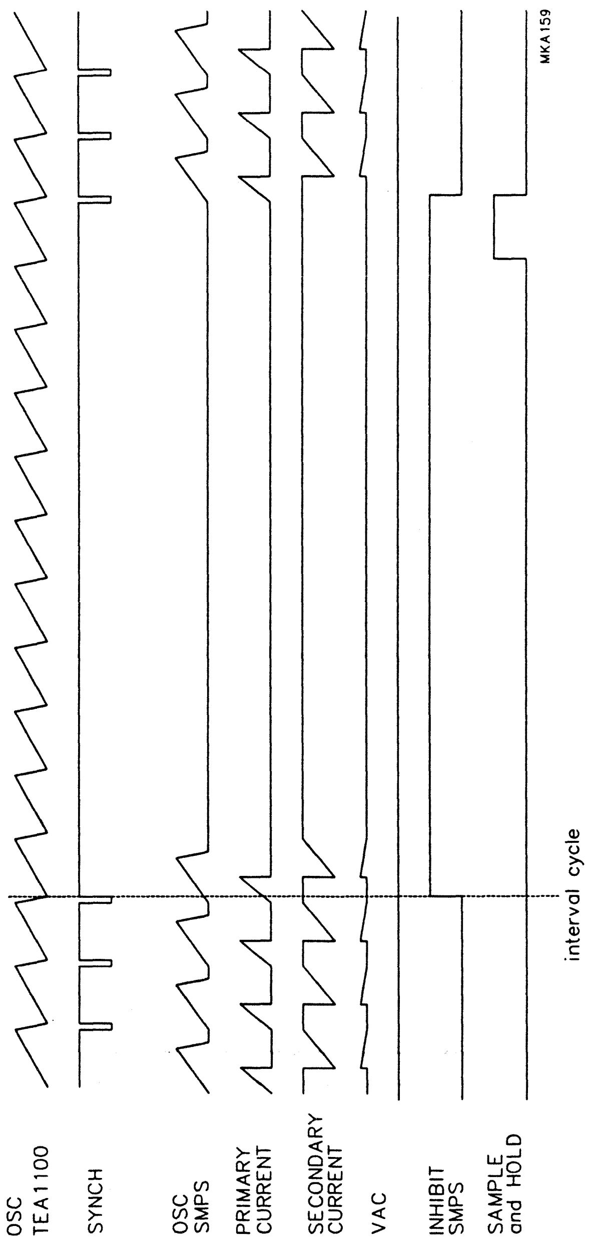
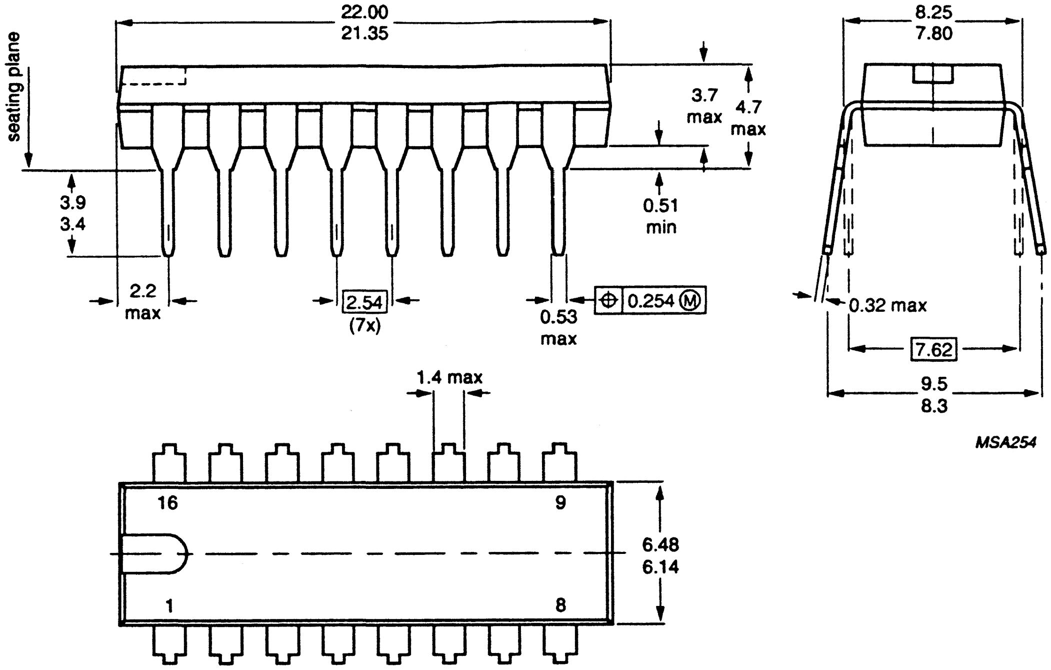


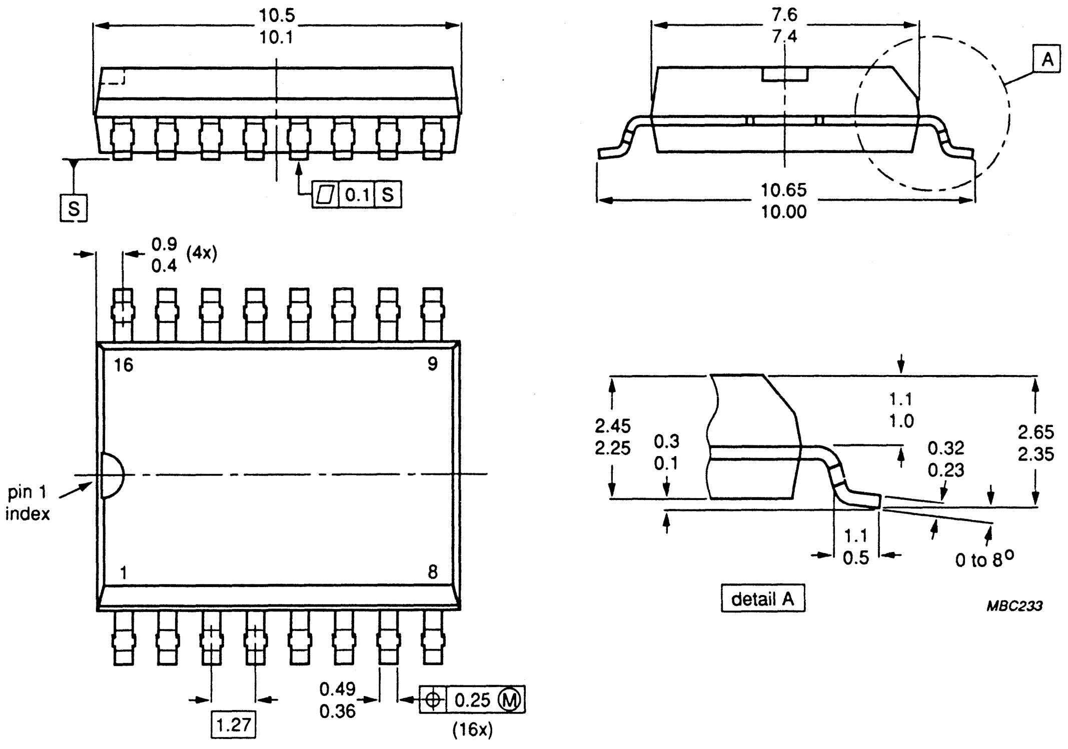
Fig.8 Synchronization waveforms.

PACKAGE OUTLINES



Dimensions in mm.

Fig.9 16-lead dual in-line; plastic (SOT38G).



Dimensions in mm.

Fig.10 16-lead mini-pack; plastic (SO16L; SOT162A).

SOLDERING

Plastic dual in-line packages

BY DIP OR WAVE

The maximum permissible temperature of the solder is 260 °C; this temperature must not be in contact with the joint for more than 5 s. The total contact time of successive solder waves must not exceed 5 s.

The device may be mounted up to the seating plane, but the temperature of the plastic body must not exceed the specified storage maximum. If the printed-circuit board has been pre-heated, forced cooling may be necessary immediately after soldering to keep the temperature within the permissible limit.

REPAIRING SOLDERED JOINTS

Apply the soldering iron below the seating plane (or not more than 2 mm above it). If its temperature is below 300 °C, it must not be in contact for more than 10 s; if between 300 and 400 °C, for not more than 5 s.

SOLDERING

Plastic mini-packs

BY WAVE

During placement and before soldering, the component must be fixed with a droplet of adhesive. After curing the adhesive, the component can be soldered. The adhesive can be applied by screen printing, pin transfer or syringe dispensing.

Maximum permissible solder temperature is 260 °C, and maximum duration of package immersion in solder bath is 10 s, if allowed to cool to less than 150 °C within 6 s. Typical dwell time is 4 s at 250 °C.

A modified wave soldering technique is recommended using two solder waves (dual-wave), in which a turbulent wave with high upward pressure is followed by a smooth laminar wave. Using a mildly-activated flux eliminates the need for removal of corrosive residues in most applications.

BY SOLDER PASTE REFLOW

Reflow soldering requires the solder paste (a suspension of fine solder particles, flux and binding agent) to be applied to the substrate by screen printing, stencilling or pressure-syringe dispensing before device placement.

Several techniques exist for reflowing; for example, thermal conduction by heated belt, infrared, and vapour-phase reflow. Dwell times vary between 50 and 300 s according to method. Typical reflow temperatures range from 215 to 250 °C.

Preheating is necessary to dry the paste and evaporate the binding agent. Preheating duration: 45 min at 45 °C.

REPAIRING SOLDERED JOINTS (BY HAND-HELD SOLDERING IRON OR PULSE-HEATED SOLDER TOOL)

Fix the component by first soldering two, diagonally opposite, end pins. Apply the heating tool to the flat part of the pin only. Contact time must be limited to 10 s at up to 300 °C. When using proper tools, all other pins can be soldered in one operation within 2 to 5 s at between 270 and 320 °C. (Pulse-heated soldering is not recommended for SO packages.

For pulse-heated solder tool (resistance) soldering of VSO packages, solder is applied to the substrate by dipping or by an extra thick tin/lead plating before package placement.

DEFINITIONS

Data sheet status	
Objective specification	This data sheet contains target or goal specifications for product development.
Preliminary specification	This data sheet contains preliminary data; supplementary data may be published later.
Product specification	This data sheet contains final product specifications.
Limiting values	
Limiting values given are in accordance with the Absolute Maximum Rating System (IEC 134). Stress above one or more of the limiting values may cause permanent damage to the device. These are stress ratings only and operation of the device at these or at any other conditions above those given in the Characteristics sections of this specification is not implied. Exposure to limiting values for extended periods may affect device reliability.	
Application information	
Where application information is given, it is advisory and does not form part of the specification.	

NOTES

Philips Semiconductors – a worldwide company

Argentina IEROD, St. Juramento 1991 - 14 B, Buenos Aires.
Zip Code 1428, Phone/Fax: 54 1 7869367
Australia 34 Waterloo Road, NORTH RYDE, NSW 2113,
Tel, (02)4393322, Fax. (02)8054466
Austria Triester Str. 64, 1101 WIEN,
Tel. (0222)60 101-0, Fax. (0222)60 101-1975
Belgium 80 Rue Des Deux Gares, B-1070 BRUXELLES,
Tel. (02)5256111, Fax. (02)5257246
Brazil Rua do Rocia 220, SAO PAULO-SP, CEP 4552,
P.O. Box 7383, CEP 01051, Tel. (011)829-1166,
Fax. (011)829-1849
Canada DISCRETE SEMICONDUCTORS, 601 Milner Ave,
SCARBOROUGH, ONTARIO, M1B 1M8,
Tel. (416)292-5161
INTEGRATED CIRCUITS, 1 Eva Road, Suite 411, ETOBICOKE,
Ontario, M9C 4Z5, Tel. (416)626-6676
Chile Av. Santa Maria 0760, SANTIAGO,
Tel. (02)773816
Colombia Carrera 21 No. 56-17, BOGOTA, D.E., P.O. Box 77621,
Tel. (01)2497624
Denmark Prags Boulevard 80, PB 1919, DK-2300 COPENHAGEN S,
Tel. 32-883333, Fax. 32-960125
Finland Sinikalliontie 3, SF-2630 ESPOO,
Tel. 358-0-50261, Fax. 358-0-520039
France 117 Quai du Président Roosevelt, 92134 ISSY-LES-MOULINEAUX Cedex, Tel. (01)40938000,
Fax. (01)40938127
Germany Burchardstrasse 19, D-2 HAMBURG 1,
Tel. (040)3296-0, Fax. (040)3296213
Greece No. 15, 25th March Street, GR 17778 TAVROS,
Tel. (01)4894339/4894911
Hong Kong 15/F Philips Ind. Bldg., 24-28 Kun Yip St., KWAI CHUNG,
Tel. (0)-4245121, Fax. (0)4806960
India Shivsagar Estate 'A' Block, P.O. Box 6598, 254-D Dr. Annie
Besant Rd., BOMBAY-40018, Tel. (022)4921500-4921515,
Fax. (022)494 19063
Indonesia Setiabudi 11 Building, 6th Fl., Jalan H.R. Rasuna Said,
P.O. Box 223/KBY, Kuningan, JAKARTA 12910,
Tel. (021)517995
Ireland Newstead, Clonskeagh, DUBLIN 14,
Tel. (01)693355, Fax. (01)697856
Italy Piazza IV Novembre 3, 1-20124 MILANO,
Tel. (02)67522642, Fax. (02)67522648
Japan Philips Bldg 13-37, Kohnan 2-chome, Minato-ku, TOKIO 108,
Tel. (03)813-3740-5101, Fax. (03)81337400570
Korea (Republic of) Philips House, 260-199 Itaewon dong,
Yongsan-ku, SEOUL, Tel. (02)794-5011, Fax. (02)798-8022
Malaysia 3 Jalan SS15/2A SUBANG, 47500 PETALING JAYA,
Tel. (03)7345511, Fax. (03)7345494
Mexico Paseo Triunfo de la Republica, No. 215 Local 5, Cd Juarez
CHI HUA HUA 32340, Tel. (16)18-67-01/02
Netherlands Postbus 90050, 5600 PB EINDHOVEN,
Tel. (040)783749, Fax. (040)788399
New Zealand 2 Wagener Place, C.P.O. Box 1041, AUCKLAND,
Tel. (09)894-160, Fax. (09)897-811
Norway Box 1, Manglerud 0612, OSLO,
Tel. (02)748000, Fax. (02)748341
Pakistan Philips Markaz, M.A. Jinnah Rd., KARACHI-3,
Tel. (021)725772
Peru Carretera Central 6.500, LIMA 3, Apartado 5612,
Tel. 51-14-350059
Philippines PHILIPS SEMICONDUCTORS PHILIPPINES Inc,
106 Valero St. Salcedo Village, P.O. Box 911, MAKATI,
Metro MANILA, Tel. (63-2)810-0161, Fax. (63-2)8173474
Portugal Av. Eng. Duarte Pacheco 6, 1009 LISBOA Codex,
Tel. (019)683121, Fax. (019)658013
Singapore Lorong 1, Toa Payoh, SINGAPORE 1231,
Tel. 3502000, Fax. 2516500
South Africa 195-215 Main Road, JOHANNESBURG 2000,
P.O. Box 7430, Tel. (011)8893911, Fax. (011)8893191
Spain Balmes 22, 08007 BARCELONA,
Tel. (03)3016312, Fax. (03)3014243
Sweden Tegeluddsvägen 1, S-11584 STOCKHOLM,
Tel. (0)8-7821000, Fax. (0)8-6603201
Switzerland Allmendstrasse 140-142, CH-8027 ZÜRICH,
Tel. (01)4882211, FAX. (01)4828595
Taiwan 581 Min Sheng East Road, P.O. Box 22978,TAIPEI 10446,
Tel. 886-2-5097666, Fax. 886 25005899
Thailand 283 Silom Road, P.O. Box 961, BANGKOK,
Tel. (02)233-6330-9
Turkey Talatpasa Cad. No. 5, 80640 LEVENT/ISTANBUL,
Tel. (01)1792770, Fax. (01)1693094
United Kingdom Philips Semiconductors Limited, P.O. Box 65,
Philips House, Torrington Place, LONDON, WC1E 7HD,
Tel. (071)4364144, Fax. (071)3230342
United States INTEGRATED CIRCUITS, 811 East Arques Avenue,
SUNNYVALE, CA 94088-3409, Tel. (800)227-1817, Ext. 900,
Fax. (408)991-3581
DISCRETE SEMICONDUCTORS, 2001 West Blue Heron Blvd.,
P.O. Box 10330, RIVIERA BEACH, FLORIDA 33404,
Tel. (407)881-3200, Fax. (407)881-3300
Uruguay Coronel Mora 433, MONTEVIDEO,
Tel. (02)70-4044
Venezuela Calle 6, Ed. Las Tres Jotas, CARACAS, 1074A,
App. Post. 78117, Tel. (02)2417509
Zimbabwe 62 Mutare Road, HARARE, P.O. Box 994,
Tel. 47211

For all other countries apply to: Philips Semiconductors,
International Marketing and Sales, Building BAF-1,
P.O. Box 218, 5600 MD, EINDHOVEN, The Netherlands,
Telex 35000 phtcnl, Fax. +31-40-724825

Printed in the Netherlands

9397 349 90011

Philips Semiconductors

PHILIPS

PHILIPS

DATA SHEET

TEA1101; TEA1101T
Battery monitor for NiCd and NiMH chargers

Preliminary specification
Supersedes data of February 1993
File under Integrated Circuits, IC03

May 1993

Philips Semiconductors

PHILIPS

FEATURES

- Accurate detection of fully charged batteries by currentless –dV sensing
- Digital filtering of the battery voltage to avoid false –dV triggering
- Minimum and maximum temperature guarding by means of an NTC resistor
- Battery checking to protect against short-circuit and open batteries
- Battery monitor allows recharging different battery-pack voltages
- Tracking of maximum fast charging time with fast charging current level
- Accurate regulation of charge current settings in co-operation with a switched mode power supply or DC current source
- Both DC and PWM outputs with polarity switch
- Adjustable fast charge level (1 C to 5 C)
- Adjustable pulsating trickle charge level (0.05 C to 0.25 C)
- Large operating temperature range.

APPLICATIONS

- Charge systems for NiCd and NiMH batteries.

GENERAL DESCRIPTION

The TEA1101 is manufactured in a BICMOS process intended to be used as a battery monitor circuit in charge systems for NiCd and NiMH batteries.
The circuit has to be situated on the secondary side in mains-isolated systems where it monitors the battery voltage and the charge current. The circuit drives, by means of an opto-coupler or a pulse transformer interface, an SMPS circuit, situated on the primary side of the system, thus controlling the charge current of the batteries.
The circuit can drive the external power transistor in switched mode systems, which have a DC power source, via a driver stage.

QUICK REFERENCE DATA

SYMBOL	PARAMETER	CONDITIONS	MIN.	TYP.	MAX.	UNIT
V_P	positive supply voltage		5.65	–	11.5	V
I_P	supply current	outputs off	–	–	4.3	mA
V_{VAC}	voltage range of battery-full detection		0.385	–	3.85	V
dV_{VAC}/V_{VAC}	–dV detection level w.r.t. top value	note 1	–	0.25	–	%
I_{VAC}	input current battery monitor		–	–	1	nA
V_{VAC}	voltage protection					
	battery low		–	0.3	–	V
	battery high		–	4.25	–	V
	charging level	$I_{charge} = R1/R_s \times I$; see Fig.3				
I_{ref}	fast	$I = I_{ref}$	20	–	100	µA
I_n	normal	$I = 1/p \times 0.1 \times I_n$ (p = prescale factor)	10	–	50	µA
f_{OSC}	oscillator frequency		10	–	100	kHz

Note

1. The –dV detection level can be adjusted by use of an external voltage regulator diode to increase the sensitivity.

ORDERING INFORMATION

EXTENDED TYPE NUMBER	PACKAGE			
	PINS	PIN POSITION	MATERIAL	CODE
TEA1101	16	DIL	plastic	SOT38G
TEA1101T	16	SO16L	plastic	SOT162A

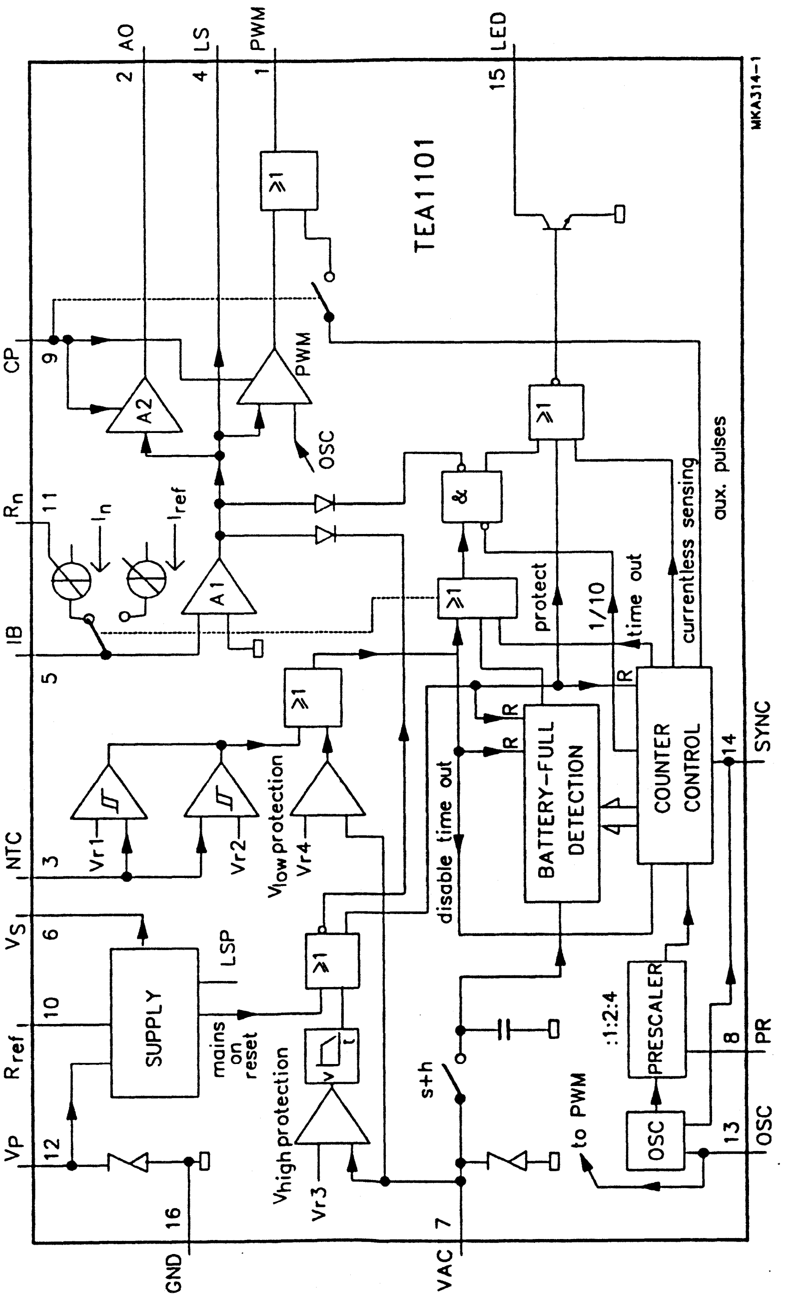


Fig.1 Block diagram.

PINNING

SYMBOL	PIN	DESCRIPTION
PWM	1	pulse width modulator
AO	2	analog output
NTC	3	temperature sensor input
LS	4	loop stability
IB	5	charge current
V_S	6	stabilized supply voltage
VAC	7	battery voltage
PR	8	prescaler
CP	9	change polarity
R_{ref}	10	reference resistor
R_n	11	normal charge reference resistor
V_P	12	positive supply voltage
OSC	13	oscillator input
SYNC	14	synchronization input
LED	15	LED output
GND	16	ground

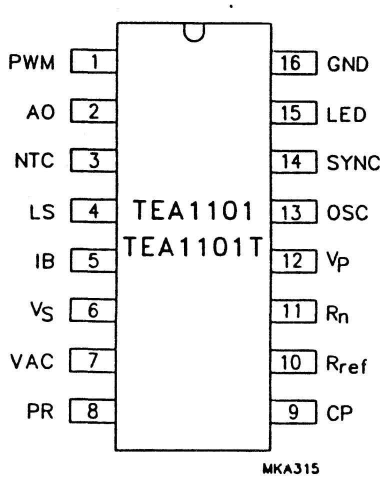


Fig.2 Pin configuration.

FUNCTIONAL DESCRIPTION

The operation of the circuit will be explained with the aid of Fig.1 (block diagram) and Fig.3 (application diagram). The circuit is divided into several blocks which are described separately.

Supply block

The circuit requires a supply voltage on pin V_P with a value between 5.65 and 11.5 V. Above 6.4 V typical, the circuit starts up assuming that mains is connected to the system and the charge session begins. This supply can be generated by a separate winding on the transformer, as shown in Fig.3, in either the flyback or the forward stroke. Another possibility is rectification from the mains secondary winding (at the connection D1 and L2). Considerations for choosing the method of supplying the IC are:

- Supply voltage range of 5.65 to 11.5 V under all circumstances (also during the 90% pause at normal charging, the standby current then is 1 mA typical)
- Maximum battery voltage (flyback stroke)
- Minimum power delivered by the primary SMPS (normal charging)

The supply block delivers the following outputs:

- By using an external resistor R_{ref} at pin 10 a reference current is obtained which defines all external related currents (charge reference currents, oscillator)
- Externally available 4.25 V stabilized voltage source (V_S). This source is used internally for a large part of the circuit and can be used to set the NTC biasing and to supply other external circuitry. V_S is cut off in the 90% pause during normal charging
- Low Supply voltage Protection signal (LSP). When the supply voltage is lower than 5.25 V typical, there is enough supply voltage left to switch off the power regulation and hereafter the IC current is limited to the start level of 35 µA typical
- Mains on reset pulse resets all digital circuitry after a start or restart due to an interrupted supply (V_P).

Charge current regulation

The charge current has to be sensed by means of a low-ohmic resistor in series with diode D1. The waveform on resistor R_s (see Fig.5 for a flyback converter) has the form of a negative-going ramp and after filtering a negative DC voltage is obtained. A positive voltage across resistor R1 is created by means of the current sources set by the pins R_{ref} and R_n. The error amplifier A1 references the result to ground and via the regulation loop of the SMPS, the secondary current will be regulated to a value which is defined by:

$I_{ch} \times R_s = R1 \times I_{ref}$ (fast charge) or,
$I_{ch} \times R_s = R1 \times I_n$ (normal charge)

The I_{ref} current is the fast charging reference current, the I_n current is used for regulation after a full battery is detected. The I_{ref} current is the reference current set by R_{ref} while I_n is dependent on the resistor at pin R_n. With no resistor on pin R_n, the I_n current has a default value which is half the I_{ref} current.
By choosing the correct resistor values R_s, R1, R_{ref} and R_n, a wide range of charge currents can be set plus a wide range of the ratio fast charge current as a function of normal charge current. For determination of the normal charge current the 1:10 duty cycle and the programmable prescale factor (p) should be taken into account (see Logic block); $I_n = 1/p \times 0.1 \times I_n$. The output of amplifier A1 is available at the loop stability pin (LS), so the time constant of the SMPS loop can be set at the secondary side of the system.

NTC block

The voltage at the NTC pin is compared with two reference voltages. When the NTC voltage is between V_{r1} and V_{r2}, the charge current regulation is unaffected. When the NTC voltage is outside this window, the power of the SMPS is reduced to the normal charge level.
The NTC input can be used for temperature protection as shown in Fig.3 (application diagram) by using a suitable NTC resistor. To avoid switching on and off with temperature, a hysteresis is built in for both levels.

Output drivers

The SMPS regulation signal is available at different pins:

- Analog voltage output (push or pull) at AO (pin 2) to drive an opto-coupler in mains separated applications when an external resistor is connected between AO and the opto-coupler. The maximum current through the opto-coupler diode is 2 mA. The voltage gain of amplifier A2 is: $A = (V_{LS} - 1.4) \times 4$ and is typically 12 dB. The voltage at AO can also be used to directly drive a PWM input of an SMPS circuit. During 'inhibit SMPS' the AO output is fixed to zero charge current for currentless sensing
- The LS voltage is compared internally with the oscillator voltage to deliver a pulse width modulated output at PWM (pin 1) to drive an output device in a DC/DC converter application via a driver stage. The PWM output is latched to prevent multi-pulsing. Moreover with the latch a kind of current mode control is possible. The maximum duty factor is internally fixed to 78% (typical). The 'PWM' output can be used for synchronization and duty factor control of a primary SMPS via a pulse transformer (the SMPS inhibit and auxiliary pulses are also available at pin PWM)
- The AO and PWM outputs can be changed in polarity by programming the change polarity pin CP. The PWM output in the on-state pushes current (CP = 0) or pulls current (CP = 1). The appearance of the auxiliary pulses at pin PWM can also be programmed with CP.

The 'LED' output pin offers the following output signals:

- 10/90% signal for driving a LED when the duty factor is too small during the 10% time. This occurs when there is a large difference between fast and normal charge currents. The LED frequency is $f_{LED} = 2^{-12} \times 1/p \times f_{OSC}$
- An SMPS inhibit period (duration 10 OSC pulses) for currentless VAC sensing
- VAC high voltage protection signals.

Battery monitor

Two batteries can be connected directly to the 'VAC' pin (Voltage ACcumulator). At higher battery voltages it is advised to divide the battery voltage with a factor by an external resistor tap, before offering this to pin VAC. It is also possible to take a tap on the chain of batteries. The VAC voltage range has to be between 0.385 V and 3.85 V.
The VAC voltage is sampled at a low cycle frequency ($f_{cycle} = 2^{-16} \times f_{OSC}$) and the analog value of VAC is digitized and stored in a register. One cycle later, the digitized value is converted back to the analog value and compared with the actual value of VAC. If the actual value is higher, then the new VAC voltage is stored in the register, otherwise no conversion is carried out. Thus the VAC top value is stored and it is possible to detect an increasing VAC indicating 'not yet full batteries' or decreasing VAC indicating that the batteries are probably fully charged. The circuit waits until the battery voltage has dropped 0.25% below the top value before indicating 'full batteries'. However, by applying a voltage regulator diode in the battery voltage sense-line (see Fig.8) an increased sensitivity of the –dV detection level can be obtained, e.g. 0.125%.
In Fig.6 the battery voltage as

function of the charging time is shown. The negative slope depends on the charge current and the battery type.

The switching of the SMPS can cause interference on the battery voltage and therefore it has been necessary to stop the SMPS during the inhibit time (see Fig.9). This can be achieved automatically via the regulation pins AO and PWM or by using the SYNC output of the logic block. The SMPS is stopped for 10 periods at the end of which sampling is carried out. The VAC voltage will now be sensed currentless. To avoid false decisions concerning a falling VAC voltage, VAC is digitally filtered and analog stored in a sample-and-hold circuit. This approach ensures, even at very high –dV sensitivity (<0.25%) accurate detection of the battery full condition. Immediately after decisions and VAC digitizing takes place. The benefit of a sample-and-hold circuit is that at high frequencies the noise on the VAC voltage is filtered and the VAC manipulations like decisions and digitizing are carried out on the same VAC voltage available in the sample-and-hold circuit.

When a –dV is detected, the reference current I_{ref} is switched off, the normal current I_n is switched on during 10% of regulation and the outputs are high-ohmic during 90%. This 1:10 ratio in active regulation, together with the ratio in reference currents (I_n as a function of I_{ref}), ensures that the resulting charge current is low enough to be allowed to flow through the batteries for a long time to overcome the self-discharge of the batteries without causing memory effects. If the prescale factor p is programmed, the I_n current has to be lowered with the p factor, so $I_n = 1/p \times 0.1 \times I_n$.

Protections

- The circuit goes into standby (not active, low current consumption) when the supply voltage is less than 5.25 V (LSP).
- When the divided battery voltage exceeds the V_{r3} level (nominal 4.25 V) this is recognized as open or removed batteries and the output control signals terminate to stop the SMPS operation. This over-voltage sensing is digitally filtered. In above events the 'battery full detector' and the 'counter/control' will be reset.
- When the divided battery voltage is less than V_{r4} (0.3 V), the circuit assumes short-circuited batteries, the charge current is reduced to the normal charge level. As soon as the voltage exceeds V_{r4}, the fast charging starts.
- The temperature protections are already mentioned in section NTC. In the event of short-circuited batteries or active temperature protections the 'battery full detector' is reset and the 'counter/control' is stopped.

Oscillator and control logic

The complete timing of the circuit is controlled by the oscillator.
The period time is defined by:
$T_{OSC} = 0.93 \times R_{ref} \times C_{OSC}$.

The counter block defines a maximum fast charge time called 'Time Out' (TO). As the charge current and the oscillator frequency (and thus the TO) are both set by R_{ref}, changing one affects the other. Initially the oscillator capacitor can be chosen such that the fast charge time is half the TO time. This means that in the event of a one hour (1C) charger, the TO signal occurs at 2 hours, in the event of a quarter of an hour (4C) charger, the TO signal is active after half an hour. After that the circuit switches over to normal charging.

To adapt the SMPS switching frequency in the synchronized mode to the required oscillator frequency of the timing logic, the timer logic is preceded by a programmable divider. The divider ratio can be set to 1, 2 or 4 (p factor) by means of the PR pin. Doing this means that the oscillator frequency can be increased with the factor p without changing TO.

Fast charging current:
$I_{ch} = R1/R_s \times V_{ref}/R_{ref}$
Time out:
$TO = 2^{26} \times 0.93 \times R_{ref} \times C_{OSC} \times p$
Normal charging current:
$I_{ch} = R1/R_s \times 1/p \times 0.1 \times V_{ref}/R_n$

The control block determines the following timing sequences:

- VAC sampling; this takes 1 clock pulse every interval cycle.

  The power converter is switched off during VAC sampling. As there are several types of converters, there also are several control signals available at :

  pin 'SYNC' for synchronization in analog voltage controlled primary SMPS circuits

  pin 'PWM' for digital controlled primary SMPS and DC/DC converters

  pin 'LED' in special applications
- Disabling –dV during $2^{-5} \times TO$ (3% of TO) for correct start with flat or inversely polarized batteries. Disabling is active at each fast charge cycle
- Maximum fast charging time (TO): the maximum timer is stopped during VAC low voltage protection and outside temperature range

- The normal charge duty cycle is 1/p x 0.1
- Auxiliary pulses to support the supply voltage of the primary SMPS circuit via pin PWM: the pulses can be programmed on and off at an appearance rate of $f_{OSC}/8$ with a duty cycle of 14%; programming is achieved by activating CP.

The timing logic and the –dV recognition circuitry are reset after each supply voltage failure and after a battery over-voltage recognition. The –dV circuit is also reset during normal charging.

The SYNC output delivers negative-going synchronization pulses which are suppressed during the sampling of the battery voltage. With these sync pulses the SMPS can be synchronized.The polarity of the sync pulses is chosen so that in the event of an open SYNC pin in the synchronization mode, the power is regulated to a minimum. During the VAC sampling the absense of sync pulses causes the SMPS to stop thus minimizing interference (see Fig.9, synchronization waveforms).

During the 90% pause, only the oscillator and the control logic are operative to save current. In the pause V_P is never allowed to become less than V_{LSP}. This would cause a 'mains-on-reset' and thus fast charging.

Programming

With pins 'CP' (change polarity) and 'PR' (prescaler) several functions can be programmed.

By defining the current (V_{ref}/R_{CP}) at pin CP, the following functions can be activated :

1	change polarity	CP = 0, normal polarity CP = 1, changed polarity
2	no auxilliary pulses at PWM	aux = 0
3	auxilliary pulses at f/8	aux = 8

CP PIN	FUNCTIONS	
	CP	aux
Open pin	0	0
10 µA	0	8
22 µA	1	0
57 µA	1	8

By defining the voltage at pin PR, the following functions can be activated :

PR PIN	FUNCTIONS
V_S	prescaler divide by 1
Open pin	prescaler divide by 2
Ground	prescaler divide by 4

Table 1 Formulae.

DESCRIPTION	SYMBOL	FORMULA	FUNCTION
Timing	T_{OSC}	$0.93 \times R_{ref} \times C_{OSC}$	repetition
	TO	$2^{-26} \times p \times T_{OSC}$	duration
	$T_{disable}$	$2^{-5} \times TO$	duration
	$T_{LED} = T_{trickle}$	$2^{12} \times p \times T_{OSC}$	repetition
	$T_{LED} = T_{trickel}$	$3/4 \times 2^9 \times T_{OSC}$	duration
	$T_{inhibit}$	$2^{16} \times T_{OSC}$	repetition
	$T_{inhibit}$	$10 \times T_{OSC}$	duration
Charge currents	I_{fast}	$R1/R_s \times V_{ref}/R_{ref}$	
	I_{normal}	$R1/R_s \times 1/p \times 0.1V_{ref}/R_n$	

LIMITING VALUES

In accordance with Absolute Maximum Rating System (IEC 134); note 1.

SYMBOL	PARAMETER	MIN.	MAX.	UNIT
Voltages				
V_P	positive supply voltage (pin12)	−0.5	13.2	V
V_{LED}	LED voltage (pin 15)	−0.5	13.2	V
$V_{1,8,4,3}$	voltage at PWM (pin 1), PR (pin 8), LS (pin 4), NTC (pin 3)	−0.5	V_P	V
V_{IB}	voltage at IB (pin 5)	−0.5	+1	V
Currents				
I_{VS}	current at V_S (pin 6)	−3	+0.01	mA
I_{LED}	current at LED (pin 15)	−	25	mA
I_{AO}	current at AO (pin 2)	−5	+5	mA
I_{PWM}	current at PWM (pin 1)	−15	+15	mA
I_{SYNC}	current at SYNC (pin 14)	−2	+2	mA
$I_{11,10,9}$	current at R_n (pin 11), R_{ref} (pin 10), CP (pin 9)	−1	+0.01	mA
$I_{4,5,7}$	current at LS (pin 4), IB (pin 5), VAC (pin 7)	−1	+1	mA
I_P	current at V_P (pin 6)	−	15	mA
Dissipation				
P_{tot}	total power dissipation at T_{amb} = 85 °C			
	SOT38G	−	0.6	W
	SOT162A	−	0.3	W
Temperatures				
T_{amb}	operating ambient temperature	−20	+85	°C
T_j	junction temperature	−	+150	°C
T_{stg}	storage temperature	−55	+150	°C

Note

1. All voltages with respect to ground; positive currents flow into the IC; all pins not mentioned in the voltage list are not allowed to be voltage driven. The voltage ratings are valid provided other ratings are not violated; current ratings are valid provided the power rating is not violated.

CHARACTERISTICS

V_P = 10 V; T_{amb} = 25 °C; R_{ref} = 33 kΩ; R_n = 68 kΩ; C_{OSC} = 1 nF; CP open-circuit; PR connected to V_S; unless otherwise specified.

SYMBOL	PARAMETER	CONDITIONS	MIN.	TYP.	MAX.	UNIT
Supply (V_P, V_S, R_{ref})						
V_P	supply voltage range		5.65	–	11.5	V
V_{PC}	clamping voltage	I_{PC} = 10 mA	11.5	–	12.8	V
V_{PS}	start voltage		6.1	6.4	6.7	V
V_{PLSP}	low supply protection level		4.85	5.25	5.65	V
V_{PLSPH}	hystereris of V_{PLSP}		0.5	0.95	–	V
I_P	supply current	outputs off	–	–	4.3	mA
I_{PP}	supply pause current	V_P = 6 V	–	–	1.71	mA
I_{PSB}	standby current	V_P = 4 V	–	35	45	μA
V_S	source voltage (stabilized)	I_S = 1 mA	4.03	4.25	4.46	V
V_{ref}	reference voltage	I_{ref} = 20 μA	1.18	1.25	1.31	V
TC	temperature coefficient of V_{ref}	T_{amb}= 0 to 45 °C	–	±100	±200	ppm/K
PSRR	power supply rejection ratio of V_{ref}	f = 100 Hz; dV_P = 2 V (p-p); V_P = 8 V	−46	–	–	dB
ΔV_{ref}	voltage difference	dI_S = 1 mA	–	–	5	mV
I_{Rref}	current range of R_{ref}		10	–	100	μA
Charge current regulation (IB, R_n, R_{ref})						
V_n	voltage at pin R_n	I_n = 10 μA; I_{ref} = 20 μA	1.17	1.25	1.32	V
I_n	current range at R_n		5	–	50	μA
I_{IB}/I_{ref}	input current ratio normal charging fast charging	R_n not connected V_{IB} = 0 V_{IB} = 0	 0.475 0.95	 0.5 1	 0.525 1.05	
I_{IB}/I_n	input current ratio normal charging	R_n connected	 0.90	 0.97	 1.04	
V_{thIB}	threshold voltage at IB T_{amb} = 25 °C T_{amb} = 0 to 45 °C		 −2 −3	 – –	 +2 +3	 mV mV
NTC input						
V_{NTCSPH}	switching protection voltage on high temperatures		0.75	0.81	0.87	V
V_{NTCHH}	hysteresis of V_{NTCSPH}		60	90	120	mV
V_{NTCSPL}	switching protection voltage on low temperatures		2.78	3.00	3.20	V
V_{NTCHL}	hysteresis of V_{NTCSPL}		65	100	135	mV
I_{NTC}	input current	V_{NTC} = 2 V	−5	–	+5	μA

SYMBOL	PARAMETER	CONDITIONS	MIN.	TYP.	MAX.	UNIT
Output drivers (AO, LS, PWM, LED)						
$I_{AOsource}$	source current	V_{AO} = 3 V; CP = 0	–	–	–2	mA
I_{AOsink}	sink current	V_{AO} = 0.5 V; CP = 1	2	–	–	mA
g_m	transconductance A1	V_{IB} = 50 mV	–	300	–	µS
G_{v1}	voltage gain A1 x A2	V_{AO} = 2 V (p-p)	–	72	–	dB
G_{v2}	voltage gain A2	V_{AO} = 2 V (p-p)	–	12	–	dB
$I_{LSsource}$	maximum source current	V_{LS} = 2.25 V	–25	–21	–16	µA
I_{LSsink}	maximum sink current	V_{LS} = 2.25 V	16	21	25	µA
I_{PWMH}	HIGH level output current	V_{PWM} = 3 V	–18	–14	–10	mA
I_{PWML}	LOW level output current	V_{PWM} = 0.5 V	7	12	17	mA
$I_{PWMleak}$	leakage current	V_{PWM} = 4.25 V	–	0.2	10	µA
δ_{PWM}	maximum duty cycle		70	78	86	%
δ_{PWMaux}	auxiliary pulse duty cycle		12.6	14	15.4	%
V_{LEDsat}	saturation voltage	I_{LED} = 15 mA	–	–	600	mV
$I_{LEDleak}$	leakage current	V_{LED} = 10 V	–	–	5	µA
Battery monitor (VAC)						
I_{VAC}	input current	V_{VAC} = 4.25 V	–	1	–	nA
V_{VAC}	voltage range of –dV detection		0.385	–	3.85	V
dV_{VAC}/V_{VAC}	–dV detection level w.r.t. top level	V_{VAC} = 2 V	–	0.25	–	%
ΔV_{VAC}	resolution –dV		0.42	0.6	0.78	mV
T_{-dV}	temperature range of –dV detection		0	–	50	°C
Protections (VAC)						
V_{VACLBP}	low battery voltage protection		–	0.3	0.33	V
V_{VACHBP}	high battery voltage protection	with respect to V_P	–	0	150	mV
Oscillator, logic (OSC, SYNCH)						
V_{OSCH}	oscillator switching level HIGH		–	2.5	–	V
V_{OSCL}	oscillator switching level LOW		–	1.5	–	V
K	period time	$T_{OSC} = K \times R_{ref} \times C_{OSC}$	0.84	0.93	1.02	
f_{OSC}	oscillator frequency range		10	–	100	kHz
V_{SYNCH}	SYNC output level HIGH	I_{SYNCH} = –0.4 mA	3.4	–	–	V
V_{SYNCL}	SYNC output level LOW	I_{SYNCL} = 0.4 mA	–	–	0.85	V

SYMBOL	PARAMETER	CONDITIONS	MIN.	TYP.	MAX.	UNIT
Programming (CP)						
I_{CP}	programming currents					
	CP = 0; aux = 0	R_{CP} = 330 kΩ	–	–	4.2	µA
	CP = 0; aux = 8	R_{CP} = 120 kΩ	9.4	10.4	11.4	µA
	CP = 1; aux = 0	R_{CP} = 56 kΩ	20.0	22.3	24.5	µA
	CP = 1; aux = 8	R_{CP} = 22 kΩ	51.1	56.8	62.5	µA

QUALITY SPECIFICATION

General quality specification for integrated circuits: UZW-B0/FQ-0601.

Note: For the synchronization pin (14), the ESD positive zap voltage is restricted to a maximum of 1000 V.

TEST AND APPLICATION INFORMATION

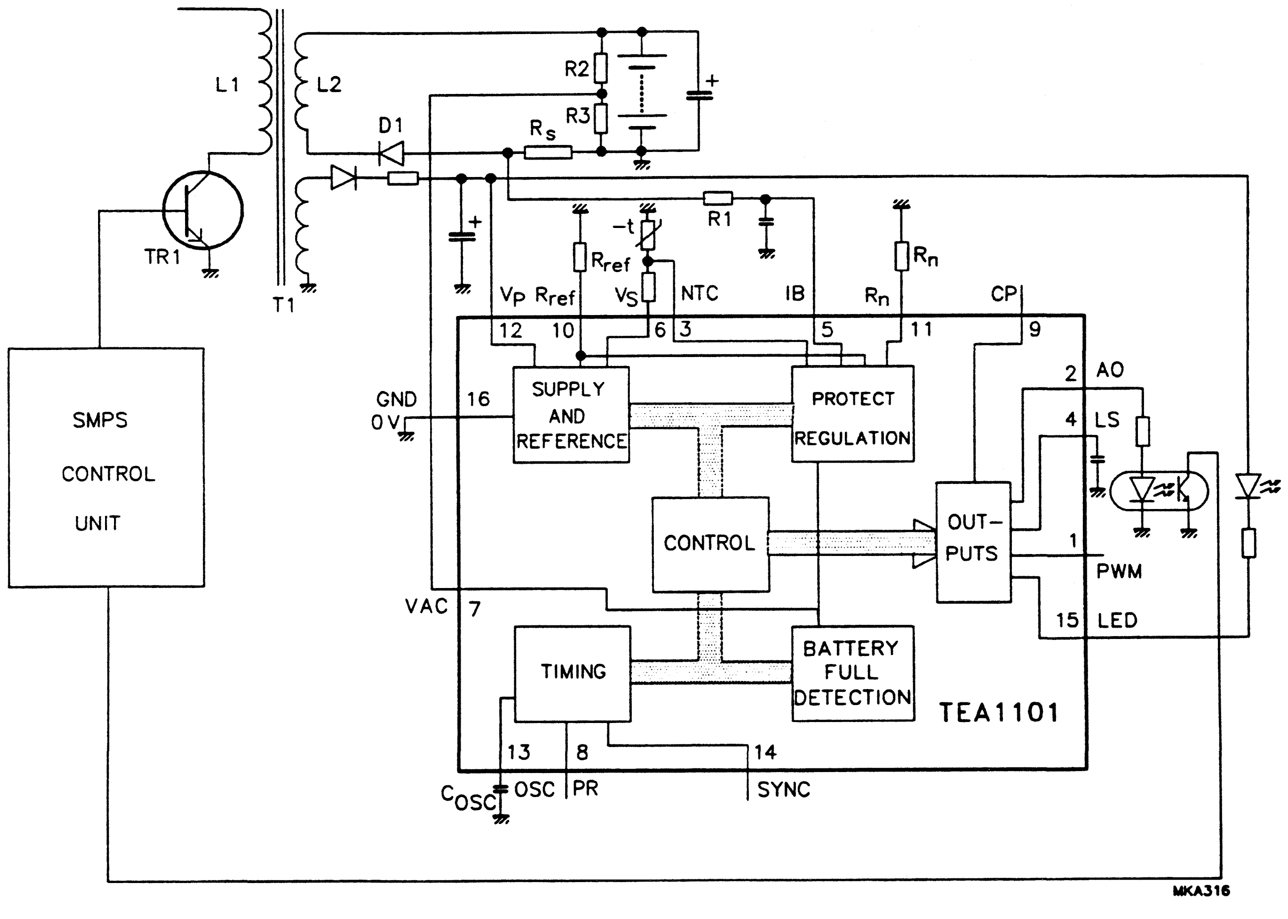


Fig.3 Application diagram.

Notes to Fig.3

1. Signaling the status of the charging session can be achieved by an LED-diode-resistor combination parallel to L2 (transformer T1). During the fast charging period the LED will burn continuously. During normal charging the LED will switch with the 10/90% rhythm. With mains-off the LED is off, thus not discharging the batteries. If at normal charging the duty cycle is too low during the 10% because of a very large difference between the fast charge and the normal charge levels, the LED can be driven by the LED pin.
2. With R_s = 50 MΩ and a required fast charging current level of 6 A (5C for 1.2 Ah batteries), the average current sense level is 300 mV. Power dissipation in R_s = 1.8 W.

   With a 3 kΩ resistor for R1, the required I_{ref} current is 300 mV/3 kΩ = 100 µA. For a normal charge level of 0.25C (300 mA) the voltage drop over R_s is 15 mV. Taking into account the duty cycle of 10%, the voltage drop over R1 = 150 mV. Thus the I_{Rn} current has to be 150 mV/3 kΩ = 50 µA (p = 1).

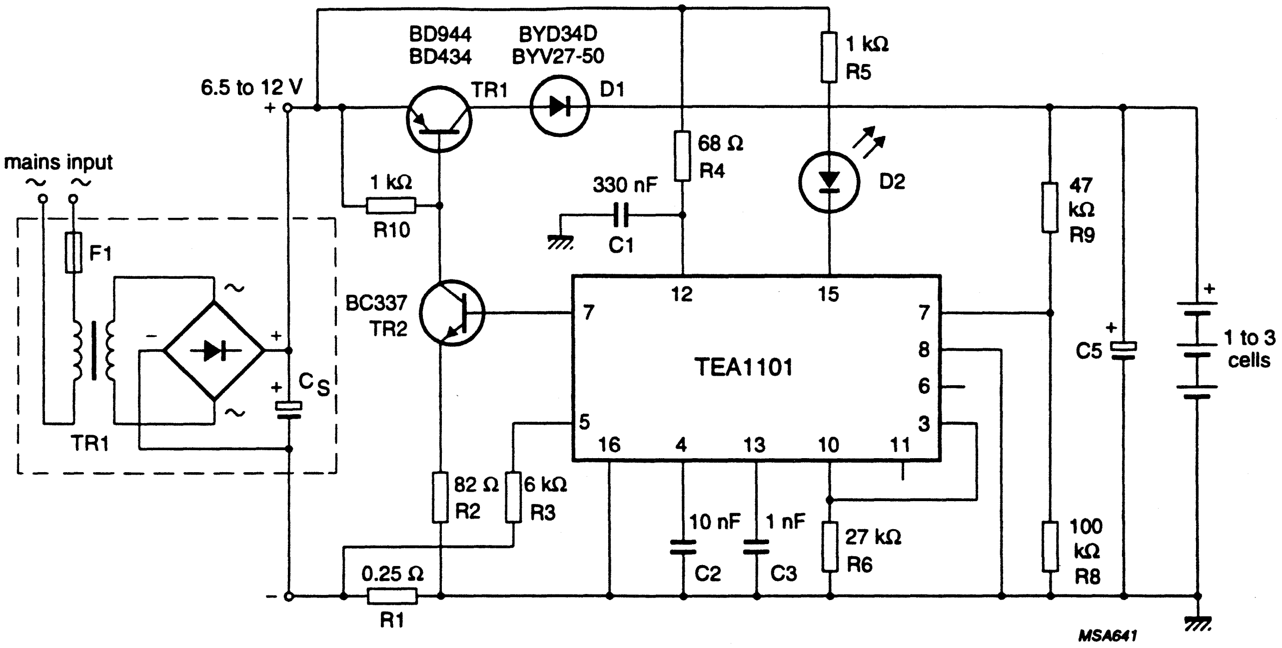


Fig.4 Battery charger with linear regulator.

Method to increase –dV sensitivity

The basic, direct battery sensing via a resistive divider, which adapts the battery voltage within the V_{AC} range, is shown in Fig.7. Detection occurs at –dV = 0.25% of $V_{B(max)}$.

The position of the Zener diode is shown in Fig.8. The TEA1101 now senses the voltage V_R, which is the battery voltage minus the Zener-diode voltage ($V_R = V_B - V_Z$).

Detection occurs at $-dV_R$ = 0.25% of V_{Rmax}. If the Zener voltage is half the maximum battery voltage, the dV_B detection will be at –0.125%.

Design example for six-cell battery and 0.125% –dV cut-off

Conditions:

- Maximum battery voltage (1.7 V/cell) = 10.2 V
- Sense network current ≈ 300 μA
- Maximum monitor sense voltage V_{AC} = 3.6 V (< 3.85 V).

For –dV ≈ 0.125%, a Zener voltage of about half the battery voltage is required; choose V_Z = 5 V. Now V_R at top level is ≈ 5.2 V and the required divider factor (V_{AC}/V_R) is 0.69. R1 and R2 become 5.6 kΩ and 12 kΩ respectively (see Figs 6 and 7).

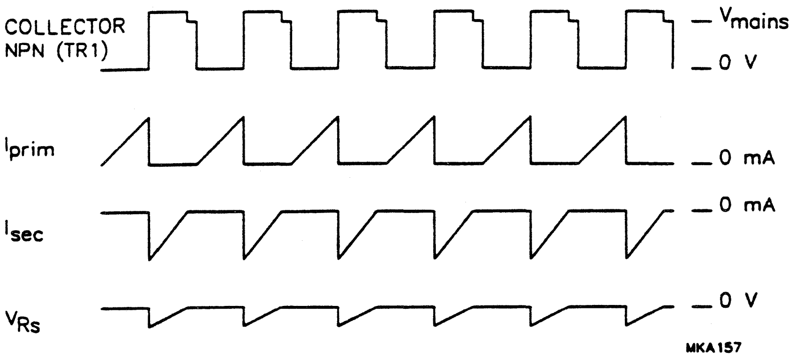


Fig.5 SMPS voltage and current waveforms.

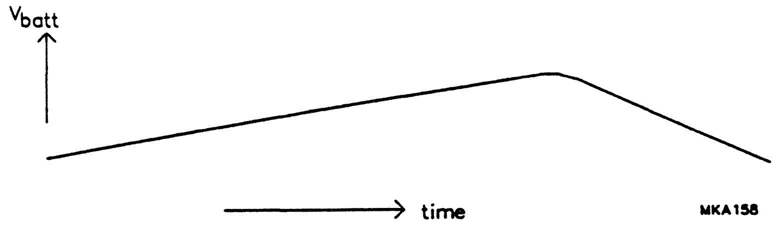


Fig.6 Battery voltage during charging.

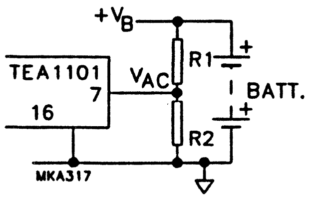


Fig.7 Basic circuit for direct battery sensing.

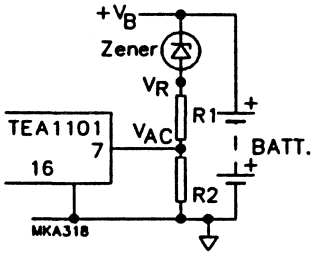


Fig.8 Basic circuit showing the use of a Zener diode.

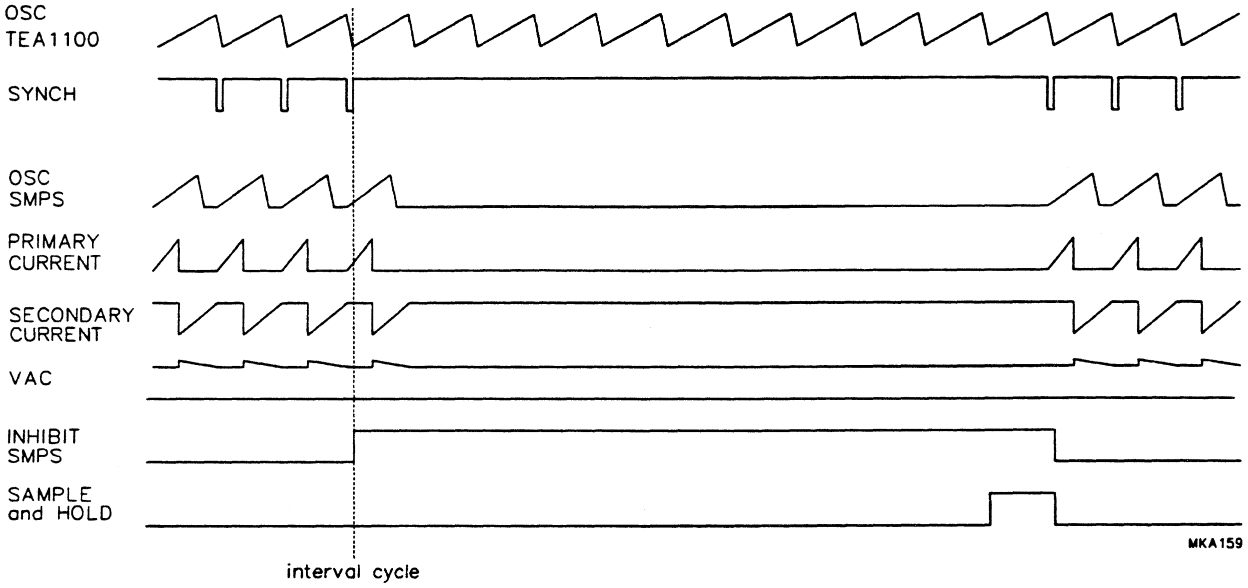
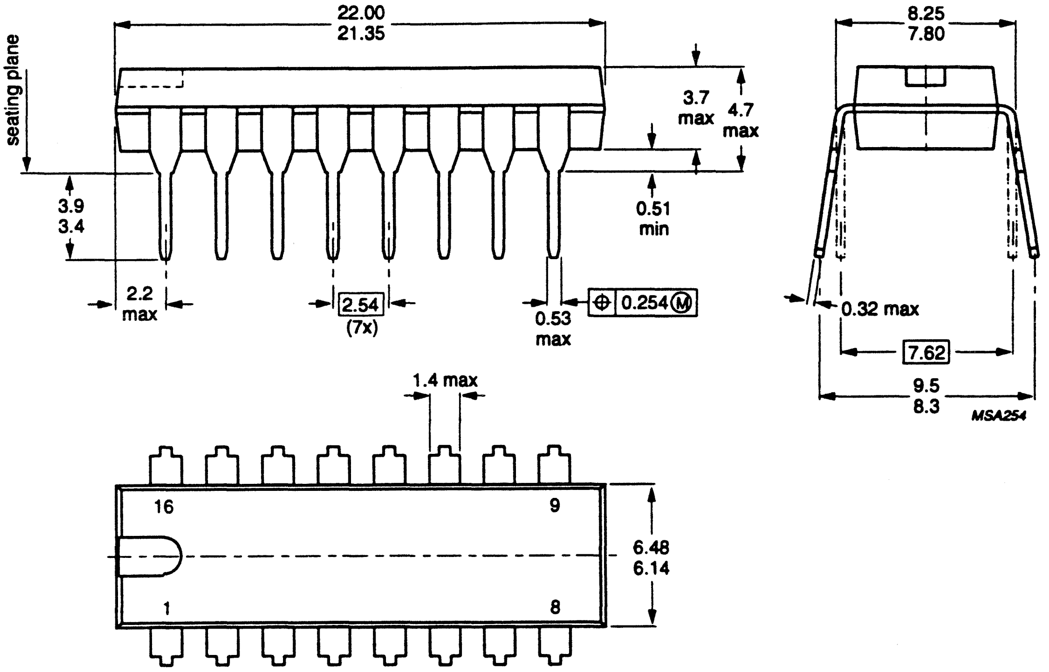


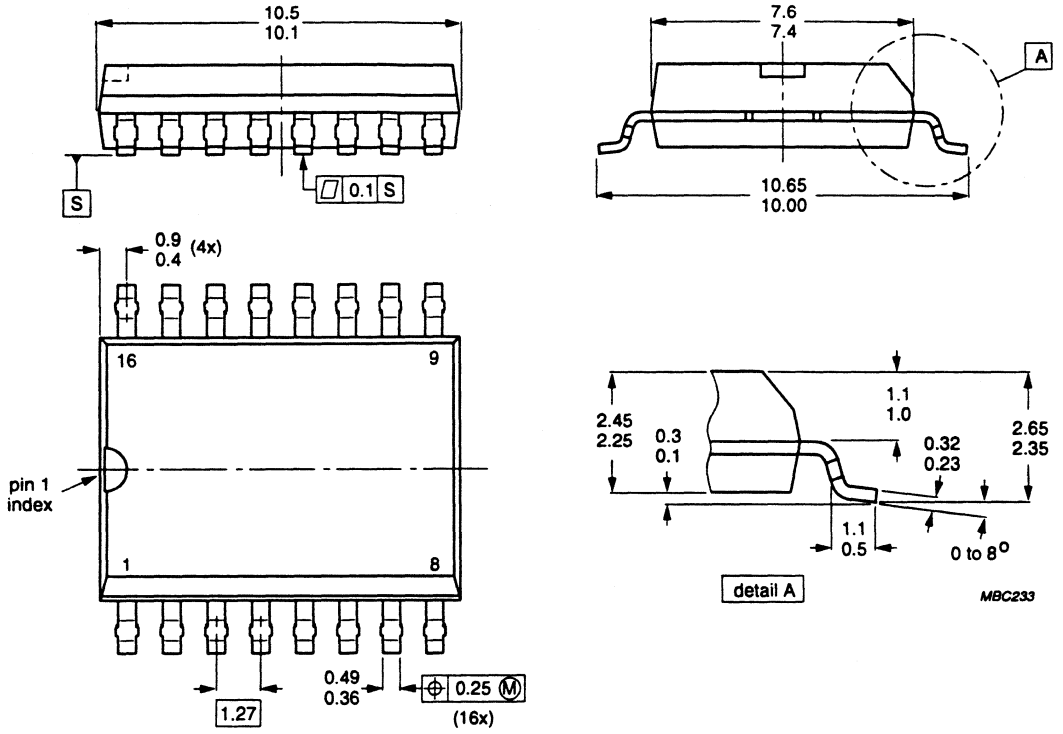
Fig.9 Synchronization waveforms.

PACKAGE OUTLINES



Dimensions in mm.

Fig.10 16-lead dual in-line; plastic (SOT38G).



Dimensions in mm.

Fig.11 16-lead mini-pack; plastic (SO16L; SOT162A).

SOLDERING

Plastic dual in-line packages

BY DIP OR WAVE

The maximum permissible temperature of the solder is 260 °C; this temperature must not be in contact with the joint for more than 5 s. The total contact time of successive solder waves must not exceed 5 s.

The device may be mounted up to the seating plane, but the temperature of the plastic body must not exceed the specified storage maximum. If the printed-circuit board has been pre-heated, forced cooling may be necessary immediately after soldering to keep the temperature within the permissible limit.

REPAIRING SOLDERED JOINTS

Apply the soldering iron below the seating plane (or not more than 2 mm above it). If its temperature is below 300 °C, it must not be in contact for more than 10 s; if between 300 and 400 °C, for not more than 5 s.

SOLDERING

Plastic mini-packs

BY WAVE

During placement and before soldering, the component must be fixed with a droplet of adhesive. After curing the adhesive, the component can be soldered. The adhesive can be applied by screen printing, pin transfer or syringe dispensing.

Maximum permissible solder temperature is 260 °C, and maximum duration of package immersion in solder bath is 10 s, if allowed to cool to less than 150 °C within 6 s. Typical dwell time is 4 s at 250 °C.

A modified wave soldering technique is recommended using two solder waves (dual-wave), in which a turbulent wave with high upward pressure is followed by a smooth laminar wave. Using a mildly-activated flux eliminates the need for removal of corrosive residues in most applications.

BY SOLDER PASTE REFLOW

Reflow soldering requires the solder paste (a suspension of fine solder particles, flux and binding agent) to be applied to the substrate by screen printing, stencilling or pressure-syringe dispensing before device placement.

Several techniques exist for reflowing; for example, thermal conduction by heated belt, infrared, and vapour-phase reflow. Dwell times vary between 50 and 300 s according to method. Typical reflow temperatures range from 215 to 250 °C.

Preheating is necessary to dry the paste and evaporate the binding agent. Preheating duration: 45 min at 45 °C.

REPAIRING SOLDERED JOINTS (BY HAND-HELD SOLDERING IRON OR PULSE-HEATED SOLDER TOOL)

Fix the component by first soldering two, diagonally opposite, end pins. Apply the heating tool to the flat part of the pin only. Contact time must be limited to 10 s at up to 300 °C. When using proper tools, all other pins can be soldered in one operation within 2 to 5 s at between 270 and 320 °C. (Pulse-heated soldering is not recommended for SO packages.)

For pulse-heated solder tool (resistance) soldering of VSO packages, solder is applied to the substrate by dipping or by an extra thick tin/lead plating before package placement.

DEFINITIONS

Data sheet status	
Objective specification	This data sheet contains target or goal specifications for product development.
Preliminary specification	This data sheet contains preliminary data; supplementary data may be published later.
Product specification	This data sheet contains final product specifications.
Limiting values	
Limiting values given are in accordance with the Absolute Maximum Rating System (IEC 134). Stress above one or more of the limiting values may cause permanent damage to the device. These are stress ratings only and operation of the device at these or at any other conditions above those given in the Characteristics sections of the specification is not implied. Exposure to limiting values for extended periods may affect device reliability.	
Application information	
Where application information is given, it is advisory and does not form part of the specification.	

LIFE SUPPORT APPLICATIONS

These products are not designed for use in life support appliances, devices, or systems where malfunction of these products can reasonably be expected to result in personal injury. Philips customers using or selling these products for use in such applications do so at their own risk and agree to fully indemnify Philips for any damages resulting from such improper use or sale.

Philips Semiconductors – a worldwide company

Argentina IEROD, Av. Juramento 1991 - 14.B, (1428) BUENOS AIRES,
Tel. (541)541 4261/541 4106, Fax. (541)786 7635

Australia 34 Waterloo Road, NORTH RYDE, NSW 2113,
Tel. (02)805 4455, Fax. (02)805 4466

Austria Triester Str. 64, A-1101 WIEN, P.O. Box 213,
Tel. (01)60 101-1236, Fax. (01)60 101-1211

Belgium Postbus 90050, 5600 PB EINDHOVEN, The Netherlands,
Tel. (31)40 783 749, Fax. (31)40 788 399

Brazil Rua do Rocia 220 - 5th floor,
CEP:04552-000-SÃO PAULO-SP, Brazil.
P.O. Box 7383-CEP:01051.
Tel. (011)829-1166, Fax. (011)829-1849

Canada INTEGRATED CIRCUITS:
Tel. (800)234-7381, Fax. (708)296-8556
DISCRETE SEMICONDUCTORS: 601 Milner Ave,
SCARBOROUGH, ONTARIO, M1B 1M8,
Tel. (0416)292 5161 ext. 2336, Fax. (0416)292 4477

Chile Av. Santa Maria 0760, SANTIAGO,
Tel. (02)773 816, Fax. (02)777 6730

Colombia Carrera 21 No. 56-17, BOGOTA, D.E., P.O. Box 77621,
Tel. (01)249 7624, Fax. (01)217 4549

Denmark Prags Boulevard 80, PB 1919, DK-2300 COPENHAGEN S,
Tel. (032)88 2636, Fax. (031)57 1949

Finland Sinikalliontie 3, SF-02630 ESPOO,
Tel. (9)0-50261, Fax. (9)0-520971

France 4 rue du Port-aux-Vins, BP317,
92156 SURESNES Cedex,
Tel. (01)4099 6161, Fax. (01)4099 6431

Germany Burchardstrasse 19, D-2 HAMBURG 1,
Tel. (040)3296-0, Fax. (040)3296 213

Greece No. 15, 25th March Street, GR 17778 TAVROS,
Tel. (01)4894 339/4894 911, Fax. (01)4814 240

Hong Kong 15/F Philips Ind. Bldg., 24-28 Kung Yip St.,
KWAI CHUNG, Tel. (0)4245 121, Fax. (0)4806 960

India PEICO ELECTRONICS & ELECTRICALS Ltd.,
Components Dept., Shivsagar Estate, Block 'A',
Dr. Annie Besant Rd., Worli, BOMBAY-400 018,
Tel. (022)4938 541, Fax. (022)4938 722

Indonesia Philips House, Jalan H.R. Rasuna Said Kav. 3-4,
P.O. Box 4252, JAKARTA 12950,
Tel. (021)5201 122, Fax. (021)5205 189

Ireland Newstead, Clonskeagh, DUBLIN 14,
Tel. (01)640 000, Fax. (01)640 200

Italy Viale F. Testi, 327, 20162-MILANO,
Tel. (02)6752.1, Fax. (02)6752.3350

Japan Philips Bldg 13-37, Kohnan 2-chome, Minato-ku, TOKIO 108,
Tel. (03)3740 5101, Fax. (03)3740 0570

Korea (Republic of) Philips House, 260-199 Itaewon-dong,
Yongsan-ku, SEOUL, Tel. (02)794-5011, Fax. (02)798-8022

Malaysia No. 76 Jalan Universiti, 46200 PETALING JAYA,
SELANGOR, Tel. (03)7755 1088, Fax. (03)757 4880

Mexico Philips Components, 6D Founder Blvd.,
EL PASO, TX 79906, Tel. 9-5(800)234-7381, Fax. (708)296-8556

Netherlands Postbus 90050, 5600 PB EINDHOVEN,
Tel. (040)78 37 49, Fax. (040)78 83 99

New Zealand 2 Wagener Place, C.P.O. Box 1041, AUCKLAND,
Tel. (09)849-4160, Fax. (09)849-7811

Norway Box 1, Manglerud 0612, OSLO,
Tel. (22)74 8000, Fax. (22)74 8341

Pakistan Philips Markaz, M.A. Jinnah Rd., KARACHI-3,
Tel. (021)577 039, Fax. (021)569 1832

Peru Carretera Central 6.500, LIMA 3, Apartado 5612,
Tel. (014)350 059, Fax. (014)468 999/468 949

Philippines PHILIPS SEMICONDUCTORS PHILIPPINES Inc,
106 Valero St. Salcedo Village, P.O. Box 911, MAKATI,
Metro MANILA, Tel. (02)810 0161, Fax. (02)817 3474

Portugal Av. Eng. Duarte Pacheco 6, 1009 LISBOA Codex,
Tel. (01)683 121, Fax. (01)658 013

Singapore Lorong 1, Toa Payoh, SINGAPORE 1231,
Tel. (65)350 2000, Fax. (65)251 6500

South Africa 195-215 Main Road, Martindale,
P.O. Box 7430,JOHANNESBURG 2000,
Tel. (011)470-5433, Fax. (011)470-5494

Spain Balmes 22, 08007 BARCELONA,
Tel. (93)301 6312, Fax. (93)301 4243

Sweden Kottbygatan 7, Akalla. Post: S-164 85 STOCKHOLM,
Tel. (0)8-6321 2000, Fax. (0)8-632 2745

Switzerland Allmendstrasse 140-142, CH-8027 ZÜRICH,
Tel. (01)488 2211, Fax. (01)482 8595

Taiwan 69, Min Sheng East Road, Sec 3, P.O. Box 22978,
TAIPEI 10446, Tel. (2)509 7666, Fax. (2)500 5899

Thailand PHILIPS ELECTRONICS (THAILAND) Ltd.,
60/14 MOO 11, Bangna - Trad Road Km. 3
Prakanong, BANGKOK 10260,
Tel. (2)399-3280 to 9, (2)398-2083, Fax. (2)398-2080

Turkey Talatpasa Cad. No. 5, 80640 LEVENT/ISTANBUL,
Tel. (01)279 2770, Fax. (01)269 3094

United Kingdom Philips Semiconductors Limited, P.O. Box 65,
Philips House, Torrington Place, LONDON, WC1E 7HD,
Tel. (071)436 41 44, Fax. (071)323 03 42

United States INTEGRATED CIRCUITS:
811 East Arques Avenue, SUNNYVALE, CA 94088-3409,
Tel. (800)234-7381, Fax. (708)296-8556
DISCRETE SEMICONDUCTORS: 2001 West Blue Heron Blvd.,
P.O. Box 10330, RIVIERA BEACH, FLORIDA 33404,
Tel. (800)447-3762 and (407)881-3200, Fax. (407)881-3300

Uruguay Coronel Mora 433, MONTEVIDEO,
Tel. (02)70-4044, Fax. (02)92 0601

Venezuela Calle 6, Ed. Las Tres Jotas, CARACAS, 1074A,
App. Post. 78117, Tel. (02)241 7509, Fax. (02)241 4518

For all other countries apply to: Philips Semiconductors,
International Marketing and Sales, Building BAF-1,
P.O. Box 218, 5600 MD, EINDHOVEN, The Netherlands,
Telex 35000 phtcnl, Fax. +31-40-724825

SCD20

Printed in The Netherlands 9397 716 40011

Philips Semiconductors

PHILIPS

DATA SHEET

SAA1500T
State-of-charge indicator for NiMH and NiCd powered applications

Objective specification
File under Integrated Circuits, IC11

July 1993

Philips Semiconductors

PHILIPS

State-of-charge indicator for NiMH and NiCd powered applications

SAA1500T

FEATURES

- 5-segment state-of-charge indication for LED or LCD displays
- Numerous display facilities to indicate the operational modes
- Designed for constant charge and varying discharge currents
- Large dynamic range of discharge currents
- Independent setting for charge and discharge efficiency
- Battery self-discharge compensation
- Automatic switch-over from fast to trickle charge (to prevent overcharging)
- Low standby current for permanent integration into a battery pack.

APPLICATIONS

- Intelligent battery powered, portable, applications with 'remaining energy' indication and fast charge control.

QUICK REFERENCE DATA

SYMBOL	PARAMETER	CONDITIONS	MIN.	TYP.	MAX.	UNIT
V_{CC}	supply voltage		1.8	–	7.0	V
I_{CC}	supply current	V_{CC} = 2.6 V; V_{CI} = 0 V	–	–	90	µA
f_{OSC}	fixed frequency	charging	–	4.2	–	kHz
V_{CI}	input sense voltage	discharging	20	–	200	mV
T_{amb}	operating ambient temperature		0	–	+70	°C

ORDERING INFORMATION

EXTENDED TYPE NUMBER	PACKAGE			
	PINS	PIN POSITION	MATERIAL	CODE
SAA1500T	20	SO20L	plastic	SOT163AH

INTRODUCTION

More and more portable appliances are being used because they can operate independently from power sources such as mains or car batteries (see Fig.1). In many cases, NiCd (powerful appliances) and NiMH (environment) rechargeable batteries are used. Because of the constant voltage of these batteries, it is not the energy (Wh) but the charge amount (Ah) that has to be known to enable the state-of-charge indication. In systems with known charge and discharge currents, time is an equivalent for the charge amount in the battery ($t = Q/I$). The charge time can be registered easily on a counter.

The system can be made universal by adapting the counting frequency to the used battery type, thereby making the counter contents a reflection of the energy state of the battery. The requirements of a battery state-of-charge indicator are in many ways similar to those of a fuel gauge for a car.

The SAA1500T is designed for enhanced systems with varying discharge currents. A variety of loads can discharge the battery without disturbing the charge account. If the equivalent of the charge counter contents is known in the SAA1500T, then the following information can be made available:

- The state-of-charge, indicated by LED or LCD displays
- The battery low state, indicated by LED and acoustic alarms
- The full state. A control signal can automatically switch the charge current from a fast to a trickle level to prevent overcharge.

The state-of-charge indication is an important and useful addition to any rechargeable battery pack for a variety of reasons. An underlying problem is the common tendency to recharge batteries many times, in fact more often than actually necessary. People waste time and effort in this way because they are always afraid of being caught out with flat batteries. The state-of-charge indicator helps to overcome this problem and brings other benefits such as:

- Increased battery cycling and consequently improved battery performance and lifetime
- Enhanced customer satisfaction
- Excellent selling feature.

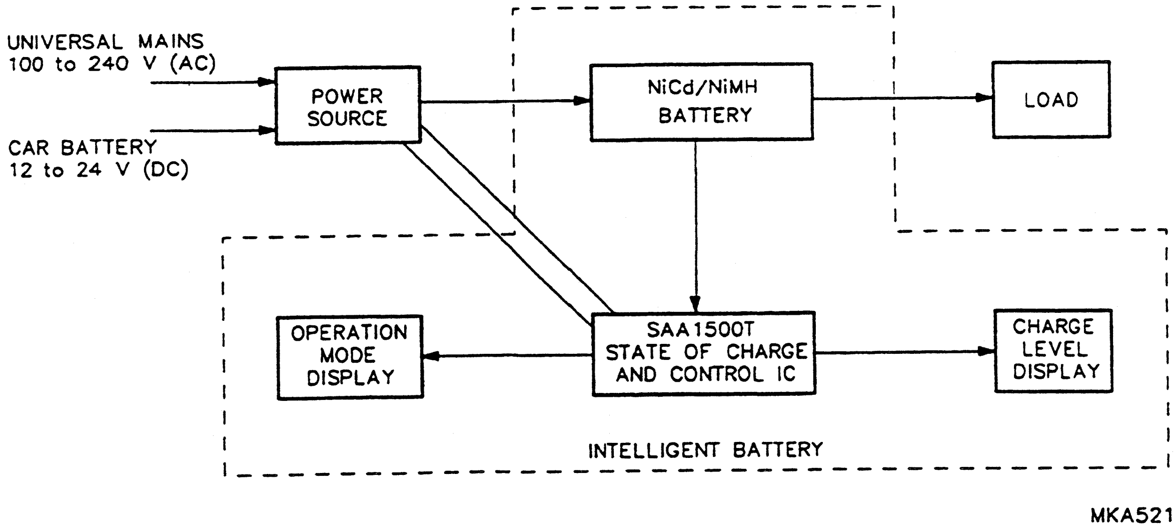


Fig.1 System diagram.

GENERAL DESCRIPTION

The SAA1500T is manufactured in a low voltage SACMOS process and has been designed for use as a battery state-of-charge indicator and as a battery charge controller for rechargeable batteries. In principle only four states exist (but because of two charging levels, fast and trickle charge), six states can be considered for the batteries; fast and trickle charge, fast and trickle charge and discharge at the same time, standby and battery discharge (see Fig.2 State diagram).

During charging, activated by power-on, the charge rate increases linearly with the charge time (assuming constant charge current). During discharge, activated by the 'load switch ON', the charge rate decreases linearly with the discharge time and the discharge current level. During charge/discharge, the charge rate is kept constant. In standby, the charge rate is decreased to compensate for the battery self-discharge.

In the SAA1500T, the known charge current and the measured discharge current are transferred in up/down oscillator pulses. By book-keeping the pulses, an image of the battery charge is created in the counter. The contents of the counter are output via 5 LEDs or a 6 segment LCD bar graph display. Two signal outputs are available to drive the LEDs to indicate whether the batteries are being charged or to indicate the nearly empty state.

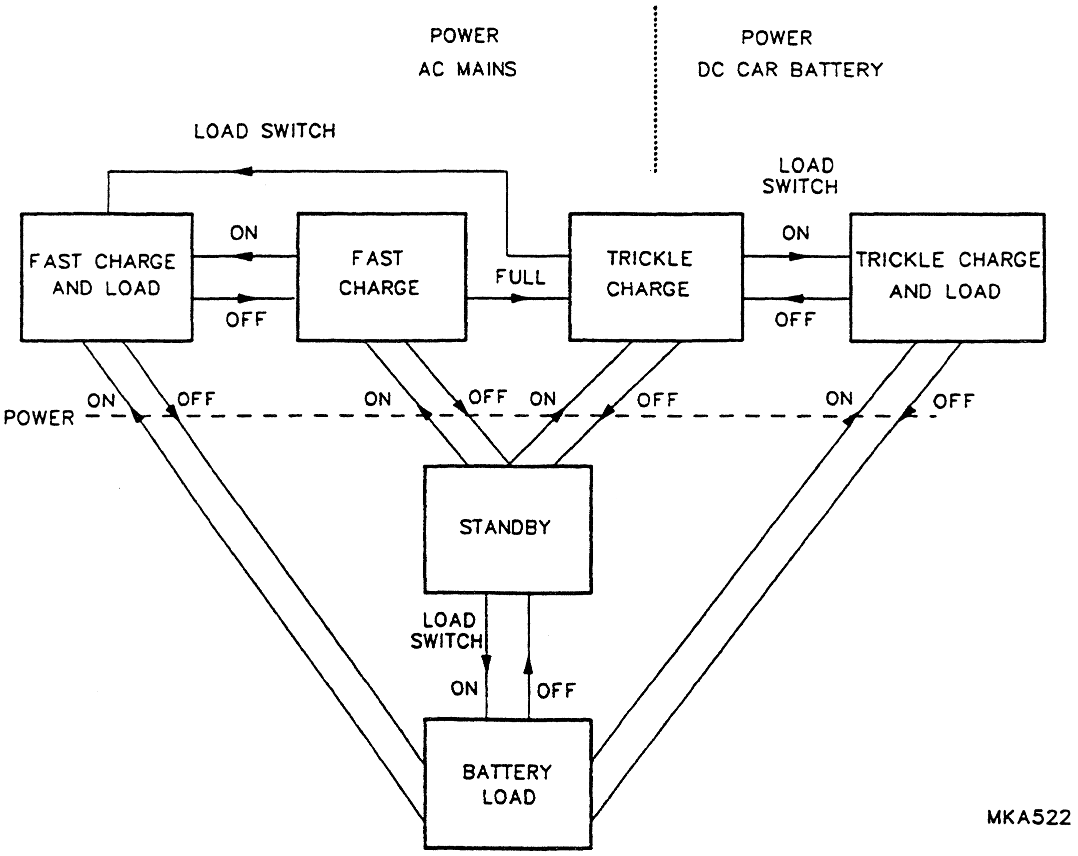


Fig.2 State diagram.

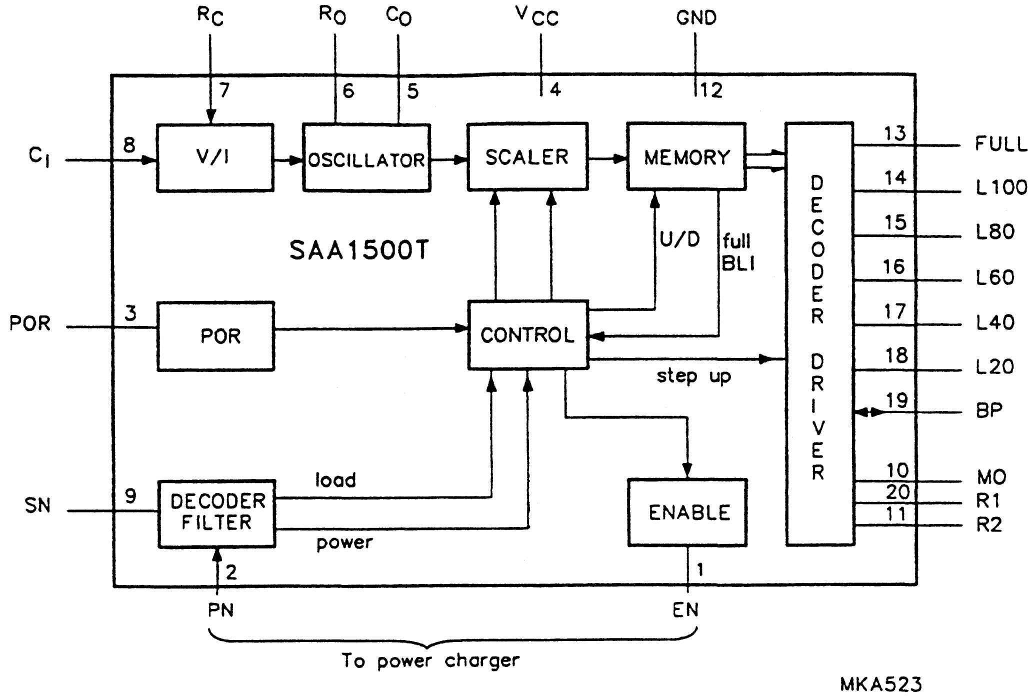


Fig.3 Block diagram.

PINNING

SYMBOL	PIN	DESCRIPTION
EN	1	enable control signal for battery charge unit
PN	2	power NOT mode detection
POR	3	power-on-reset, reset at LOW battery voltage
V_{CC}	4	supply voltage
C_O	5	capacitor for oscillator frequency
R_O	6	resistor for charge or self-discharge oscillator frequency
R_C	7	resistor to convert sense input voltage
C_I	8	discharge current sense input
SN	9	switch NOT, load switch ON detection
MO	10	mains ON state indication
R2	11	battery LOW drive signal for external buzzer
GND	12	ground
FULL	13	battery FULL indication, only LCD
L100	14	100% indication, LCD or LED driven
L80	15	80% indication, LCD or LED driven
L60	16	60% indication, LCD or LED driven
L40	17	40% indication, LCD or LED driven
L20	18	20% indication, LCD or LED driven
BP	19	backplane, (LCD), LCD/LED mode detection input
R1	20	battery LOW indicator, LOW drive

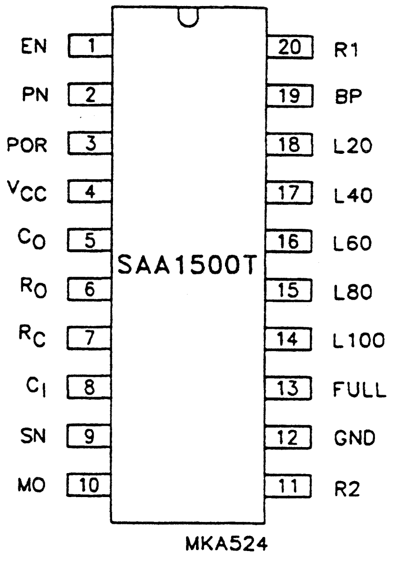


Fig.4 Pin configuration.

FUNCTIONAL DESCRIPTION

Power-on-reset

The POR circuit resets all counters if the supply voltage rises from 0 to V_{CC}. Reset is performed before $V_{CC} = 1.8$ V.
If the rise time of V_{CC} is fast, an external resistor and capacitor ensure the minimum pulse width of the internal reset pulse. If the rise time of V_{CC} is slow, the POR circuit acts as a level detector. The counters are also reset before $V_{CC} = 1.8$ V.

Oscillator (see Fig.5)

An RC type oscillator is used in the SAA1500T. The HIGH and LOW switching levels, and thus the voltage swing of the oscillator, are derived from V_{CC}. The timing components are designed so that the oscillator frequency is at V_{CC} and temperature independent.

The rise and fall times of the oscillator period are:

$t_r = R_O \times C_O \times \ln 2$; $t_f = R_O // 10\ k\Omega \times C_O \times \ln 2$;
$t_u \approx 15\ \mu s$ (9.1)

t_u is the undershoot time. The timings given in the specification are referenced to a fixed frequency of 4.2 kHz. Table 1 gives the modes at which the fixed frequency is active.

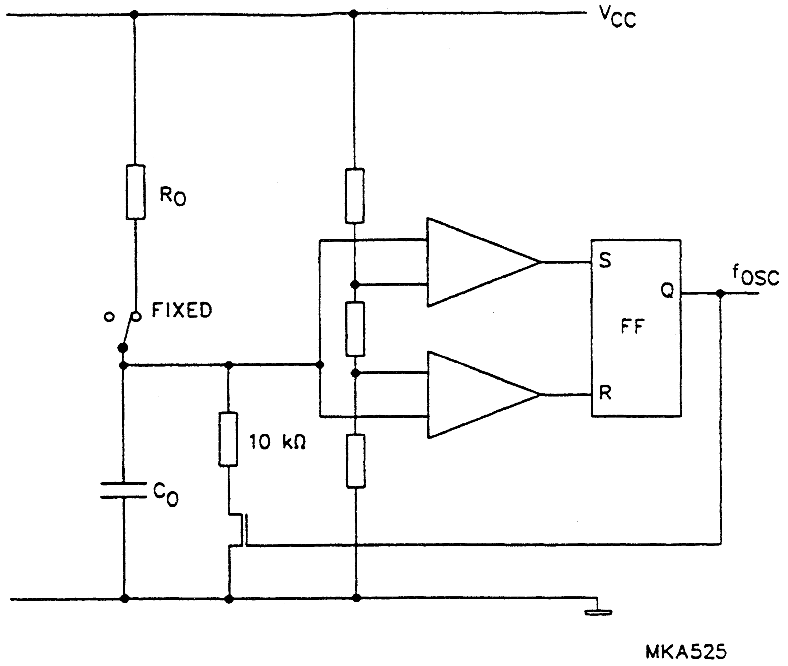


Fig.5 Fixed oscillator frequency.

V/I convertor (variable frequency)

The discharge current is sensed over a low ohmic (less dissipation) resistor R_{sense} (70 mΩ). In the voltage-to-current block (V/I), the sense voltage is converted into a current and this current is used to charge the oscillator capacitor C_O, (see Fig.6).

The period time can be calculated using the following equation:

$t_r = (C_O \times V_{CC} \times R_{conv})/(3 \times I_{load} \times R_{sense})$;
$t_f \approx 10\ k\Omega \times C_O \times \ln 2$; $t_u \approx 15\ \mu s$ (9.2)

The rise time is the dominating factor, and so it is clear that the countdown frequency is proportional to the discharge current .
In the discharge mode the frequency is V_{CC} dependent. However, as the battery voltage is rather insensitive to the battery charge, this influence is low.
The input voltage to the V/I block ranges from 20 mV to 200 mV. Outside of this range it is less accurate (see Fig.7).

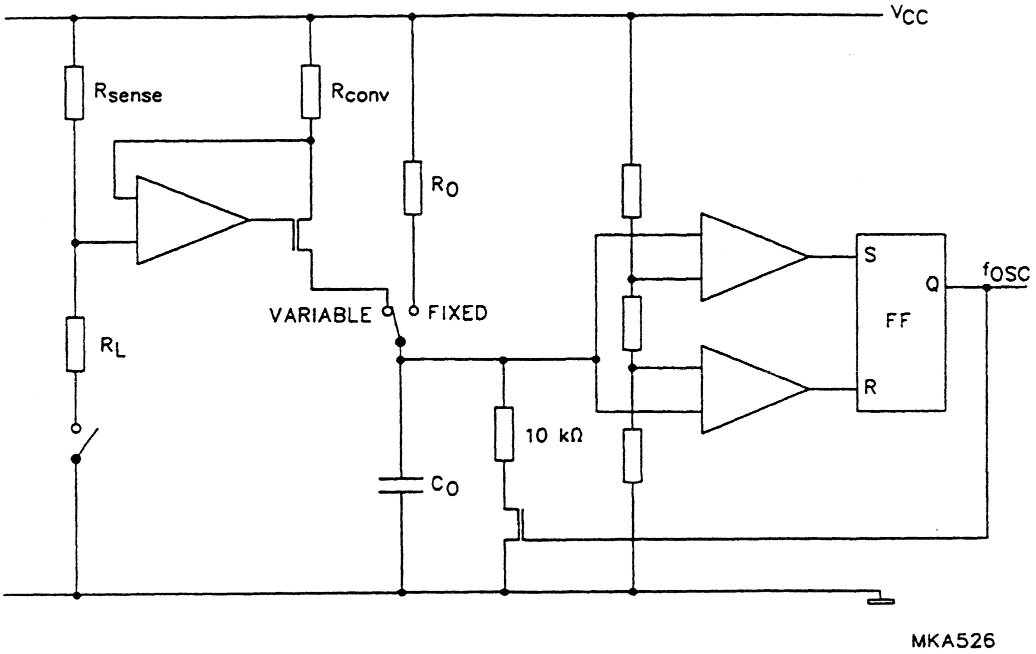


Fig.6 Variable oscillator frequency.

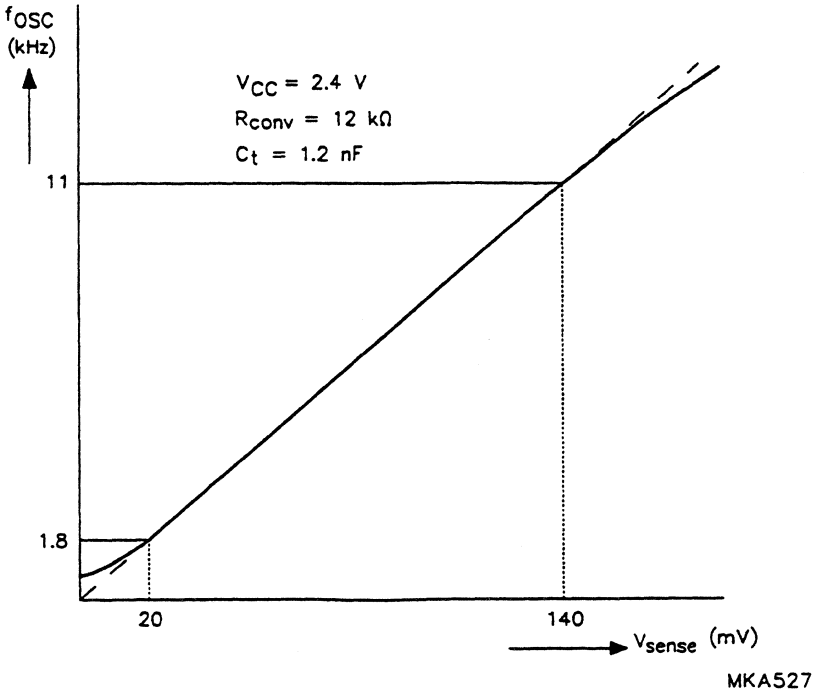


Fig.7 Voltage to frequency conversion.

Decoder filter

In the decoder filter block, the two inputs PN (power NOT) and SN (switch NOT) are tested. SN is tested on two logic states (HIGH and LOW) and PN on four states (HIGH, LOW, >5Ffixed and <3Ffixed). The inputs are digitally filtered to make them insensitive to external interference caused by motor commutation currents. The PN input informs the SAA1500T about the power charge unit state. The most important states are; fast charge, trickle charge and no power.
The SN input informs the SAA1500 about the load state. Two states are possible, load or no load. The load switch is sensed for that information.

Memory

The memory has a fixed volume of ten times. The memory contents (= battery charge) will be output via the decoder/driver block.

Scaler

In the scaler, a scale factor is introduced to scale down the memory counting frequency. In the trickle and standby modes, the battery current is adapted and so the memory counting frequency is changed accordingly. For the following modes, fixed scale factors are programmed:

- fast charge = 7.37×10^5
- trickle charge = 32 x fast charge
- standby = 150 x trickle charge
- discharge mode = 8.85×10^5.

Control

In the control block, in co-operation with the decoder/filter, six user modes and two test modes are provided together with appropriate signals to select the correct scale factors. The pin control signals, the different modes of operation, the oscillator frequency and scale factors are given in Table 1. The direction of the scale factor, up or down, is dependent on the mode, and thus determined in the control block. Four user modes have been explained previously.

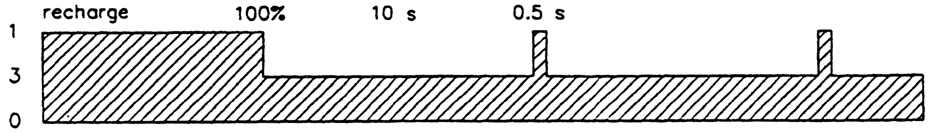
In the 'trickle charge and load mode', the discharge current is nearly always much higher than the trickle charge level, consequently the scale factor is chosen to be the same as in the discharge mode.
In many cases a motor is driven in the 'fast charge and load mode'. The control pin EN switches the system from a current source to a voltage source regulation to keep the motor speed constant. This ensures that only the current demanded by the motor will be delivered and the batteries will remain unaffected. For this reason a scale factor of zero is selected in this mode.
During the two test modes the scale factor is set at 1 to speed up the counter checks.
Step-up means that LEDs are switched on successively at the beginning of an LED display. If the LEDs are already active, there will be no step-up. Only those LEDs that are concerned with the charge status are activated. The step up frequency is $2 \times 10^{-3} \times f_{OSC}$ (see appendix B).
The LED and LCD display, during charge and discharge, is illustrated in Fig.12 (Appendix A) and in Fig.13 (Appendix B).
The memory block passes information to the control unit concerning the extreme states of the battery charge. At batteries FULL, a signal is sent to the enable block to terminate the fast charge session. At batteries almost empty, a Battery Low Indication (BLI) signal is sent to the decoder/driver (see also appendix A).

Table 1 Modes of operation.

INPUTS		MODES OF OPERATION	INDEX	OSCILLATOR FREQUENCY	SCALE FACTOR
PN	SN				
L	L	test, very fast down count	TD	fixed	512
<3Ffix	L	trickle charge and load, count down	TCL	variable	885k
>5Ffix	L	fast charge and load, no count	FCL	fixed	0
H	L	battery load, count down	BL	variable	885k
L	H	test, very fast up count	TU	fixed	512
<3Ffix	H	trickle charge, slow up count	TC	fixed	23M6
>5Ffix	H	fast charge, fast up count	FC	fixed	737k
H	H	standby, self discharge, POR active	SB	fixed	7G1
H	H	standby, self discharge, after POR	SB	fixed	3G5

Enable

It is possible to use the enable output (EN) to control the fast charge on the batteries. The EN signal is HIGH when the indicator shows not full. When an LED bar-graph is used and it shows FULL, then the EN signal will pulsate between HIGH and OPEN states. This provides a pulsating trickle charge waveform with a 5% duty factor for a 10 second period to compensate for the LED currents. If an LCD bar-graph is used, then at the full state, the EN output goes into a continuous high impedance state (see Fig.8).



0 = V_{SS}
1 = V_{CC}
3 = 3–state; high impedance
EN signal with LED bar graph applied.
With LCD bar graph, the signal is continuously 3–state after 100%.

MKA528

Fig.8 Waveform of enable signal with full batteries.

Decoder driver

The SAA1500T is intended to display the charge of the batteries in a six segment LCD bar graph or five segment LED display. LED indication is also given for batteries nearly empty, batteries on recharge and batteries full during recharge. Outputs L100, L80, L60, L40, L20 and FULL are designed to drive an LCD bar-graph with output BP connected to the backplane of the LCD. The LCD segment is active if the segment voltage is in antiphase with the BP voltage. If, however, BP is connected to ground (V_{SS}), then outputs L100, L80, L60, L40 and L20 may drive the LEDs directly with their anodes connected to the positive supply (V_{CC}). Outputs MO and R1 can drive LEDs directly. Output MO indicates batteries on recharge (LED constantly ON) and batteries full while recharging (LED flashes). Output R1 indicates that the battery is nearly empty. Output R2 provides a second battery nearly empty signal which can be used to drive a simple electroacoustic transducer with an audio tone via an external transistor. Circuit waveforms are illustrated in Fig.9.

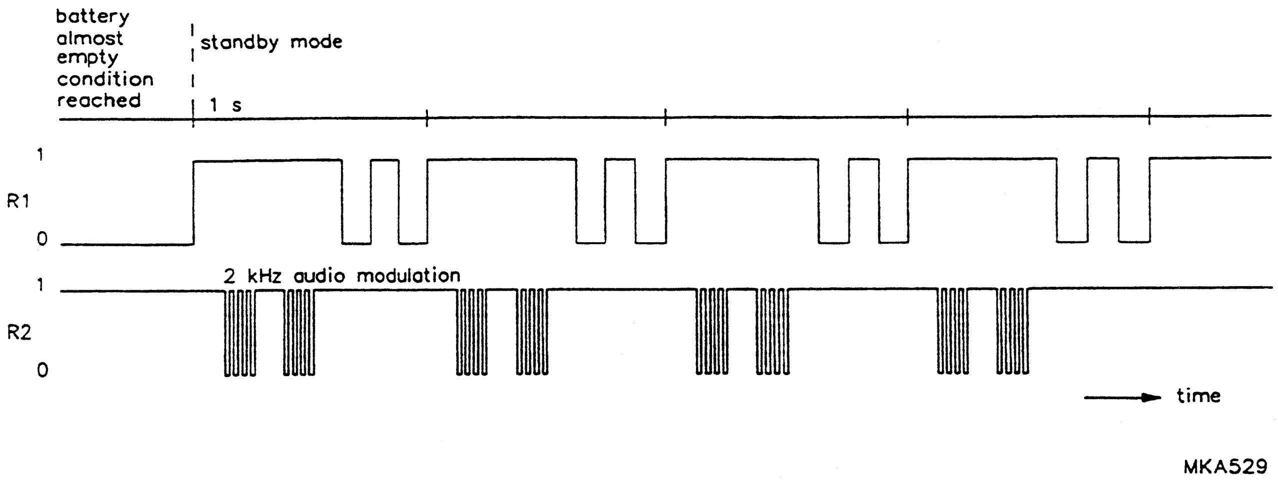


Fig.9 Waveform of R1 and R2 signalling empty batteries.

LIMITING VALUES

In accordance with the Absolute Maximum Rating System (IEC 134).

SYMBOL	PARAMETER	CONDITIONS	MIN.	MAX.	UNIT
V_{CC}	supply voltage		−0.5	7.0	V
V_I	input voltage	note 1	−0.5	V_{CC}+0.5	V
V_O	output voltage	note 1	−0.5	V_{CC}+0.5	V
I_{CC}	supply current		−	40	mA
I_{SS}	supply current		−	−120	mA
I_I	input current		−	±10	mA
I_O	output current		−	±20	mA
T_{amb}	operating ambient temperature		0	+70	°C
T_{stg}	storage temperature		−65	+150	°C

Note

1. V_{CC} + 0.5 V must not exceed 7.0 V.

DC CHARACTERISTICS

T_{amb} – 25 °C; unless otherwise specified.

SYMBOL	PARAMETER	CONDITIONS	MIN.	TYP.	MAX.	UNIT
V_{OH}	HIGH level output voltage (pins 13 and 19)	V_{CC} = 2.4 V; I_O = –650 µA	2.0	–	–	V
V_{OH}	HIGH level output voltage (pins 1 and 11)	V_{CC} = 2.4 V; I_O = –1 mA	2.0	–	–	V
V_{OH}	HIGH level output voltage (pins 10, 14 to 18 and 20)	V_{CC} = 2.4 V; I_O = –300 µA	2.0	–	–	V
V_{OL}	LOW level output voltage (pins 13 and 19)	V_{CC} = 2.4 V; I_O = 650 µA	–	–	0.4	V
V_{OL}	LOW level output voltage (pin 11)	V_{CC} = 2.4 V; I_O = 1 mA	–	–	0.4	V
V_{OL}	LOW level output voltage (pins 10 14 to 18 and 20)	V_{CC} = 3.8 V; I_O = 15 mA	1.8	–	–	V
		V_{CC} = 2.4 V; I_O = 5 mA	–	–	0.55	V
$\lvert I_{LO} \rvert$	output leakage current (pins 1 and 19)	V_{CC} = 3.8 V; V_O = V_{CC} or 0 V	–	–	±1	µA
$\lvert I_{LI} \rvert$	input leakage current (pin 3)	V_{CC} = 3.8 V; V_3 = 0 V	–	–	1	µA
$\lvert I_{LI} \rvert$	input leakage current (pin 2)	V_{CC} = 3.8 V	–	–	1	µA
I_9	input current (pin 9)	V_{CC} = 2.4 V	30	–	80	µA
V_{ihs}	Schmitt trigger HIGH (pins 2, 9 and 19)	V_{CC} = 2 V	1.6	–	–	V
		V_{CC} = 4 V	2.2	–	–	V
V_{ils}	Schmitt trigger LOW (pins 2, 9 and 19)	V_{CC} = 2 V	–	–	0.4	V
		V_{CC} = 4 V	–	–	0.18	V
V_{hys}	hysteresis voltage (pins 2, 9 and 19)	V_{CC} = 2 V	0.4	–	–	V
		V_{CC} = 4 V	0.5	–	–	V
V_{IH}	HIGH level input voltage (pin 3)	V_{CC} = 3.8 V; I_3 = 1 µA	0.8	–	1.5	V
		V_{CC} = 1.8 V; I_3 = 1 µA	–	–	1.55	V
V_{IL}	LOW level input voltage (pin 3)	V_{CC} = 1.8 V; I_3 = 1 µA	–	–	1.55	V
I_5	input current (pin 5)	V_{CC} = 2.4 V; V_5 = 2.4 V	150	–	360	µA
I_{CC}	supply current	V_{CC} = 2.6 V; V_3 = V_8 = 0 V; V_5 = V_7 = 0 V	–	–	90	µA

AC CHARACTERISTICS

T_{amb} = 25 °C; unless otherwise specified.

SYMBOL	PARAMETER	CONDITIONS	MIN.	TYP.	MAX.	UNIT
f_{OSC}	fixed oscillator frequency	V_{CC} = 2.4 V; R_6 = 260 kΩ C_5 = 1.2 nF; $V_8 = V_2 = V_9$ = 0 V	3.81	4.24	4.67	kHz
Δf_{OSC}	variable oscillator frequency	V_{CC} = 2.4 V; V_8 = 56 mV; R_7 = 12 kΩ C_5 = 1.2 nF; PN = 1; SN = 0	4.04	4.65	5.02	kHz
		V_{CC} = 2.4 V; V_8 = 35 mV; R_7 = 12 kΩ C_5 = 1.2 nF; PN = 1; SN = 0	2.47	2.94	3.24	kHz
SVRR	supply voltage ripple rejection	V_{CC} = 2.2 to 2.6 V; V_8 = 49 mV	–	–	5	%/0.1V

QUALITY SPECIFICATION

General quality specification for integrated circuits: UZW - BO/FQ - 601.

APPLICATION INFORMATION

The SAA1500T can be used in many applications with different types and sizes of rechargeable batteries. In the Introduction it is stated that the discharge/charge time is the equivalent of the battery charge.
The formula of the elapsed discharge/charge time and thus the collected charge is (see Fig.3):

discharge/charge time = t_{OSC} x scale factor x memory

(1) The fixed time period of the oscillator during charging is given in the following equation:

t_{OSC} = (Charge Time) / (scale factor x 10)

Where: the Charge Time (CT) is the time taken to fully charge empty batteries.

Therefore: f_{OSC} = 7370 / (3600 x CT [h]) [kHz]; (exclusive efficiency corrections) (13.1)

From formula (13.1) it is clear that the fixed oscillator frequency is dependent on the charge time (and so the charge current) and not on the battery size.
If the charge time and the oscillator frequency are fixed, the external components C_O and R_O can be calculated with formula (9.1). Some examples are given in Table 2.

(2) Many timing functions are related to the fixed oscillator frequency. The most important are the trickle charge and the self-discharge times (see also Table 2).

(3) The variable time period of the oscillator during discharge is given in the following equation:

t_{OSC} = (discharge time) / (scale factor x 10)

Therefore: f_{OSC} = 8850 / (3600 x DT) [kHz] (13.2)

Furthermore, formulas (9.1) and (9.2) are valid and, combining them with formula (13.2), the external components R_{conv}, and R_{sense} can be calculated:

R_{conv}/R_{sense} ' 500 x 10^3 x Q [Ah] / C_O [nF]; (V_{CC} = 2.4 V) (13.3)

As can be seen from formula (13.3), it is the battery size that is of importance and not the discharge time. Some examples are given in Table 3.

Table 2 Charge components.

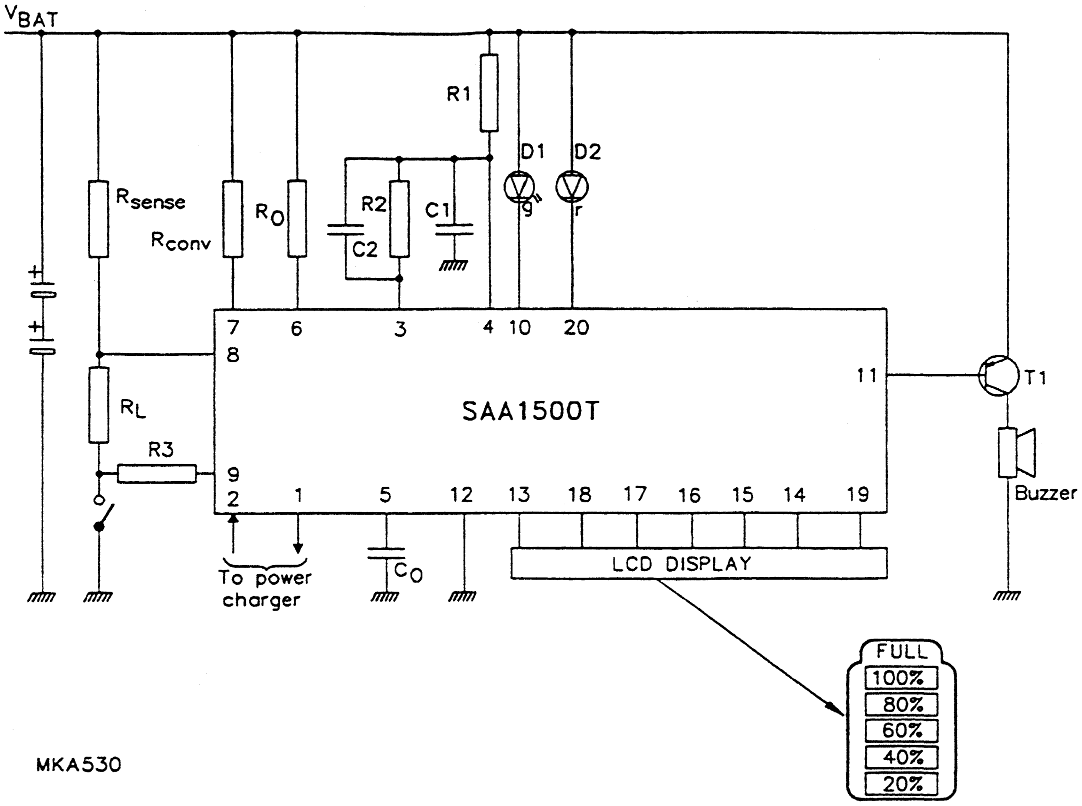
	NiXX / xxAh			
charge time (h)	1/3	1/2	1	
charge current (CA)	3	2	1	
fixed frequency (kHz)	6.1	4.1	2.05	formula 13.1
R_O (kΩ) (C_O = 1.2 nF); note 1	180	264	556	formula 9.1
R_O (kΩ) (C_O = 2.4 nF); note 1	90	132	277	
trickle charge current (CA)	1/32 x I_{charge}			
self-discharge current (CA)	1/150 x $I_{trickle}$			

Note

1. $C_O > 1.2$ nF because of the parasitic capacitance influence on the printed-circuit board.

Table 3 Discharge components.

	NiCd		NiMH	
	AA(600mAh)	**Sub C(1.2Ah)**	**AA(1.1Ah)**	**AA(1.7Ah)**
R_{conv}/R_{sense}(C_O = 1.2 nF)	252k	499k	459k	706k
R_{conv}/R_{sense}(C_O = 2.4 nF)	126k	252k	229k	352k
R_{conv} (C_O = 1.2 nF) (R_{sense} = 70 mΩ)	17k6	34k8	32k	49k3
R_{conv} (C_O = 2.4 nF) (R_{sense} = 70 mΩ)	8k8	17k4	16k	24k6



R1 = 100 Ω; R2 = 1.5 MΩ; R3 = 5.1 kΩ; C1 = 100; C2 = 1.2 nF.

Fig.10 Application diagram for LCD display.

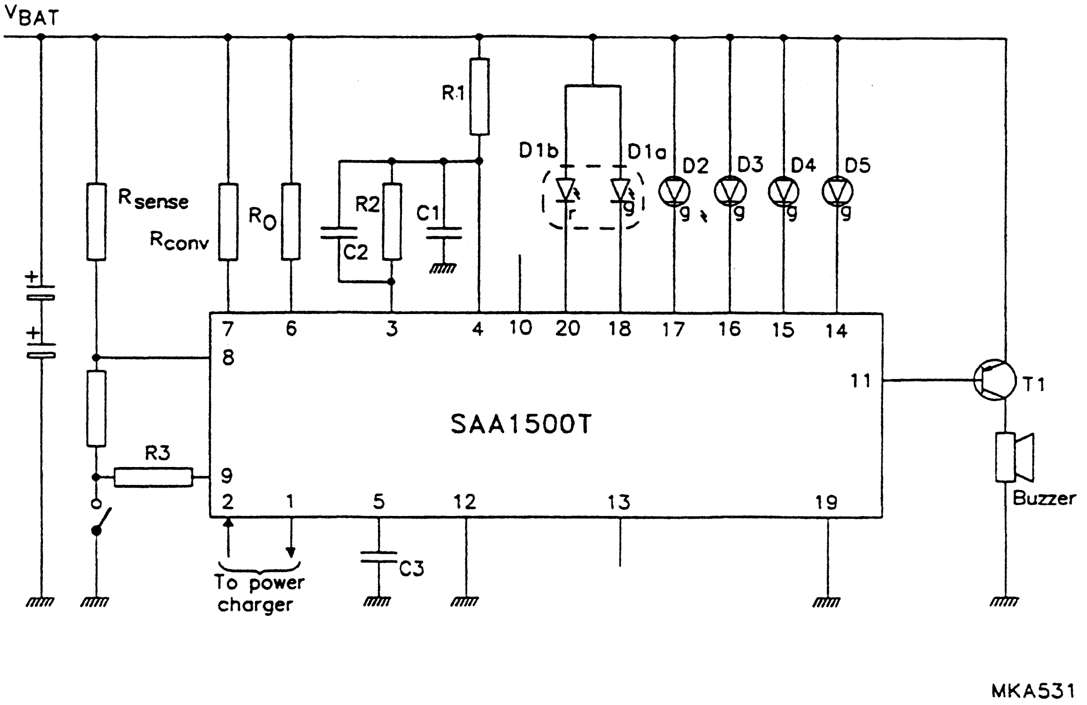


Fig.11 Application diagram for LED display.

APPENDIX A

MKA532

○ LED is OFF

⊜ LED (green) is ON

● LED (red) is ON

✹ LED (green) is ON, blinking at 2 Hz during charging
LED MO starts blinking at > 100%

▭ LCD is ON

▭ LCD is ON, blinking at 2 Hz during charging

[20] Blinking at 1 Hz during discharge or rest.

→ Time sequence.
(frequencies are referenced to a fixed oscillator frequency of 4.2 kHz)

Fig.12 Legend for figure 13.

APPENDIX B

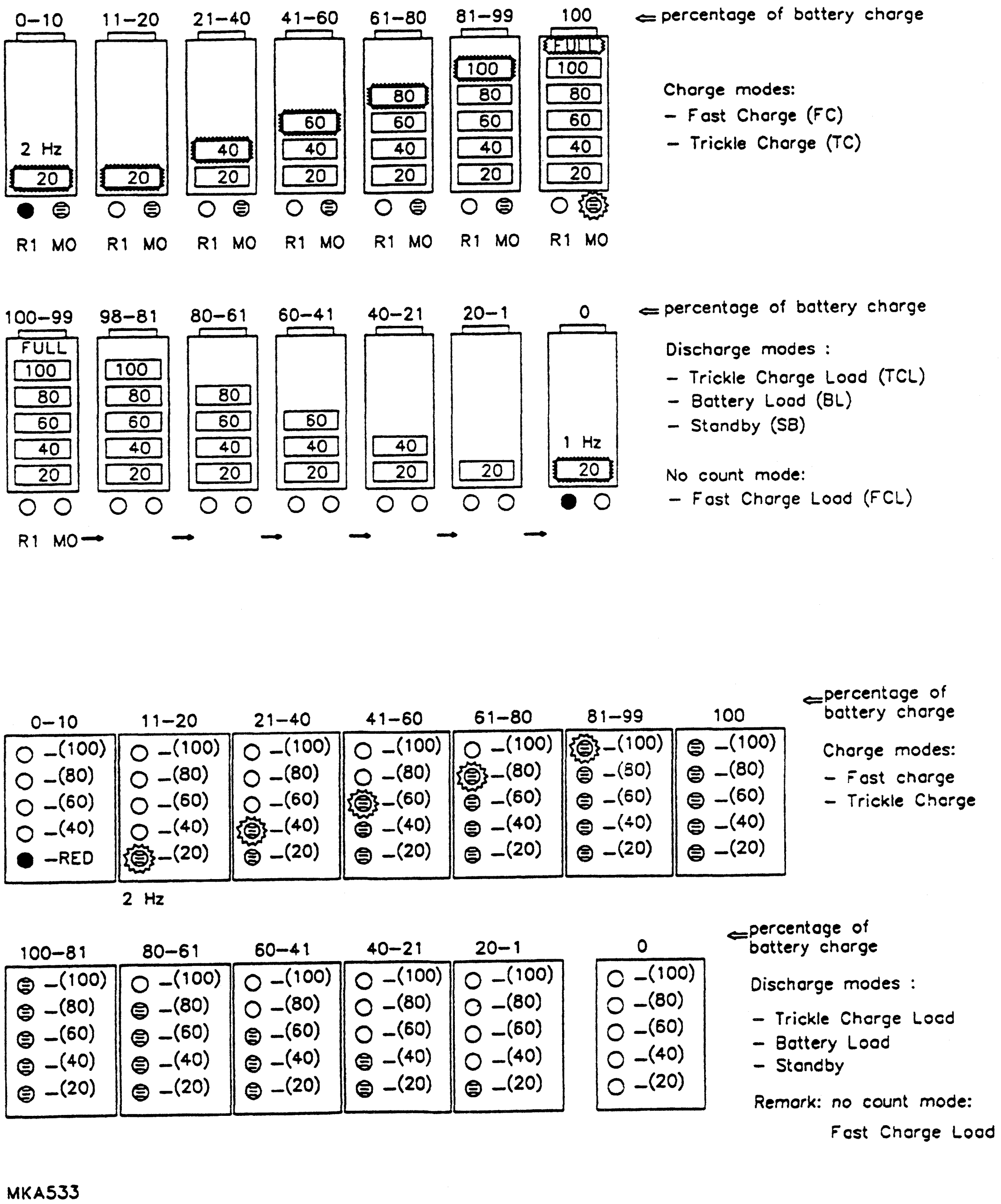


Fig.13 Battery charge displayed by LCD and LED.

Remarks to figure 13

LED mode

The LEDs are activated at every operational mode change. If the LEDs were already active before the mode change, the display will not be changed. If the LEDs were not active before the mode change, a step-up pattern is generated. Step up means that the LEDs are activated successively recording the charge status every 1/8 second. In the Battery Load, Trickle Charge Load and Standby modes, the LEDs are switched off after 8 second. In the Fast Charge, Trickle Charge and Fast Charge Load modes, the LEDs remain on.

Battery Low Indication is active if;

The charging (FC,TC) is stopped (==> standby) before 10% charge is reached.
The discharging (BL, TCL) is started below 10% and stopped (==> SB) above 0%.
The discharging is stopped (==>) at 0%.
For BLI waveforms (see Fig.9) R1 is active before the mode is switched over to standby.

Battery Low Indication is not active if;

Discharging is started above 10% and stopped 10-1%. Instead, the green L20 LED will be active for 8 seconds. If recharging is started during 1-10%. The L20 LED is blinking at 2 Hz, the red LED is not active.

LCD mode

The LCD display, as against the LED mode, is always visible.

Battery Low Indication is active if:

The charging (FC, TC) is stopped (==> standby) before 10% charge is reached.
The discharging (BL, TCL) is stopped (==> SB) at 0% charge.

PACKAGE OUTLINE

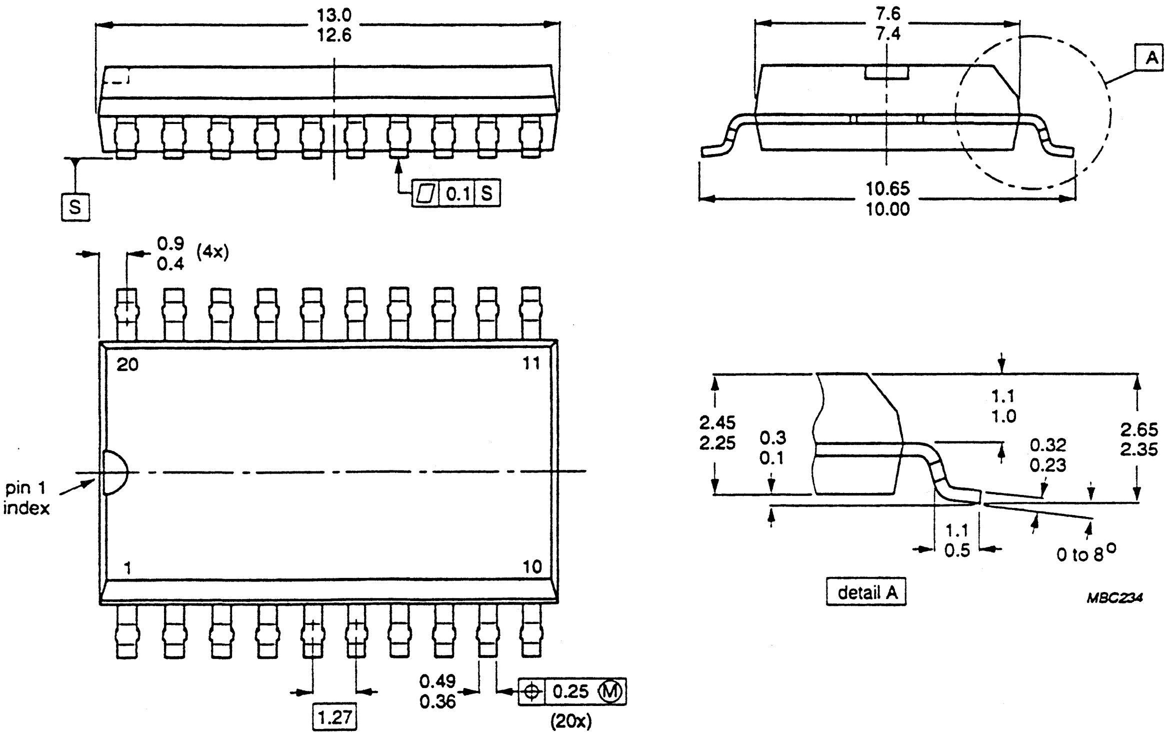


Fig.14 20-lead mini-pack; plastic (SO20L; SOT163AH).

SOLDERING

Plastic mini-packs

BY WAVE

During placement and before soldering, the component must be fixed with a droplet of adhesive. After curing the adhesive, the component can be soldered. The adhesive can be applied by screen printing, pin transfer or syringe dispensing.

Maximum permissible solder temperature is 260 °C, and maximum duration of package immersion in solder bath is 10 s, if allowed to cool to less than 150 °C within 6 s. Typical dwell time is 4 s at 250 °C.

A modified wave soldering technique is recommended using two solder waves (dual-wave), in which a turbulent wave with high upward pressure is followed by a smooth laminar wave. Using a mildly-activated flux eliminates the need for removal of corrosive residues in most applications.

BY SOLDER PASTE REFLOW

Reflow soldering requires the solder paste (a suspension of fine solder particles, flux and binding agent) to be applied to the substrate by screen printing, stencilling or pressure-syringe dispensing before device placement.

Several techniques exist for reflowing; for example, thermal conduction by heated belt, infrared, and vapour-phase reflow. Dwell times vary between 50 and 300 s according to method. Typical reflow temperatures range from 215 to 250 °C.

Preheating is necessary to dry the paste and evaporate the binding agent. Preheating duration: 45 min at 45 °C.

REPAIRING SOLDERED JOINTS (BY HAND-HELD SOLDERING IRON OR PULSE-HEATED SOLDER TOOL)

Fix the component by first soldering two, diagonally opposite, end pins. Apply the heating tool to the flat part of the pin only. Contact time must be limited to 10 s at up to 300 °C. When using proper tools, all other pins can be soldered in one operation within 2 to 5 s at between 270 and 320 °C. (Pulse-heated soldering is not recommended for SO packages.)

For pulse-heated solder tool (resistance) soldering of VSO packages, solder is applied to the substrate by dipping or by an extra thick tin/lead plating before package placement.

DEFINITIONS

Data sheet status	
Objective specification	This data sheet contains target or goal specifications for product development.
Preliminary specification	This data sheet contains preliminary data; supplementary data may be published later.
Product specification	This data sheet contains final product specifications.
Limiting values	
Limiting values given are in accordance with the Absolute Maximum Rating System (IEC 134). Stress above one or more of the limiting values may cause permanent damage to the device. These are stress ratings only and operation of the device at these or at any other conditions above those given in the Characteristics sections of the specification is not implied. Exposure to limiting values for extended periods may affect device reliability.	
Application information	
Where application information is given, it is advisory and does not form part of the specification.	

LIFE SUPPORT APPLICATIONS

These products are not designed for use in life support appliances, devices, or systems where malfunction of these products can reasonably be expected to result in personal injury. Philips customers using or selling these products for use in such applications do so at their own risk and agree to fully indemnify Philips for any damages resulting from such improper use or sale.

NOTES

NOTES

NOTES

Philips Semiconductors – a worldwide company

Argentina IEROD, Av. Juramento 1991 - 14.B, (1428) BUENOS AIRES,
Tel. (541)541 4261/541 4106, Fax. (541)786 7635

Australia 34 Waterloo Road, NORTH RYDE, NSW 2113,
Tel. (02)805 4455, Fax. (02)805 4466

Austria Triester Str. 64, A-1101 WIEN, P.O. Box 213,
Tel. (01)60 101-1236, Fax. (01)60 101-1211

Belgium Postbus 90050, 5600 PB EINDHOVEN, The Netherlands,
Tel. (31)40 783 749, Fax. (31)40 788 399

Brazil Rua do Rocia 220 - 5th floor,
CEP:04552-000-SÃO PAULO-SP, Brazil.
P.O. Box 7383-CEP:01051.
Tel. (011)829-1166, Fax. (011)829-1849

Canada INTEGRATED CIRCUITS:
Tel. (800)234-7381, Fax. (708)296-8556
DISCRETE SEMICONDUCTORS: 601 Milner Ave,
SCARBOROUGH, ONTARIO, M1B 1M8,
Tel. (0416)292 5161 ext. 2336, Fax. (0416)292 4477

Chile Av. Santa Maria 0760, SANTIAGO,
Tel. (02)773 816, Fax. (02)777 6730

Colombia Carrera 21 No. 56-17, BOGOTA, D.E., P.O. Box 77621,
Tel. (01)249 7624, Fax. (01)217 4549

Denmark Prags Boulevard 80, PB 1919, DK-2300 COPENHAGEN S,
Tel. (032)88 2636, Fax. (031)57 1949

Finland Sinikalliontie 3, SF-02630 ESPOO,
Tel. (9)0-50261, Fax. (9)0-520971

France 4 rue du Port-aux-Vins, BP317,
92156 SURESNES Cedex,
Tel. (01)4099 6161, Fax. (01)4099 6431

Germany Burchardstrasse 19, D-2 HAMBURG 1,
Tel. (040)3296-0, Fax. (040)3296 213

Greece No. 15, 25th March Street, GR 17778 TAVROS,
Tel. (01)4894 339/4894 911, Fax. (01)4814 240

Hong Kong 15/F Philips Ind. Bldg., 24-28 Kung Yip St.,
KWAI CHUNG, Tel. (0)4245 121, Fax. (0)4806 960

India PEICO ELECTRONICS & ELECTRICALS Ltd.,
Components Dept., Shivsagar Estate, Block 'A',
Dr. Annie Besant Rd., Worli, BOMBAY-400 018,
Tel. (022)4938 541, Fax. (022)4938 722

Indonesia Philips House, Jalan H.R. Rasuna Said Kav. 3-4,
P.O. Box 4252, JAKARTA 12950,
Tel. (021)5201 122, Fax. (021)5205 189

Ireland Newstead, Clonskeagh, DUBLIN 14,
Tel. (01)640 000, Fax. (01)640 200

Italy Viale F. Testi, 327, 20162-MILANO,
Tel. (02)6752.1, Fax. (02)6752.3350

Japan Philips Bldg 13-37, Kohnan 2-chome, Minato-ku, TOKIO 108,
Tel. (03)3740 5101, Fax. (03)3740 0570

Korea (Republic of) Philips House, 260-199 Itaewon-dong,
Yongsan-ku, SEOUL, Tel. (02)794-5011, Fax. (02)798-8022

Malaysia No. 76 Jalan Universiti, 46200 PETALING JAYA,
SELANGOR, Tel. (03)7755 1088, Fax. (03)757 4880

Mexico Philips Components, 6D Founder Blvd.,
EL PASO, TX 79906, Tel. 9-5(800)234-7381, Fax. (708)296-8556

Netherlands Postbus 90050, 5600 PB EINDHOVEN,
Tel. (040)78 37 49, Fax. (040)78 83 99

New Zealand 2 Wagener Place, C.P.O. Box 1041, AUCKLAND,
Tel. (09)849-4160, Fax. (09)849-7811

Norway Box 1, Manglerud 0612, OSLO,
Tel. (22)74 8000, Fax. (22)74 8341

Pakistan Philips Markaz, M.A. Jinnah Rd., KARACHI-3,
Tel. (021)577 039, Fax. (021)569 1832

Peru Carretera Central 6.500, LIMA 3, Apartado 5612,
Tel. (014)350 059, Fax. (014)468 999/468 949

Philippines PHILIPS SEMICONDUCTORS PHILIPPINES Inc,
106 Valero St. Salcedo Village, P.O. Box 911, MAKATI,
Metro MANILA, Tel. (02)810 0161, Fax. (02)817 3474

Portugal Av. Eng. Duarte Pacheco 6, 1009 LISBOA Codex,
Tel. (01)683 121, Fax. (01)658 013

Singapore Lorong 1, Toa Payoh, SINGAPORE 1231,
Tel. (65)350 2000, Fax. (65)251 6500

South Africa 195-215 Main Road, Martindale,
P.O. Box 7430,JOHANNESBURG 2000,
Tel. (011)470-5433, Fax. (011)470-5494

Spain Balmes 22, 08007 BARCELONA,
Tel. (93)301 6312, Fax. (93)301 4243

Sweden Kottbygatan 7, Akalla. Post: S-164 85 STOCKHOLM,
Tel. (0)8-6321 2000, Fax. (0)8-632 2745

Switzerland Allmendstrasse 140-142, CH-8027 ZÜRICH,
Tel. (01)488 2211, Fax. (01)482 8595

Taiwan 69, Min Sheng East Road, Sec 3, P.O. Box 22978,
TAIPEI 10446, Tel. (2)509 7666, Fax. (2)500 5899

Thailand PHILIPS ELECTRONICS (THAILAND) Ltd.,
60/14 MOO 11, Bangna - Trad Road Km. 3
Prakanong, BANGKOK 10260,
Tel. (2)399-3280 to 9, (2)398-2083, Fax. (2)398-2080

Turkey Talatpasa Cad. No. 5, 80640 LEVENT/ISTANBUL,
Tel. (01)279 2770, Fax. (01)269 3094

United Kingdom Philips Semiconductors Limited, P.O. Box 65,
Philips House, Torrington Place, LONDON, WC1E 7HD,
Tel. (071)436 41 44, Fax. (071)323 03 42

United States INTEGRATED CIRCUITS:
811 East Arques Avenue, SUNNYVALE, CA 94088-3409,
Tel. (800)234-7381, Fax. (708)296-8556
DISCRETE SEMICONDUCTORS: 2001 West Blue Heron Blvd.,
P.O. Box 10330, RIVIERA BEACH, FLORIDA 33404,
Tel. (800)447-3762 and (407)881-3200, Fax. (407)881-3300

Uruguay Coronel Mora 433, MONTEVIDEO,
Tel. (02)70-4044, Fax. (02)92 0601

Venezuela Calle 6, Ed. Las Tres Jotas, CARACAS, 1074A,
App. Post. 78117, Tel. (02)241 7509, Fax. (02)241 4518

For all other countries apply to: Philips Semiconductors,
International Marketing and Sales, Building BAF-1,
P.O. Box 218, 5600 MD, EINDHOVEN, The Netherlands,
Telex 35000 phtcnl, Fax. +31-40-724825

SCD20

Printed in The Netherlands 9397 717 60011

Philips Semiconductors

PHILIPS

PRELIMINARY DEVICE SPECIFICATION

TYPE COMMERCIAL : SAA1501t

STATE-OF-CHARGE MONITOR IC FOR INTELLIGENT NICD AND NIMH BATTERY PACKS

1. Market introduction : End 1993.
This information applies to a product under development. Its specification and characteristics are subject to changes without notice.

2. Features

- drive signals for a LED bargraph or LCD; 5-level indication of battery charge, 20%, 40%, 60%, 80% and 100%.
- direct drive for LEDs ('battery low' and 'batteries under charge') and buzzer.
- large dynamic range of charge and discharge currents.
- accurate sensing charge/discharge currents due to BICMOS process technology.
- independent setting of charge and discharge efficiency.
- automatic switch-over between charge/discharge and selfdischarge mode.
- charge current regulation drive signals.
- temperature compensated self-discharge account.
- temperature protection via adjustable Tabsolute.
- large operating temperature range.
- low standby current drain to allow integration in batterypack.

3. Applications

Intelligent battery packs / Intelligent battery powered applications.

4. General description

The SAA1501t is intended to be used as a battery monitor and control circuit in rechargeable systems for NiCd and NiMH batteries. The primary function is to provide a readout of instantaneous charge in the battery. Secondary function is to control the external current source via charge current regulation drive signals.
The SAA1501t accurately determine the state-of-charge (remaining energy) of a battery at any instant by keeping account of the charge/discharge time and the charge/discharge current, - via the analogue V/I convertors with large dynamic range -, in a 'coulomb' counter.
This information is used to drive a LED bargraph or liquid crystal 'remaining energy' display and to regulate the external current source.

5. Pinning

The pin numbering is according to the standard numbering of pins of the package S(mall) O(utline) 24L(arge) SO24L.

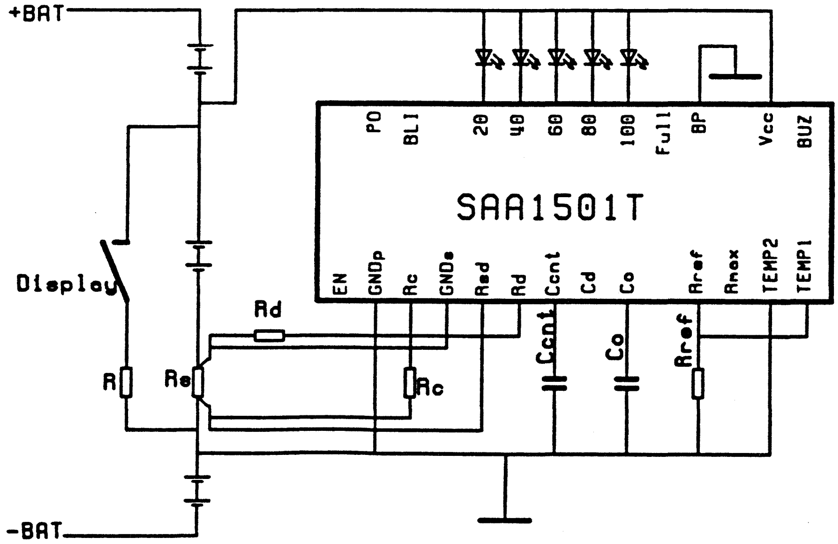
Pinnumber	Pin Name	Description
1	Vcc	Supply voltage
2	EN	Enable
3	Cd	Duty cycle capacitor
4	Ccnt	Charge counter capacitor
5	Rmax	Maximum average charge current setting
6	Rref	Current reference resistor
7	Rd	Discharge current conversion resistor
8	Rc	Charge current conversion resistor
9	GNDs	Charge sense input
10	Rsd	Discharge sense input
11	TEMP1	Temperature sensing resistor
12	TEMP2	Temperature setting resistor
13	Co	Oscillator capacitor
14	BUZ	Buzzer
15	FULL	Full battery indication
16	100	100% segment indication
17	80	80% segment indication
18	60	60% segment indication
19	40	40% segment indication
20	20	20% segment indication
21	BP	LCD back plane drive
22	BLI	Battery Low Indication LED
23	PO	Power On LED
24	GNDp	Power ground

6. Quick reference data

Parameter	Min	Typ	Max	Unit
Supply voltage range	2.0		4.5	V
Supply current		0.5		mA
Stand by current			100	uA
V/I converter range	2mV		Vcc−1.6V	
V/I linearity @ 3mV	−33	1.8	37	%
@ 10mV	−11	0	11	%
@ 30mV	−4.5	−0.5	3.5	%
@ 100mV	−2	−0.5	1	%
@ 300mV	−2	−1.0	0	%

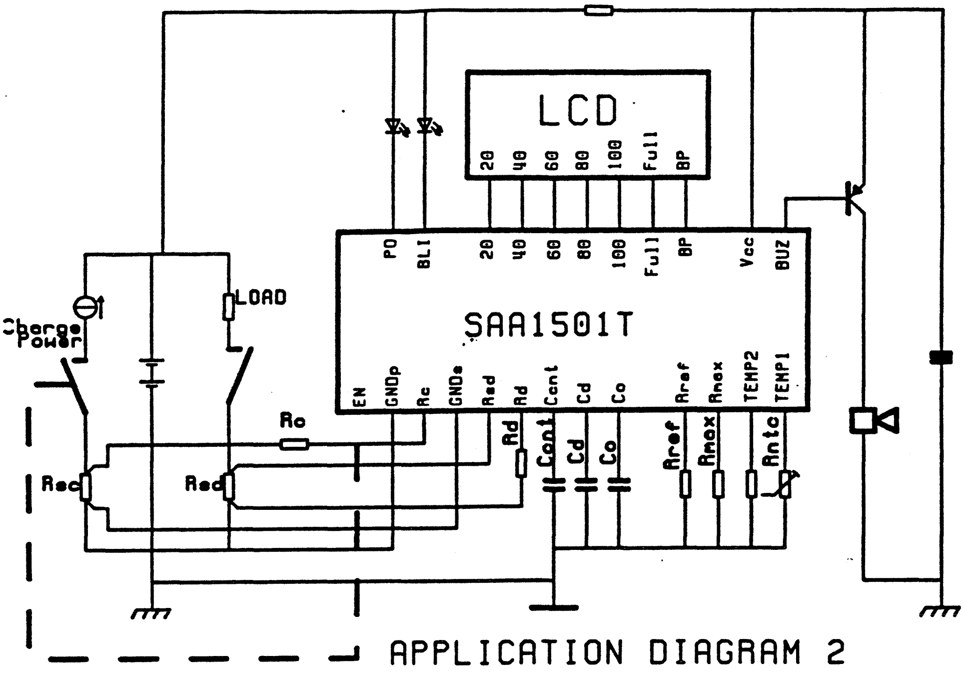
7. Application– information

A. Intelligent battery pack.



APPLICATION DIAGRAM 1

B. Intelligent battery pack controlling external current source.



APPLICATION DIAGRAM 2

APPLICATION NOTE

ICs for Battery Management

TEA1100 Battery Charger with flyback SMPS current regulator from 12 volt input

Report No: NPO/AN9201

W. Bosboom

Customer & Product Support Group Consumer ICs Nijmegen, the Netherlands.
Fax: +31 80 533659

Keywords
Power Control & Misc.
Battery management
NiCd/NiMH fast charger
TEA1100

Date : 21 oct. 1992
Pages: 23

SUMMARY

A design description is given of a NiCd or NiMH battery charger with the TEA1100 charge and regulation control IC.
The current supply stage is a highly efficient flyback SMPS circuit which includes a current mode slave IC UC3843 and a BUK443-60B power Mosfet switch.
The charger is applicable for a wide range of different voltage batteries and the upper limit is determined by the SMPS design power limitation. A 7.2V battery is maximum with the given charger design.

The charger is outlined for a 12V power source such as a car battery which makes it specially suitable for field use.

The design calculations are extensively given to support redesign on chargers with different requirements.

TABLE OF CONTENTS

1 INTRODUCTION

This paper gives a practical example of a fast charger design for charging , within 1 hour , a NiCd or NiMH battery pack containing up to six , "AA" type cells (500mAhr with NiCd).
This example is a follow up to the charger design for four cells which is described in the TEA1100 applicaton note NPO/AN9102.
The flyback topology applied has the advantage that it provides a large degree of design freedom in charge current magnitude, number of battery cells and input voltage especially. A large number of cells can be charged with a relatively low input voltage (12V). The use of a buck converter, which is popular because of the circuit symplicity, is restricted in the input/output voltage ratio due to the step down principle.

The charge current supply is a discontinuous flyback regulator which is powered from a nominal 12V dc source. It operates according to the current mode principle and is master controlled by the TEA1100 charge management IC. This IC monitors the battery voltage for the battery full event, stabilizes the charge current and determines safe operation in the fast or trickle charge mode.

A full calculation procedure is given for the SMPS power transformer and components design. This simplifies adaptation of the design example to different charger requirements and it might be worthwile to apply the formulae in an electronic spreadsheet.
For the design-in of TEA1100, its time settings and related components can quickly be derived from given graphs.

2 CHARGER SPECIFICATION

- Input voltage range : 9 - 20V dc, 12V nominal

- Output voltage range : 0 - 10.2V.
- Open output voltage : protection level ≥ 13V.

- Battery type : 7.2V ; 500mAh (6 NiCd cells).

- Fast charge time : within 1 hour.
- Abs.max. fast charge time : 1.5 hour time out.

- Charge currents : 625mA (1.25C) fast charge.
  30mA (0.06C) trickle charge,
  (6 - 60mA setting range).

- Charge control methods : -dV , timer , temperature optional.

- Switch mode frequency : 60kHz.

- Maximum continuous output power : 6.4W.

- Charger efficiency : >75% at maximum output power.

3 BASIC CIRCUIT OUTLINE

3.1 Block diagram

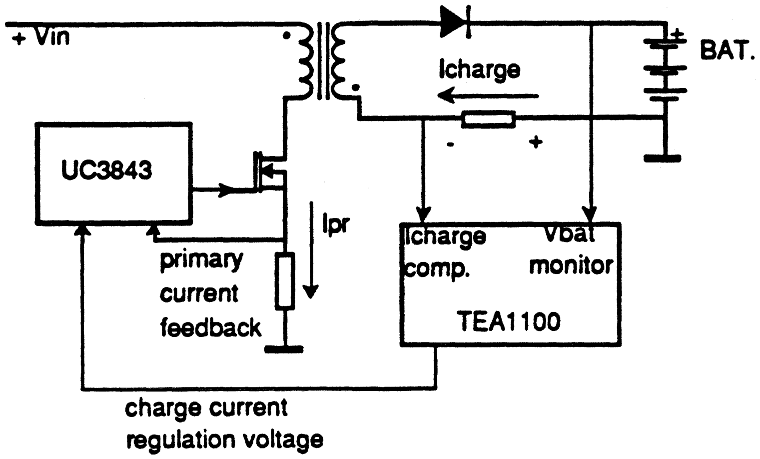


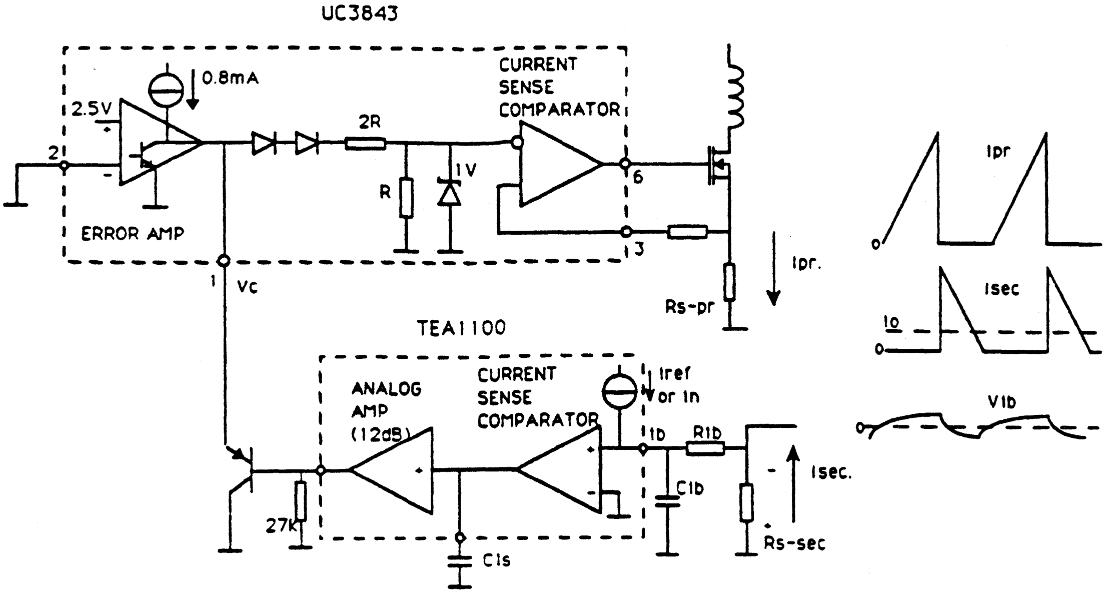
Fig. 1. Basic circuit diagram of the charger system example.

The essential parts are the supply flyback converter with its transformer, power switch and the IC UC3843 which drives the switch and controls the primary input current. The IC follows the primary current flow by sensing the voltage across a resistor in the primary loop.
At the secondary output side the battery is under the control of the TEA1100. This IC senses the battery voltage for save charge operation and it senses the output current via a current sense resistor in the secondary current loop. The current magnitude is referred by TEA1100 which provides an error feedback voltage for the stabilization of the charge current. This error voltage is fed back to the primary controller which translates the info into a correct duty cycle of the MOSFET drive pulses.

3.2 Current regulation loop

See figure 2.
The secondary output current is sensed at the sense resistor ,Rs-sec, and via voltage level shifting and RC smoothing, the residual voltage is compared to ground level by TEA1100. The dc error voltage is build up across the loop stability capacitor Cls. This voltage is, 4x amplified, available as output at the AO pin of the IC.
The small signal bandwidth of the AO signal has to be a distance below the switching frequency to ensure stability and therefore the error loop signal has a relative slow response time.

The overall response time can then be improved again by applying a feedforward, cycle by cycle, approach within the primary switching loop. Current mode controllers do respond to this as they drive the power switch under control of interacting error and primary current signals. The improved response time applies however only to rapid changes in the input voltage. A sudden change in the output condition is only corrected via the slower secondary to primary feedback.



$$V_{Rs\,pr(pk)} = \frac{V_c - 1.4}{3}$$

Fig. 2 The current regulation feedback loop diagram

The UC3843 is selected for primary control and power mosfet drive. The current comparitor which delivers the duty cycle controlled drive pulses, compares the primary peak current at Rs-pr with the error feedback voltage from TEA1100.
Peak current protection occurs when the peak voltage across the sense resistor reaches the UC3843 internal zener value, V_Z, which can range between 0.9 and 1.1V. Below the peak voltage value the duty cycle depends on the magnitude of error voltage at the Vc pin. The voltage error amp of UC3843 is set to function as a current source of typ 0.8mA with input pin 2 to common.
At the Vc pin, the voltage feedback active range is from 1.4V(diodes) to [3 x V_Z + 1.4V(diodes)]. The high level can so vary between 4.1 and 4.7V, (typ. 4.4V) depending on V_Z.
When TEA1100 is not initialized or is in the off period within the pulsating trickle charge mode, then its AO output floats. In this case or an other case of feedback absense the Vc pin is drawn to common via the PNP buffer which disables converter operation. On feedback signal from TEA1100 the converter is able to run.
In the non invert mode the AO amp sources current only and the maximum output voltage is 3.6V. Therefore the Vc voltage can reach a maximum of 3.6V plus the V_{BE} of the buffer: total 4 - 4.3V, which coincides with a comparitor sense level of 0.87 - 0.97V. Here an advantage appears of a tighter tolerance on the maximum Vc- or switch current peak level.

3.3 Current sensing parts

- Primary current sense resistor value: Rs-pr ≤ 0.87V/Ipr(pk)max. (Ipr(pk)abs max = 0.97V/Rs-pr).
- Secondary current sense resistor value : Rs-sec ≥ Iref x Rib / Io charge.

For sensing the correct output current, the triangle pulse sense voltage, analogue to the current shape, must be averaged and the residual ripple should be less than 160mV. Further smoothing of the error voltage occurs in C_{LS}. (ref 1).

The required smoothing Rib x Cib network can be derived from the formula;

$$Rib \cdot Cib \geq \frac{1}{Vripple} \cdot \frac{(V_{pk} - V_{avg})^2}{2 \cdot V_{pk}} \cdot \delta sec \cdot T$$

were: $Vripple \leq 0.16V$

$V_{pk} = I_{sec\ pk} \times Rsense$
$V_{avg} = Isec_{avg} \times Rsense$
$\delta sec \cdot T =$ Isec pulse time

The highest ripple voltage will be at maximum output power.

The ripple voltage on the LS capacitor is by approximation;

$$V_{Cls}\ ripple = \frac{V_{ib\ ripple} \cdot g_m}{8 \cdot f_{osc} \cdot C_{ls}}$$

In section 5 the formulae above are used with the determined data.

4 POWER STAGE AND TRANSFORMER DESIGN

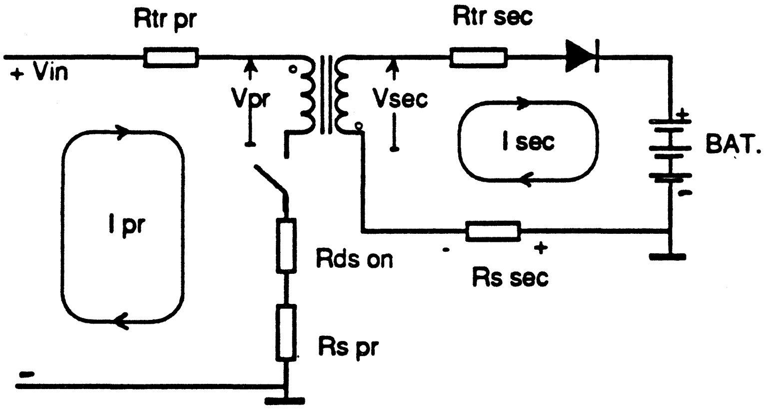
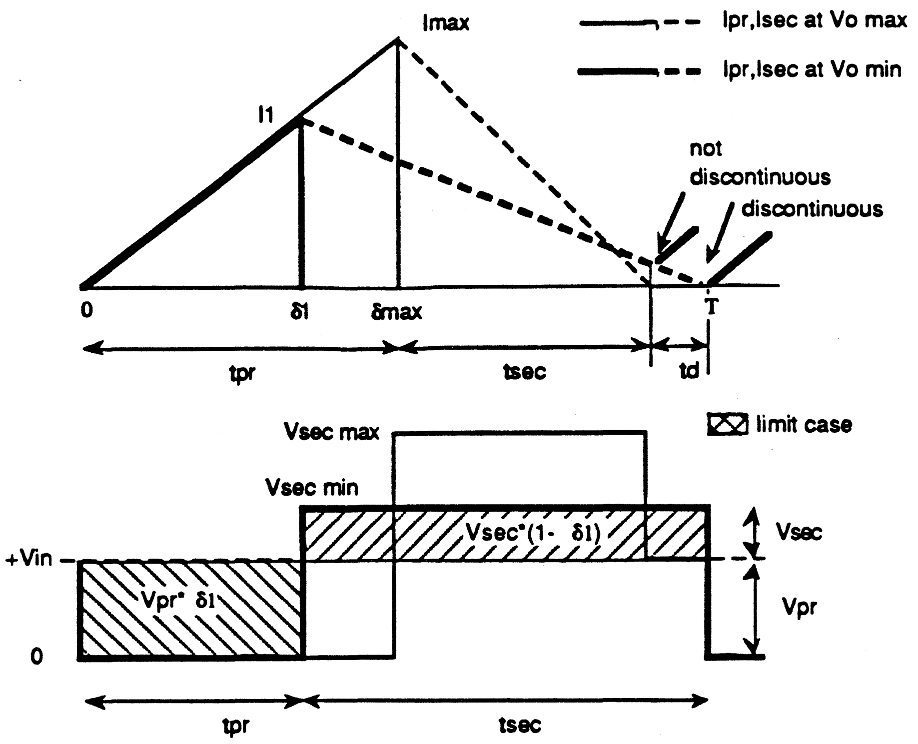


Fig. 3 The large signal power stage configuration

4.1 Transformer operation

The discontinuous current mode has been chosen because constraints on the feedback loop stability are minimal. Also the supply output power source characteristic is easier to handle than an output voltage source which is inherent to the continuous current mode operation.
At fixed frequency the limit case of discontinuous current operation, whereafter it changes to continuous current operation, is however different between a voltage regulated and a current regulated supply. With constant output voltage the limit case occurs at maximum power while with constant output current the limit occurs at minimum power, i.e. the lowest output voltage.
Figure 4 shows the current and voltage waveforms of the transformer at both maximum power and limit case with reduced power for the constant output current condition.



Transformer current and voltage waveforms at constant output current

Fig. 4 Transformer current (a) and voltage (b) waveforms at constant current output

At maximum power the waveforms shows that the transformer is not fully utilised due to the introduced dead time in current. (Only a Self Oscillating Power Supply, SOPS, mode can prevent this dead time periode). Nevertheless, for a battery charger the transformer and supply performance with fixed frequency will hardly differ from the SOPS performance while interference-wise the fixed frequency is preferable.

4.2 Basic formulae for transformer operation.

Energy in transformer: $E = \frac{1}{2} \cdot L \cdot I^2$ [1]

Peak current: $I_{pk} = \frac{V_{pr}}{L_{pr}} \cdot t_{pr}$ [2]

Primary on time: $t_{pr} = \delta \cdot T \qquad (T = \frac{1}{f})$ [3]

Throughput power: $P_{th} = E \cdot f$ [4]

substitute 1,2,3,4: $P_{th} = \frac{(V_{pr} \cdot \delta)^2}{2 \cdot L_{pr}} \cdot T$ [5]

Limit discontinuous condition / constant output current;
(occurs at minimum V_{sec} and minimum V_{pr})

see shaded parts in fig. b $V_{pr(min)} \cdot \delta_1 = N \cdot V_{sec(min)} \cdot (1-\delta_1)$ were ($N = \frac{N_{pr}}{N_{sec}}$)

Duty cycle relation with secondary voltage; $\frac{\delta_1}{\delta_{max}} = \sqrt{\frac{V_{sec(min)}}{V_{sec(max)}}}$ [7]

Secondary fraction at maximum power: $\delta_{sec} = \delta_{max} \cdot \frac{V_{pr(min)}}{N \cdot V_{sec(max)}}$

or: $\delta_{sec} = \frac{V_{sec(min)}}{V_{sec(max)}} \left(\sqrt{\frac{V_{sec(max)}}{V_{sec(min)}}} - \delta_{max} \right)$ [8]

Dead fraction at maximum power: $\delta_d = 1 - \delta_{max} - \delta_{sec}$ [9]

Minimum duty cycle: $\delta min = \delta max \cdot \frac{Vpr(min)}{Vpr(max)} \cdot \sqrt{\frac{Vsec(min)}{Vsec(max)}}$ [10]
(at Vpr(max) and Vsec(min))

4.3 First shot at transformer size and loss

Let us assume that a feasable SMPS efficiency,η, is in the order of 70%. The maximum output power is 10.2V x 0.625A = 6.4W (Vo= 1.7V/cell x 6).
Core size can quickly be determined by the energy figure $L \cdot I^2$.

The required $L \cdot I^2 = 2 \times Po / \eta / f = 2 \times 6.4W / 0.7 / 60kHz = 3 \times 10^{-4}$ J.

- Size: Out of the RM cores*) data in fig.4 a selection is made for a RM6/i core with a standard range airgap of 0.15mm which maximum $L \cdot I^2$ value is 3.2×10^{-4}. [A_L = 250nH]

- Material: For 60kHz frequency a suitable core material is 3C85.

- Loss: The total dissipation in the transformer is the sum of the windings' copper and core losses. The thermal resistance of RM6 is about 75°C/W and when a temperature rise of 40°C can be accepted then the maximum dissipation is 0.53W.

  The core loss can be derived from the data in fig.5. which give watts per core volume with peak flux density B and frequency. This peak flux density refers to bidirectional flux swing. For the discontinuous operation the flux swing goes in one direction only and therefore the flux density is half the top flux which can be up to 320mT at 100°C. If B peak density is limited to 150mT then at 60kHz the core loss outcome is 120kW/m^3. The effective volume of RM6 is 1325mm^3 which gives a maximum core loss of 159mW.

  The copper loss is now restricted to 374mW which can equally be divided amongst the two windings.

---

*) In this SMPS example the RM core choice does not imply that it is the best overall choice. Other types should be considered as well. Acquaintance and possesion of RM types was the authors' selection reason.
For any customized supply, Philips magnetics department can offer the best and cheapest solution for your transformer requirement.

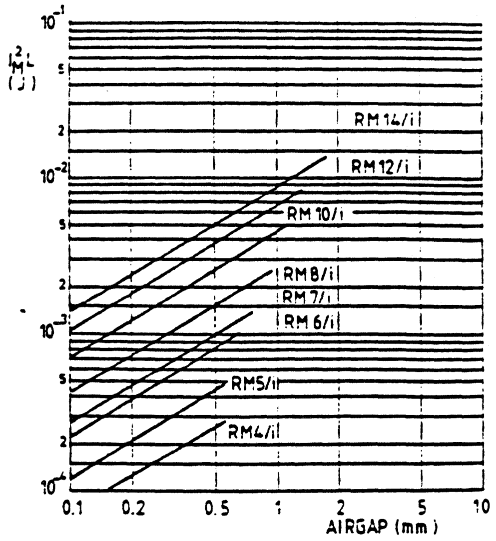


Fig. 5 Maximum energy versus airgap for RM range.

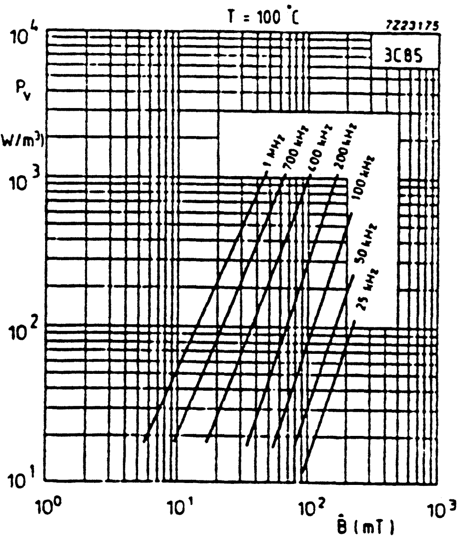


Fig. 6 Specific power loss versus peak flux density with frequency as parameter

4.4 Setting the design start parameters.

A completely empty cell can have a voltage that is near zero or even slightly negative but as soon as charging starts, a healthy cell will quickly restore its voltage to beyond 0.9V.
It is therefore unrealistic to include a very low output voltage condition into the transformer design limits as this might lead to a more bulky transformer due to less transformer utilization (increased dead time) at the high power operation.

- The design example is based on a transformer secondary voltage range factor of: $V_{sec(max)}/V_{sec(min)} = 2$
- A maximum duty cycle has to be asigned. Chosen is : $\delta_{max} = 0.5$
- Derived from [9] $\delta_{sec} \leq 0.45$. To allow for tolerances a wider margin is included into δ_d, so: $\delta_{sec} = 0.4$
- Secondary current sense resistor : $R_{s\ sec} = 0.2\Omega$

4.5 Calculations on the power stage

Contrary to that in a mains operated supply, the resistive losses in a low input voltage supply will have a higher influence on the transformation conditions. Therefore a more detailed calculation approach is followed which includes all power stage components. The calculation approach goes from secondary to primary side and applies to the limit case of maximum power conversion.
Series resistances with the transformer are taken not to affect the current ramp linearity in both primary and secondary currents.

Refer to the requirements data of the supply example (section 2) and the design start parameters (section above) to work out using a formula, the result which is given adjacent;

Maximum output power	$P_o = V_{o(max)} \cdot I_o$	6.4W
Average secondary pulse current	$I_{sec\,(pulse)} = \frac{I_o}{\delta_{sec}}$	1.56A
Peak secondary pulse current	$I_{sec\,(pk)}= 2 \cdot I_{sec\,(pulse)}$ [triangle waveform]	3.13A
RMS value of output current	$I_{RMS} = I_{sec\,(pk)} \sqrt{\frac{\delta_{sec}}{3}}$	1.14A
Square of RMS output current	$I_{RMS}^2 = \frac{4 \cdot I_o^2}{3 \cdot \delta_{sec}}$	1.3A
Loss secondary sense resistor	$P_{R\,sense} = I_{RMS}^2 \cdot R_{sense}$	0.26W
Loss secondary winding [section 4.3]	$P_{tr\,Rsec} = \frac{P_{copper\,max}}{2}$	0.19W
Max. resistance secondary winding	$R_{tr\,sec} = \frac{P_{tr\,Rsec}}{I_{RMS}^2}$	0.15Ω
Secondary diode loss	$P_{D\,sec} = I_o \cdot V_{D(max)}$ [$V_{D(max)}$= 1V]	0.63W
Transformer secondary power	$P_{tr\,sec} = P_o + P_{tr\,Rsec} + P_{R\,sec} + P_{D\,sec}$	7.45W
Transformer secondary voltage	$V_{tr\,sec}= V_o + V_{D(max)} + \{I_{sec\,(pulse)} \cdot (R_{s\,sec}+R_{tr\,sec})\}$	11.75W
Transformer primary power (Leakage inductance loss is neglected)	$P_{tr\,pr}= P_{tr\,sec} + P_C$	7.61W
Transformer primary power also	$P_{tr\,pr} = V_{in(min)} \cdot I_{in} - P_{Rpr(tot)}$	

The primary current has to be deducted and known are; $V_{in(min)}$, primary winding loss and the peak voltage across the primary sense resistor. $R_{DSon(max)}$ of the MOSFET is also required.
For the resistive losses the RMS current is essential.

For a MOSFET selection the approximate primary input current is:

$$I_{in} = Po / Vin\,(min) / \eta = 6.4 / 9 / 0.7 = 1A$$

$$I_{pr\,RMS}^2 = 4 x I_{in}^2 / 3 / \delta_{max} = 2.6\ A^2$$

If MOSFET dissipation ≤ 0.05x Pin = 0.45W then its on resistance ≤ 0.45/2.6 = 0.18Ω at Tj = 100°C.

At Tj = 25°C $R_{DSon} \leq 0.18/1.4 = 0.13\Omega$.

Within the 60V range, BUK442 ,452 or 472 can be applied.

To continue, first some formulae to enable further calculations.
Energy is built up during the on time period, δ_{max}, so;

Transformer primary power $\quad P_{tr\,pr}= I_{pr(pulse)} \cdot V_{tr\,pr(pulse)} \cdot \delta_{max}$ [11]

where	$V_{tr\,pr(pulse)} = V_{in} - V_{Rpr\,tot(pulse)}$	[12]
RMS value of primary current	$I_{prRMS} = 2\cdot I_{pr(pulse)}\cdot\sqrt{\frac{\delta_{max}}{3}}$ (triangle waveform	[13]
Maximum primary winding resistance	$R_{trpr} \le \frac{3\cdot P_{Rtr\,pr}}{4\cdot I_{pr(pulse)}^{2}\cdot\delta_{max}}$	[14]
Average primary winding pulse voltage	$V_{Rtr\,pr(pulse)} = \frac{3\cdot P_{Rtr\,pr}}{4\cdot I_{pr(pulse)}\cdot\delta_{max}}$	[15]
Average primary sense pulse voltage	$V_{Rs\,pr(pulse)} = \frac{V_{Rs\,pr(pk)}}{2} = 0.45V$ [section]	[16]

Substituting [11 - 16];

Average primary pulse current	$I_{pr\,(pulse)} = \frac{P_{trpr}/d_{max}}{(V_{in}-0.45)-I_{pr(pulse)}\cdot R_{DSon}-\frac{3\cdot P_{Rtrpr}}{4\cdot I_{pr(pulse)}\cdot\delta_{max}}}$	[17]
solve I_{pr} (pulse) by	$I_{pr\,(pulse)} = \frac{-b\pm\sqrt{b^{2}-4\cdot a\cdot c}}{2\cdot a}$ were: $a = R_{DSon(max)}$ $b = -V_{in}+0.45$ $c = \frac{\frac{3}{4}P_{Rtrpr}+P_{trpr}}{d_{max}}$	1.89A
Primary peak current	$I_{pr\,(pk)} = 2\cdot I_{pr\,(pulse)}$ (triangular waveform)	3.78A
RMS value of input current	$I_{prRMS} = I_{pr\,(pk)}\sqrt{\frac{\delta_{max}}{3}}$	1.54A
Primary sense resistor	$R_{s\,pr} \le \frac{0.9}{I_{pr(pk)}}$	0.24A
Primary winding resistance	$R_{tr\,pr} \le$ [14]	79mΩ
Primary resistive loss	$P_{Rtot\,pr} = I_{prRMS}^{2}\cdot(R_{s\,pr}+R_{DSon}+R_{tr\,pr})$	1.18W
Input power	$P_{in} = P_{tr\,pr}+P_{R\,pr}$	8.79W
Transformer primary voltage	$V_{tr\,pr} = V_{in\,(min)} - \{I_{pr\,(pulse)}\cdot R_{pr(tot)}\}$	8.06V
Transformer winding ratio	$N = \frac{N_{pr}}{N_{sec}} \ge \frac{d_{max}\cdot V_{trpr}}{d_{sec}\cdot V_{trsec}}$	≥0.86
Primary inductance	$L_p \le \frac{V_{trpr}\cdot\delta_{max}}{I_{pr(pk)}\cdot f}$	17.8µH
Transformer windings	$N_{pr} \le \sqrt{\frac{L_p}{A_L}}$; $N_{sec} \le \frac{N_{pr}}{N}$ resp.	< 8.44 ; < 9.8

The transformer windings must consist of whole turns numbers:	Npr; Nsec = 8 ; 9
The practical primary inductance with 8 turns:	N_{pr} = 16µH,
The lower N_{pr} changes the primary peak current to maximum:	$I_{pr\,pk} = 3.78\sqrt{\frac{17.8}{16}} =$ 4A
and the primary sense resistor becomes:	$R_{s\,pr} \leq 0.9/4 =$ 0.22Ω

Primary blocking voltage across	$V_{pr}\,max = V_{in\,(max)} + V_{tr\,sec} \cdot N + \Delta V$	30.5V + ΔV
the switch	$(\Delta V = I_{pr\,(pk)}\sqrt{\frac{L_L}{C_{tr}}})$; for ΔV=20V, L_L/C_{tr}	< 27.7 nH/nF

The transformer windings were bifilar wound which gave a very low leakage inductance. The ΔV was 20V at maximum power and a snubber could be omitted. In the case a snubber is required then a zener diode, V_Z = ΔV max, in series with a reverse blocking diode can be applied across the primary winding.

5 TEA1100 SETTINGS

In this system the current feedback signal is derived from the analog output of TEA1100, so there is no link between the SMPS frequency and the TEA1100 frequency like in the system as in ref.1. (chpt.8)

Here following the TEA1100 calculation procedure with the formulae is given . It is also possible to get results quickly from the nomograph given in fig. 8.

5.1 Time settings

Start with the requirement for the maximum fast charge time, TO, which should be about 25 - 50% above the maximum expected 1Hr charge time by -ΔV cut-off. A 50% longer TO = 1.5Hr. (This can be altered by changing the oscillator capacitor value).

Directly related to TO are:

• Monitor disable time at start of charge	$t_{disable}$	$= 2^{-5} \times TO$	= 2.8 min.
• Trickle charge repetition time	Ttrickle	$= 2^{-14} \times TO$	= 330 ms.

Now a frequency has to be chosen using the pre scaler factor p = 1, 2 or 4.
Choosing p=1 by setting (PR [pin8] to Vsource [pin6]);

• Oscillator frequency	f_{osc}	$= \frac{2^{26} \times p}{TO}$	= 12.4kHz
• Selection of Rref and Cosc by	R x C	$= \frac{1}{0.93 \times f_{osc}}$	$= 87 \cdot 10^{-6}$
• 12.5k < Rref < 125k	Rref selected		= 27kΩ
	Cosc		= 3.3 nF

From fosc or Tosc derive;

• Monitor sampling repetition	Tsampling $= 2^{16} \times$ Tosc	= 5.3s
• Monitor sampling charge inhibit period	t inhibit = 10 x Tosc	= 805µs
• Trickle charge ON duration	t trickle $= 0.75 \times 2^{9} \times$ Tosc	= 31ms

- Duty cycle trickle charge pulses $\delta\,\text{trickle} = \frac{\text{t trickle}}{\text{Ttrickle}}$ = 9.4%

5.2 Charge current settings

- Current reference resistor;

$$\text{Rib} = \frac{\text{Ich fast x Rref x Rsec sense}}{1.25} = \frac{625\text{mA x } 27\text{k x } 0.2\Omega}{1.25} = 2.7\text{k}\Omega$$

- Trickle charge current;

Rn not applied   Ich trickle $= \frac{\text{Ich fast}}{2} \times \delta\,\text{trickle}$ = 30mA

Rn applied (≥Rref)   Ich trickle $= \text{Ich fast} \times \frac{\text{Rref}}{\text{Rn}} \times \delta\,\text{trickle} =$

25k < Rref < Rn < 250k

Rn = Rref ; 60mA
Rn = 250k ; 7mA

5.3 Current sense input filter

The Cib, Rib time constant should be as small as possible to ensure quick stabilization of the charge current following an interrupt. This occurs at start up, monitor sampling and trickle charge. Therefore allow for the maximum ripple of 0.16V at the IC sense input, ib.
With the formula in section 3.3 the RC can be determined by filling in the derived data (section 4.4, 4.5)

$$\text{Rib}\cdot\text{Cib} \geq \frac{1}{0.16}\,\frac{(0.626 - 0.125)^2 \cdot 0.4}{2 \cdot 0.626 \cdot 60 \cdot 10^3} \geq 8.4 \cdot 10^{-6}$$

Rib = 2k7 so Cib is minimal 3.1nF and a practical value is 3.3nF.

5.4 Vac Setting

The battery voltage has to be adapted to within the Vac monitor input voltage range via a divider.
In appendix 1 divider resistor values with applied number of battery cells are listed for quick resistor selection.
In this design a divider ratio p = 0.33 is chosen (56kΩ and 27kΩ) which allows recording up to a maximum battery voltage of 11.84V (~2V/cell). At an open output, charge supply protection activates at and beyond 13V.
The -ΔV cut-off sensitivity of 1% can be made more sensitive if required in case of NiMH battery use. A value adjustment of the 56KΩ resistor which is connected in series with a low voltage avalanche diode out of the Philips PLVA400A range is all that is needed to achieve a -ΔV of about 0.5%. (Details in ref.1 or 2)
.

5.5 Temperature sensing

An extra protection against fast charge operation when the battery temperature is out of a safe range can be added. A NTC thermistor characteristic is adapted to the sense input of TEA1100 by one or two resistors and a bias voltage. Details are given in Appendix 2.

6 FEEDBACK LOOP STABILITY

In figure 2, chapter 4, the various stages within the current regulation feedback loop have been given. For stability, loop compensation can be set with a capacitor at the C_{ls} node of TEA1100.

The transfer characteristics of the stages are;

6.1 SMPS stage

Based upon formula [ ] the output current can be expressed as: $I_o = \frac{L\, I_{pr}^2}{2\, V_o} f\, \eta$ [21]

The primary peak current is also : $I_{pr} = \frac{(V_c - 1.4)}{3\, R_{s\,pr}}$ [22]

Substitude 21,22: $I_o = \frac{(V_c - 1.4)^2}{V_o} \cdot \frac{L \cdot \eta \cdot f}{18 \cdot R_{s\,pr}^2}$ where the latter term is constant; [23]

k = 0.88 at this example.

For low frequencies the small signal admittance is determined by the differential $\frac{\partial I_o}{\partial V_c}$ at the adjustments of I_o and V_o ;

SMPS transfer: $Y_3 = \frac{\partial I_o}{\partial V_c} = \frac{2 \cdot (V_c - 1.4)}{V_o} \cdot k$ or $2 \cdot \sqrt{\frac{I_o \cdot k}{V_o}}$ by substitution V_c as in [23]. [24]

The SMPS stage ac voltage gain $H1 = \frac{\partial V_{Rs\,sec}}{\partial V_c} = 2 \cdot Rs\ sec \sqrt{\frac{I_o \cdot k}{V_o}}$ [25]

The highest voltage gain will occur at maximum I_o and minimum V_o.

RESULTS FOR H1 : at V_o = 2 V H1 = 0.21x or -13.6dB
at V_o = 10.2V H1 = 0.1x or - 20 dB

The voltage gain is independent of the frequency but a phase shift lag , proportional to frequency, will occur because of the converter flyback operation. By measurement the phase shift was determined to be ;

$\varphi_1 = -2^{\circ}/kHz$

6.2 TEA1100 regulation stage.

The voltage transfer of the input filter $H2 = \frac{Vib}{V_{Rs\,sec}} = \frac{1}{\sqrt{1 + (\omega Rib\ Cib)^2}}$ [26]

$\varphi_2 = -\arctan\ (\omega\ Rib\ Cib)$.

- The voltage transfer of the transconductance error amp stage with a capacitor Cls at output $H3 = \frac{Vls}{Vib} = g_m \frac{Ro}{\sqrt{1 + (\omega\ Ro\ Cls)^2}}$ [27]

(g_m = 250μA/V ; Ro = 4.5MΩ) $\varphi_3 = -\arctan\ (\omega\ Ro\ Cls)$

The voltage transfer of the AO amp $H4 = \frac{Vao}{Vls} = 4x$ or 12dB (Vao = Vc) [28]

The unity bandwidth of AO is 4MHz so $\varphi_4 = 0^\circ$ at frequency < 100kHz

6.3 Total loop response

The loop gain has to be well below zero at the SMPS switching frequency of 60kHz. The unity gain can be chosen at a frequency less than one third the switching frequency, so ≤ 20kHz if the phase margin is adequate ≈ 50 degrees.

The total gain is H1 x H2 x H3 x H4 where H1 x H4 = -2dB max. This gain is frequency independent within the relevant frequency range. At unity loop gain; H2 x H3 = 2dB.
The frequency at unity gain depends on φ_1 and φ_2 . For φ_3 -90° can be filled in when only a compensation capacitor C_{ls} is applied: single low frequency pole, no zero.
The sum of φ_1 and φ_2 is $\frac{-2 \cdot 10^3}{f_u}$ - arctan $2 \pi f_u$ Rib Cib and is allowed to be -180 + 90 + 50(margin) = -40°.

$$f_u \leq 9\text{kHz}$$

At 9kHz ; H2 = - 1.8dB and $\varphi_2 = -27^\circ$. The residual gain for H3 = 3.8dB.

Formula [27] can be rearranged to get the required $Cls = \frac{1}{\omega_u \cdot H3 \cdot Ro}\sqrt{(g_m \cdot Ro)^2 - H3^2}$. For H3 < 200x this formula can be simplified to: $Cls \approx \frac{g_m}{\omega_u \cdot H3}$. The required Cls is 3nF minimal, 3.3nF as practical value.

Figure 7 gives the bode plot of the closed loop response.

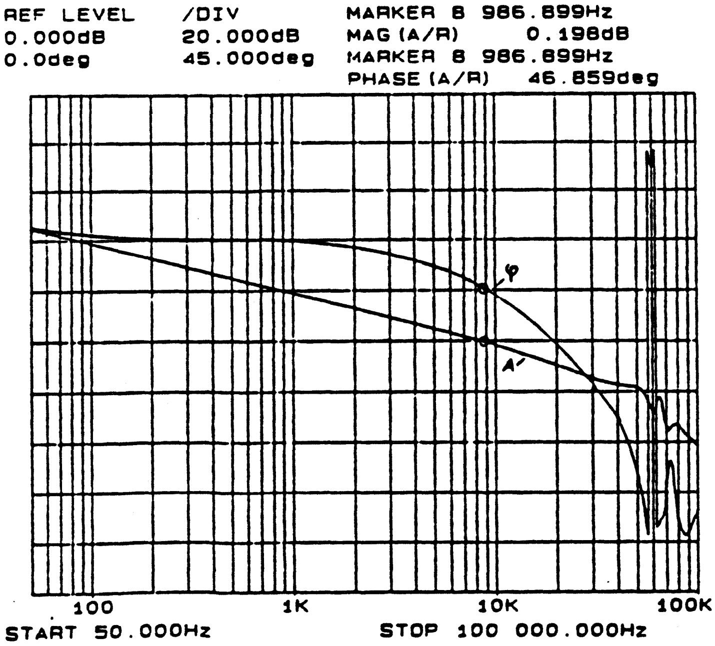


Fig. 7 Closed loop bode plot

7 DESIGN RECOMMENDATIONS

Where high rated pulse currents flow , as in the primary and secondary power loops, care has to be taken for the proper component selection and a well thought out PCB layout
Especially for the low ohmic current sense resistors, avoid a wire-wound type even when specified as low inductive because large voltage spikes is what you get at current switching. Use carbon or metal film types and use mutiple units in parallel to obtain a low value or power rating.

Buffer and decoupling capacitors have to cope with steep pulse currents and should have a low ESR value.

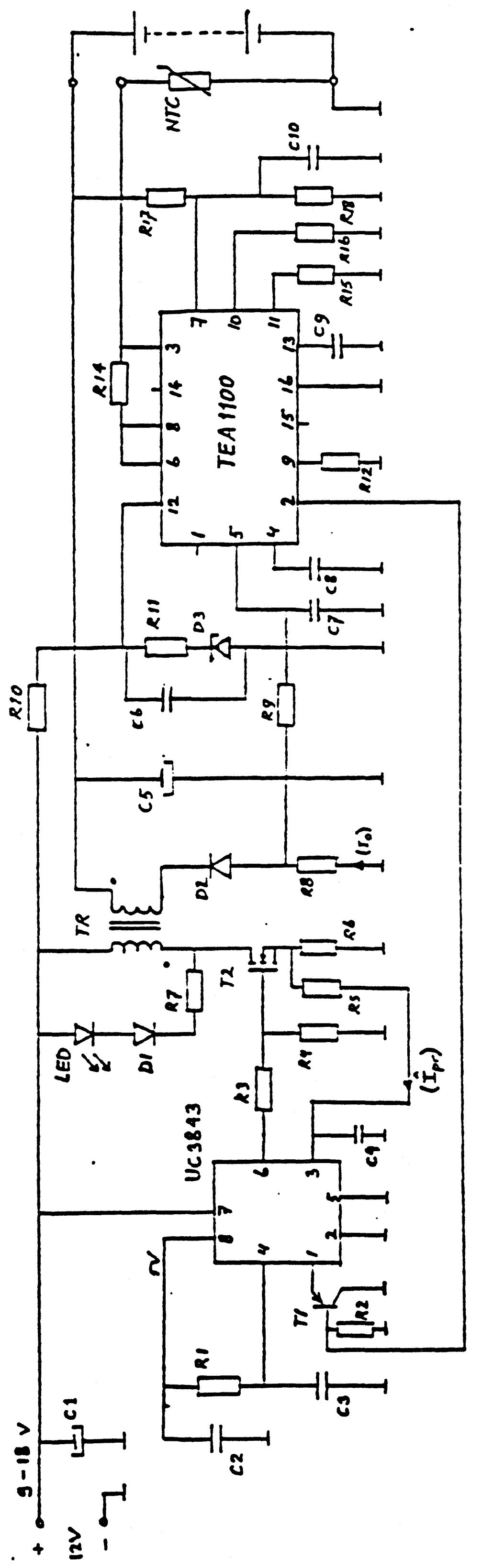
Make the PCB current power loops as close as physically possible. Prevent sharing of ground tracks between the power and small signal circuitries by using star configurations.

Place the supply voltage decoupling capacitor of TEA1100 close to the IC supply and ground pins. A great part of this IC consists of digital circuitry and possible interference from the input supply (insufficient filtering especially with a simple mains adapter) should be filtered via the capacitor.

8 REFERENCES

1 NPO/AN9102b. TEA1100, Versatile battery management IC for NiCd charge systems (updated version).
2 TEA1100(T) Preliminary specification. April 1992.
3 DATA HANDBOOK , linear products IC11; sections: AN125, AN1261, UC3843.
4 3C85 HANDBOOK First in Ferrites. 12NC: 9398 345 90011.
5 DATA HANDBOOK, Soft Ferrites,MA01 12NC: 9398 174 50011.
6 SPO/AN91007 Mains Powered Fast NiCd Battery Charger using TEA1100.
7 SPO/AN92004 TEA1100 battery Charger with Linear Regulator.

9 CIRCUIT DIAGRAM



6 NiCd CELLS (550mAhr) IC CHARGER, Vin-12V.

Fig. 7 Charger diagram

10 PARTS LIST

Resistors	Value	Series	Part no.
R1	10k	MRS16T 1%	2322 157 31003
R2,4,15,18	27k	SFR16T 5%	2322 180 53273
R3	27Ω	SFR16T 5%	2322 180 53279
R5	1K	SFR16T 5%	2322 180 53102
R6	0.22Ω		
	= 5x1.1Ω in parallel	MRS25 1%	2322 156 21108
R7	620Ω	SFR16T 5%	2322 180 53621
R8	0.2Ω		
	= 5x1Ω in parallel	MRS25 1%	2322 156 21008
R9	2.7k	MRS16T 1%	2322 157 32702
R10	300Ω	SFR16T 5%	2322 180 53301
R11	240Ω	SFR16T 5%	2322 180 53241
R14	t.b.f.	SFR16T 5%	2322 180 5
R16	27k	MRS16T 1%	2322 157 32703
R17	56k	SFR16T 5%	2322 180 53563

Capacitors

C1, C5	220µF	RLI 135	2222 135 66221
C2	10nF	ceremic	2222 629 09103
C3	2.7nF	KP 1%	2222 424 42702
C4	470pF	ceramic	2222 630 08471
C6	100nF	MKT 10%	2222 370 11104
C7,8	3.3nF	ceramic	2222.630 09332
C9	3.3nF	KP 1%	2222.424 43302
C10	680pF	ceramic	2222 630 08681

Semiconductors

T1	BC557B	9332 026 20126
T2	BUK443-60B	9332 546 60127
D1	BAV18	9331 892 00153
D2	BYD74B	9337 537 40153
D3	BZX79 C7V5	9331 177 60153
D4	LED	9338 501 20112
IC1	UC3843	
IC2	TEA1100	

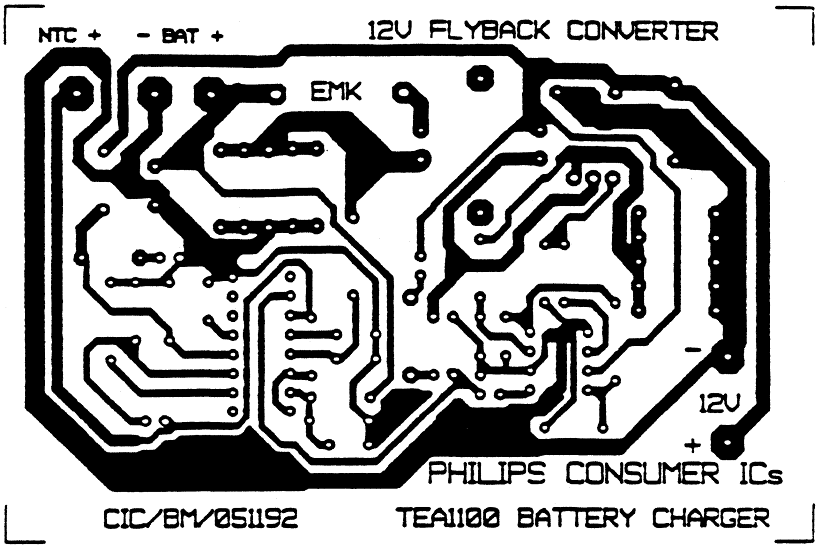
Transformer

RM6/i core set , gap= 0.15mm , 3C85 material ,(Al=250)	4322 025 05069
Coil former	4322 021 34511
Clips 2x	4322 021 34301

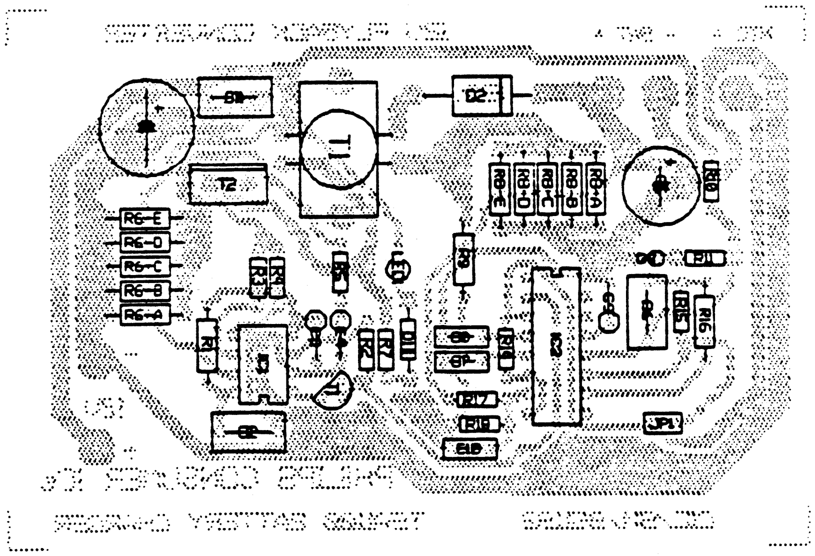
Lp=16µH , primary winding: 8 turns, 0.35mm wire
secondary winding: 9 turns, 0.35mm wire

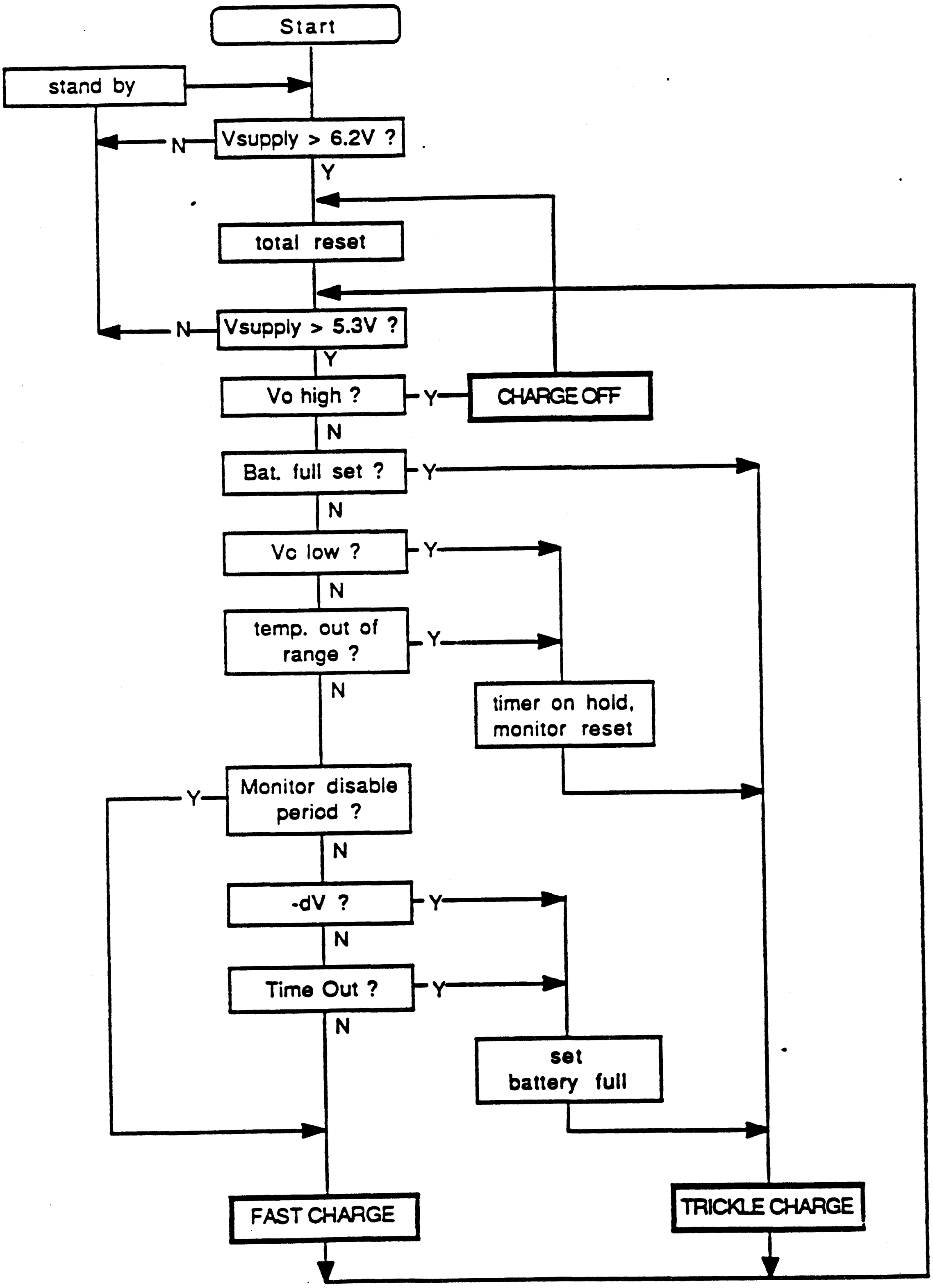
11 PCB LAYOUT

Copper side



Components side

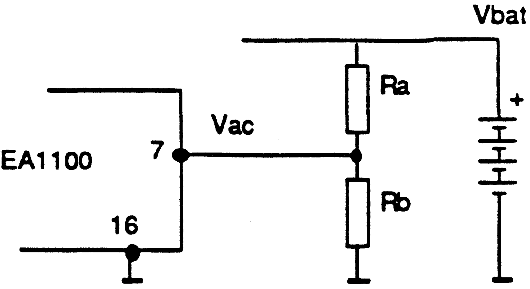




TEA1100 charger flow diagram

Appendix 2

Resistor divider selection for TEA1100 sense input adaptation to the battery.



$$\frac{Ra}{Rb} \geq \frac{V_{BATmax}}{3.85} - 1$$

$$V_{BAT} \text{ high protection} \geq 4.25\left(1+\frac{Ra}{Rb}\right)$$

Cell number	2	3	4	5	6	7	8	9	
V_{BAT} max (1.7V/cell)	3,4	5,1	6,8	8,5	10,2	11,9	13,6	15,3	V
Ra ≥	12	12	22	39	47	68	82	91	kΩ
Rb	-	27	27	27	27	27	27	27	kΩ
V_{BAT} high protection ≥	4,3	6,1	7,7	10,4	11,6	15	17,2	18,6	V

For the design has been chosen : Ra = 56kΩ ; Rb= 27kΩ
where;
V_{BAT} sense max. = 11.84V (2V/cell)
V_{BAT} protection ≥ 13V

The division ratio is the key factor and the absolute resistor values can be chosen within a broad range.
The lower impedance is limited by discharge consideration while the higher impedance should remain low compared to the input impedance of the Vac pin 7 which is more than 200MΩ.
The lower impedance can be seen in relation with the batteries' self discharge. When assumed that the maximum capacity of a unloaded battery reduces by 30% over the first month (720 hours), then the average equivalent self discharge current in 500mAhr battery is in the order of ;

$$\frac{\text{capacity x loss}}{\text{period}} = \frac{\text{500mAhr x 0.3}}{\text{720hr}} = 0.21\text{mA}.$$

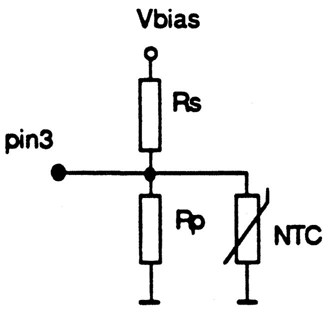
The imaginary battery discharge resistor is about; $\frac{\text{1.27V x cell number}}{\text{Idischarge}} = \frac{1.27 \times 6}{0.21} = 36\text{k}\Omega.$

The total divider load resistance in the table above is based on twice the imaginary discharge resistor in a battery of 500mAhr. For higher capacity ratings the given divider resistors can also be used.

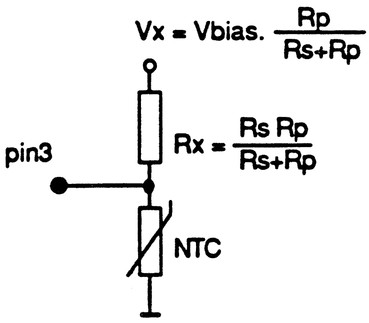
For applications with maximum 2 cells battery no Vac divider is required but a series resistor with the Vac input must be present to limit clamp current at 4.25V , protection level, in case the battery is removed while the charger is on. The limit Vac clamp current is ± 1mA.

Appendix 3

Battery NTC thermistor adaptation to TEA1100 temperature sense input.



Temperature sense network



Equivalent circuit

For the low and high temperature levels the corresponding R_h and R_l thermistor resistance values need to be derived

The following calculations lead to the required battery serie-, parallel resistor and the bias voltage;

low temperature limit , NTC resistance :	R_h	
high temperature limit, NTC resistance :	R_l	ratio; $p = \frac{R_h}{R_l}$
TEA1100 upper threshold	$Vh = 3V$	
lower threshold	$Vl = 0.81V$	range; $\Delta V_t = 2.19V$

If $k = \frac{\Delta V_t}{(p \cdot V_l) - V_h}$ then $Rx = k \cdot R_h$; $Vx = V_h\,(1+k)$

V_B = constant bias voltage ; $Rs(eries) = \frac{k}{3(1+k)} \cdot V_B \cdot R_h$

$(V_B \geq Vx)$ $Rp(arallel) = \frac{k}{V_B - 3(1+k)} \cdot V_B \cdot R_h$

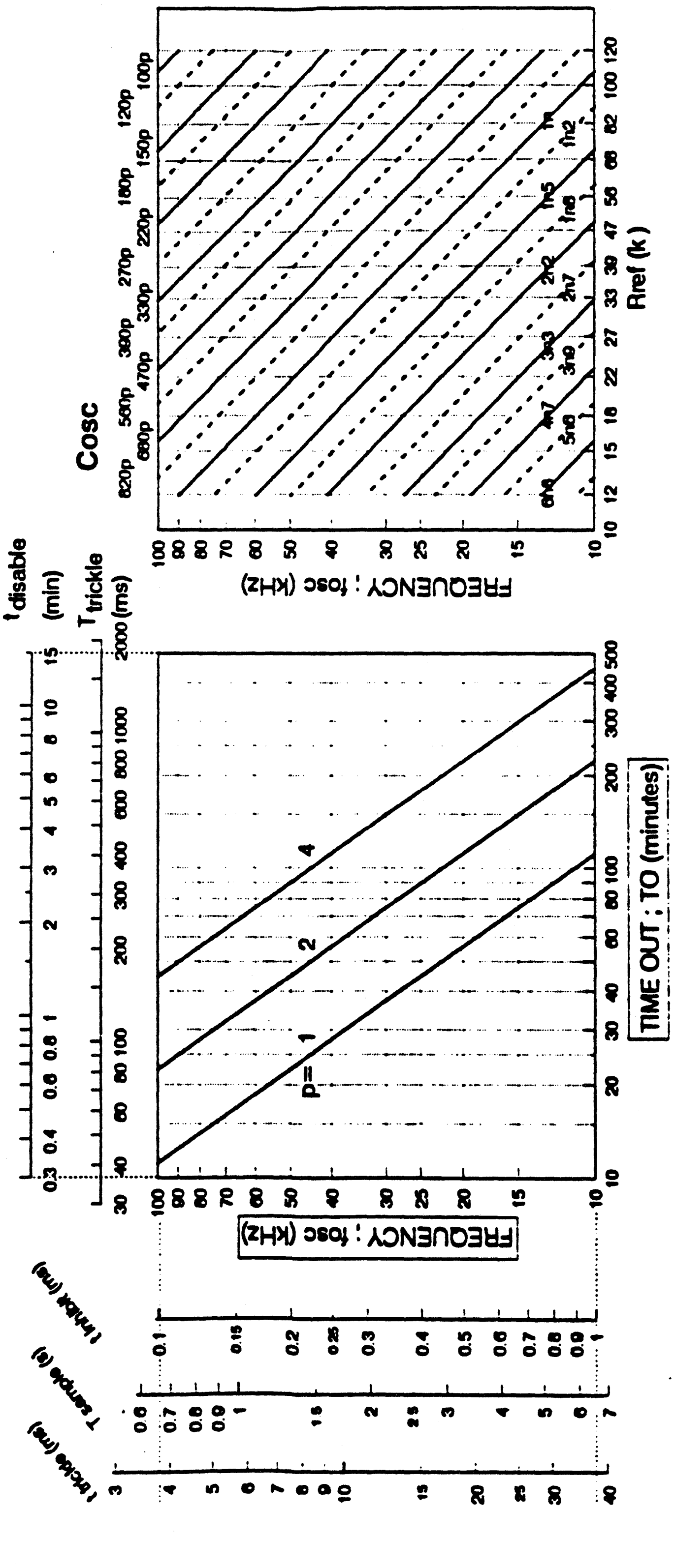
When the B value of the NTC is used then ; $p = e^{B \cdot \left(\frac{1}{Tl} - \frac{1}{Th}\right)}$ T is in Kelvin.

$$R_h = R_{25} \cdot e^{B \cdot \left(\frac{1}{Tl} - \frac{1}{298}\right)}$$

Note: Without temperature sensing ; **a) i.s.o. NTC use resistor with same value of Rx (=R14) or**
b) if resistor terminates pin11 then connect pin 3 to pin11
and skip Rx.

Appendix 4

TEA1100 timing and related components charts



NOMOGRAM OF TIME SETTINGS IN TEA1100

Application Note

ICs for BATTERY MANAGEMENT

TEA1100 Battery Charger with Linear Regulator

Report No: SPO/AN92004 a

R Verney

Product Concept & Application Laboratory Southampton, England.

Keywords
TEA1100
Fast Battery Charger
NiCd
NiMH
Linear Regulator
Cordless Telephone

Date : 08 OCT 1992
Pages : 18

Summary:

The fast NiCd and NiMH battery charger described in this report uses a TEA1100 control circuit, with a simple linear current regulator to provide precise battery charging. It is suitable for small to medium sized battery packs with charge times between 30 minutes and 5 hours. Typical applications include Cordless Telephone, Personal Audio and small domestic appliances.

Table of Contents Page Number

List of Tables

List of Figures

1. INTRODUCTION

The TEA1100 fast charge monitor and control circuit features -dV and timer charge cutoffs, with over and under temperature protection of the batteries. The control circuit is a versatile feature which is suited to all applications; PWM (switch mode), linear regulated supply, or simple non-regulated supply.

Where charge currents are low, or where the electrical noise of a SMPS circuit is unacceptable, a linear regulator circuit provides the best solution to current control with TEA1100.

Key Features:

- TEA1100 provides all battery management functions.
- Fast battery charging.
- Long battery life.
- Simple circuitry - easy to design.
- Low cost.

2. CIRCUIT DESIGN

The linear regulator is formed using the analogue control signal available from the AO pin (pin 2) of the TEA1100, a drive transistor, and a series pass transistor.

A circuit diagram is shown in Fig 1.

2.1 Linear Regulating Pass Transistor and Driver

T1, T2:

T1 and T2 are chosen appropriate to the voltage and power levels present. A heatsink must usually be fitted to T1. At the higher power levels, a large heatsink will be required. A switch mode solution should be carefully considered when power output exceeds 5 W.

D1:

Diode D1 prevents forward bias of the collector base junction of T1 occurring when batteries are inserted but no power applied. This prevents battery discharge.

R2:

R2 is chosen to provide sufficient base drive current to T1.

Table 1 Component Choice

Max no of cells	Max charge current	T1	T2	D1	R2
10	350 mA	BD132	BC549	IN4001	180R
10	1 A	BDT60A	BC635	IN4001	180R
10	1.2 A	BDT60A	BC635	IN5401	180R
6	2.2 A	BDT60A	BC635	IN5401	180R
3	3 A	BDT60A	BC635	IN5401	100R

2.2 Setting the System Timings

The system timings can be freely chosen to set the maximum time spent in fast charge, termed 'Time Out' period.

C3:

C3 sets the clock frequency, and hence all system timings.

$\text{Tosc} = 0.93 \times \text{R6} \times \text{C3}$

$\text{Time Out} = \text{Tosc} \times 2^{26} \times p$
$\text{Sample repetition} = \text{Tosc} \times 2^{16}$
$\text{Trickle repetition} = \text{Tosc} \times 2^{12} \times p$

where p is the prescaler factor set by the PR pin.

$p = 1$ PR = Vsource (pin 6), $p = 2$ PR = open, $p = 4$ PR = GND.

The prescaler setting affects the ratio of fast charge to trickle charge currents. With $p = 1$, the range of ratios is 10:1 to 20:1, $p = 2$, 20:1 to 40:1, $p = 4$, 40:1 to 80:1.

2.3 Setting the Charge Currents

R1, R3, R6, R7:

R1, the current sensing resistor should be chosen to give a voltage between 50 and 200 mV. This ensures accuracy, while keeping power loss in the resistor to a minimum.

R6, R7 determine the fast and trickle charge reference currents respectively. In turn, these reference currents, in combination with R1 and R3, set the charge currents.

For R6 a value of 27 k can usually be used, R3 and R7 can then be chosen appropriately.

I fast = Reference current (fast) × resistor ratio.

$$I\,fast = \frac{1.25 \times R3}{R6 \times R1}$$

Therefore, having set R6 and R1:

$$R3 = \frac{R6 \times R1 \times I\,fast}{1.25}$$

Values for R3 will usually be between 1 k and 10 k.

$$I\,trickle = \frac{1.25 \times R3 \times 0.094}{R7 \times R1 \times p}$$

Therefore:

$$R7 = \frac{1.25 \times R3 \times 0.094}{I\,trickle \times R1 \times p}$$

where p is 1, 2, 4 set by PR pin. See Section 2.2.

NB: 1 If R7 is greater than twice R6, it can be omitted, and a default reference current in trickle charge will be taken of half that used in fast charge.

2 R7 should generally not be less than R6, because the pulse of trickle current should not be of greater instantaneous magnitude than the fast charge current. In certain circumstances, such as where a supply of large enough current capability is available, this may be permissible however.

2.4 Control Loop Components

R2, R10:

To prevent excessive loop gain, local negative feedback is applied to T2 by R2. A value is chosen to ensure sufficient base current for T1 with AO at its maximum of 3.6 V, and allowing for minimum gain in T1. R10 ensures correct turnoff of T1.

C2:

C2 is connected to maintain loop stability. For the linear regulator its value is noncritical and can be found empirically. The linear regulator is inherently very stable with rapid transient response. A value of 10 nF will usually suffice.

NB: The capacitor on the IB pin (pin 5), fitted in SMPS circuits, must be omitted with the linear regulator circuit. Instability will result if an IB pin capacitor is used.

2.5 Designing the Battery Voltage and -dV Detection Circuit

R9, R8:

R9 and R8 form the VAC (accumulator voltage) divider chain.

The VAC thresholds are:

0.3 V assume short circuit batteries. Trickle charge.
0.385 V minimum voltage for -dV detection.
3.85 V maximum voltage for -dV detection.
4.25 V assume open circuit batteries. Zero current.

R9 and R8 must be carefully chosen to keep the battery voltage within the -dV detection range, and to make sure that at open circuit the input voltage to the regulator is sufficient to cause an open circuit shutdown.

Table 2 lists appropriate values. These values apply equally well for all applications of TEA1100.

Table 2 VAC Divider Resistor R9, with R8 = 100 k

Number of cells	R9
2	10 k
3	47 k
4	100 k
5	150 k
6	180 k
7	220 k
8	270 k
9	330 k
10	390 k

C5:

Output capacitor C5 must be present (typically 100 µF or more) to stop rapid oscillation and LED flicker in the case of open circuit batteries.

2.6 Using the Temperature Protection Block

NTC input:

Where temperature protection is not required, the NTC pin (pin 3) can be tied to pin 10 or 11, reference current generator pins.

For connection of an NTC, see Application Note SPO/AN91007, page 37, 38, 45. The NTC circuit around pin 3 and calculations are correct for all applications of TEA1100.

2.7 TEA1100 Supply Connections

The voltage supply to TEA1100 should not exceed 12V. A supply decoupling RC filter should be applied to limit the voltage rise to ≤0.6V / µs at switch-on to ensure correct power-on reset. This filter also suppresses supply interference.

Where the DC input voltage to the circuit can exceed 12V, the supply to TEA1100 can either be voltage regulated or current-fed via a dropping series resistor.
A current supply to TEA1100 is allowed up to 15mA, while the IC internally zener clamps its supply voltage at 12V.

- Input supply voltage till 13V.

Apply at pin 12 a RC network R4 / C1: 68Ω / 330nF.
The operational DC input voltage range is between 5.9 and 13V.
At switch-on the unloaded input voltage should reach 6.7V to initialize TEA1100.

- Input supply voltage beyond 13V.

a) A 8V, 3-terminal positive regulator IC or a discrete series voltage limiter, (NPN transistor, zener, resistor) can be applied together with the decoupling filter R4 / C1.

b) TEA1100 supply adaptation can be simplified by using R4 as supply current-fed dropping resistor. R4 is determined by;

R4 = (Vin max - 12) / 15 (kΩ) where Vin max is maximum average input voltage, not charge current loaded.

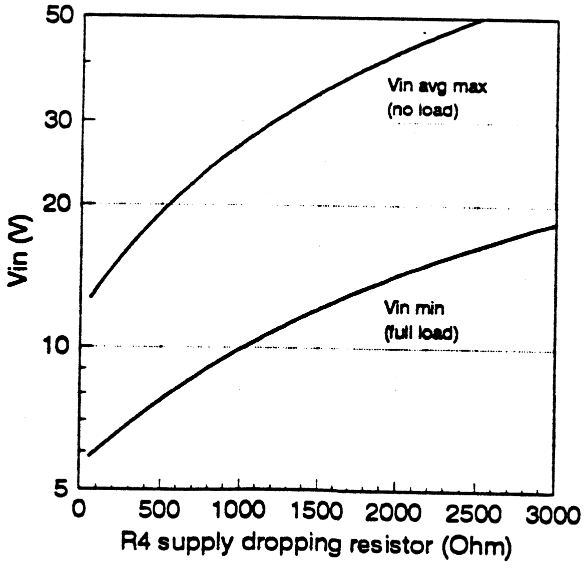
For continuous operation the loaded input voltage must not drop below;

Vin minimum = 4.3 x R4(kΩ) + 5.6V.

or else system reset might occur.

*) Note: If current-fed and input voltage >13V then via R5 and the LED, when not driven, additional unwanted current can be bypassed through the internal protection diode of pin 15 to the 12V zenered supply. To prevent this a resistor Rp can be added across the node R5 / LED anode and ground.
Resistor Rp value; ≤ 13 x R5 / (Vin max - 13).

Input voltage limits with TEA1100 supply drop resistor.



3. MAINS INPUT

To use the linear regulator TEA1100 battery charger from mains requires an AC/DC convertor circuit. Either an existing power supply, a dedicated SMPS or low frequency transformer can be used.

The most simple low frequency transformer circuit is described here. The circuit consists of a step down transformer, a DC rectifier and reservoir capacitor. For economic design, the voltages and ratings of these parts must be chosen quite carefully.

A circuit diagram of the mains input section is shown in Fig 2.

The maximum charge current (Iout), and the number of cells to be charged must be known. The charge current is found from the battery capacity, and required fast charge time. A multiplying factor of 1.4 allows for the charge acceptance factor (which is less than unity) of the battery.

$$Iout = \frac{Ah \times 60 \times 1.4}{Charge\,Time\ (minutes)}$$

3.1 Reservoir Capacitor Values and Ratings

The minimum capacitor value required is calculated from Table 3 as the charger output current times the capacitance per ampere figure given. This gives the capacitor value for 30% ripple. The next larger value capacitor should be chosen. Using an excessively large capacitor value will cause a large ripple current to flow and is to be avoided. Smaller capacitors should not be used, as this could lead to the charger dropping out of regulation, leading to erratic performance.

The capacitor voltage ratings quoted in the table allow for 10% mains voltage variation, and 20% load regulation of the mains transformer. Also it is assumed that a transformer with the correct output voltage is used. If the load regulation of the mains transformer is greater than 20%, or a higher voltage transformer is used, then a higher voltage capacitor may be needed.

The ripple rating (in amps RMS) of the reservoir capacitor must be at least 2 times the maximum DC output current of the battery charger. Philips 056 series capacitors will prove suitable in most cases.

3.2 Transformer Voltages and Ratings

The VA rating of the mains transformer should be found as:

$$VA = 4 \times Iout\,(n + 1.5)$$

Where Iout is the output current of the regulator, and n is the number of cells.

This calculation assumes that the correct voltage mains transformer is used.

Overloading the mains transformer will lead to overheating and a possible fire hazard.

The secondary voltage of the mains transformer quoted in Table 3 is carefully chosen to allow for ±10% mains variation, diode drops in the circuit, and to keep power dissipation in the linear regulator to a minimum. Take care that the transformer is designed to give the recommended secondary voltage at maximum loading.

3.3 Rectifier Diode Ratings

The rectifier diodes must be chosen so that their I_F (av) rating is greater than the maximum output current, and that their V_{RRM} rating is at least equivalent to the DC voltage rating of the reservoir capacitor.

For up to 1 A, IN4001 are suitable, and up to 3 A, IN5401.

Table 3 Mains Input Section Components

Number of cells to be charged	Transformer secondary voltage V RMS, full load	Capacitor value µF / A	Capacitor voltage rating V DC
2	7.1	4000	16
3	9.1	3000	25
4	11.2	2400	25
5	13.2	2000	35
6	15.2	1700	35
7	17.3	1500	40
8	19.3	1300	40
9	21.2	1200	50
10	23.3	1100	50

For further details on TEA1100 please refer to Philips Semiconductors Notes NPO/AN9102 and SPO/AN91007.

4. DESIGN EXAMPLE

4.1 Design Specification

Charger requirement: 1 hour charger, 3 AA cells, UK mains input.

4.2 Design Procedure

Step 1. How much charge current is required?

$$Iout = \frac{Ah \times 60 \times 1.4}{Charge\,Time\,(minutes)}$$

To charge a 600 mAh battery in 1 hour or less requires typically:

$$Iout = \frac{600\,mAh \times 60 \times 1.4}{60} = 840\,mA$$

Step 2. Select mains components.

Transformer VA rating = 4 × Iout × (n + 1.5) = 4 × 0.84A × (3 + 1.5) = 15 VA

Transformer primary voltage = 240 V RMS

Transformer secondary voltage (from Table 3) = 9.1 V RMS, full load.

Reservoir capacitor value (from Table 3) = 3000 µF / A × 0.84 A = 2520 µF

Next larger capacitor = 3300 µF

Reservoir capacitor voltage (from Table 3) = 25 V

Reservoir capacitor ripple rating = 2 × Iout = 2 × 840 mA = 1680 mA

Chosen capacitor: Philips 056 series type 2222 056 56332

NB: Check ripple rating.

Rectifier diodes: IN4001

Step 3. Select regulating transistor and driver.

Using Table 1:

T1 = BDT60A, T2 = BC635, D1 = IN4001, R2 = 180R

Step 4. Calculate heatsink size.

The maximum power dissipation in the series pass transistor will be:

Pdiss = 1.30 × Iout × (V sec - 2.0)

where V sec is the RMS AC secondary voltage of the mains transformer, from Table 3.

Pdiss = 1.30 × 0.84 × (9.1 - 2.0) = 7.8 W

for a maximum temperature rise of 55 °C, the required heatsink size is:

55 °C / 7.8 W = 7.0 °C / W

Step 5. Evaluate viability.

Having completed steps 1 to 5, all the major cost and size determining items in the circuit have been specified. From this it can be decided whether or not the linear regulator is suitable, or whether a switch mode solution is required.

Step 6. Setting the charge currents.

Starting with R6 = 27 k,

Choose R1 to give between 50 and 200 mV.

For 840 mA, a value of R1 = 0R1 will give a voltage of 84 mV, inside the desired range.
R1 = 0R1

$$R3 = \frac{R6 \times R1 \times Ifast}{1.25}$$

$$R3 = \frac{27\,k \times 0R1 \times 0.84}{1.25} = 1814\,R = 1k8$$

Trickle charge is chosen as one twentieth Ifast, hence prescaler factor = 1 with R7 open. Therefore PR connected to Vsource.

Trickle charge current = 42 mA

Step 7. Setting system timings.

Time out can be set at 60 minutes.

R6 = 27 k, p = 1

$$C3 = \frac{60 \times TimeOut\ (minutes)}{0.93 \times R6 \times p \times 2^{26}}$$

$$C3 = \frac{60 \times 60}{0.93 \times 27000 \times 1 \times 67108864} = 2n2$$

Step 8. Battery voltage and -dV detection circuit.

From Table 2, R8 and R9 are chosen.

With n = 3, R8 = 100 k, R9 = 47 k

Step 9. Voltage Regulator bias.

R11 is chosen for a minimum of 0.2 mA in Zener D3. Minimum input voltage is 10 V, therefore R11 = 0.9 V / 0.2 mA = 4k7.

Table 4 List of Parts, Linear Regulator Design Example

R1	0R1
R2	180R
R3	1k8
R4	link
R5	2k2
R6	27 k
R7	open
R8	100 k
R9	47 k
R10	1 k
R11	4k7
C1	open
C2	10 nF
C3	2n2
C4	-
C5	100 μ, 25 V
C6	10 n
D1	IN4001
D2	3 mm LED
D3	BZX79C9V1
T1	BDT60A
T2	BC635
T3	BC549
IC1	TEA1100

Table 5 List of Parts, Mains Input Section Design Example

TR1	240 V primary, 9.1 V secondary, 9 VA
D1-D4	IN4001
C1	3300 μ, 25 V Philips 056 series
F1	500 mA fast blow

Table 6 List of Parts, Linear Regulator Section

R1	See Section 2.3
R2	100R / 180R. See Section 2.1
R3	See Section 2.3
R4	68R. See Section 2.7
R5	2k2
R6	See Section 2.3
R7	See Section 2.3
R8	100 k
R9	See Table 2 and Section 2.5
R10	1 k
R11	See Section 2.7
C1	330 n. See Section 2.7
C2	10 n. See Section 2.4
C3	See Section 2.2
C4	-
C5	100 µ DC voltage as mains reservoir capacitor. See Section 2.5
C6	10 n. See Section 2.7
D1	IN4001/IN5401. See Section 2.1
D2	3 mm LED
D3	9V1 Zener. See Section 2.7
T1	BD132/BDT60A. See Section 2.1
T2	BC549/BC635. See Section 2.1
T3	BC549. See Section 2.7
IC1	TEA1100/TEA1100T

Table 7 List of Parts, Mains Input Section

TR1	Mains transformer. See Section 3.2
D1-D4	Rectifier diodes. See Section 3.3
C1	Reservoir capacitor. See Table 3, Section 3.1
F1	500 mA fast blow

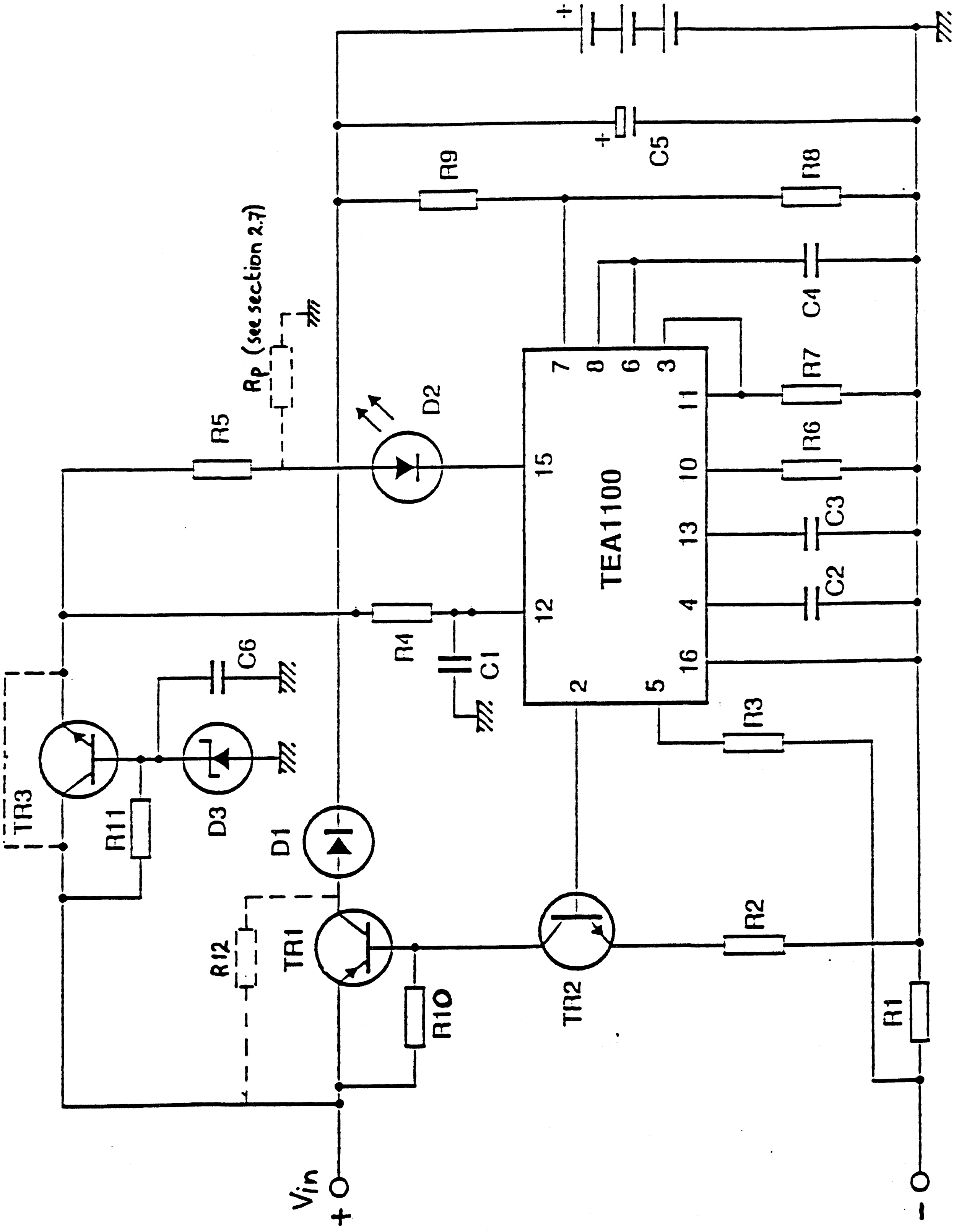


Fig 1   TEA1100 Battery Charger with Linear Regulator Circuit

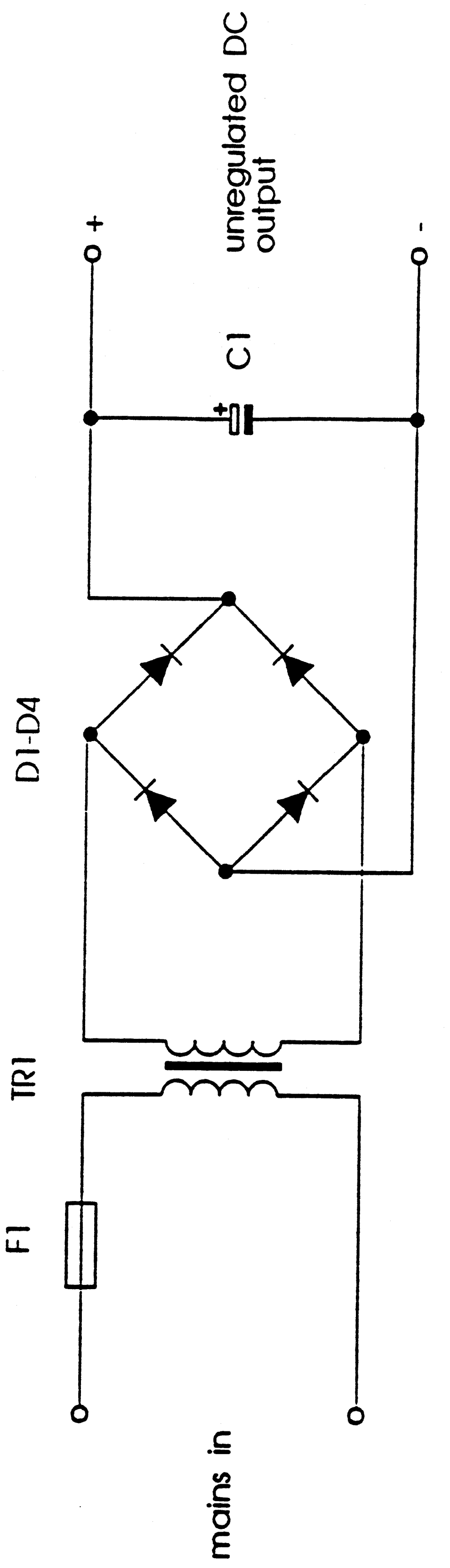
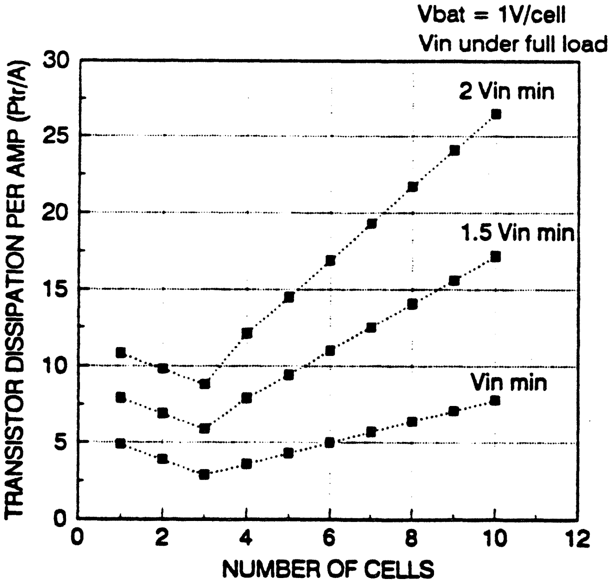


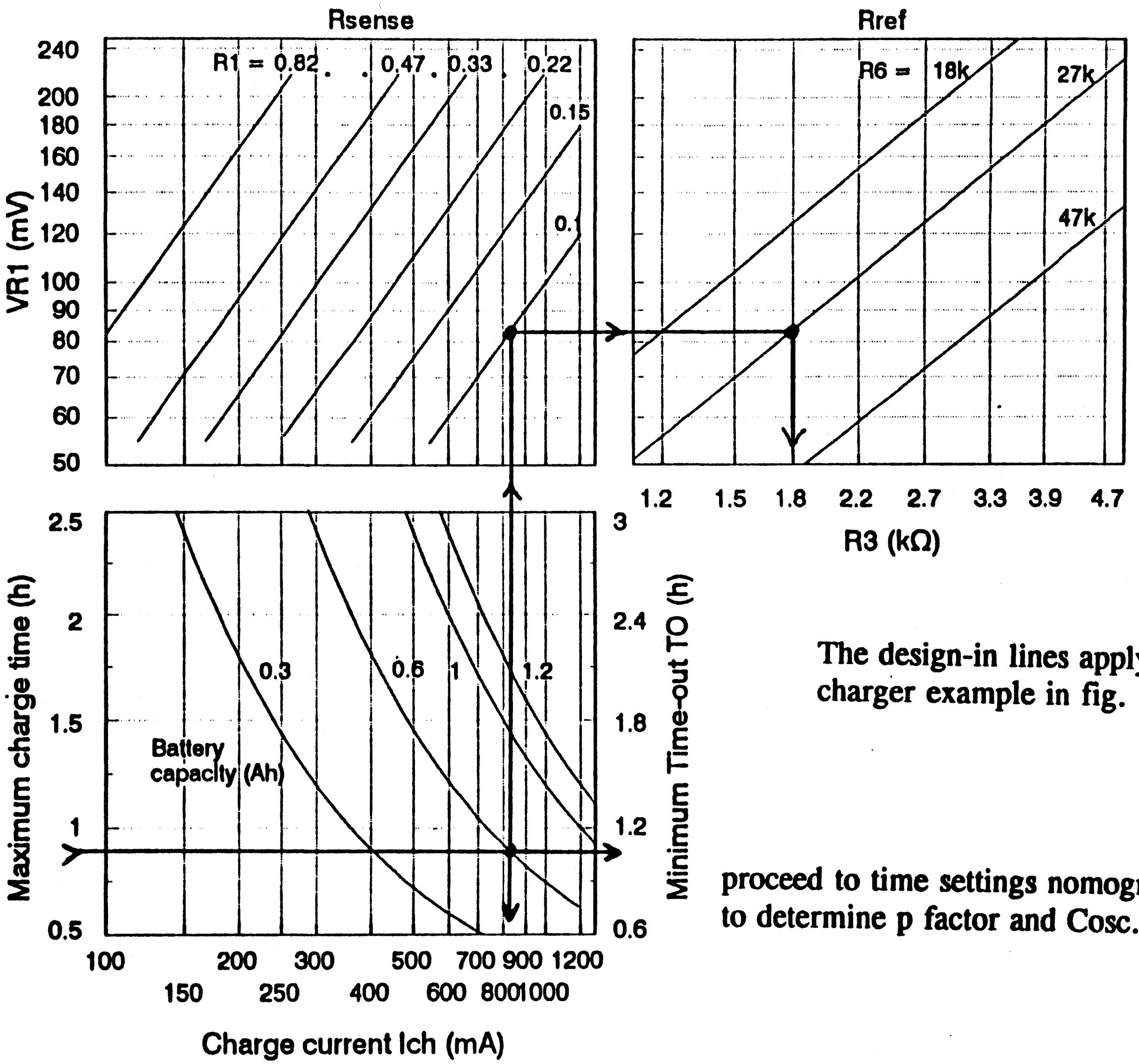
Fig 2 Mains Input Circuit

Survey of settings for 1 - 10 battery cells application.

		Vcell max = 1.7V			T1;Vce sat max = 0.8V			Vd1 max = 0.9V			
number of cells		1	2	3	4	5	6	7	8	9	10
D1 required		no	no	no	yes	yes	yes	yes	yes	yes	yes
Vin min - full load	V	5.9	5.9	5.9	8.5	10.2	11.9	13.6	15.3	17	18.7
Vin min - no load	V	5.9	5.9	5.9	8.7	10.4	13.	14.9	17.2	19.7	22.5
Vin avg max - no load	V	13	13	13	20	27	30	39	45	45	53
Vbat max (Ifast)	V	1.7	3.4	5.1	6.8	8.5	10.2	11.9	13.6	15.3	17
Vbat full (Itr)	V	1.4	2.8	4.1	5.5	6.9	8.3	9.7	11.	12.4	13.8
Vbat high prot.=>	V	4.3	4.3	5.7	7.7	10.6	11.9	13.6	15.7	18.3	20.8
Vbat low prot.=<	V	0.3	0.3	0.4	0.55	0.75	0.84	0.96	1.11	1.29	1.47
R4 (supply)	Ω	68	68	68	560	1000	1200	1800	2200	2200	2700
R12 (Rshunt)	kΩ	10	8.2	8.2	15	18	22	27	33	33	39
R8 (Rvac-)	kΩ			100	100	100	100	100	100	100	100
R9 (Rvac+)	kΩ	12	12	33	82	150	180	220	270	330	390

Dissipation in regulator power transistor.

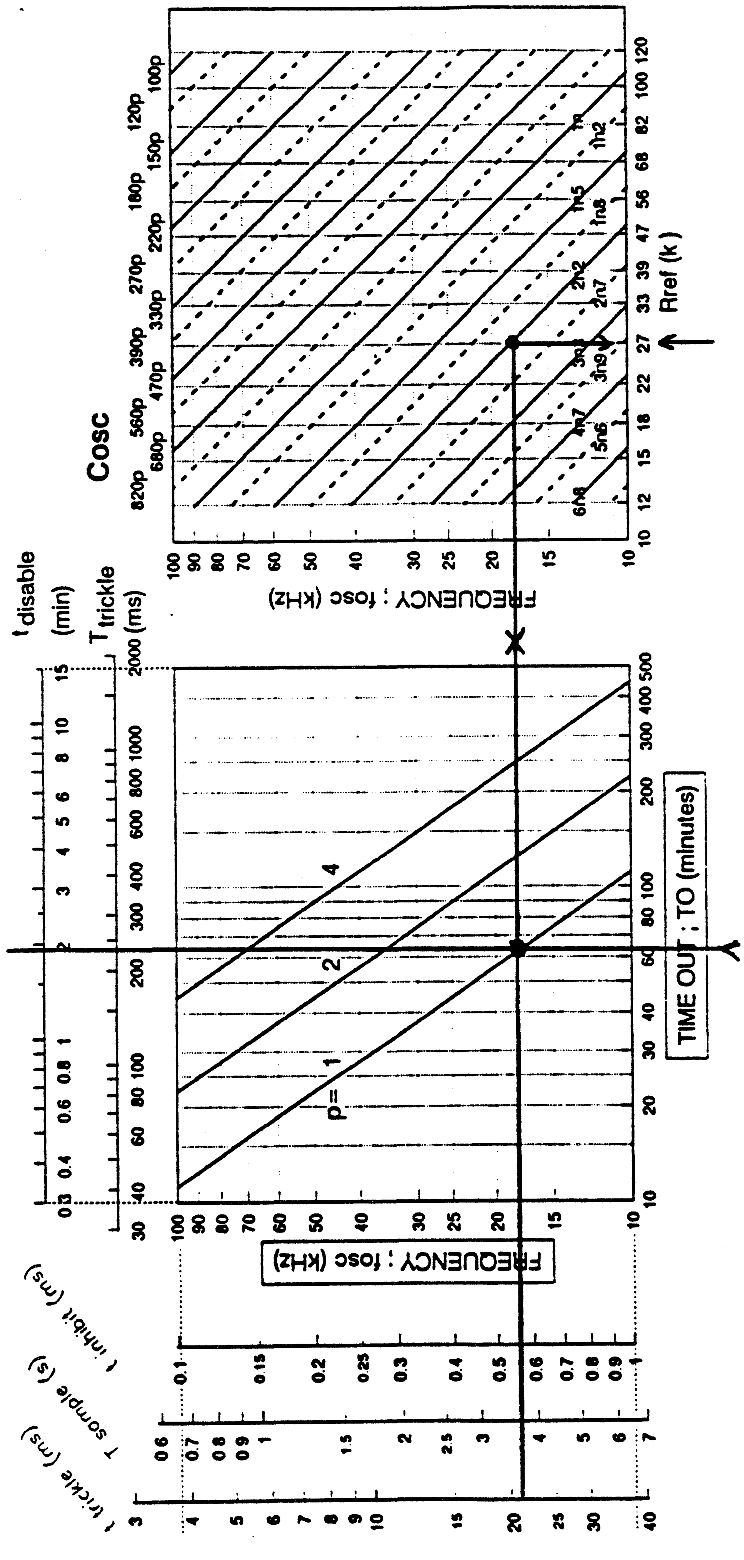




The design-in lines apply for charger example in fig. 1.

proceed to time settings nomogram to determine p factor and Cosc.

NOMOGRAM OF CHARGE CURRENT SETTING WITH TEA1100



NOMOGRAM OF TIME SETTINGS IN TEA1100

APPLICATION NOTE

ICs for Battery Management

TEA1100
Versatile battery management IC for NiCd charge systems

Report No: NPO/AN9102b

W. Bosboom

Customer & Product Support Group Consumer ICs Nijmegen, the Netherlands.

Keywords
Power Control & Misc.
Battery management
NiCd/NiMH fast charger
TEA1100

Date : 04-11-1992
Pages: 33

SUMMARY:

The TEA1100 is a battery management circuit intended for fast and very fast charging of Ni-Cd batteries. The I.C. detects full charge by means of - V and it incorporates temperature and time-out overcharge protections. The I.C. can control either a d.c. or switched mode power supply directly.
The IC is versatile in use and can be applied for a wide range of charge requirements.
This note is intented as an initial users guide to support the introduction of TEA1100.

TABLE OF CONTENTS: PAGE NUMBER

LIST OF TABLES AND FIGURES PAGE NUMBER

TABLES;

FIGURES;

1. INTRODUCTION

Rechargeable Ni-Cd batteries are undoubtedly becoming the common standard power source in portable electric and electronic appliances. Not only are Ni-Cd batteries advantageous economically and environmentally compared to throw away batteries, but they are essential in the expanding market of power equipment.
Obviously connected with rechargeable batteries is the need for charging units. As wide as the application areas for Ni-Cd batteries are; from stand-by low power use to heavy power tools, so too are the criteria for recharging methods. Similarity in requirement aspects are; reliabilty, long life time, high number of recharging cycles and of course low cost.

The complexity of the charger unit will increase with demand for shorter charge time and higher charge capacity cells. Charge rates presently vary from 0.1C normal recharge up to 4C ultra fast recharge with cell capacity ranging between 0.5 and 4 AHr.
Therefore the power needed from a fast charge supply unit will in most cases reach such a magnitude that switch mode conversion is the only solution to reach an acceptable efficiency and volume.

For charging systems, especially quick charge up to ultra fast charge, we have designed a management I.C. TEA1100, which monitors and controls charge in the Ni-Cd batteries under recharge. This I.C. offers great versatility and design freedom which, in combination with a charge supply, can ease the designers task to realise the optimum system with a minimum of engineering time.

The most important features of TEA1100 are;

- Accurate detection of fully charged batteries using -dV sensing.
- Adjustable absolute maximum fast charge period.
- Adjustable pulsating trickle charge with full batteries.
- Temperature guarding and out of range voltage protection.
- Charge current regulation control for a charge supply (e.g.SMPS).

A full listing of features can be found in table 1.

This paper gives a summary of existing charge methods with charge rate capacity as a review for the applied charge approach in TEA1100 and its charge regime.
Further on, design orientated functionallity and characteristics of TEA1100 will be explained with ideas for circuit design. Finally, a design example of a 12V input fast charge system is described.

2. CHARGING METHODS.

The nominal charging rate for batteries is 0.1C, at which it requires about 14 hours to fully recharge flat batteries. This charge rate is harmless when maintained for a far longer period as no considerable heat or pressure rise will occur in the batteries. Although safe, continuously applied normal charge will result in degradation of the battery's maximum attainable life time. Therefore very long overcharge should be limited by switch off or switch over to a lower rate only required to counteract the worst case self discharge in the batteries.

For all charging rates above normal charge it is necessary that the charge progress is controlled. The methods used vary from timing control, (the simplest but not the optimum manner with quick charge) to a combination of time, temperature and voltage control with very fast charge. When the batteries reach full charge then the charge rate must be reduced to a safe trickle charge level otherwise overheating will occur with danger of venting cells or even explosion (very fast charge).

Figure 1 shows the different charge cutoff methods.

a) Time control is a safeguard against too long an overcharge period. It is the simplest applicable method but it is static and does not take into account the state of charge of the batteries. For maximum battery life the use of this cutoff alone at rates up to quick charge (0.3C) is not optimal and is certainly unsuitable at higher charge rates.

b) Temperature rise indicates full batteries when the charge energy is no longer stored but being dissipated in the cells. It is measured by a thermistor close to the batteries. The response time of this sensing is relatively slow and also ambient temperature affects the outcome. This method can be used at charge rates between quick and fast charge (<1.5C).

c) Absolute maximum voltage control is too complex. The battery sense voltage needs a type of temperature compensation but the changing level of the charge voltage profile of the battery during its lifetime is unpredictable .
This cutoff method is not recommended.

d) Relative voltage control using the -ΔV principle has long been proven to be a very reliable cutoff method. It makes use of the characteristic phenomenom that the battery voltage drops slightly when the inside temperature increases due to the nearly fully charged condition. Compared to outside temperature control this method is direct with a fast response and therefore it is the better method with fast charge and the only reliable cutoff approach with very fast charge.
The drawback of this method has been the complexity in the realisation of a very secure and sensitive -ΔV detector.

With all the cutoff methods mentioned above it is essential that the current supply is sufficiently constant neither to interfere with the charge time nor to cause battery voltage fluctuation under sense.

3. CUTOFF CONTROL AND CHARGE REGIME WITH TEA1100

3.1 Cutoff methods

The main cutoff principle in TEA1100 is -ΔV detection. This method has been proven to be reliable and applicable over a wide range of charge rates. It is also based on design experience of such a detector and its practical use.

The detector features a currentless wide input voltage range which adapts, with or without an additional resistive divider, to any number of fixed cells or a variable range of cells.
The detection is related to the top voltage level of the batteries and therefore independent of the absolute battery voltage. Detection occurs when the battery voltage drops 1% below its maximum recorded value.

The second cutoff method incorporated as a safety guard is time out; a wide maximum charge period setting for the rare case of no -ΔV occurrence.(faulty batteries).

3.2 Protections

TEA1100 contains conditional protective charge control circuits;

For fast charge release the battery voltage must be within a certain range. In case of a too low battery voltage, reconditioning of the batteries will be attempted with trickle charge. When the voltage happens to be too high it is seen as a non connected battery which stops and resets system operation, followed by restart.

The second optional protection is under and over temperature, measured via a thermistor on the battery. This protection prevents fast charge when the battery temperature is outside a specific range which is safe for the applied battery type (10°- 45°C typical).
Outside that range the trickle charge mode will take over as a high charge rate will certainly damage the batteries.
A too high battery temperature can indicate that the cells have been under strain in use, for instance by high discharge. Directly trying to recharge excessively hot batteries will then only occur at trickle charge, allowing the batteries to cool down. When thereafter fast charge takes over, the battery temperature will drop even faster due to the endothermic effect.
If temperature protection occurs within a charge session then the timing state is set on hold while the voltage monitor is reset.

3.3 Charge system characteristics

With TEA1100 linked as the control stage to the charge current supply, the system performance obtained is shown by the state diagram in figure 2.

4. TEA1100 FUNCTIONAL BLOCKS AND USE

This section explains the functional blocks of TEA1100.
For design in, useful characteristics will be given as far as it is not detailed in the data specification. The adjustable parameters receive full attention and system approaches are indicated with simplified diagrams.

For the terminology of the I.C.s' input, output and program pins, reference can be made to the list with figure 3.

As shown in figure 3, TEA1100 consists of the following blocks;

1. Supply
2. Monitor
3. Timing
4. Protection
5. Regulation control

4.1. The supply block

This block provides bandgap stabilised voltages for all circuitry and outputs 1.25V at Rref and 4.25V at Vsource.
Nearly all adjustable features of the I.C. are related to currents which are drawn from the Rref voltage. The output current from Rref to common, via a resistor, is the basic reference. Vsource makes a reference supply for external use.
The supply input Vs is provided with an under voltage lockout with hysteresis while the maximum voltage is zener protected. The graph in figure 4 gives supply characteristics under various operating modes.
At initialisation a reset pulse is generated for all the internal logic.

4.2 The monitor block

The monitor tracks the (divided) battery voltage over a wide input voltage range, 0.38- 3.8V, and it makes the decision when the batteries are fully charged. When that occurs it will trigger the permenant switch-over from fast charge to trickle charge. Fast charge can then only resume following a reset.
From a very high impedance sense input, VAC, the monitor receives the "battery" voltage via a filter and a sample and hold circuit. The repetitively sampled sense voltage is a/d converted and the digital output value is then compared with the previously stored value in the register. The register value will and can only be replaced by a new, higher value. When subsequent sampled values decline because of battery voltage drop (fully charged), then those values are compared with the last maximum stored. If a measured value is $\geq 1\%$ below the maximum stored, then the monitor's last remaining action is to set the trickle charge mode.

False voltage sampling must be avoided. It could be caused by fluctuation or interference on the charge current, bad battery connection, shock or

vibration. The safest method of sampling is to interrupt charge during the sampling interval. This is realised by an inhibit signal on all the outputs for supply control use. Within the inhibit period the charge current should decay to zero so that at the end of this period sampling of currentless battery voltage takes place.
At the beginning of the charge session, the sample and hold will already be active. The monitor however, only starts recording after a certain time, **the disable time**, has elapsed. This period allows the battery to format or recondition.

The required clock and control pulses are directed by the timing block and the designer has to reckon with the settings of;

- disable period
- sample rate and duration
- charge supply inhibit during sampling

The time setting formulae are given in the timing block section 4.3.

The direct output of the monitor block is the **LED** pin. This pin is grounded when charge is enabled and floats when not. This signal can be used as on/off remote control for a stand alone charge supply.
The LED signal can also be mixed externally with a current reference signal from the control block to provide a 3 state control signal for the supply;

- on and high current reference ; fast charge
- on and lower current reference; trickle charge
- off

It requires an interface to adapt the reference signals to the current reference control stage of the supply.
A block configuration of this suggestion is shown in figure 5.

4.3. The timing block

This block is the operating heart of the IC. All the timing signals are extracted from a counter control stage which receives its clock signal via a prescaler from the oscillator. Therefore the timing signals are proportional to the oscillator period time, **Tosc** and the prescale factor.
The prescaler with its programmable dividing factor $p = 1, 2$ or 4 presets the ratio between fast and trickle charge rates. For a required oscillator frequency setting to be used for SMPS control, the p factor setting can be used to adjust the **maximum** fast charge time (**TO**) to an acceptable value.

In the trickle charge mode, an on/off timing signal is sent to all the output stages so that charge supply operation is pulsating. During the on and off periods the maximum IC supply current requirement characteristics are 4.1mA and 1.7mA respectively.(Fig.4).
If a charge light indicator is used, e.g. a LED, then the trickle charge repetition time should be longer than 20ms to ensure that light flashing remains visible. If necessary the trickle or LED repetition time can be scaled down by adjustment of the prescaler p factor.

A LED can be connected at the supply output or driven via the LED pin. The time settings and formulae are as follows;

- Oscillator frequency is set by;
  the current out of Rref with a resistor to common, $13k < R < 130k$,
  and the capacitor on the OSC pin by; **Tosc= 0.93 x Rref x Cosc**
- Absolute maximum fast charge period; **$t_{TO} = 2^{26} \times p \times Tosc$**
- Scale factor p is set by PR on Vsource, open or common for resp. p= 1,2 or 4.
- Monitor disable period at begin of fast charge;
  t disable= 2^{-5} x TO
- Monitor sampling repetition time; **t sample = 2^{16} x Tosc**
- Monitor sampling inhibit period; **t inhibit= 10 x Tosc**
- Monitor sampling actually takes place at the tenth oscillator cycle during t inhibit.
- trickle charge repetition time; **T trickle= 2^{12} x p x Tosc**
- Trickle charge on duration; **t trickle= 3/4 x 2^{9} x Tosc**

4.4 The protection block

The functionality of this block is already described in section 3.2, protective control circuits.

4.5 The regulation control block

This block provides additional control and consists of current control circuitry for a direct current or a switch mode power supply. These circuits are an error amp followed by both a Pulse Width Modulator and an analog amplifier, AO.
The circuits will be described in relation with the diagram in figure 6 and the waveforms in figure 7. For a current supply the flyback converter is taken as example.
Charge current is the demagnetising current of the transformer during the flyback stroke of the SMPS. Since the minus battery pole is connected to the ground of TEA1100, the sawtooth voltage pulses across the sense resistor Rs are negative going with respect to ground. The voltage pulses are averaged via the RC network, R1 and C1, and shifted positive to ground level by passing a reference current from the Ib pin through R1.
The Ib reference current can take two values during the charge cycle;

a) the current Iref, set via the Rref pin, internally activated in the fast charge mode;

$$\mathrm{Iref} = 1.25\mathrm{V} / \mathrm{Rref} \qquad (10\mu A < \mathrm{Iref} < 100\mu A)$$

b) the current In, set via the Rn pin, internally activated in the trickle charge mode;

$$\mathrm{In} = 1.25\mathrm{V} / \mathrm{Rn} \qquad (5\mu A < \mathrm{In} < 50\mu A)$$

or in default of Rn; $\mathrm{In} = 0.5 \times \mathrm{Iref}$

The error amp +input Ib refers to ground level. This amplifier has a forward transconductance characteristic which is given in figure 8 and its gain and phase response in figure 9.
In the linear operation range $g_M = 250\mu A/V$ at V_{Ib} within ± 50mV.
The output connection is the LS pin. Output saturation current is ±20µA at $V_{Ib} > \pm 150mV$ and the output voltage range is between 0.5 and 3.6V. These characteristics are important design items for correct current sensing, loop response and stabilisation.

The current sense input network must be carefully choosen.
The sense resistor Rs will be chosen as small as possible power wise, but the average voltage must be large compared to the reference tolerance; ±5mV.
The R1,C1 time constant should not be too long (response time) but sufficient to ensure that the residual ripple on the averaged sense voltage is < ±100mV in worst case. Only then will the average charge current be correctly represented at the error amp output through lineair operation.

The fast charge current is set by;
$$\mathrm{Ich} = \frac{\mathrm{Iref} \times \mathrm{R1}}{\mathrm{Rs}}$$

The average trickle charge current is set by;
$$\mathrm{Ich} = \frac{\mathrm{In} \times \mathrm{R1}}{\mathrm{Rs}} \times \frac{0.094}{p}$$
(where the last term states the charge current duty factor)

On the LS output a capacitor, possibly with a series resistor, must be connected to ground to obtain the required feedback error voltage. *)
The output load network adds compensation to the small signal loop. In table 2, given with figure 9, compensation circuits with resulting gain response are shown and an example is plotted in figure 9.

The error amp output contains a series switch which is open during both the monitor inhibit period and the trickle charge off period. In these periods the charge current interrupts causing the error amp output to go high. The feedback regulation voltage, present before those interrupts, remains stored by the LS capacitor due to the open switch. When charging resumes, the current will quickly stabilise at its regulation level with

---

*) Note: A resistor on LS to ground, e.g. an 1MΩ probe, causes an input offset dependent on the output voltage according to:
$\mathrm{Vin} = \mathrm{Vout} / 250\mu A \times \mathrm{Rload}$. So at Vout= 2V, 1MΩ gives 8mV offset.

minimum current overshoot.
At initialisation or after a protection occurence a slow start is also achieved by the LS capacitor which at protection is discharged via the LS pin to ground. Figure 10 shows the start cycle.

The error amp output can be used via a buffer, the analog output AO.
The voltage gain of this amplifier is 4x (12dB) with a gain bandwidth of 4MHz. The transfer characteristics are shown in figure 11.
By means of current programming at the CP pin the amplifier can be set either inverting or non inverting with respectively a sink or source output.

AO; non invert; CP pin left open
invert; CP pin via R=56kΩ to ground

The AO output can be used to control a dc supply or a stand alone SMPS where an optocoupler is used for mains isolation .

For duty cycle regulation the error amp output is passed to the pulse width modulator PWM. The error voltage is the slicing level which is compared with the oscillator waveform to produce rectangular pulses as shown in figure 12.
Via a latch, only one pulse per cycle starting in the oscillator flyback is enabled as output (anti-multiple pulsing).

The error voltage can be modulated with the supply current waveform. See figures 13 and 14. This type of current control or feed forward adds cycle by cycle regulation to the relatively slow loop regulation. This enhances the response to sudden changes in supply operation with minimum current overshoots. Modulation can be achieved via the LS capacitor which instead of being grounded is connected to a negative going sense voltage of the duty switching current (flyback and forward SMPS).
It can also be achieved via the error amp by means of a resistor R_{LS} in series with the LS capacitor (forward SMPS only). By this manner the sense voltage at the error amp input should not be filtered while the ac ripple voltage must be well within the linear input voltage range. The added modulation voltage on LS is 250μA x R_{LS} x Vi, and it reflects the ac current waveform and partly an error in the dc supply current. Together with a relatively small C_{LS} fast tracking of the dc loop is realised.

Similar to the analog output the polarity of the PWM output pulses can be inverted by current programming at the CP pin. In either mode the output can sink or source current.

In the trickle charge off period auxillary output pulses can be produced, programmed by the CP pin . These pulses have a 14% duty factor and a repetition rate of Tosc x 8. They are intended to support primary and secondary supply recovery in an isolated supply.

non invert; CP pin via R=120kΩ to ground
invert; CP pin via R= 22kΩ to ground

The PWM output can be used to control the drive of a SMPS switching device directly or a mains isolated supply via a pulse transformer (master-slave principle).

5. EVALUATION BOARD OF A DC/DC SMPS CHARGER

An experimental design of a 40 minute charger is shown in figure 15. The circuit applies TEA1100 in combination with a low voltage forward converter switching at 40kHz. This system will charge a maximum of four, fast charge type, AA cells (550mAHr) from a dc input voltage between 11 and 16V. The fast charge rate is 2C while the trickle charge rate is 0.05C by means of 550mA current bursts.
The board layout is shown in figure 16.

5.1 Design specification

Fast charge current: 1.1A.	Abs. max. charge time: 45- 60min.
Trickle charge current: ≈28mA.	Trickle repetition time: ≥ 80ms.
Output overvoltage protection: < 9V.	
Max. cell voltage under charge: 1.7V.	

5.2 Circuit description

The main parts of the forward converter are the swiching device T1, the flywheel diode D2 and the choke L1 in series with the battery load. The switch is driven by transistor T2 which receives duty cycle controlled pulses from the control circuit TEA1100. The supply is current regulated and TEA1100 senses the current at resistor R4.
The battery voltage is continuously monitored via divider R14,R15 and temperature guarded via divider R11,R_{NTC}. The charge mode is indicated by a LED, D5, which is on continuously during fast charge and flashes during trickle charge.
The network with T3 adapts the supply input to the I.C. supply.
The residual components around the I.C. program the timing control, the charge current and its regulation.

5.3. SMPS operation and coil design

5.3.1. Operating limits

The forward converter regulation equation is; $Vo' = \delta\ Vi'$

where: Vo' = the voltage across L during the off time of T1 and hence
$Vo' = Vbattery + V_F(d2) + V_{R4}$.

Vi' = the supply voltage at L during the on time of T1 and hence
$Vi' = Vs - V_F(D1) - Vsat(T1) - V_{R4}$.
δ = duty factor ton/T; TEA1100 δ cycle range= 0 to 80%.

The limits in supply operation are;

a) maximum Vi' and minimum Vo' e.g. battery short (trickle charge).
At this condition the δ cycle is minimal and is 6%.

b) minimum Vi' and maximum Vo' when the battery is nearly fully charged
The maximum δ cycle restricts the last condition so if taken;
maximum battery voltage at end of charge= 4x1.7V= 6.8V,
V_F of the diodes D1,D2 ≤0.8V , Vsat of T1 ≤0.4V and VR4= 0.28V then the min. input voltage Vs= ≤10.9V.

5.3.2. Determination of the coil

The maximum inductance is limited by the requirement that the maximum inductance current must decay to zero within 9 period cycles (monitor sense inhibit period). The worst case condition is at minimum Vo' which is assumed to be 4 x 1V= 4V; the reconditioned low battery voltage.
By approximation;

$$Lmax= 9\ Tosc\ (Vo + V_F)\ /\ Io\ :$$
$$Lmax= 9x25x10^{-6}x4.8\ /1.1 \approx 980\mu H.$$

The minimum inductance is limited by the condition of continuous coil current flow. The worst case situation is the maximum ripple current at the lower output current setting (i.e. trickle charge). Maximum ripple ΔI_L equals 2xIaverage (=1.1A) and this occurs at Vo_{max} with 50% δ cycle.

So according to;

$$Lmin= Vo'max(1-\delta)T\ /\ 2Iav\ :$$
$$Lmin= 7.6x0.5x25x10^{-6}/\ 2x0.55 \approx 87\mu H.$$

A compromise has to be chosen between the switching peak current I_{Lmax}, the core size determining I^2L value and the current deviation rate with respect to loop response.

The chosen inductance value is: L= 400μH.

and now derived; ΔI_L max = 238mA
I_L peak= 1.1 + 0.238/2≈ 1.22A

The abs. max. peak current is taken on the safe side: 1.25xI_L peak ≈1.5A.

- For the core a RM type is selected by;

  Abs max energy: $I^2L = 1.5^2x400x10^{-6} \approx 0.9mJ$
  Derived from core data the core choice is;
  RM7/i, 3C85 grade , airgap= 0.4mm

- The winding number of turns and wire size are;

  with a 0.4mm airgap the inductance per turn is; $A_L \approx 160nH$
  so the winding number of turns becomes; $N = (L/\ AL)^{1/2} \approx 50$

  and the wire size required is AWG 27 (≈0.35mm).

5.4. Control I.C. TEA1100 and its associated components

The peripheral components for TEA1100 will be specified by function and design order.
For settings and formulae see also sections 4.3 and 4.5.

Pin 10; Ireference setting via R13. Chosen current is 46μA so; R13=1.25V/46μA= 27kΩ. This current equals the reference current for fast charge regulation.

Pin 11; I_N reference current for trickle charge regulation. The default value 0.5xIref is chosen so this pin can be left open.

Pin 13; Oscillator. The frequency is set by C7 and the reference resistor at pin 10. For Tosc= 25μs; C7= 25μs/ 0.93x27k ≈ 1nF.

Pin 8; p scaling factor. By leaving pin 8 open, p is set for a factor 2 to obtain an abs. max. charge time $_{TO}$ of 2^{26}x2x25μs ≈ 56minutes which is required.
(With pin 8 connected to pin 6, p= 1 and TO will be 28minutes and with pin 8 to ground then p= 4 and TO becomes 112minutes).

All the other timings can now be derived;
Trickle charge: T trickle= 0.2s
t trickle= 10ms ; hence: δ factor= 5%.
Monitor part: t disable≈ 1.7min
T inhibit≈ 1.6s
t inhibit= 250μs

Pin 7; Monitor input. The input voltage is adapted to the battery voltage by divider R14,R15. A divider factor 0.5 is taken which results in battery overvoltage and max. - V detecting voltage levels of 8.5V and 7.7V respectivily.

Pin 3; Temperature protection input.
From a constant bias voltage source the divider R11, R_{NTC} adapts the batteries safe temperature range with the input voltage window of pin 3.

In appendix 3 a calculation procedure is given for the determination of the NTC network.

If temperature sensing is not used then R11 is omitted and pin 3 should be connected to pin 11.

Pin 5; Current sense input. The resistors R4, R8 and the reference currents I_F and I_N determine the charge current regulation levels. For a fast charge current= 1.1A, I_F= 46μA and R4= 0.25Ω then;

$$R8= 1.1x0.25/ 46x10^{-6} \approx 6k; \quad 6k2 \text{ taken.}$$

With the given I_L max.= 0.23A the maximum ripple voltage at pin 5 becomes: 0.23xR4= 58mV.

C5 is small and only present for hf noise filtering.

Pin 4; Loop stability. The functions achieved via this pin are; error control voltage retrieval and frequency bandwidth setting with C6 and error voltage modulation with inductance current information via R9.
C6= 680pF. This value is as small as possible. Its residual ripple p-p voltage, 90°phase lag with the input voltage, is at most 70mV.

The ac modulation voltage across R9= 47k becomes maximum;

$$V_{R9} = 250\mu \times V_{in} \times R9 = 0.68Vp\text{-}p.$$

Pin 9; Change polarity. The polarity of the duty cyle output pulses at pin 1 have to be positive. This is set by leaving pin 9 open. (A reversed polarity is set with a 56k resistor at pin 9 to ground).

Pin 15; LED output. This pin is used to switch on a LED when charge is supplied. The trickle charge mode is indicated by a flashing LED.

Pin 12; IC supply. The supply is derived from the input voltage via a voltage clamp circuit with T3 and D4. The low impedance of this circuit together with C4 prevents IC supply on/off oscillation within the low supply cutoff hysteresis band at initialisation.

Pins not used;

Pin 2 ; Analog output.
Pin 14; Synch.

6. SUBSEQUENT APPLICATION EXAMPLES

In appendix 4 some circuit design ideas are given for application of power Mosfets as SMPS switching devices.
Also given is a dc regulated charge source in case a switch mode power source is not acceptable because of interference. This dc charger with TEA1100 is the simplest of its kind.

Application notes which can be expected soon will give description of battery management systems which apply the flyback converter topology.
The system examples are;

a) a charger with low supply voltage input, 12V, capable of charging 6 AA type cells in one hour. The output can easily be adapted for other requirements.
b) a charger with world wide mains input, 90 - 265V, providing mains isolated output. It can charge a battery pack containing 8 cells with 1.5AHr capacity within one hour.

TABLE 1 TEA1100 FEATURES

TEA1100 Battery Management Systems

Application

- Monitor and Control circuit for NiCd charger systems

Features

- Accurate detection of fully charged batteries using -ΔV sensing, with 12.5 bit resolution, digital filtering, currentless voltage sensing, and nominal -1% (w.r.t. to top level) -ΔV detection level.
- Adjustable Time-Out of fast charging, -ΔV detection inhibited for the first 3.1% of Time-Out.
- Tracking of Time-Out with charge rate.
- Switch over to trickle charge following either -ΔV detection or Time-Out.
- Pulsating trickle charge current with 10, 5 or 2.5% duty factor setting to avoid cell degradation due to dendritic grows.
- Detection of low battery voltage, during which time trickle charge is applied, and timing reset.
- Protection against open circuit batteries, during which time outputs are switched off, and timing reset.
- Charge current regulation in conjunction with a SMPS via Analog and PWM outputs.
- Temperature guarding possibility by means of NTC resistor. Outside set temperature range, trickle charge takes place. Hysteresis is built in to avoid rapid on/off cycling.
- Independently adjustable ratio between fast and trickle charge rates, ratio range 10 to >500, set by two external resistors and duty factor selection.

  Default trickle rate=1/20 fast rate (d=10%) if only one resistor is used.
- Large 10:1 voltage range of operation of -ΔV circuit.
- 16 pin DIL package.

Fig.1a TIMER CUTOFF

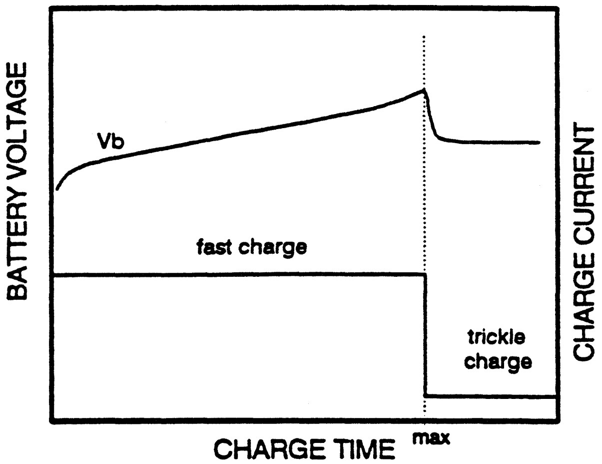


Fig.1b TEMPERATURE CUTOFF

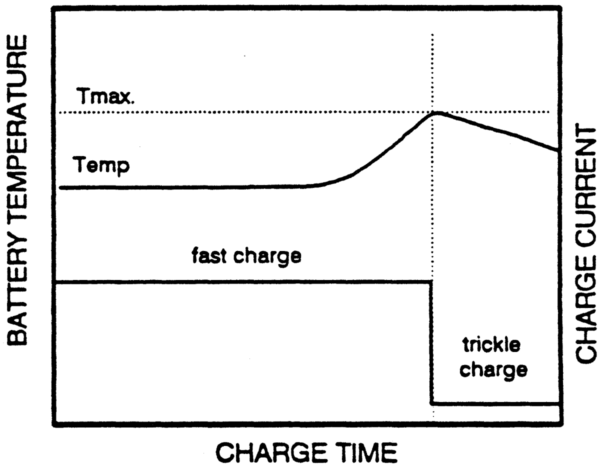


Fig.1c MINUS DELTA VOLTAGE CUTOFF

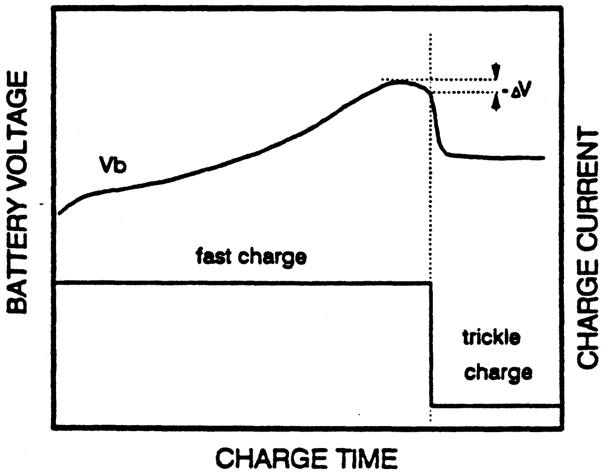
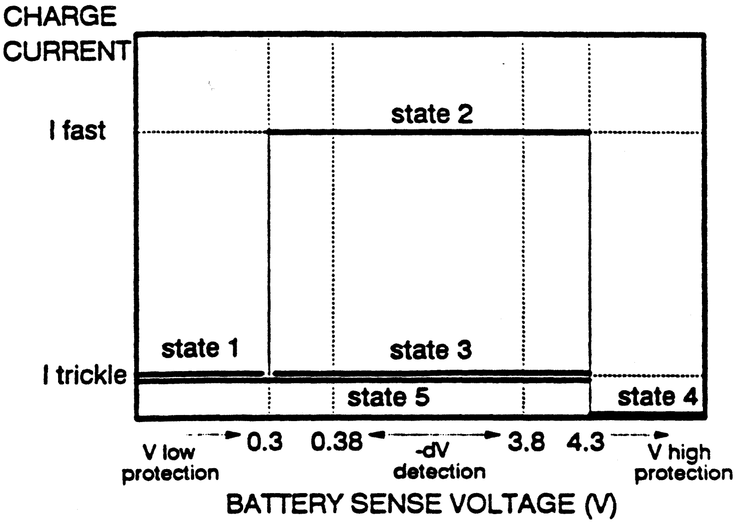


FIG.2 CHARGE STATE DIAGRAM



Operational states of charger system

state 1: Trickle charge with flat battery
state 2: Fast charge
state 3: Trickle charge, battery full detected
state 4: Disabled charge, open circuit shutdown
state 5: Trickle charge with out of range battery temperature

Fig.4 TEA1100 SUPPLY CHARACTERISTIC

(excluding IC source load currents)

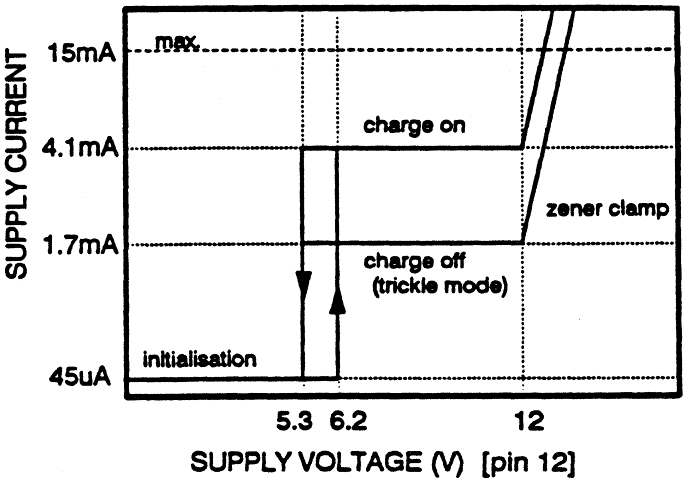
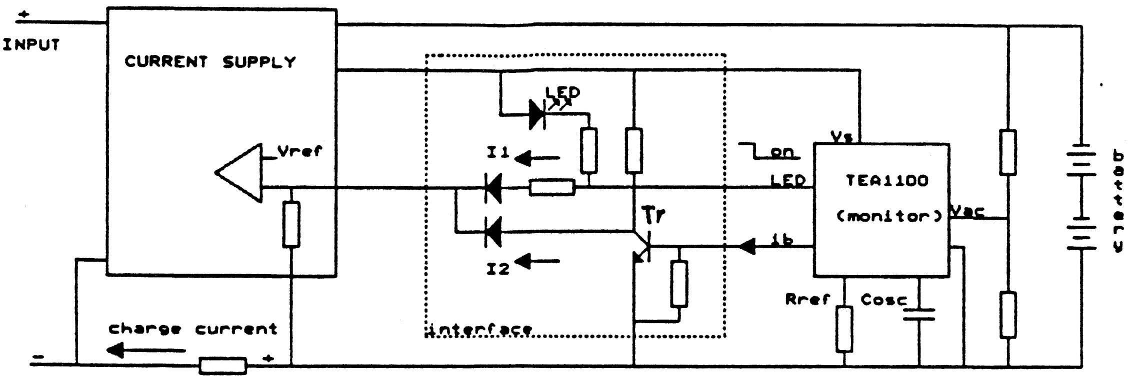


FIG.5 STAND ALONE CHARGER WITH TEA1100



This charger example shows TEA1100 as monitor only, which remotely controls the charge output current from a stand alone SMPS.

The SMPS regulates the charge current by refering to a positive going current sense voltage. Via an interface TEA1100 controls the release of charge current and reduces the trickle charge current magnitude.

This 3 state charge control is realised as follows;

Fast charge: LED output low, ib= i_{Rref} switching Tr on.

Both i1 and i2 currents are zero therefore not affecting the SMPS current sense voltage and the charge is maximum.

Trickle charge burst period:

LED output low, ib= iRref/2 switching Tr off.

i1 current is zero but i2 flows which causes a voltage step on top of the current reference voltage so the charge current magnitude is reduced.

Charge-off period (trickle charge and protection modes):

LED output high, ib= zero.

Both i1 and i2 currents flow. The sum of these currents causes a voltage step which exceeds the comparators' Vref so charge current is disabled.

FIG.6 CHARGE CURRENT REGULATION

Flyback regulator example

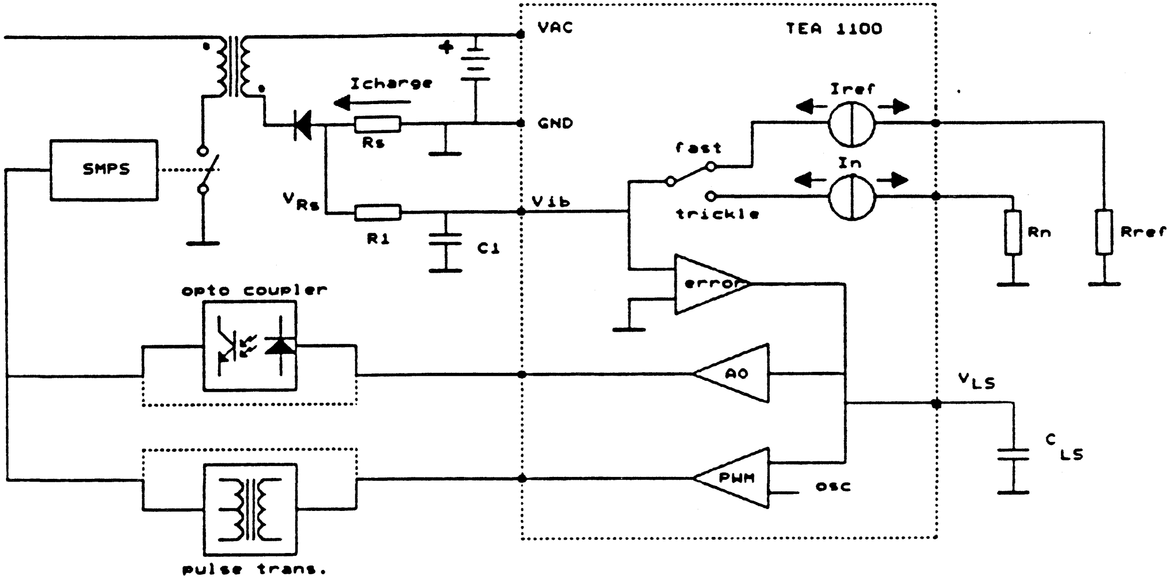


Fig.7 Waveforms related to fig.6

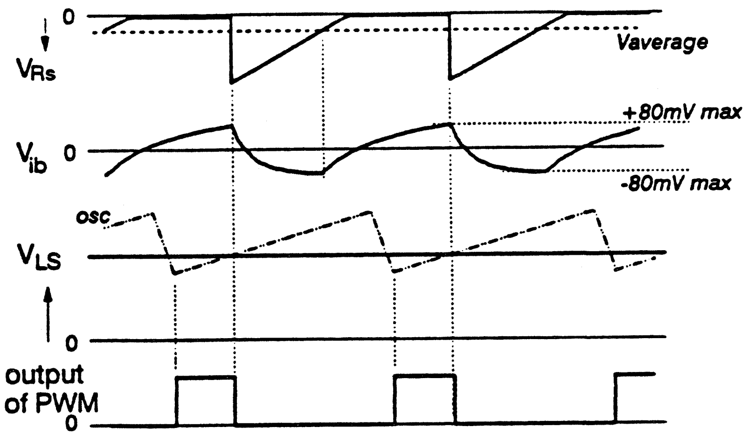


Fig.8 TRANSCONDUCTANCE ERROR AMP

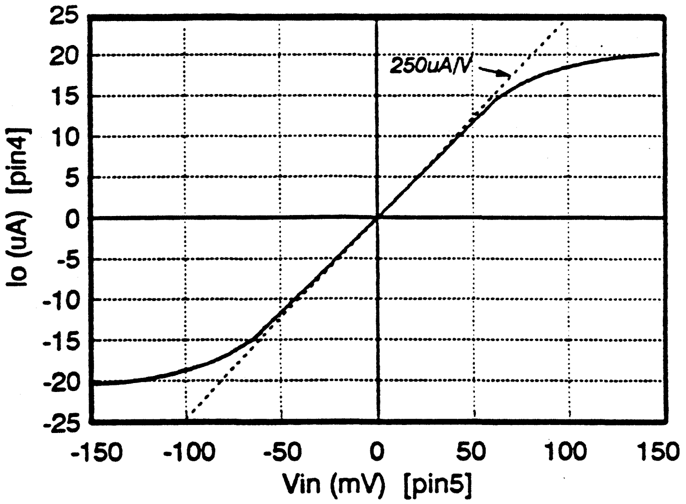


Fig.9 GAIN/PHASE RESPONSE ERROR AMP

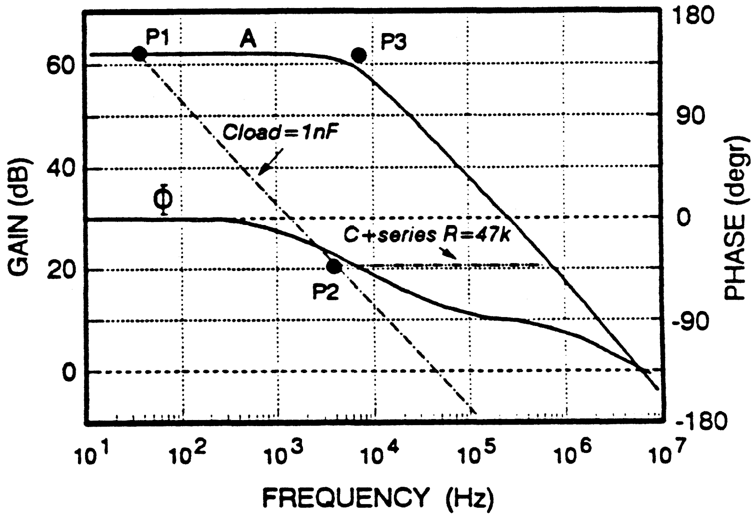


TABLE 2 LOOP COMPENSATION NETWORKS AND RESPONSE

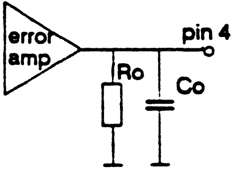
	ac small signal model	with load at LS, pin 4	
g_M=250µA/V Ro =4.5MΩ Co =4.7pF		 C>>Co	 C>>Co R<<Ro
gain	$A_{(f)} = g_M \frac{Ro}{1+sRoCo}$	$A_{(f)} = g_M \frac{Ro}{1+sRoC}$	$A_{(f)} = g_M (R+\frac{1}{sC})$
poles	$f_3 = \frac{1}{2\pi RoCo} = 8kHz$	$f_1 = \frac{1}{2\pi RoC}$	$f_1 = \frac{1}{2\pi RoC}$
zeros	-	-	$f_2 = \frac{1}{2\pi RC}$

FIG.10 PWM CONTROL AT START-UP

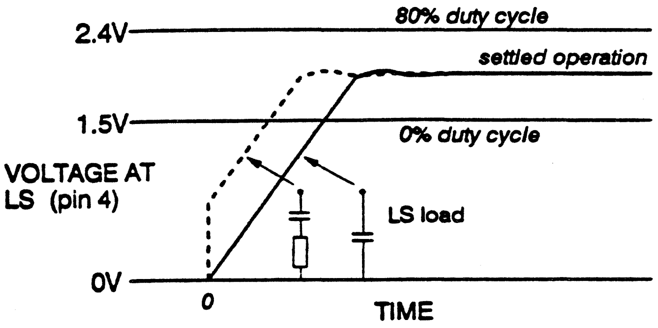
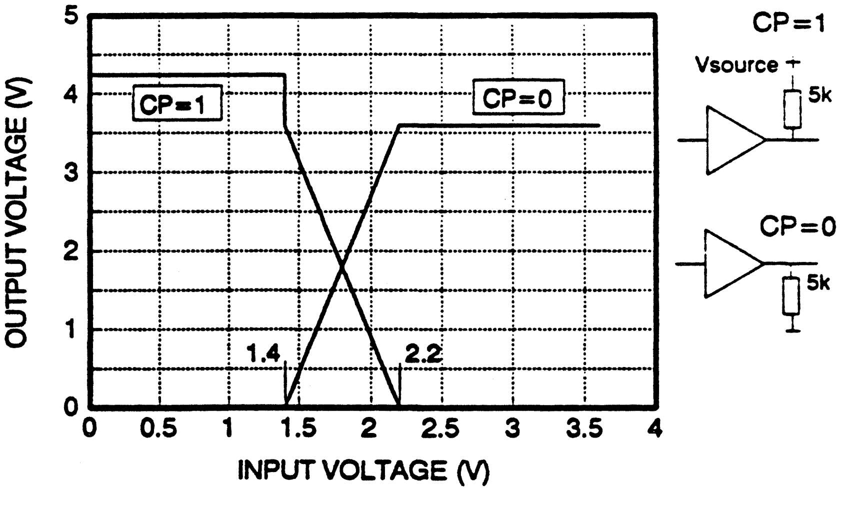


FIG.11 ANALOG AMP TRANSFER CHARACTERISTICS



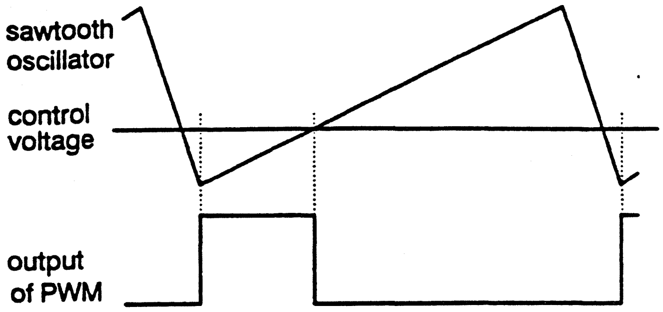


FIG.12 PULSE WIDTH MODULATION

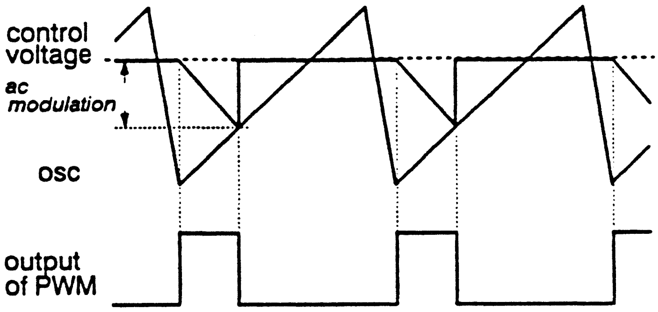


FIG.13 FEED FORWARD CONTROL (flyback SMPS)

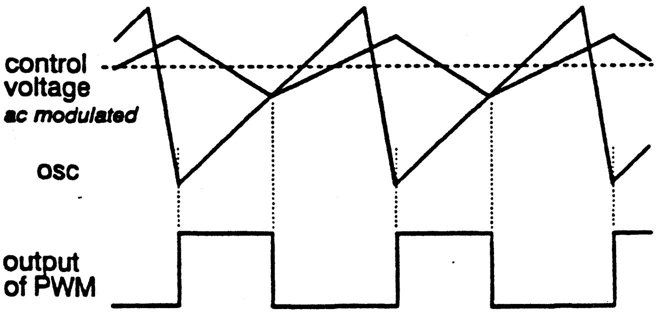


FIG.14 FEED FORWARD CONTROL (forward SMPS)

OPERATIONAL CONDITIONS;

Vin = 10- 16.5V
Vout= 0- 6.8V

Fast charge current = 1.1A
Trickle charge current= 28mA

Time out = 56min.
Frequency= 40kHz.

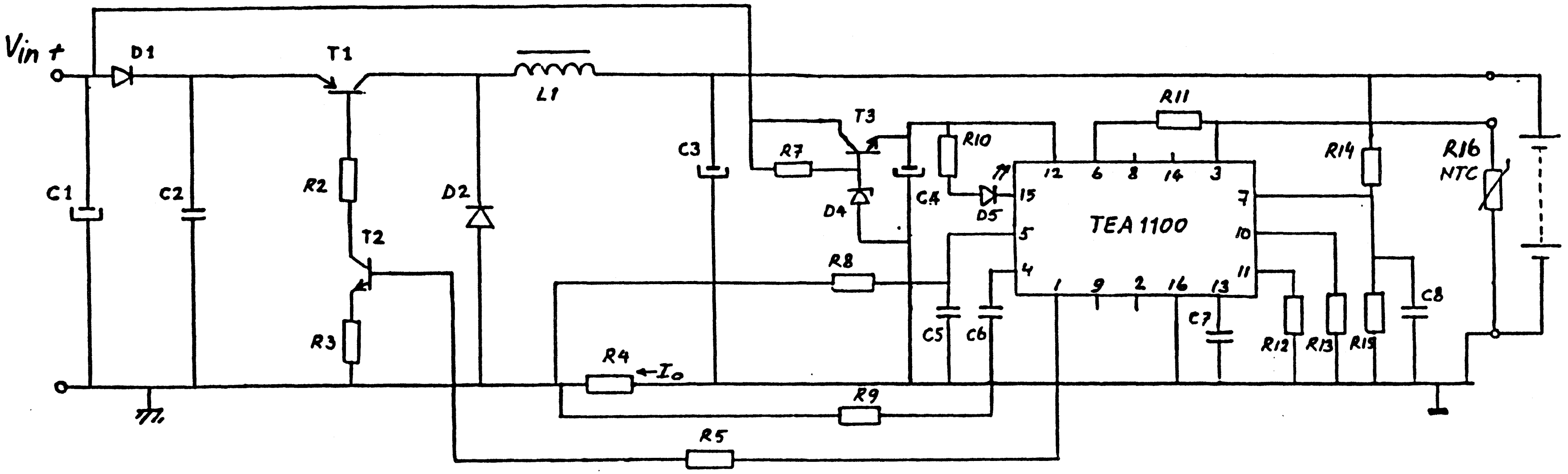


FIG.15 CIRCUIT DIAGRAM OF 12V INPUT, DC/DC CHARGER

TABLE 3. PARTS LIST OF DC/DC CHARGER

Resistors

R1- 330Ω	SFR16T	2322 180 73331
R2- 110Ω	SFR16T	2322 180 73111
R5,R10- 680Ω	SFR16T	2322 180 73629
R4- 4x 1Ω parallel	MRS25 1%	2322 156 11008
R7- 5k1	SFR16T	2322 180 73512
R8- 6k19	MRS25 1%	2322 156 16192
R9- 47k	SFR16T	2322 180 73473
R11- t.b.f.	SFR16T	
R12 not applied		
R13- 27k4	MRS25 1%	2322 156 12743
R14,R15- 27k	SFR16T	2322 180 73273

R16- NTC. If NTC not used then omit R11 and connect pin 3 to pin 11.

Capacitors

C1- 47µF/25V		2222 030 26479
C2- 47nF	MKT 10%	2222 370 11473
C3- 10µF/16V		2222 134 55109
C4- 22µF/16V		2222 134 65229
C5- 82pF	CERAMIC	2222 652 58829
C6,C8- 680pF	CERAMIC	2222 630 08681
C7- 1nF	KP 1%	2222 426 41002

Semiconductors

T1- BD227	9331 603 50127
T2- BC337	9331 492 00000
T3- BC548B	9331 977 60112
D1,D2- BYD73A	9337 537 40153
D4- BZX79 C9V1	9331 177 80000
D5- LED	9338 501 20112

Choke

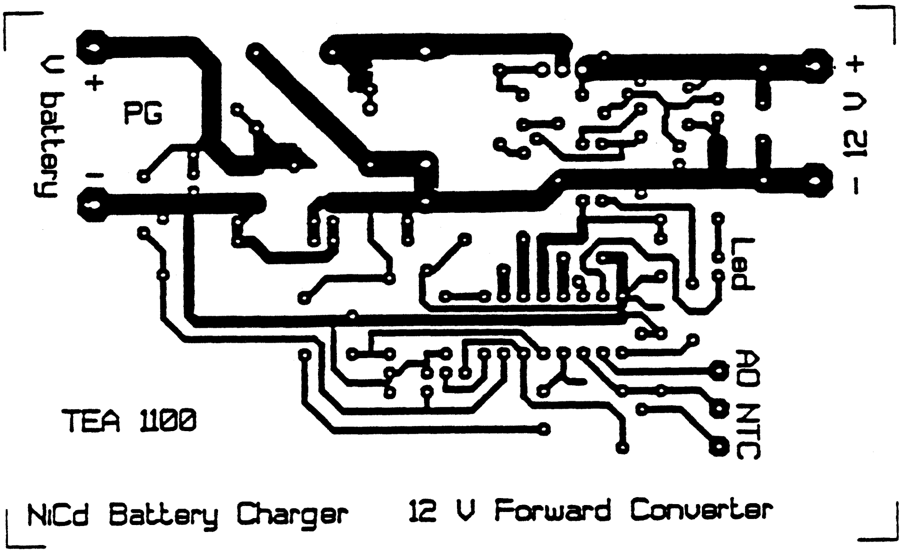
L- 400µH;	
N=50, 0.35mm wire (AWG27)	
RM7/i core set (Al=160)	4322 025 00451
Coil former	4322 021 34611
Clips 2x	4313 021 03951

Miscellaneous

IC 16pins socket	2422 549 13266
Heatsink	2422 487 89039
Pins 4x	

FIG.17 PCB LAYOUT OF THE DC/DC CHARGER

A) PCB COPPER SIDE



B) PCB COMPONENTS SIDE

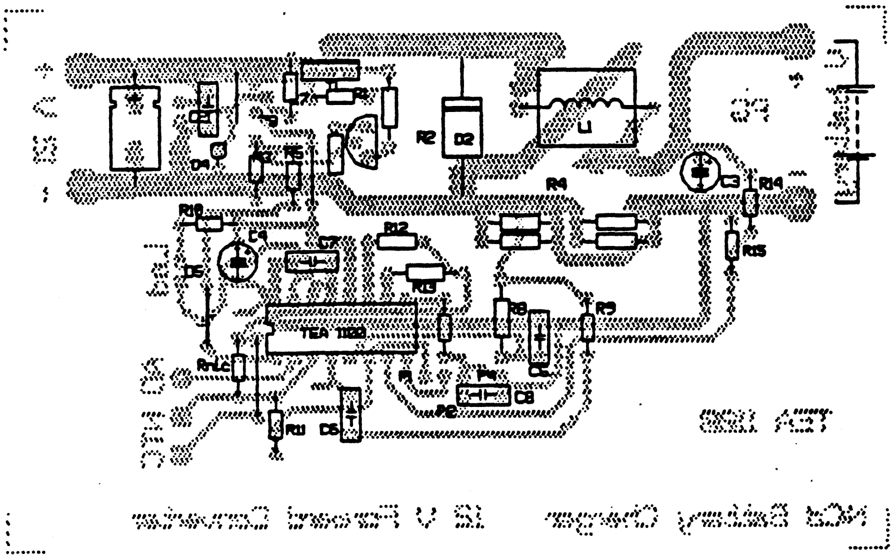
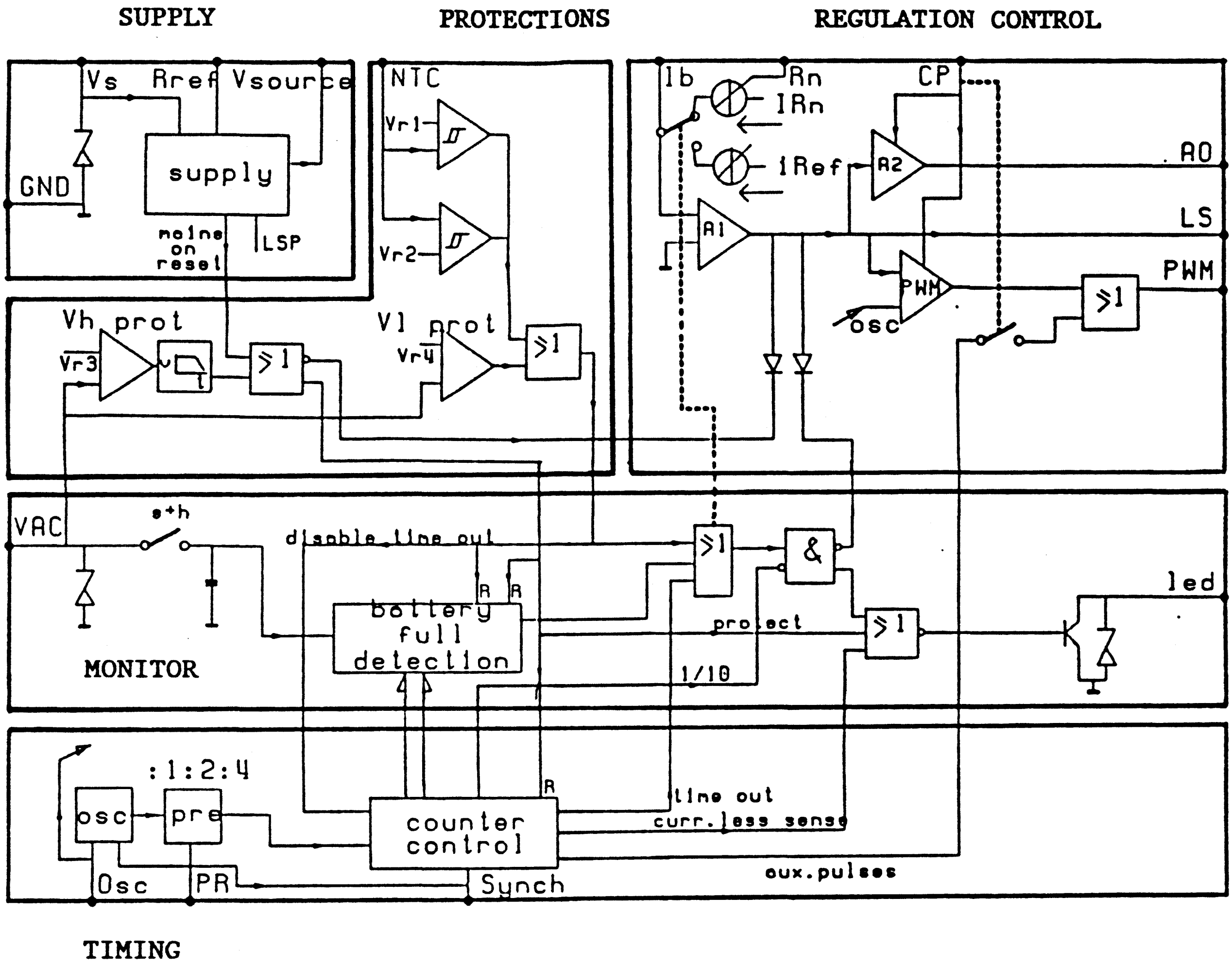
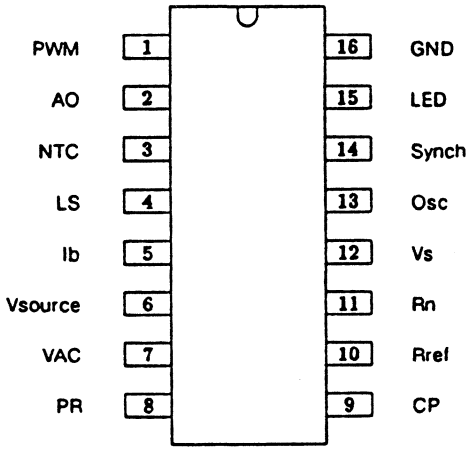
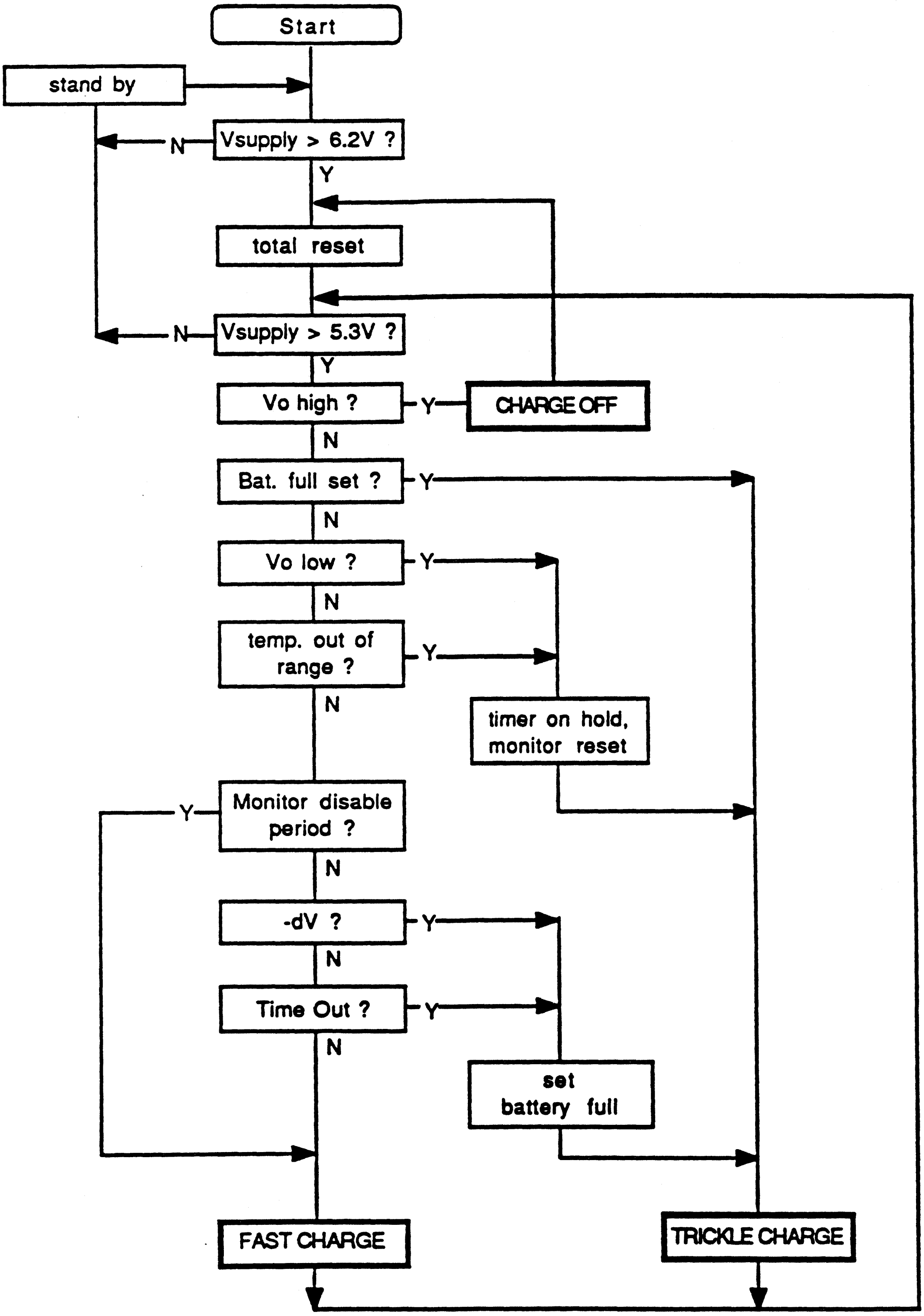


FIG.3 TEA1100 INTERNAL BLOCK DIAGRAM AND PIN CONFIGURATION





Pin number	Function	Description
1	PWM	Pulse Width Modulator
2	AO	Analog Output
3	NTC	Temp sense input
4	LS	Loop Stability
5	Ib	Charge current
6	Vsource	Stabilized supply voltage
7	VAC	Battery voltage
8	PR	Prescaler
9	CP	Change Polarity
10	Rref	Reference resistor
11	Rn	Normal charge ref resistor
12	Vs	Supply voltage
13	Osc	Oscillator
14	Synch	Synchronization
15	LED	LED output
16	GND	Ground



TEA1100 charger flow diagram

APPENDIX 2

APPLICATION NOTE

Method to increase -dV sensitivity for TEA1100 charge system.

The basic design of a high performance battery charge system with TEA1100 features battery full detection by -dV sensing; -1% of the battery top voltage level. This corresponds to a -dV of approximately 16mV per cell. The performance and sensitivity of the -dV detection is broadly accepted as well suited to the task.
In some applications (NiMH or "heavy-duty" NiCd) however an increased sensitivity of -dV detection to -0.5% is required. Although the -1% threshold in TEA1100 is fixed, it is possible to have detection at -0.5% by addition of a zener diode in the battery voltage sense line.
This method is outlined as follows;

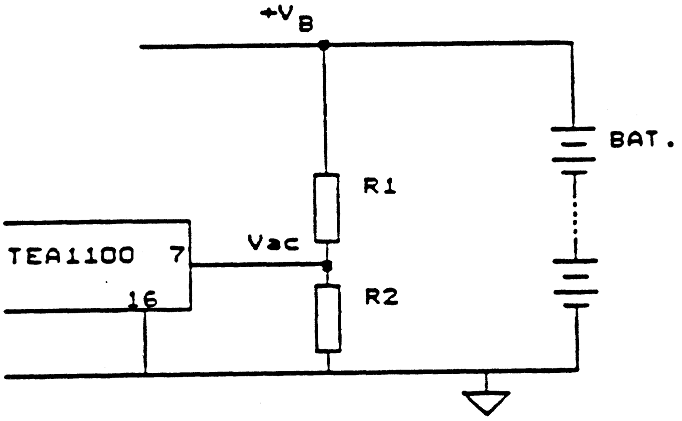


Figure 1

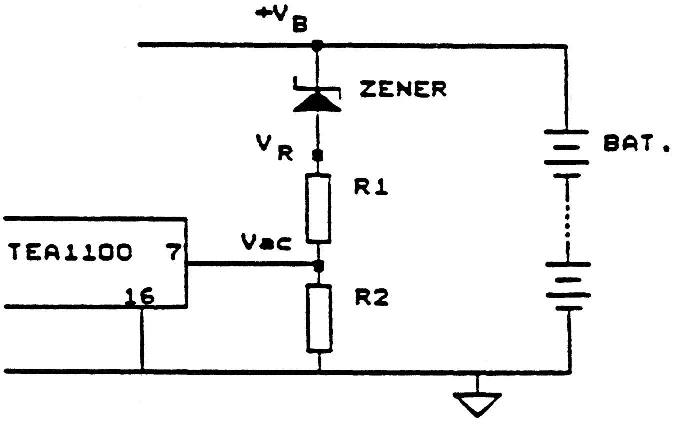


Figure 2

Figure 1 shows the basic, direct battery sensing via a resistive divider which adapts the battery voltage to within the Vac range. Detection occurs at -dV= 1% of V_B max.

Figure 2 shows the position of the zener diode. TEA1100 now senses the voltage V_R which is the battery voltage minus the zener voltage, V_B - V_R.
Detection occurs at -dV_R= 1% of V_R max. This detection corresponds with a -dV in the battery according to:

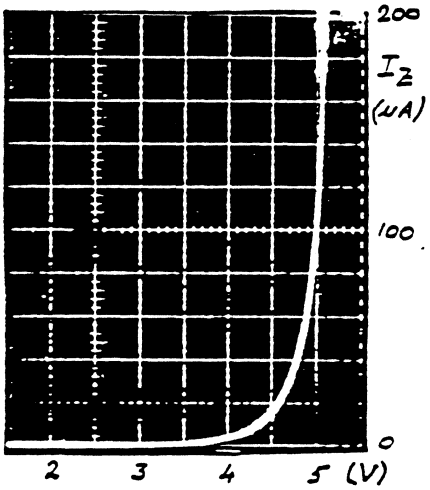
$$. \, dV_B = \frac{V_Z}{V_B \max} - 1 = \quad (\%).$$

If the zener voltage is half the maximum battery voltage then dV_B detection will be at -0.5%.

Choice of zener type.

The current flowing in the V_{AC} sense network should be small to prevent unwanted high discharge of the battery when the charge supply is switched off. A rule of thumb is that the discharge allowed falls in the range of the battery self discharge rate which is approximate -25% per month. Hence, the equivalent average discharge current for different battery capacity is than;

Battery capacity	0.5	1	1.5	(AHr)
Self discharge current	180	360	540	(µA)



At these low currents common zeners show a soft knee characteristic which is not applicable. Instead good results are obtained with the Philips low voltage avalanche diode type, the PLVA400A series, specially intended for low current use with tight tolerance. The available voltage range is 5 to 6.8V.

Figure 3 shows the zener characteristic of the 5V diode, PLVA450A.

Design example for a six cells' battery and 0.5% -dV cutoff.

Conditions:
Maximum battery voltage (1.7V/cell) = 10.2V.
Sense network current ≈ 300µA.
Maximum monitor sense voltage V_{AC} = 3.6V. (<3.85V).

For -dV ≈ 0.5% a zener voltage of about half the battery voltage is required; take V_Z = 5V. Now V_R top level is ≈ 5.2V and the divider factor, V_{AC}/V_R, needed is 0.69. R1 and R2 become respectively 5.6kΩ and 12kΩ.

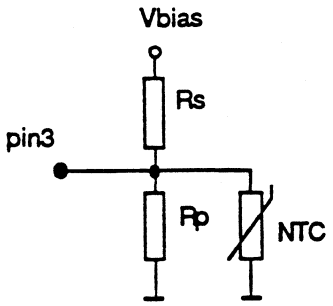
During charge the battery voltage rises with a minimum rate of about 8mV per minute for 6 cells under 1C charge. The temp. coefficient of the zener and its ambient temperature change should not cause a premature full detection.
With the following equation the allowance of zener temperature change can be derived;

$$\frac{K}{\min} < \frac{0.5\%(V_B - V_Z) + 8\text{mV/min}}{S_Z}$$

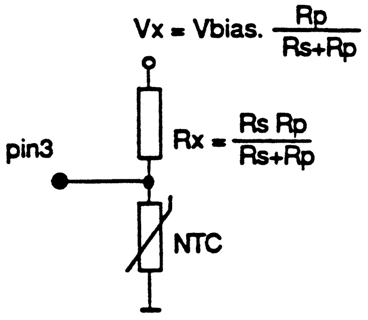
At the early stage of charge V_B is ≈ 8.4V for 6 cells, 0.5% is a safe figure for 1% detection in TEA1100 and V_Z = 5V.
If assumed a S_Z = +0.5mV/K (typ. 0.2mV/K for PLVA450A) then the outcome is 50°K/min maximum.

Appendix 3

Battery NTC thermistor adaption to TEA1100 temperature sense input.



Temperature sense network



Equivalent circuit

For the low and high temperature levels the corresponding R_h and R_l thermistor resistance values needs to be derived

The following calculations is a way to get the values of the required battery serie-, parallel resistor and the bias voltage;

low temperature limit , NTC resistance :	R_h	ratio; $p = \frac{R_h}{R_l}$
high temperature limit, NTC resistance :	R_l	
TEA1100 upper threshold lower threshold	$Vh = 3V$ $Vl = 0.81V$	range; $\Delta V_t = 2.19V$

If $k = \frac{\Delta V_t}{(p \cdot V_l) - V_h}$ then $Rx = k \cdot R_h$; $Vx = V_h\,(1+k)$

V_B = constant bias voltage ; $Rs(eries) = \frac{k}{3(1+k)} \cdot V_B \cdot R_h$

$(V_B \geq Vx)$ $Rp(arallel) = \frac{k}{V_B - 3(1+k)} \cdot V_B \cdot R_h$

When the B value of the NTC is used then ;

$$p = e^{B \cdot \left(\frac{1}{Tl} - \frac{1}{Th}\right)}$$ T is in Kelvin.

$$R_h = R_{25} \cdot e^{B \cdot \left(\frac{1}{Tl} - \frac{1}{298}\right)}$$

Note: Without temperature sensing pin 3 can directly be connected to pin 11.

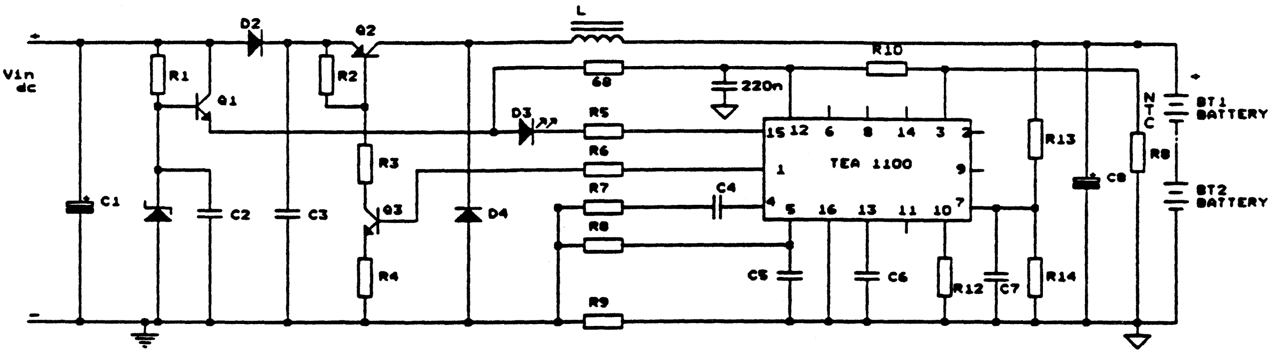


FIG. IV-a. CHARGER DIAGRAM WITH BIPOLAR SWITCH IN THE BUCK REGULATO

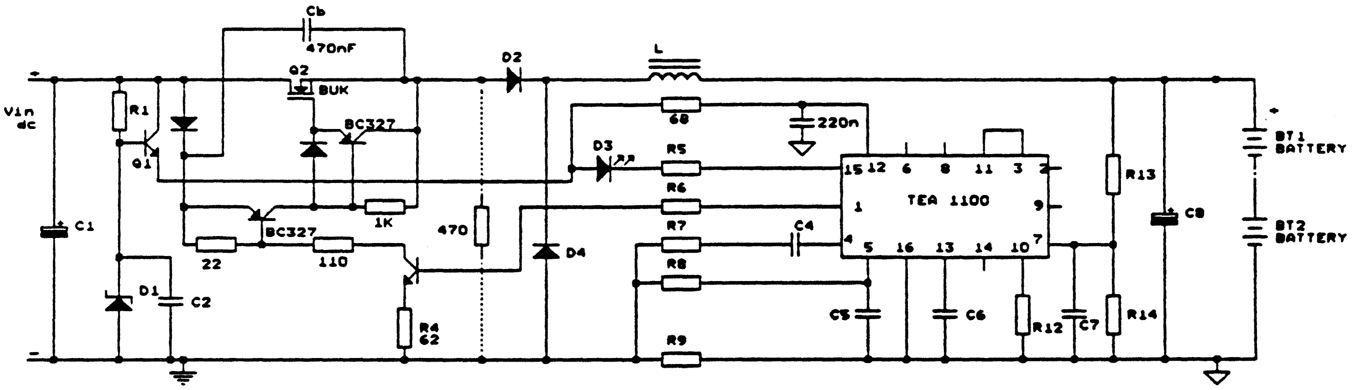


FIG. IV-b. CHARGER DIAGRAM WITH N CHANNEL POWERMOSFET SWITCH
IN THE BUCK REGULATOR

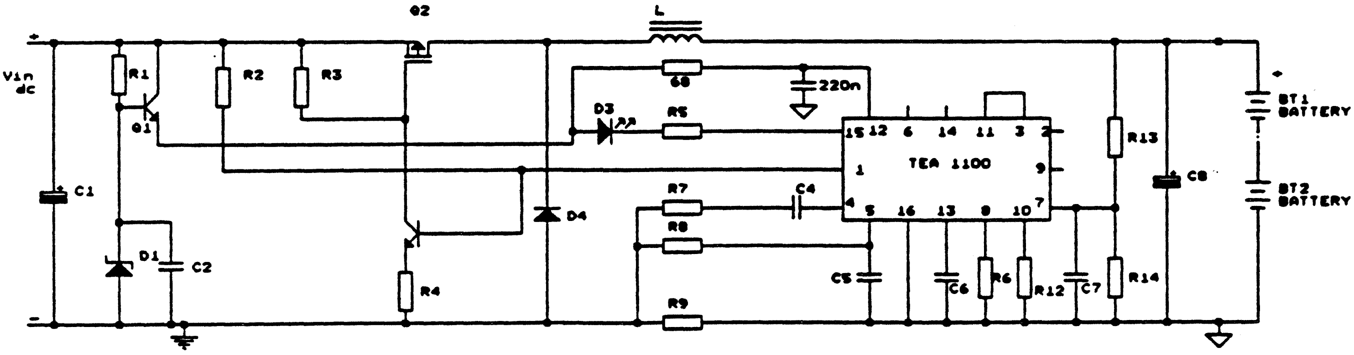
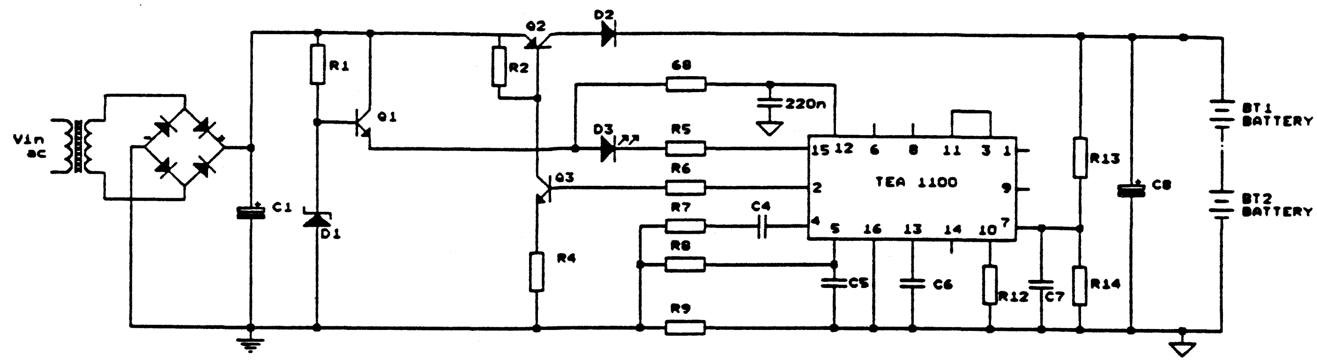


FIG. IV-c. CHARGER DIAGRAM WITH P CHANNEL POWERMOSFET SWITCH
IN THE BUCK REGULATOR



TRICKLE CHARGE BY REGULATED CURRENT BURSTS.

FIG. 4d CHARGER DIAGRAM WITH DIRECT CURRENT SERIES REGULATION.

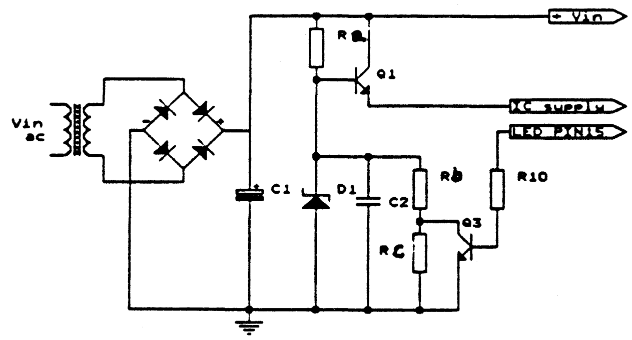
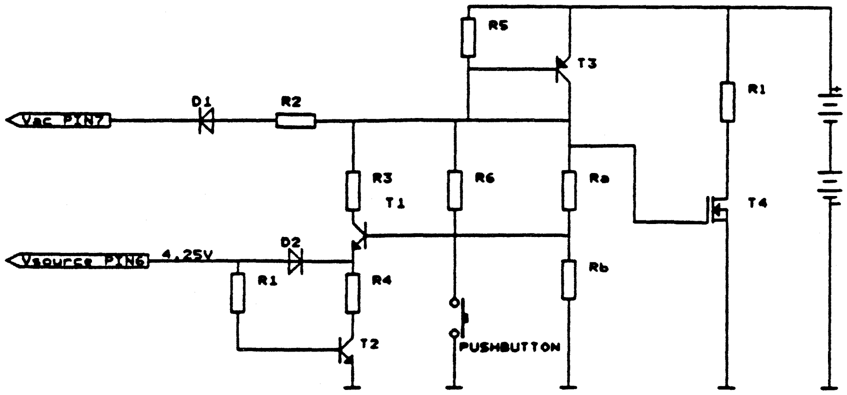


FIG. 4e SYSTEM OPERATION WITH SUPPLY HYSTERESIS CONTROL

When charge current starts flowing and internal resistance in the system input supply causes a noticeable voltage dip, then charger on/off switching can occur; this can be prevented by using a supply hysteresis control.
With the divider Ra over Rb or Ra over Rb and Rc, controlled by Q3 via the LED signal of TEA1100, a large supply hysteresis can be programmed.



DISCHARGE BEFORE CHARGE EXAMPLE

When the feature, discharge before charge is required, then this circuit can fulfil this option and is applicable for a 5 or more cell's battery.
Discharge is switched-on on demand by pressing the pushbutton, only when the charger is on. During discharge, charge is disabled by forcing the Vac high into protection and remains so until the batteries' divided voltage by Ra/Rb drops below 4.25V. This terminates the discharge latch function and restarts the charge sequence.
Instead of the discharge power Mosfet, a bipolar power transistor is an alternative.

Transistor functions;

T2: enables latching,T1 and T3, when TEA1100 is on.
T1: comparator within the latch.
T3: latch driver switch for T4 and Vac high setting via D1.

APPLICATION NOTE

ICs for Battery Management

NiMH Batteries recharge characteristics under TEA1101 charge management

Report No: NPO/AN9301

W. Bosboom

Customer & Product Support Group Consumer ICs Nijmegen, the Netherlands.
Fax: +31 80 533659

Keywords
NiMH/NiCd battery
TEA1100/TEA1101
Battery management
Fast charge systems

Date : 20 January 1993
Pages: 16

Summary

Rechargeable NiMH-batteries of the 4 major, worldwide operating suppliers (Vendor A - D) have been tested on their fast charge characteristics under control of the Philips Semiconductors IC TEA1101,- battery management IC for NiMH and NiCd chargers -.
For reference measurements of NiCd cells have also been included.

Tests were performed under several conditions;

- fast charge currents: 1CA and 1.5CA at 25°C ambient temperature
- extreme conditions of batteries' temperature and state of charge

The charge voltage profile of NiMH batteries does not exhibit a voltage drop into overcharge as pronounced as that of NiCd batteries. The evaluations were performed with -dV charge termination of 2 and 4mV/cell.

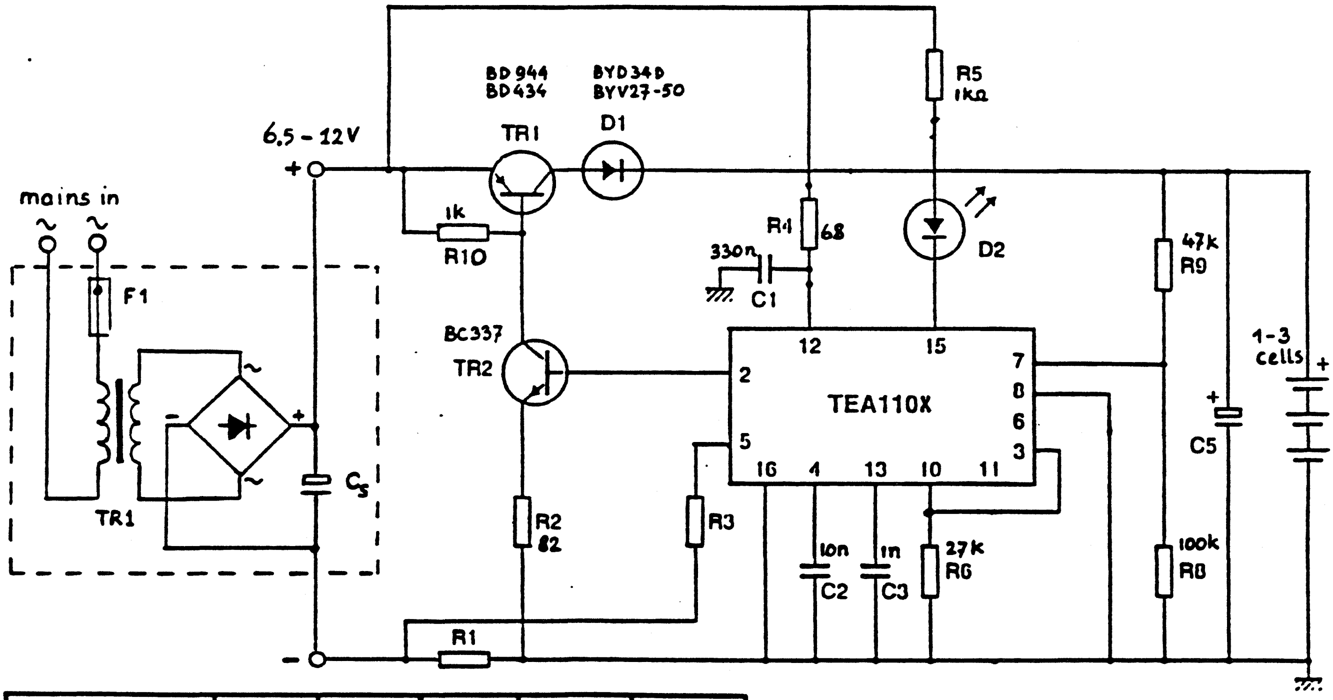
The test results show that TEA1101 is well suitable to charge NiMH and NiCd cells under both normal and stress conditions.
The implementation of the sensitive negative delta voltage (-dV) cut-off keeps the batteries' temperature rise into overcharge within acceptable limit, - important reliabilty factor and condition for good cycle performance -.
Unlike temperature based charge termination techniques the implemented charge voltage profile termination method remains reliable in case battery ambient temperature may fluctuate due to external influences; e.g. heat generating charging electronics or extreem conditions of batteries' temperature like insertion of a cold or warm battery.

1. Measurement setup

The NiMH/NiCd batteries tests were carried out on a Semi Automatic Measuring battery test gear, equiped with a fast charger unit using the Philips Semiconductors TEA1101. This monolithic Integrated Circuit features:

- Battery full detection by battery voltage sensing.
  The TEA110X-family features an accurate and reliable -dV mechanism by currentless voltage sensing and digital filtering techniques, a prevention against false -dV cutoff due to noise and other natural fluctuations during charge.
- protection against minimum and maximum cell temperatures.
- protections against open, short or faulty battery packs.
- time-out function.
- regulation of charge currents in conjunction with a switched mode or a dc power supply.
- interfacing control for a stand-alone current power supply.
- user-interfacing by LEDs

Circuit diagram of the charger unit.



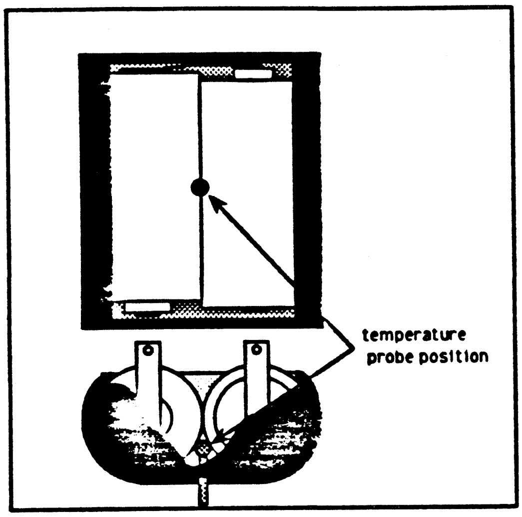
Icharge (A)	0.6	1	1.1	1.5	1.65
R1 (Ω)	0.25	0.25	0.25	0.167	0.16
R3 (kΩ)	3.24	5.4	6	5.4	6

TEA110X Battery Charger with Linear Regulator Circuit

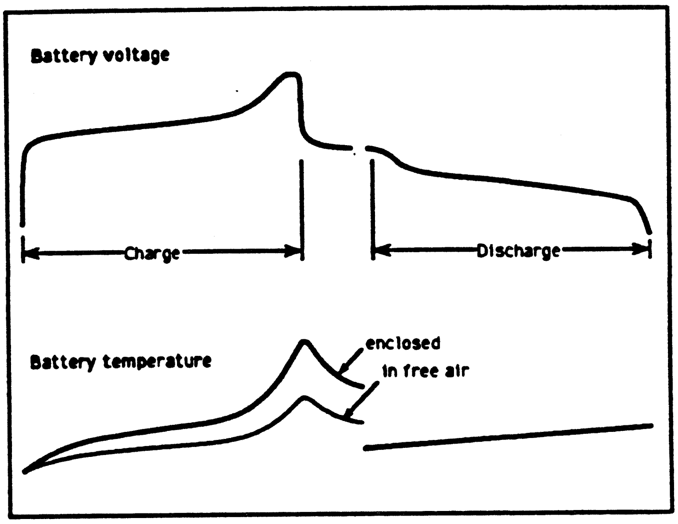
Details of this circuit and alternative charger configurations have been described in application notes mentioned in references, section 3.

Battery assembly

Temperature rise of battery cells in an enclosed plastic encapsulation (pack) will be higher than that of cells widely spaced apart in open air.
The measurements have been performed on two cells constructed side-by-side in an open top plastic case. The battery temperature was probed in between the cells, at the confined area side.



Battery assembly



Voltage and temperature characteristics

The end of charge temperature of widely space battery cells without an plastic case will be appoximately 10°C less than that of the tested enclosed cells.

2. Measurement data

Measurements at ambient temperature 25 °C

The measurement results are presented in voltage and temperature graphs where:

- a) 2 cells battery voltage.
- b) battery temperature.

NiMH

-Icharge: 1 CA, -dV: 1/4% (≈4mV/cell)	Fig. 1 - 4.
-Icharge: 1 CA, -dV: 1/8% (≈2mV/cell)	Fig. 5 - 8.
-Icharge: 1,5CA, -dV: 1/4% (≈4mV/cell)	Fig. 9 - 12.
-Icharge: 1,5CA, -dV: 1/8% (≈2mV/cell)	Fig. 13- 16.

NiCd

-Icharge: 1 CA, -dV: 1/4% (≈4mV/cell)	Fig. 17.
-Icharge: 1,8CA, -dV: 1/8% (≈2mV/cell)	Fig. 18.

Stress conditions

NiMH-battery and charger both at 40°C Tambient	Fig. 19.
NiMH-battery at 0 °C and charger at 25°C Tambient	Fig. 20.
NiMH-battery at 60 °C and charger at 25 °C Tambient	Fig. 21.
overdischarged NiMH battery	Fig. 22.
full NiMH battery	Fig. 23.

3. References

1 NPO/AN9102b. TEA1100, Versatile battery management IC for NiCd charge systems.
2 TEA1100(T) / TEA1101(T). Device specification
3 NPO/AN9201. TEA1100, Battery charger with flyback SMPS current regulator
4 SPO/AN92004. TEA1100, Battery charger with linear regulator

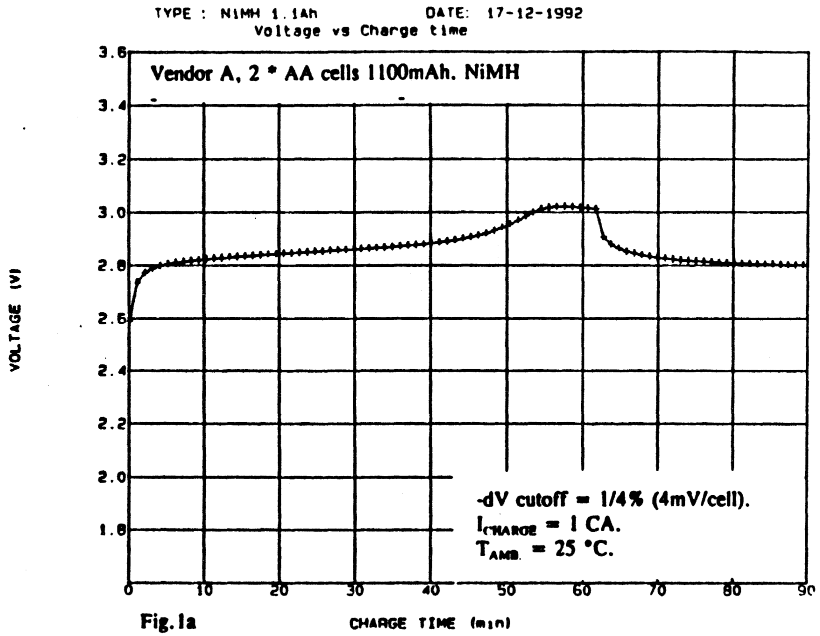


Fig.1a

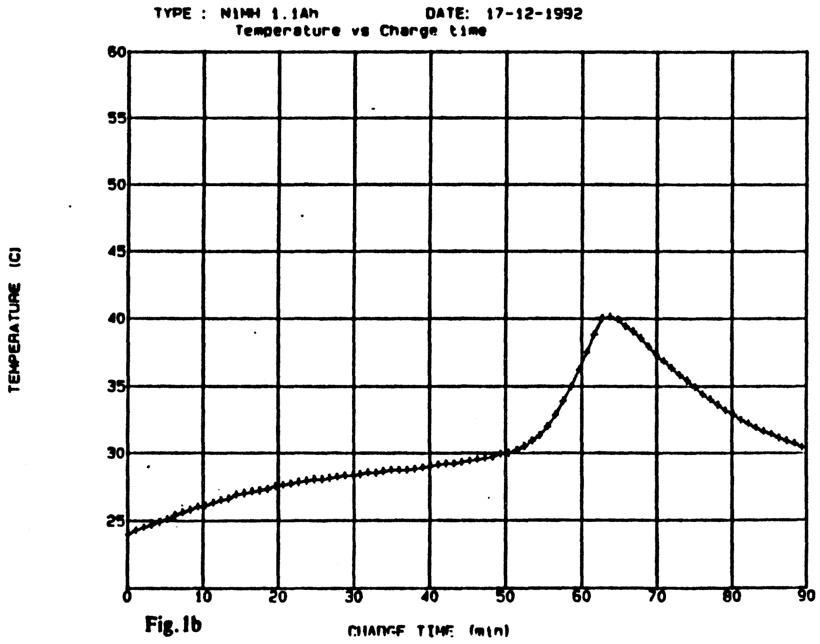


Fig.1b

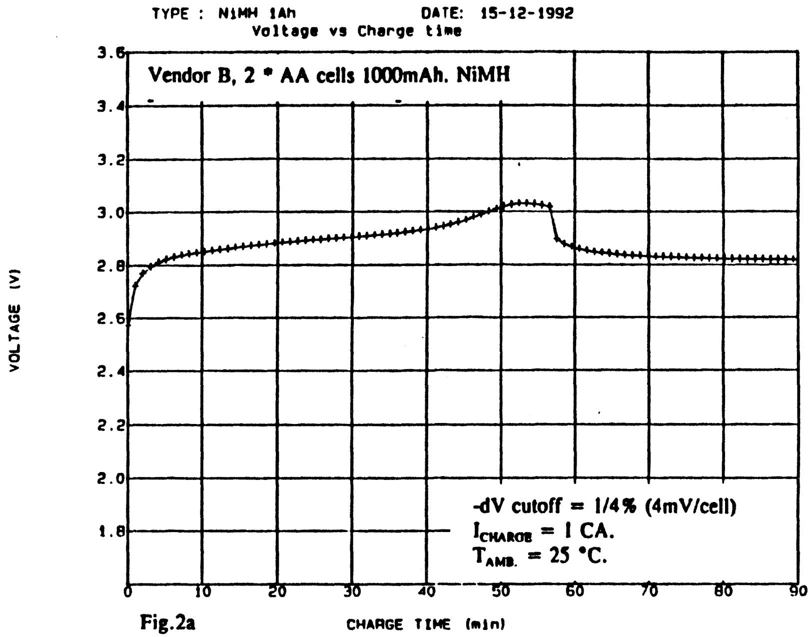


Fig.2a

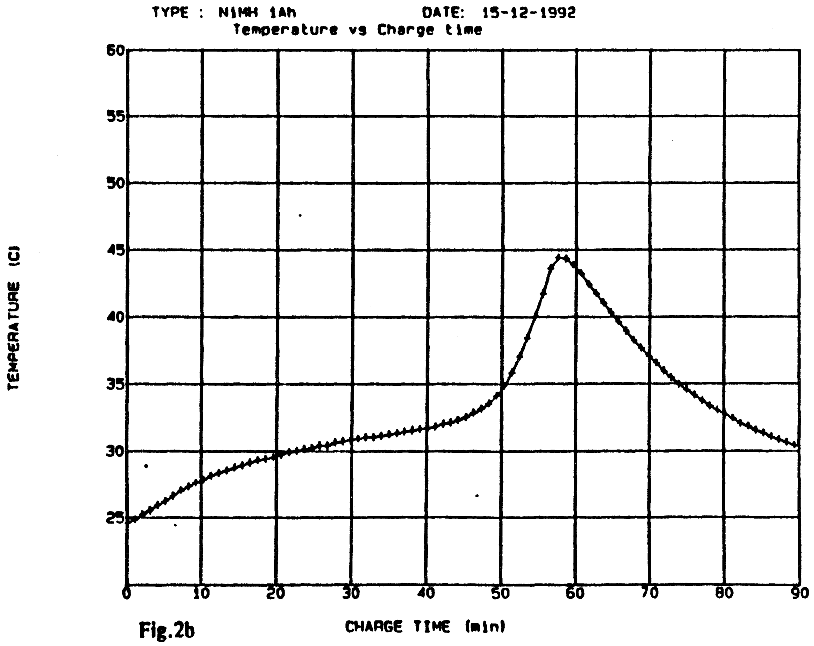


Fig.2b

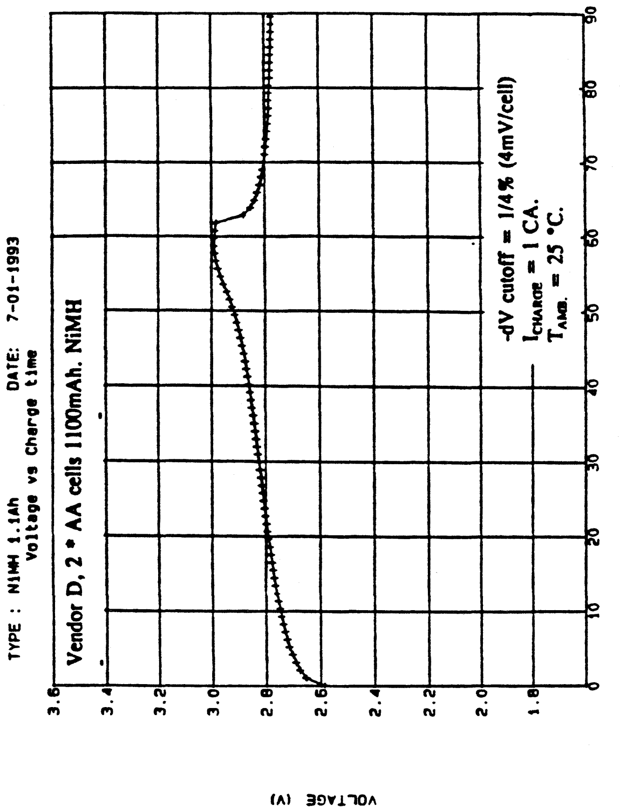


Fig.4a

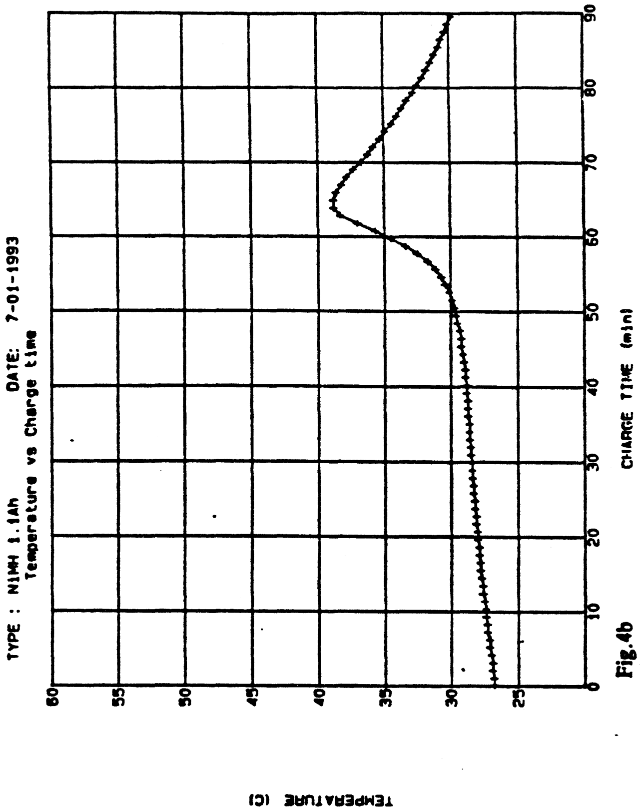


Fig.4b

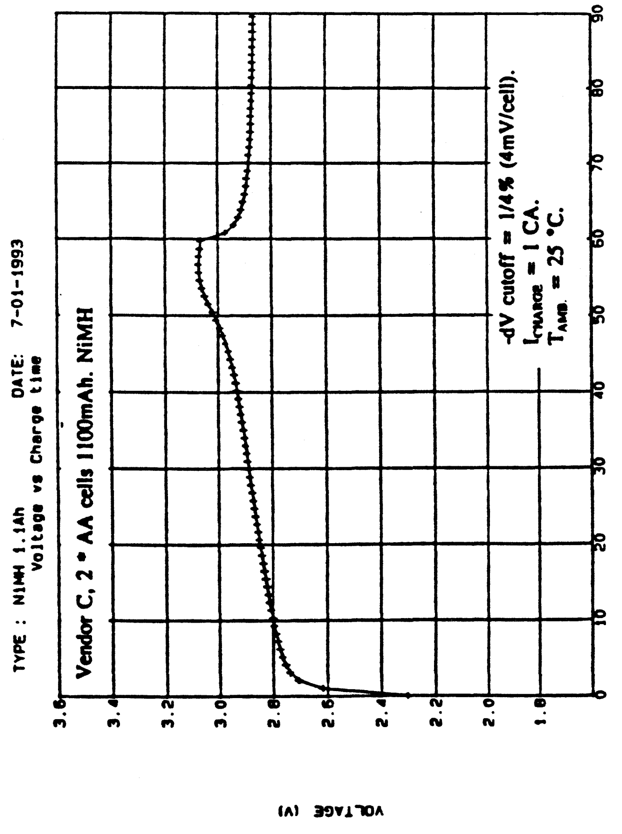


Fig.3a

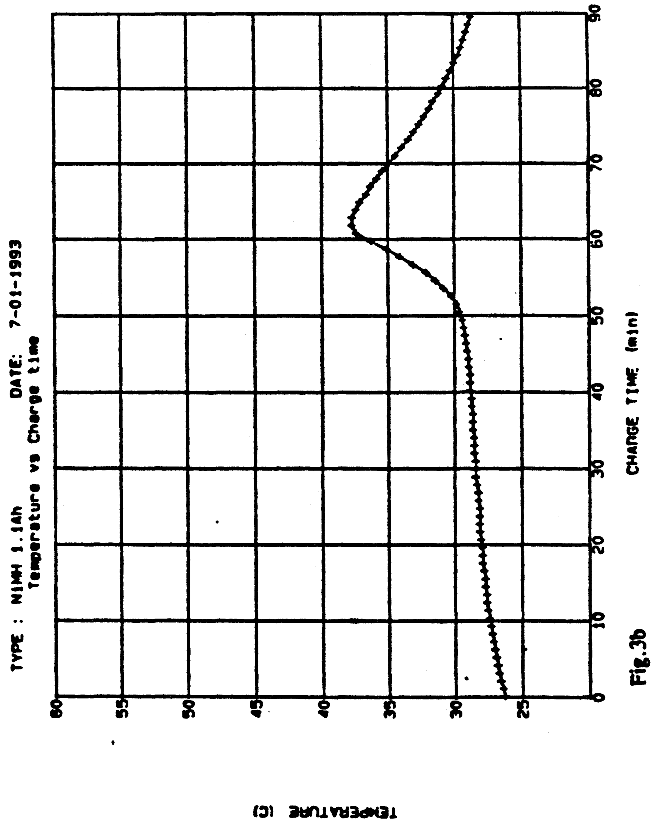


Fig.3b

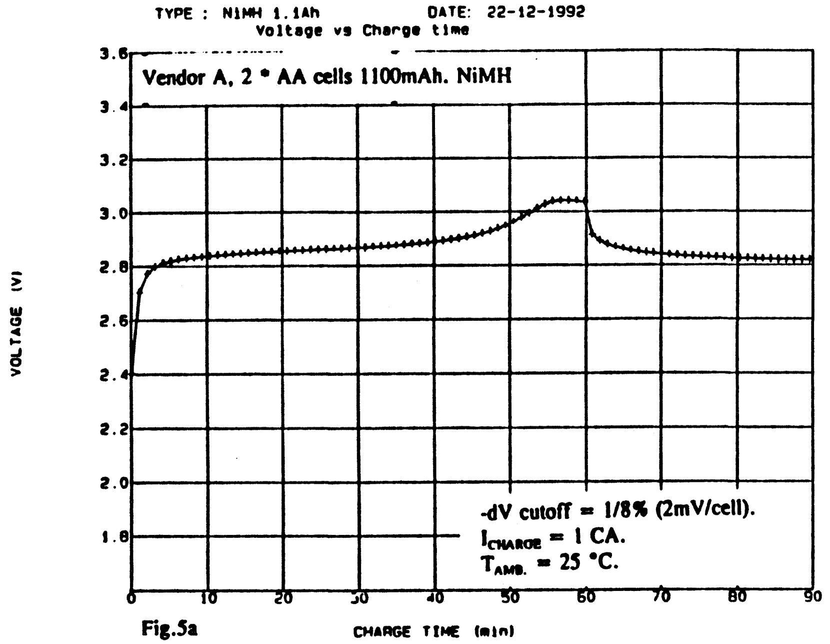


Fig.5a

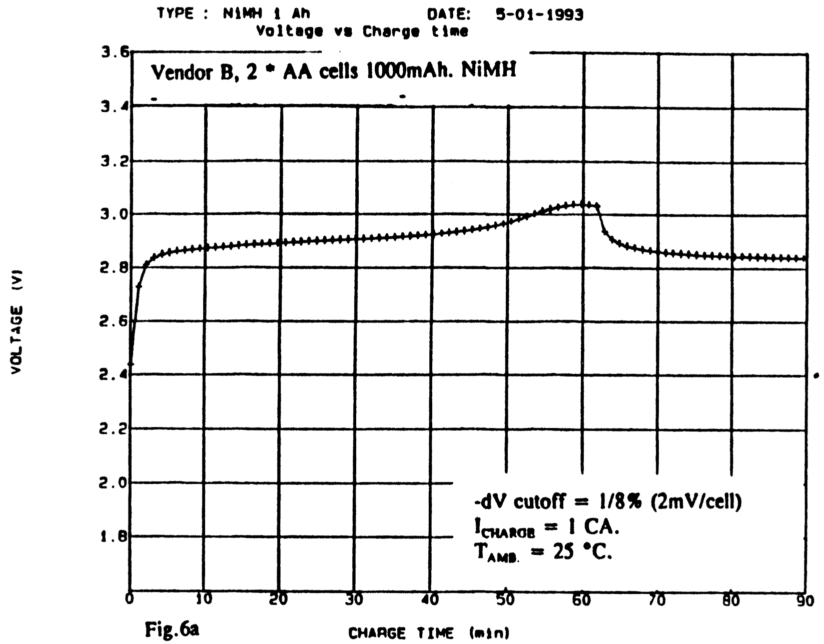


Fig.6a

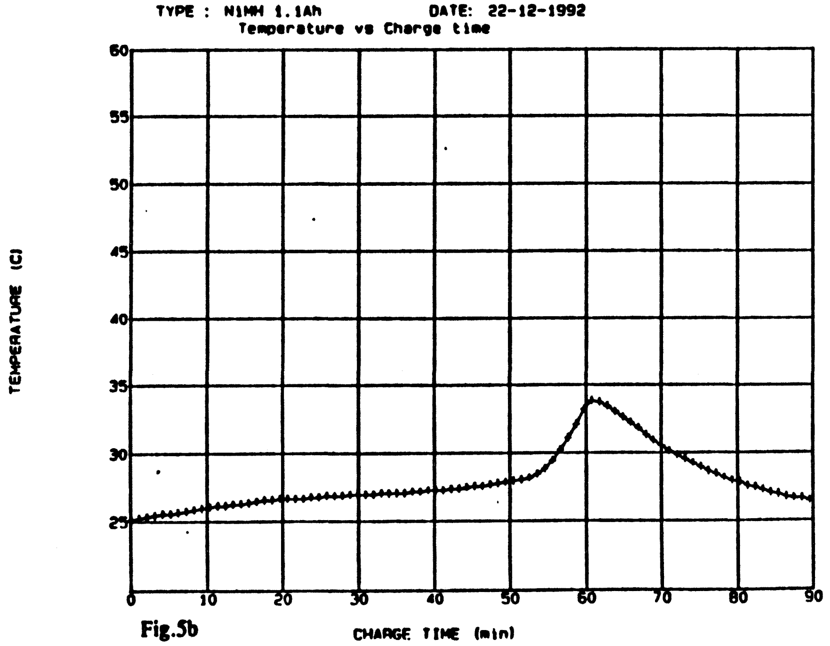


Fig.5b

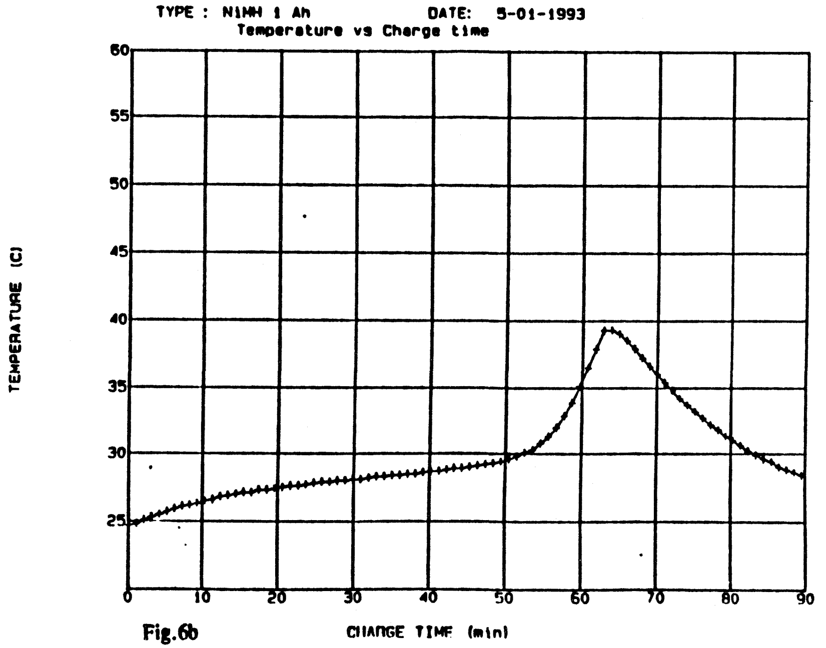


Fig.6b

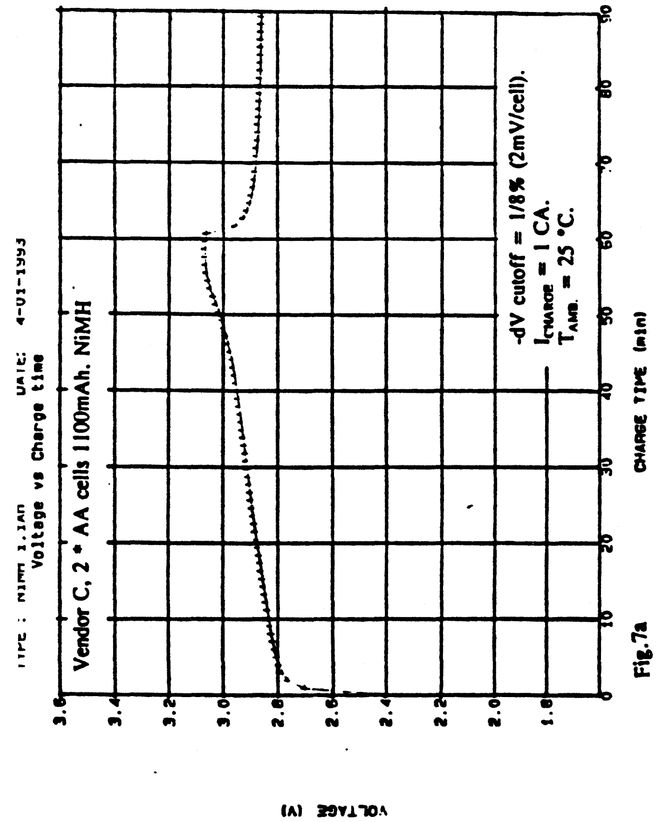


Fig.7a

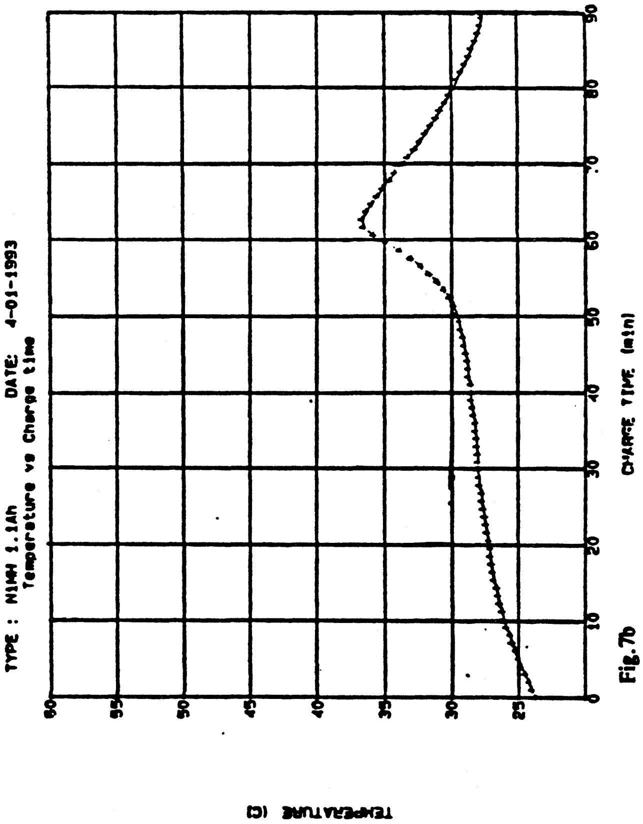


Fig.7b

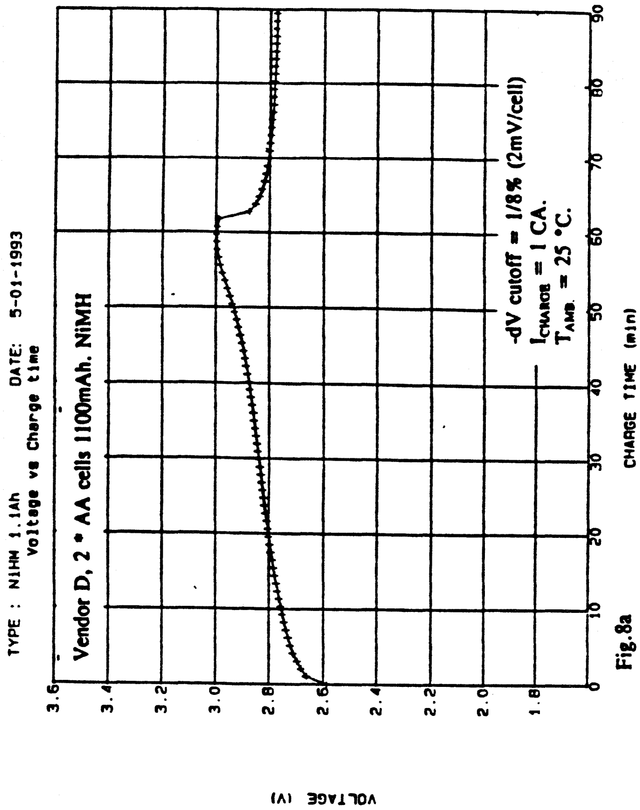


Fig.8a

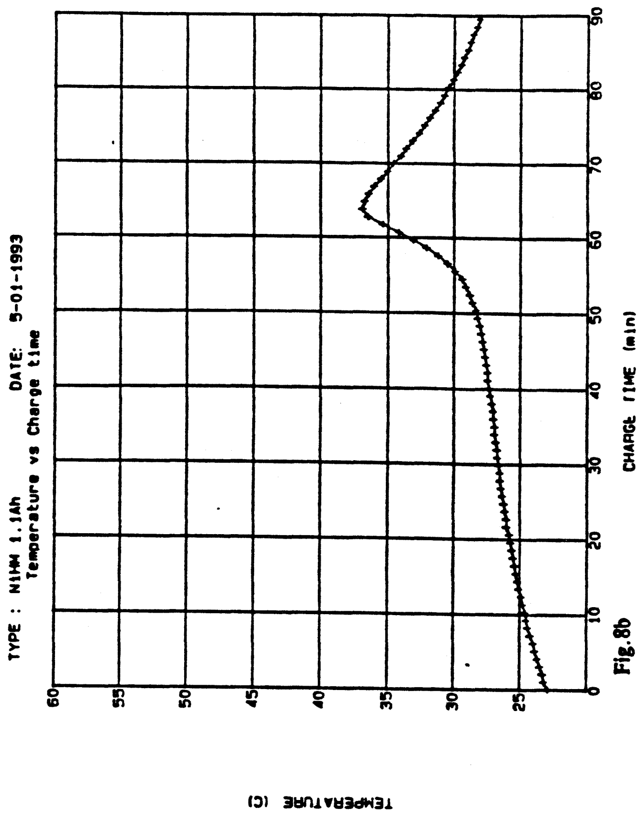


Fig.8b

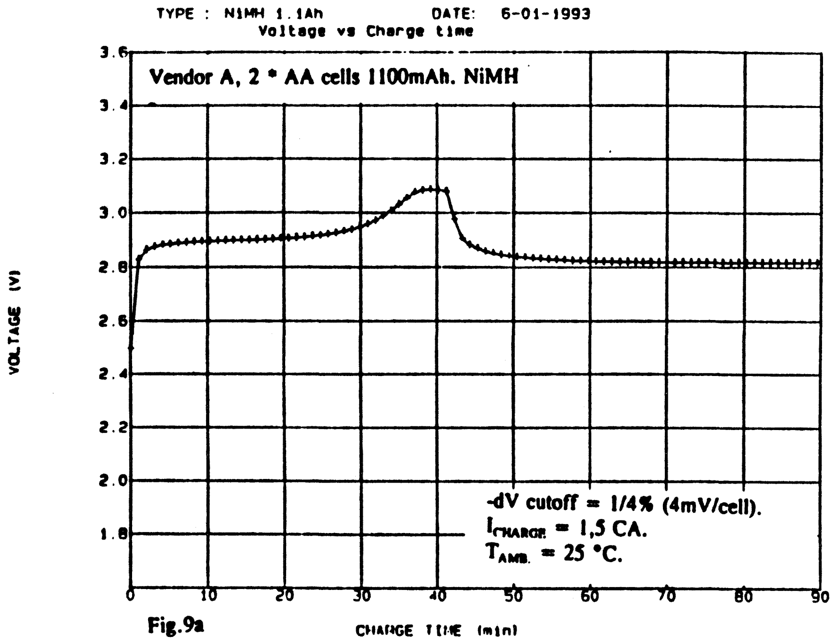


Fig.9a

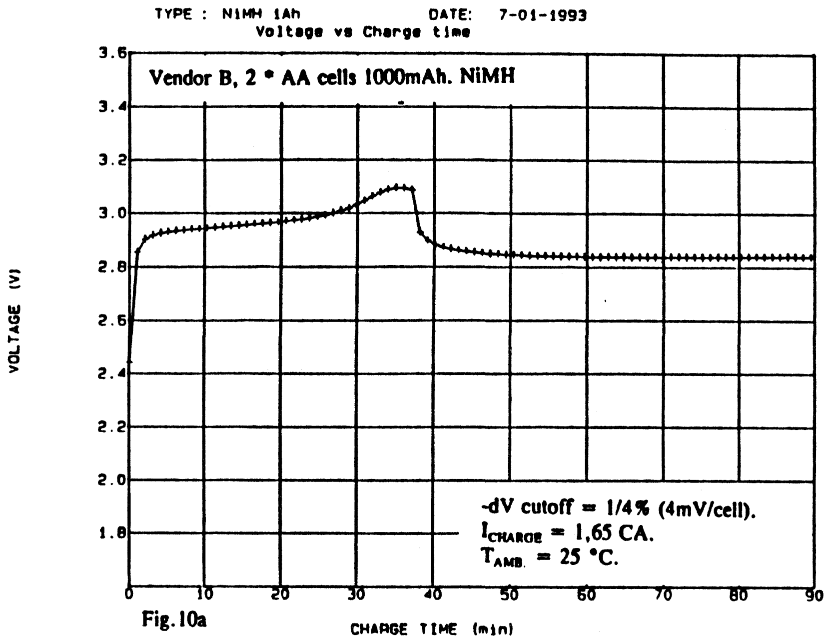


Fig.10a

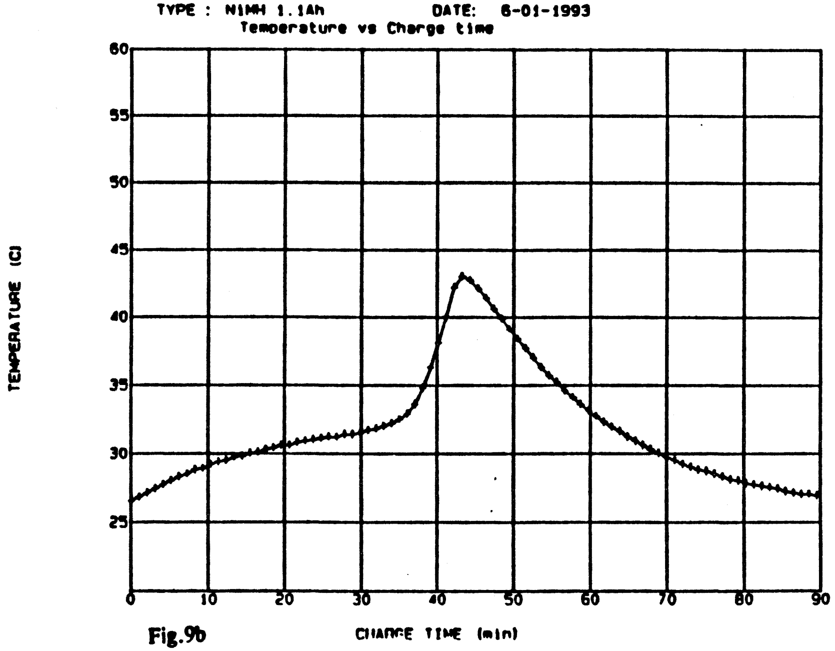


Fig.9b

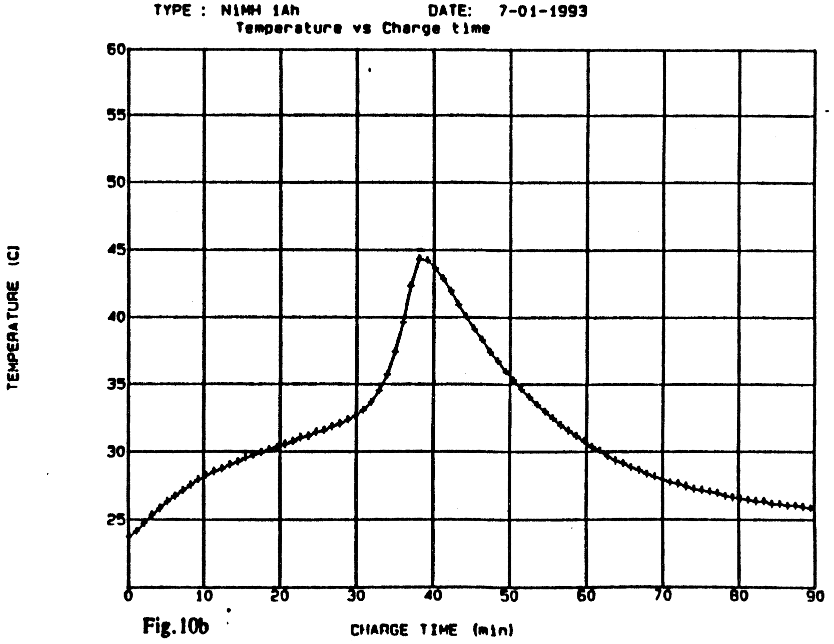


Fig.10b

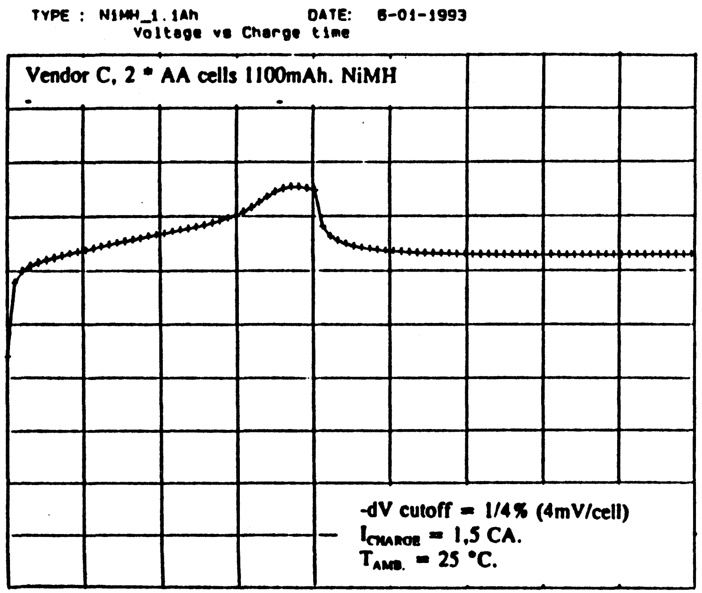


Fig.11a

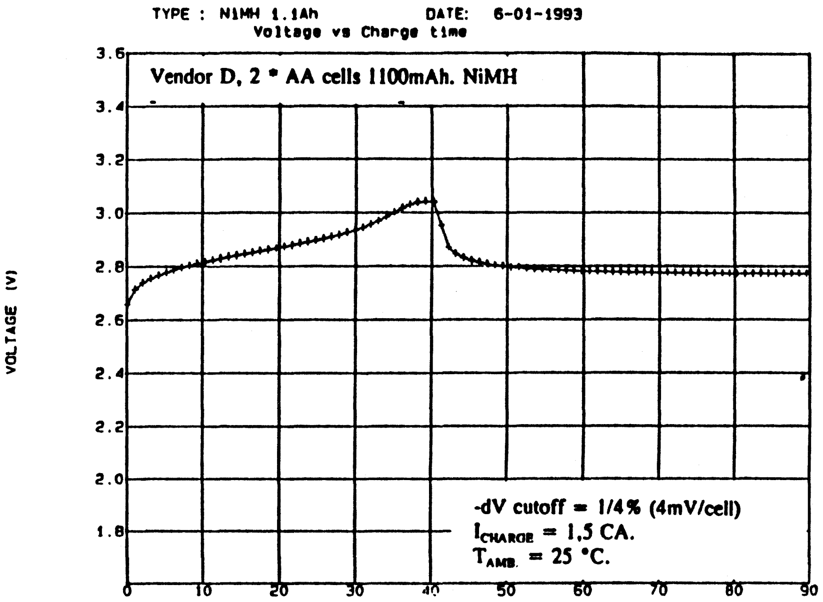


Fig.12a

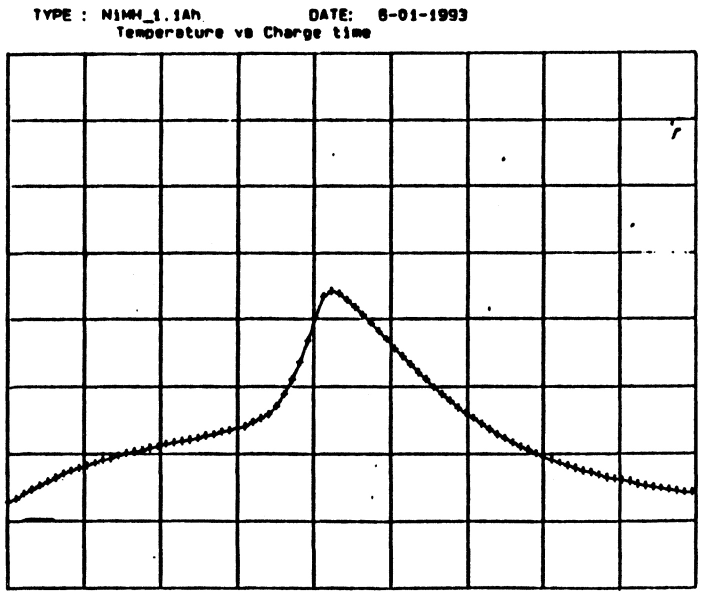


Fig.11b

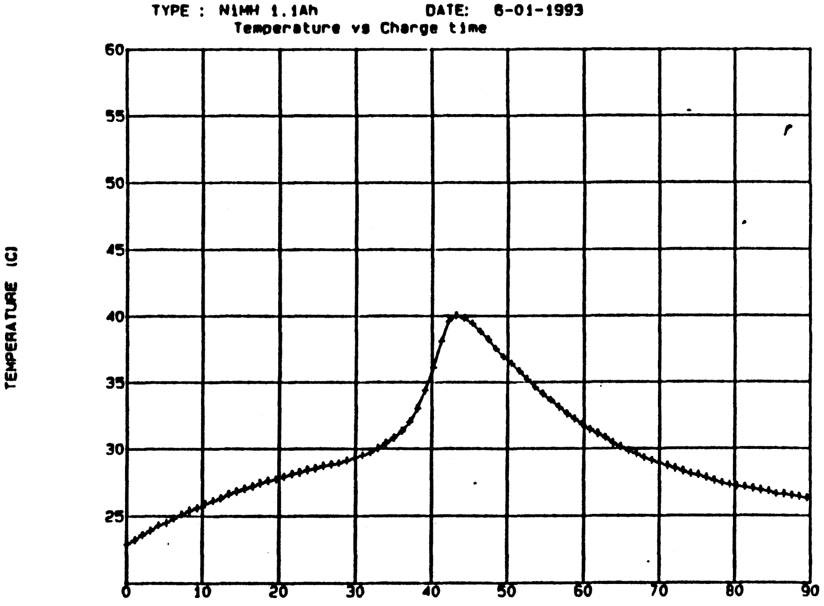


Fig.12b

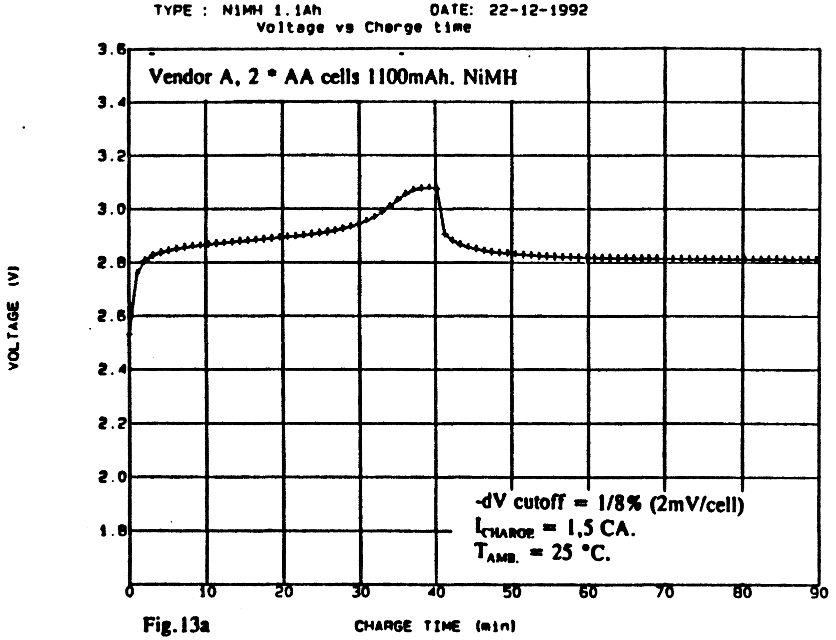


Fig.13a

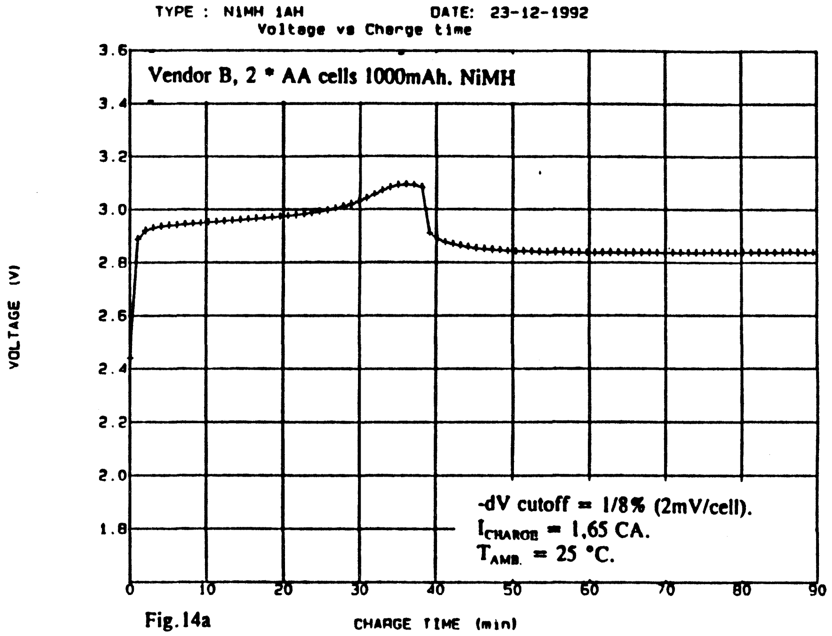


Fig.14a

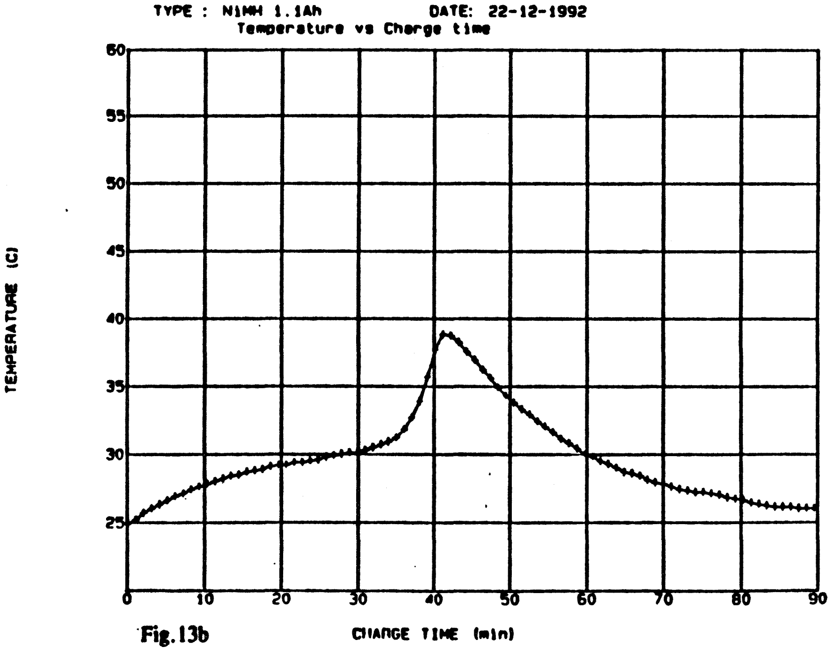


Fig.13b

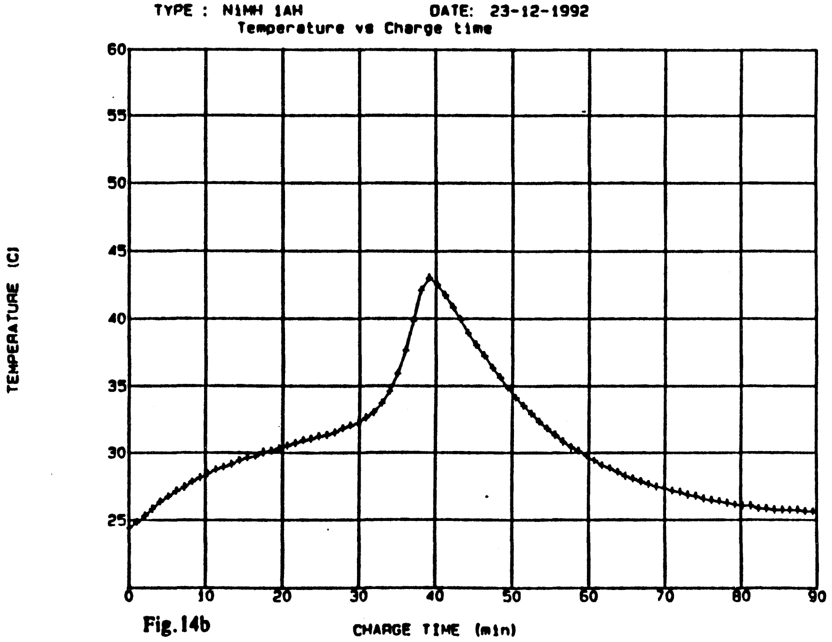


Fig.14b

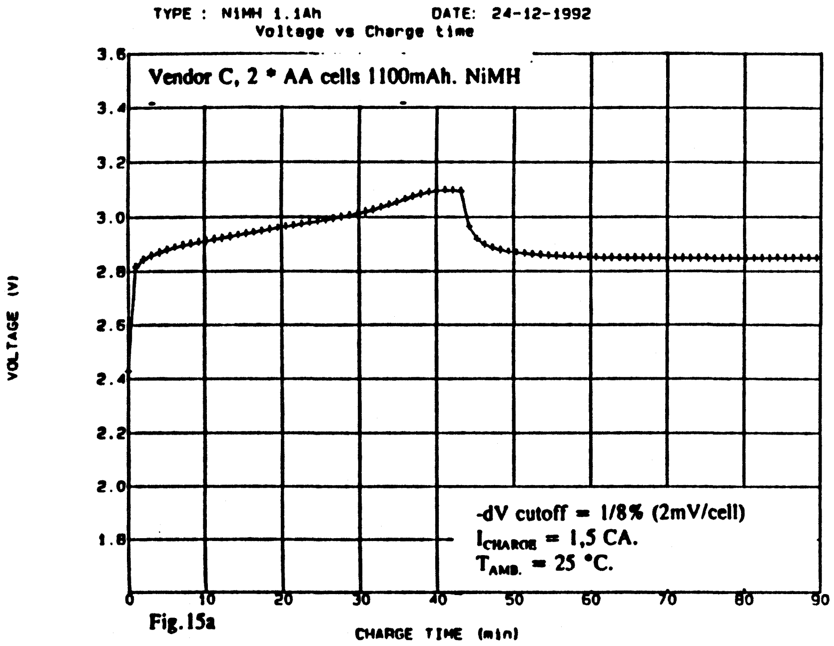


Fig.15a

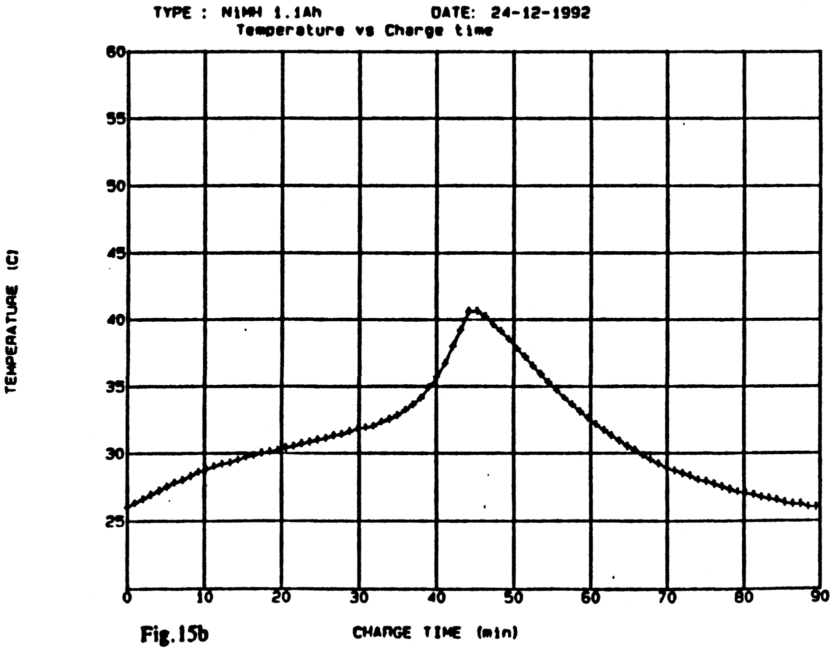


Fig.15b

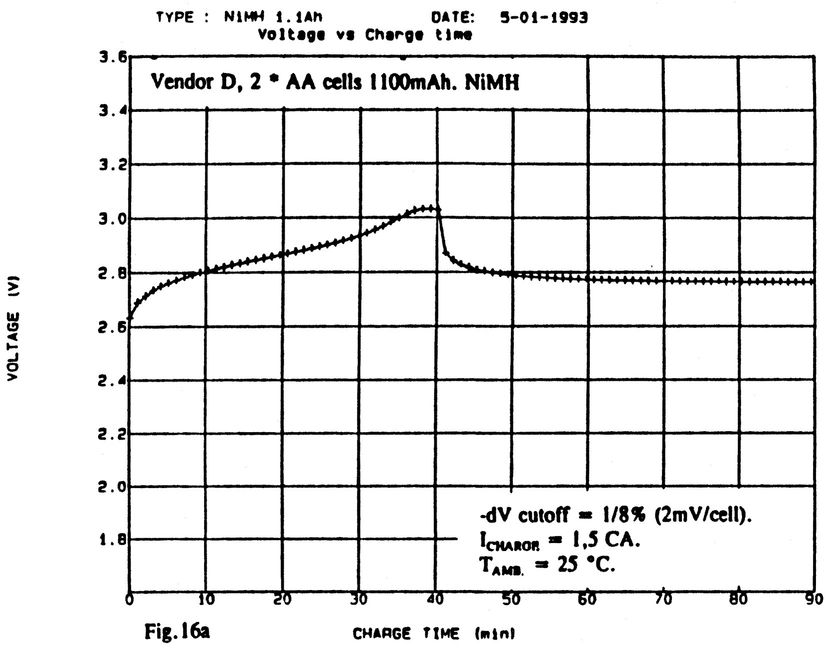


Fig.16a

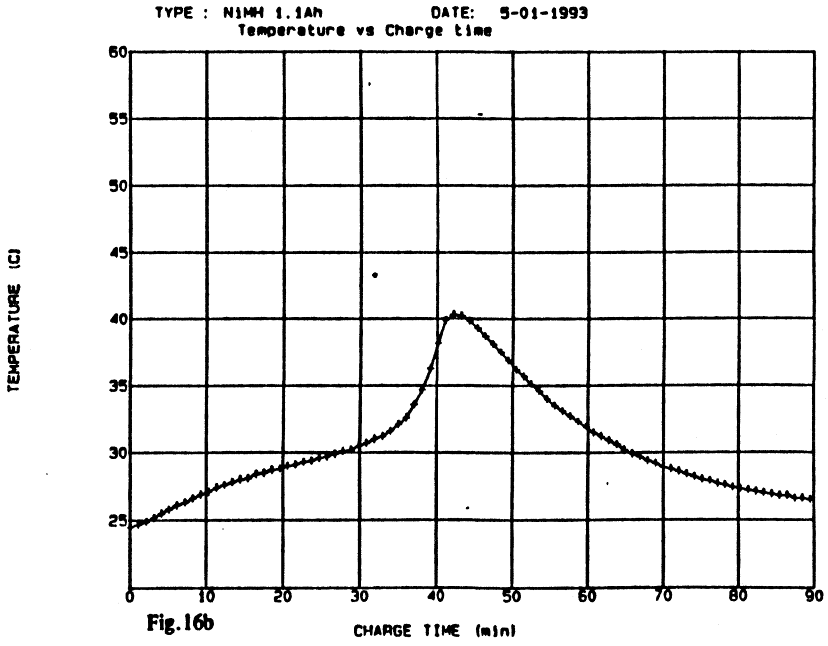


Fig.16b

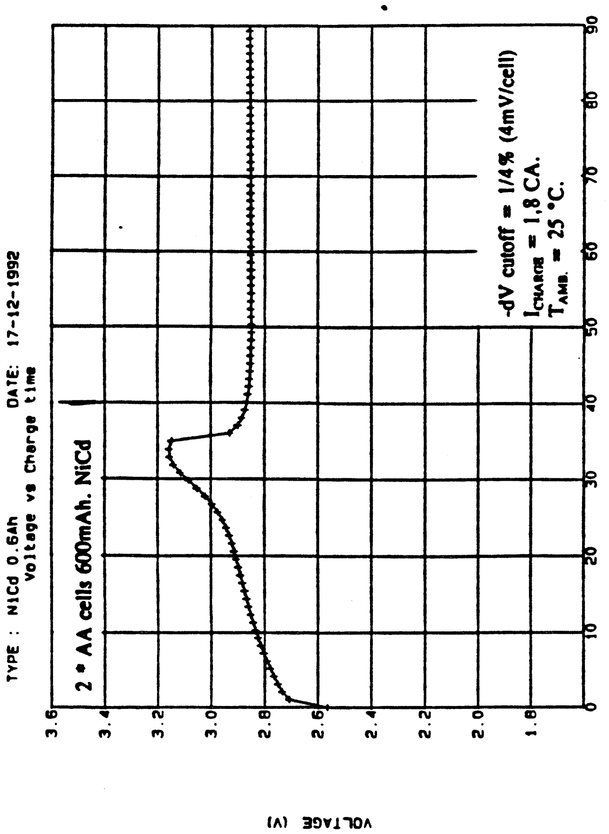


Fig.18a

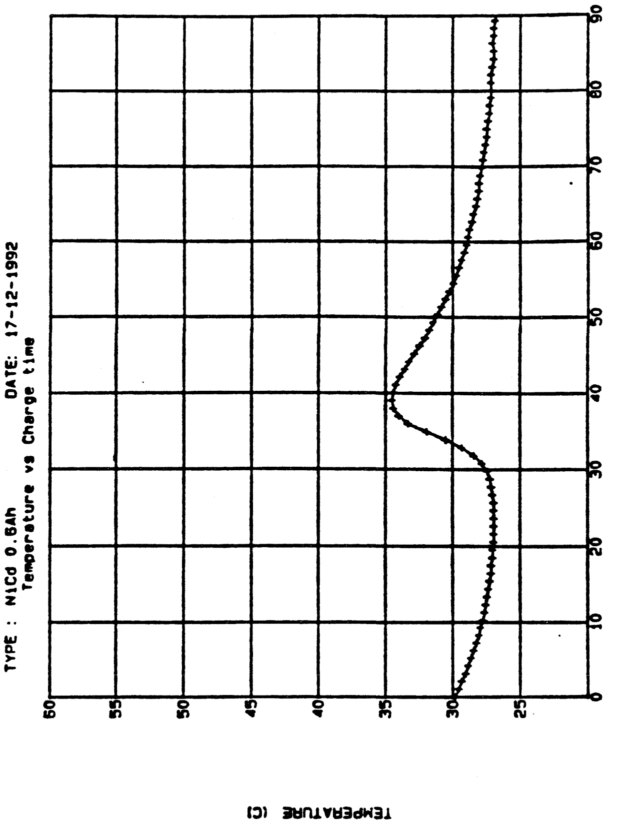


Fig.18b

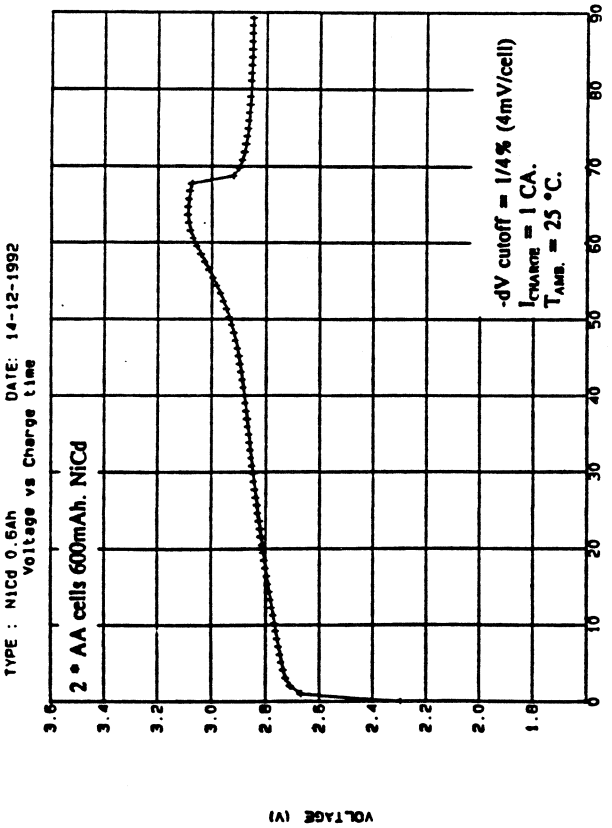


Fig.17a

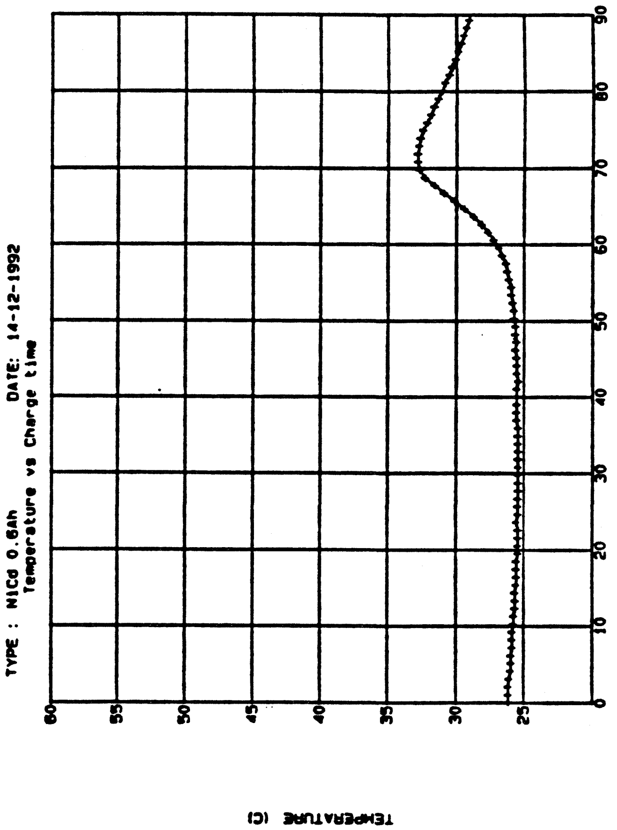
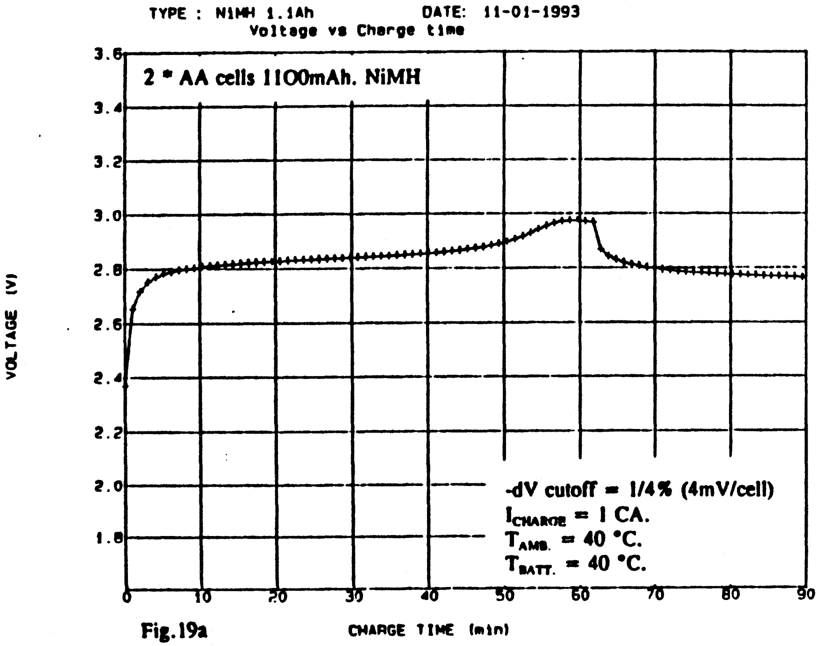


Fig.17b



-dV cutoff = 1/4% (4mV/cell)
I_{CHARGE} = 1 CA.
$T_{AMB.}$ = 40 °C.
$T_{BATT.}$ = 40 °C.

Fig.19a

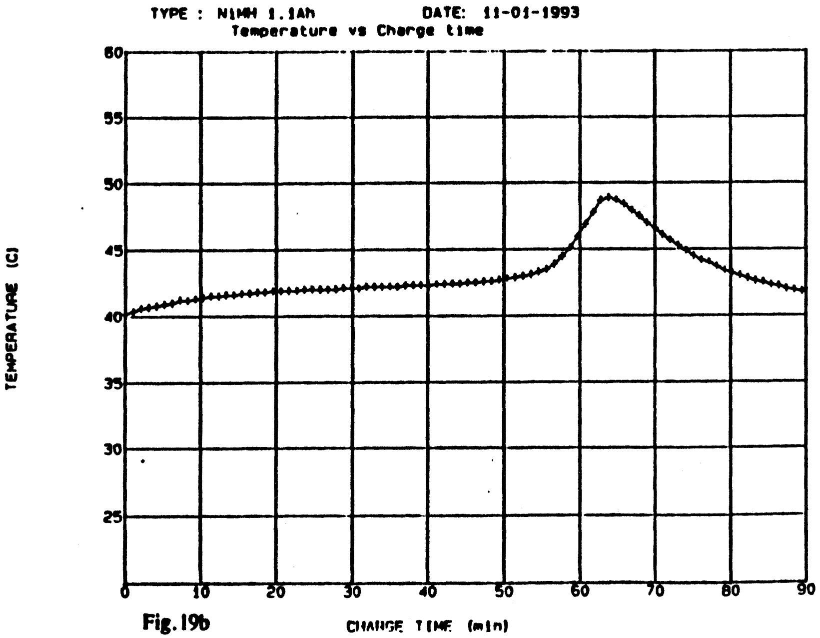
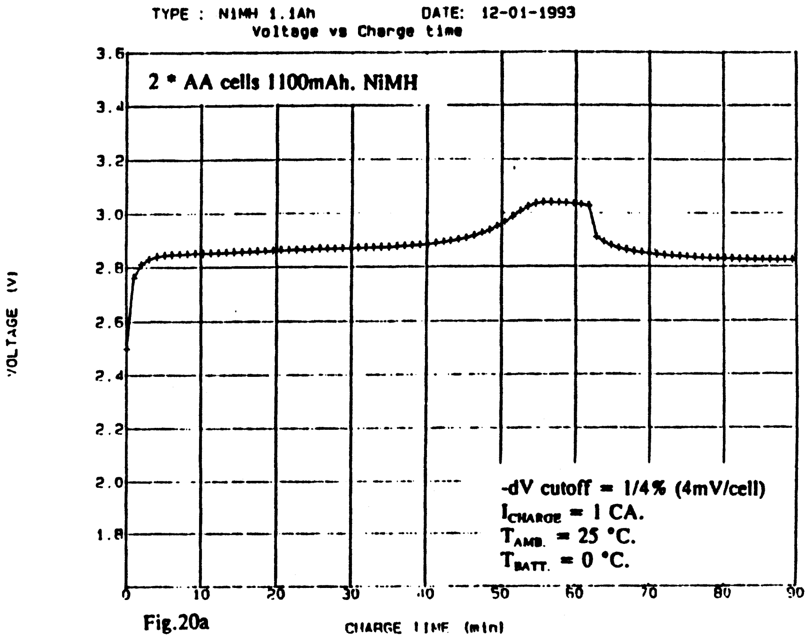


Fig.19b



-dV cutoff = 1/4% (4mV/cell)
I_{CHARGE} = 1 CA.
$T_{AMB.}$ = 25 °C.
$T_{BATT.}$ = 0 °C.

Fig.20a

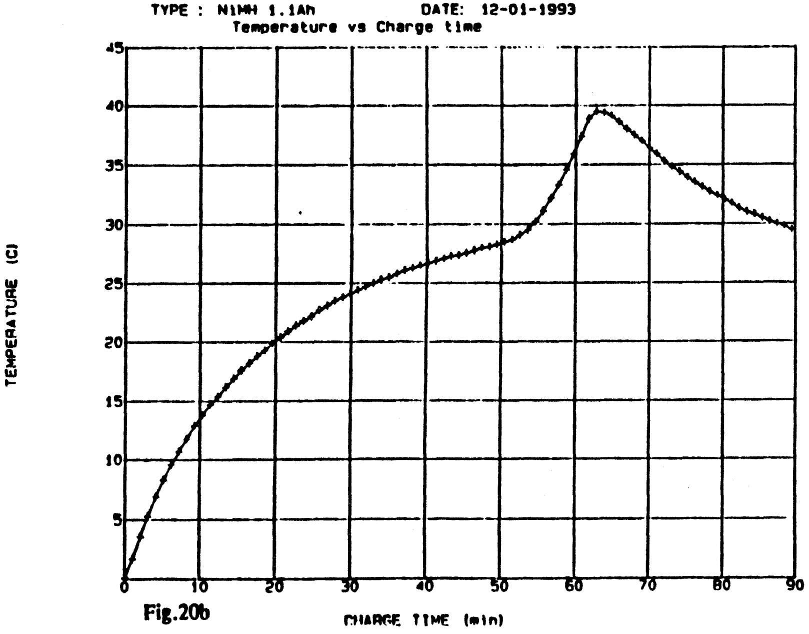


Fig.20b

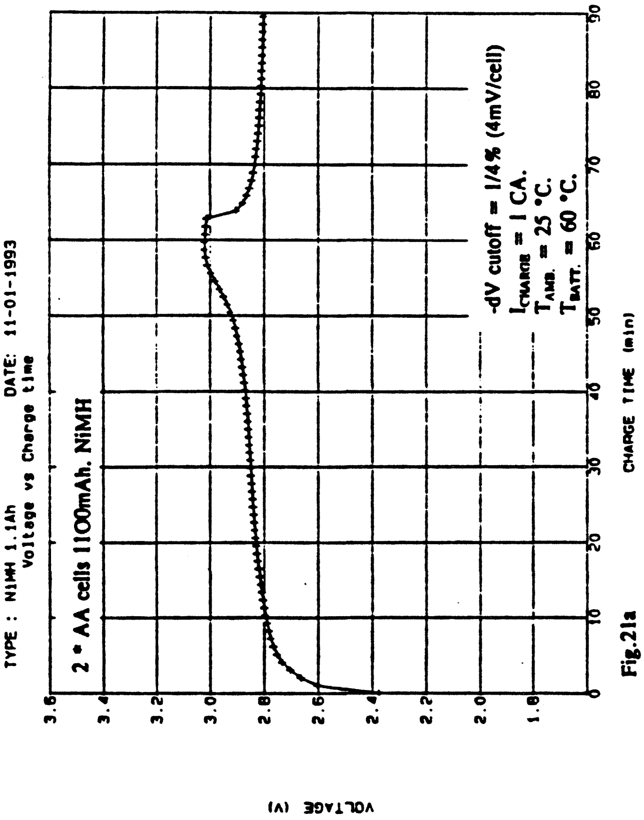


Fig.21a

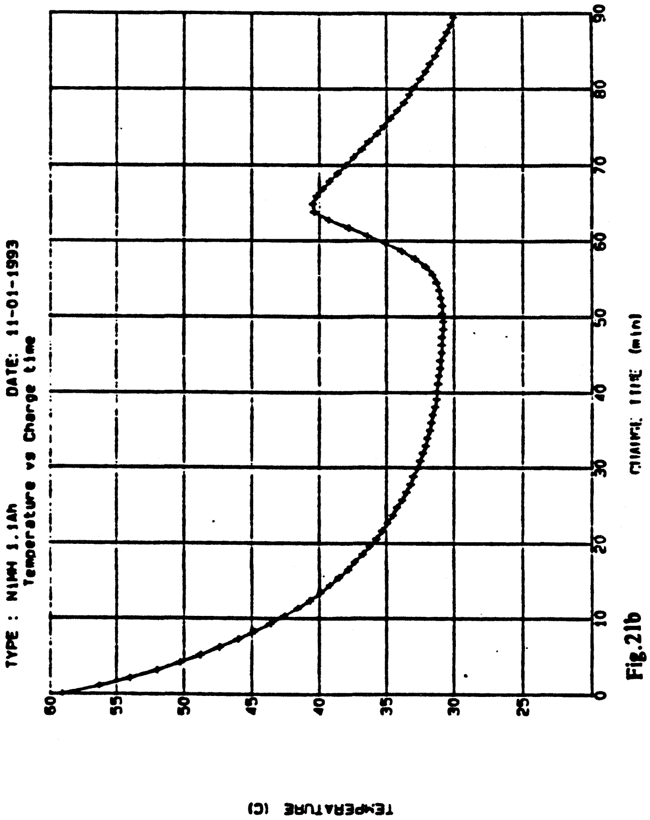


Fig.21b

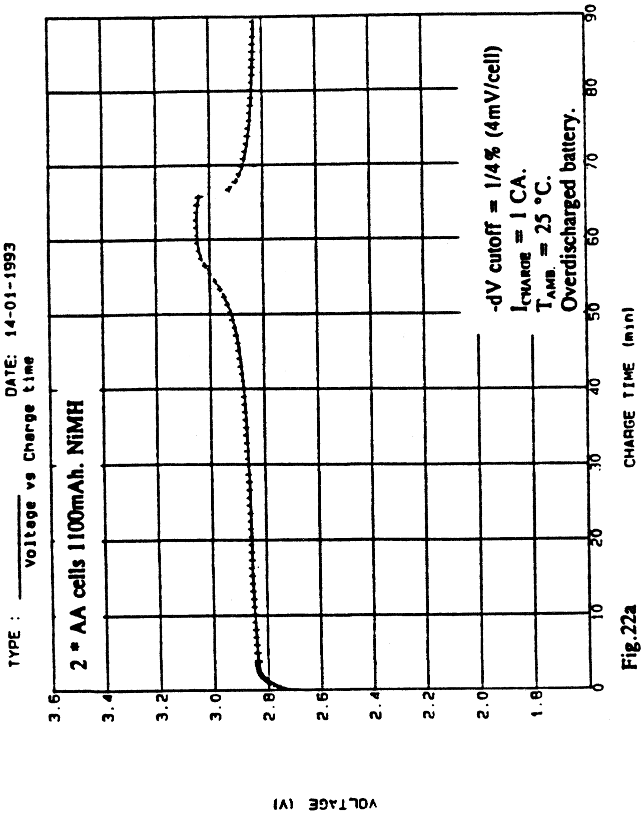


Fig.22a

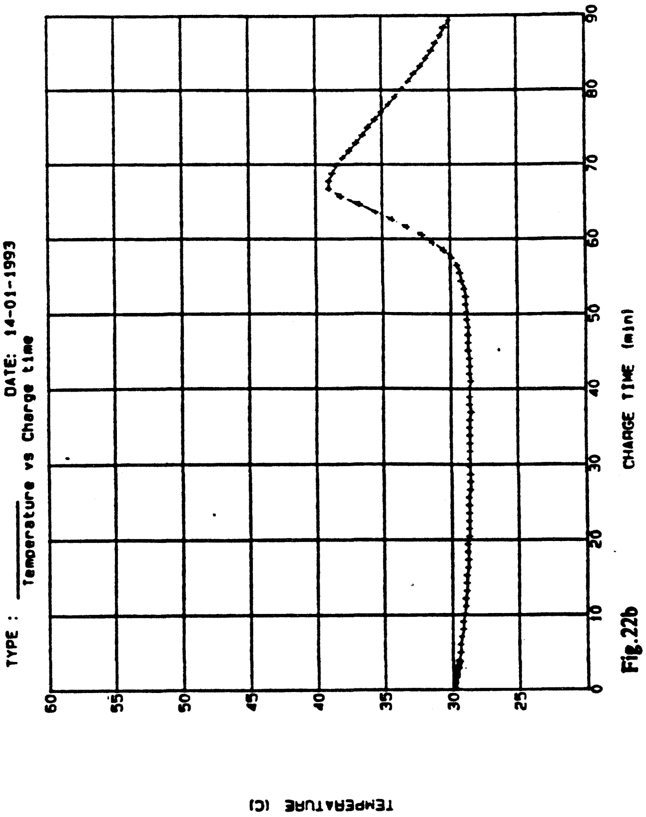


Fig.22b

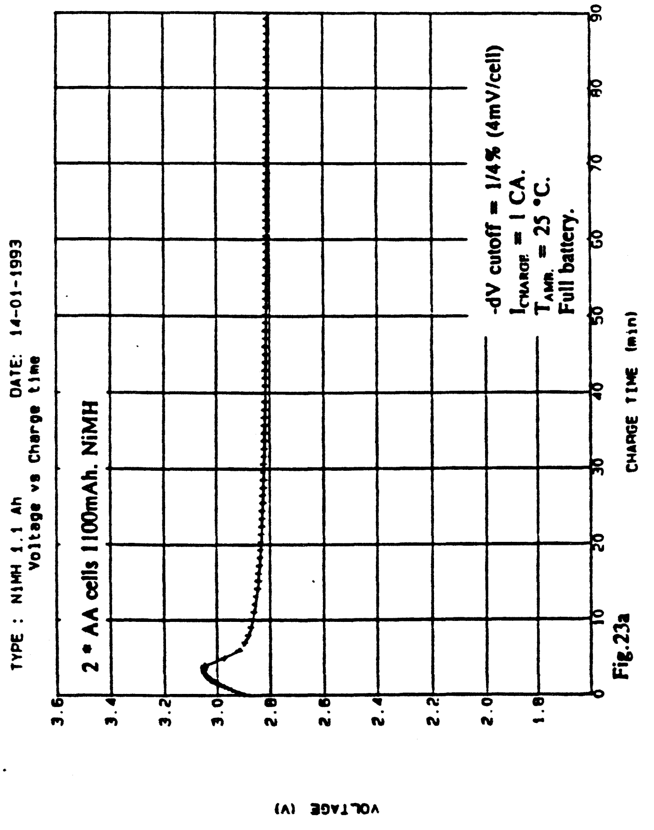


Fig.23a

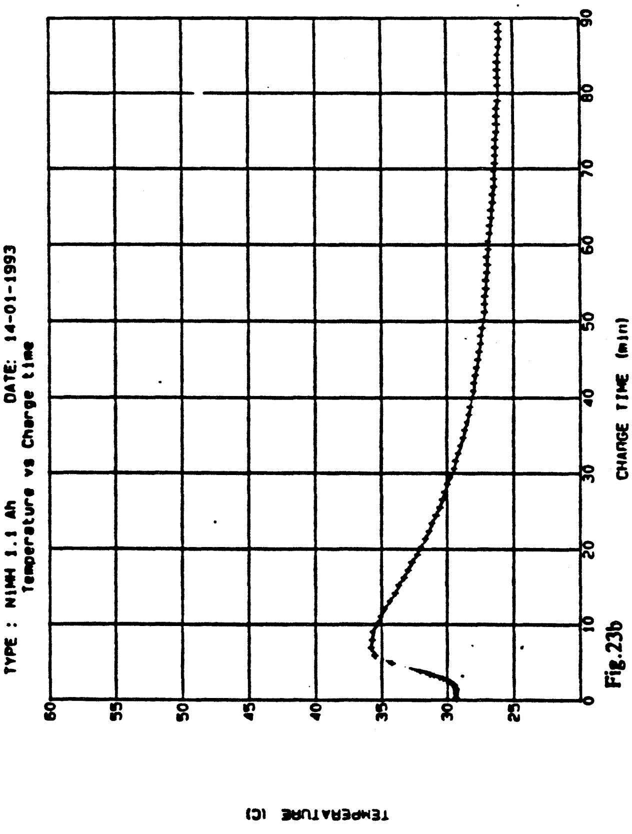


Fig.23b

Application Note

Rechargeable Battery Pack State Of Charge Indicator Using SAA1500T

R P Verney

Product Concept & Application Laboratory Southampton, England

Keywords
SAA1500T
Rechargeable
Charge Indicator

Report No : SPO/AN92001
Date : 24 FEB 1992
Pages : 28

Philips Semiconductors

PHILIPS

PHILIPS

Summary:

The state of charge indicator described in this Application Note can be incorporated into almost any (three or more cell) rechargeable battery pack. The indicator counts recharge and discharge energy to provide the user with an elegant six segment LED gauge of remaining battery capacity, together with indication of recharging. Circuit options include an LCD display and output for an audible warning device.

Table of Contents Page Number

Table of Contents Contd — Page Number

APPENDICES

List of Figures — Page Number

List of Tables

1. INTRODUCTION - STATE OF CHARGE INDICATION

State of charge indication is an important and useful addition to any rechargeable battery pack for a variety of reasons. An underlying problem is the common tendency to recharge batteries many times more often than actually necessary. People waste time and effort in this way because they are always afraid of being caught out with empty batteries. State of charge indication helps to overcome this problem and brings other benefits such as:

- Increased battery cycling and consequently improved battery performance and lifetime.
- Enhanced customer satisfaction.
- Excellent selling feature.

The requirements of a battery state of charge indicator are in many ways similar to those of a fuel gauge for a car. The most important requirement of the indicator is not accuracy per se, but that an easy to read, reliable indication is given to the end user. The possibility of the indicator showing a part full battery when the battery is completely empty must be avoided. A slightly pessimistic display at all times is therefore needed.

2. MEASUREMENT STRATEGIES

In order to give battery state of charge indication, some method is needed to measure the battery state. Two potentially practical solutions exist:

- Absolute voltage measurement.
- Charge measurement.

2.1 Absolute Voltage Measurement

With rechargeable batteries, the variability of the discharge voltage with temperature, age and loading means that state of charge indication using cell voltage is not as simple as it might first appear. With NiCd and NiMH cells, the flatness of the voltage characteristic during a single discharge cycle further complicates use of absolute voltage. Figure 1 illustrates the problem.

2.2 Charge Measurement

Measuring the charge supplied to and withdrawn from the battery pack provides us with a direct gauge of how much energy is left in the battery. The capacity of the battery does not change with temperature, age and discharge rate to the same extent as absolute cell voltage - see Figs 2a and 2b. However for the potential accuracy of the technique to be reached, two areas of design need care:

- Allowance must be made for the charge acceptance ratio of the battery, which will normally be less than unity.
- There must be a phase in the charge-discharge cycle when the pessimistic counter is brought into alignment with the battery state. If this phase does not occur, then the counters intentional pessimism will cause it to become completely out of line with battery state over a few charge-discharge cycles.

With careful design, charge measurement is an accurate and reliable method of measuring the state of charge of rechargeable batteries. This is the method used in SAA1500T.

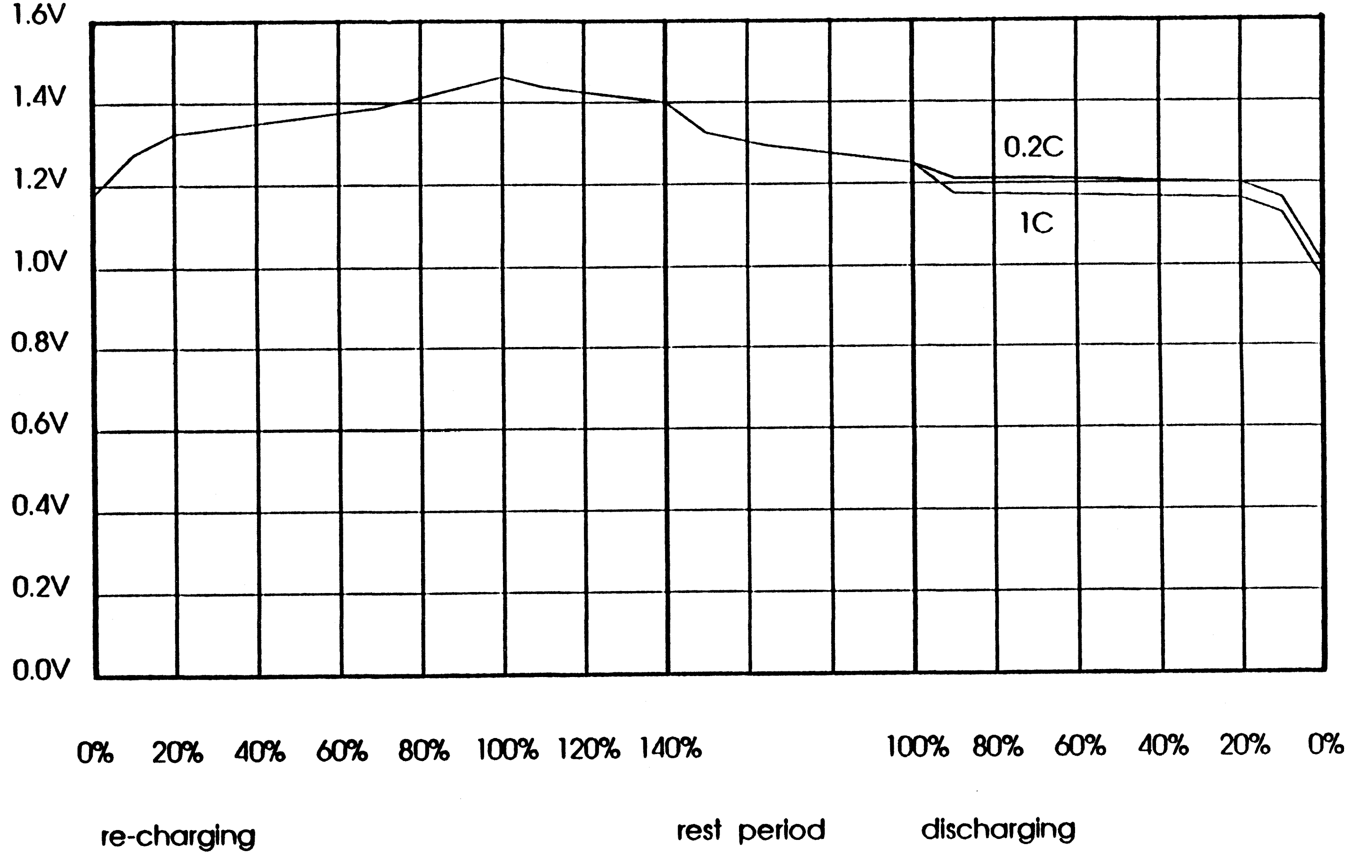


Fig 1 Typical Absolute Voltage of NiCd Cell versus State of Charge

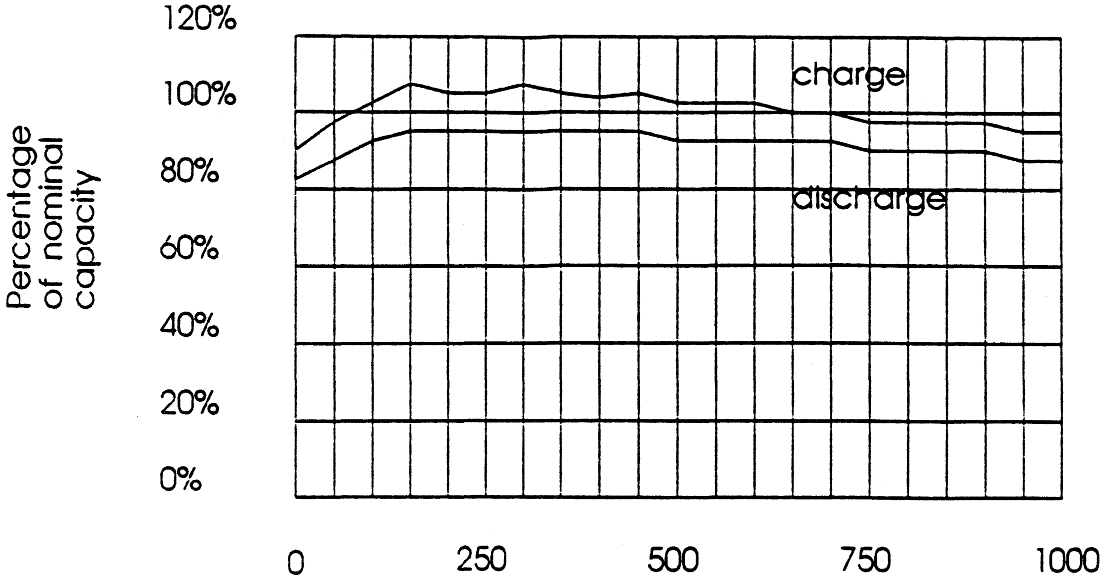


Fig 2a Typical NiCd Battery Charge/Discharge Capacity versus Number of Cycles

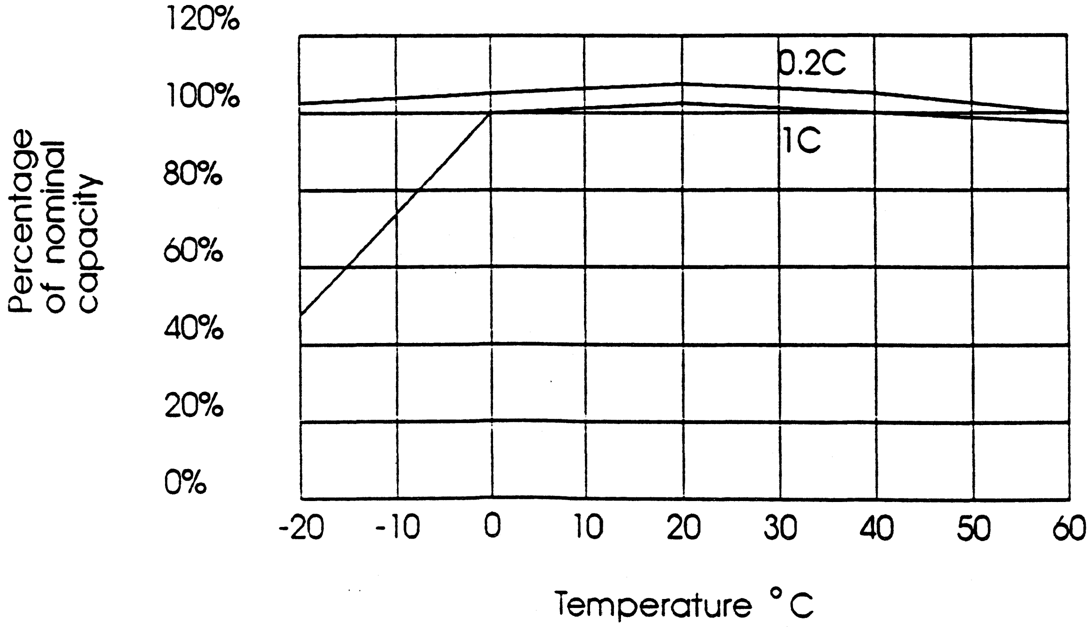


Fig 2b Typical NiCd Battery Discharge Capacity versus Temperature

3. STATE OF CHARGE INDICATOR BASED ON CHARGE COUNTING

The state of charge indicator presented here functions on the basis of counting charge (ampere hours) during discharge, and time during (fixed current) recharging and self-discharge. The circuit can be built using surface mount components to form a compact assembly permanently integrated into the battery pack.

3.1 Recharging

For the circuit design of Fig 9, the battery pack must be dedicated to a particular battery charger of known recharge current and hence recharge time. If the charge current is not known, then it is possible to vary the count rate during recharging proportionate to current using the additional circuitry shown in Appendix A.

When recharging, the rate at which the standard circuit of Fig 9 counts up corresponds to the maximum recharging time. So at the end of recharging, a short phase occurs where the battery is full, (which the battery charger may or may not detect), and the indicator shows that the battery is only partly full. In this phase the battery charger must be left connected, and the indicator allowed to continue to count up to 100%. This makes sure that when the indicator displays 100%, the battery is indeed full. This does not result in overcharging however, since the battery charger retains control over the amount of charge supplied to the battery.

Note: If the battery charger goes into a pulsed trickle charge mode, then the fixed oscillator of the SAA1500T will speed up to the rate used in self-discharge mode. This shortens the final phase of recharging where the counter is catching up with the battery state. However if pulsed trickle charge takes place for some reason other than battery full (eg: faulty battery or over-temperature), then an incorrect count will occur.

3.2 Discharge

So that an accurate gauge of battery state is kept while a variable discharge current is drawn, the count rate during discharge is made proportionate to discharge current. This is important in a number (the majority) of applications where the load placed on the battery is not constant.

The discharge counter is set up to count the number of ampere-hours which correspond with the minimum capacity of the battery. In this way, the display is never caught out by showing for example 20% full when the battery is in fact completely empty.

3.3 Self-discharge

When the battery is neither being recharged or discharged, the counter counts down at a slow rate to allow for self-discharge of the battery. This count rate is typically set for a complete discharge time of between 50 and 200 days. This feature again ensures that the indicator does not overestimate the energy left in the battery.

4. SAA1500T FUNCTIONAL DESCRIPTION

The SAA1500T is a SACMOS integrated circuit designed to provide state of charge indication for rechargeable batteries. A block diagram of the IC is shown in Fig 3, pin functions in Table 1 and electrical characteristics in Appendix A.

SAA1500T Features:

- Five segment LED or six segment LCD display of state of charge.
- LED and audio output for battery empty warning.
- LED indication of recharging.
- Discharge countdown rate proportionate to discharge current.
- Recharging, discharge and self-discharge counted to ensure accurate display.
- Output signal to control battery charger.
- Counter reset on battery connection.
- Low quiescent current (90 µA maximum) maximizes battery life.
- SO20L (SOT163.7) package.

4.1 Display Function

The SAA1500T is intended to display state of charge of the batteries in a six segment LCD bar graph or five segment LED display. Further LED indication is given for batteries nearly empty, batteries on recharge and batteries full during recharging. Outputs L100, L80, L60, L40, L20 and FULL are designed to drive a LCD bargraph with output BP connected to the backplane of the LCD. If however BP is connected to ground (Vss) then outputs L100, L80, L60, L40, L20 may directly drive LEDs with their anodes connected to the positive supply (Vcc). Outputs MO and R1 are able to drive LEDs directly, with MO indicating batteries on recharge (LED constantly lit) and batteries full while recharging (LED flashes), and R1 indicating that the battery is nearly empty. Output R2 provides a second battery nearly empty signal which can be used to drive a simple electroacoustic transducer with a 2 kHz audio tone via an external transistor. Circuit is shown in Appendix A, Fig 10.

Figures 4 and 5 chart operation of the display in recharge, discharge and self-discharge mode, and during switch-over between the modes. Figure 6 shows the signals available from outputs R1 and R2.

4.2 Counter Modes

The circuit counts cycles of a switched variable/fixed frequency oscillator in a counter which can have six user and two test modes. The mode selected is determined by the signal applied to inputs SN and PN. The SN input responds to the load state:

With the load switched OFF: SN = 1

With the load switched ON: SN = 0

The PN input responds to four different signals related solely to any external power source connected to the battery:

PN = 0 is a test mode

PN = 1 for no external power source

PN less than 14 kHz with a 'low current' supply, and PN greater than 20 kHz for a 'high current' supply.

The low and high current options can be used in special cases where for example either a mains or automotive supply is available. Usually the distinction can be ignored however, and the high current option alone used.

Combination of the two signals PN and SN produces the eight modes as shown in Table 2.

4.3 Enable Output

If the battery pack and the battery charger are contained in a single box, then it is possible to use the EN output to decide when to stop charging the batteries. When used with a LED bargraph the EN signal is high when the indicator shows not full, and provides a pulsating trickle charge waveform with 5% duty factor of ten second period when the indicator shows full. If a LCD bargraph is used, then when the indicator shows full, the EN output goes into a continuous high impedance state. See Fig 7.

4.4 Power on Reset

The Power On Reset (POR) pin responds to connection of the batteries and performs a counter reset when the POR pin is in the range of 0 to 1.8 V.

Table 1 IC Pinning

Pin No	Name	Description
1	EN	enable, controls the battery charger.
2	PN	recharge power OFF detection.
3	POR	batteries connected reset, resets counter.
4	Vcc	positive supply.
5	Ct	oscillator capacitor.
6	Ro	resistor for fixed oscillator.
7	Rc	resistor for variable oscillator.
8	Ci	current sense input (discharge).
9	SN	load switched OFF detection.
10	MO	recharge indicator.
11	R2	empty cell warning - buzzer.
12	Vss	negative supply.
13	Full	full indication after 100% charge, LCD only.
14	L100	100% charge indication, LCD and LED.
15	L80	80% charge indication, LCD and LED.
16	L60	60% charge indication, LCD and LED.

contd:

Pin No	Name	Description
17	L40	40% charge indication, LCD and LED.
18	L20	20% charge indication, LCD and LED.
19	BP	LCD backplane, LED detection (connected to Vss).
20	R1	empty cell warning - LED.

Table 2 SAA1500T Operation Modes

SAA1500T Operation Modes			
State of Inputs PN	SN	Operation Mode	Number of Oscillator Cycles for Full Count
0	0	Test, countdown fast.	5.12×10^3
<14 kHz	0	Low current supply, discharge, countdown.	8.85×10^6
>20 kHz	0	High current supply, no count.	---
1	0	No supply, discharge, countdown	8.85×10^6
0	1	Test, count up fast.	5.12×10^3
<14 kHz	1	Low current supply, recharge, count up slow.	2.36×10^8
>20 kHz	1	High current supply, recharge, count up.	7.37×10^6
1	1	Self discharge count down slow.	7.08×10^{10}
1	1	Self discharge count down slow.	3.54×10^{10} see note

Note: Self discharge counts at double normal rate for first battery discharge cycle.

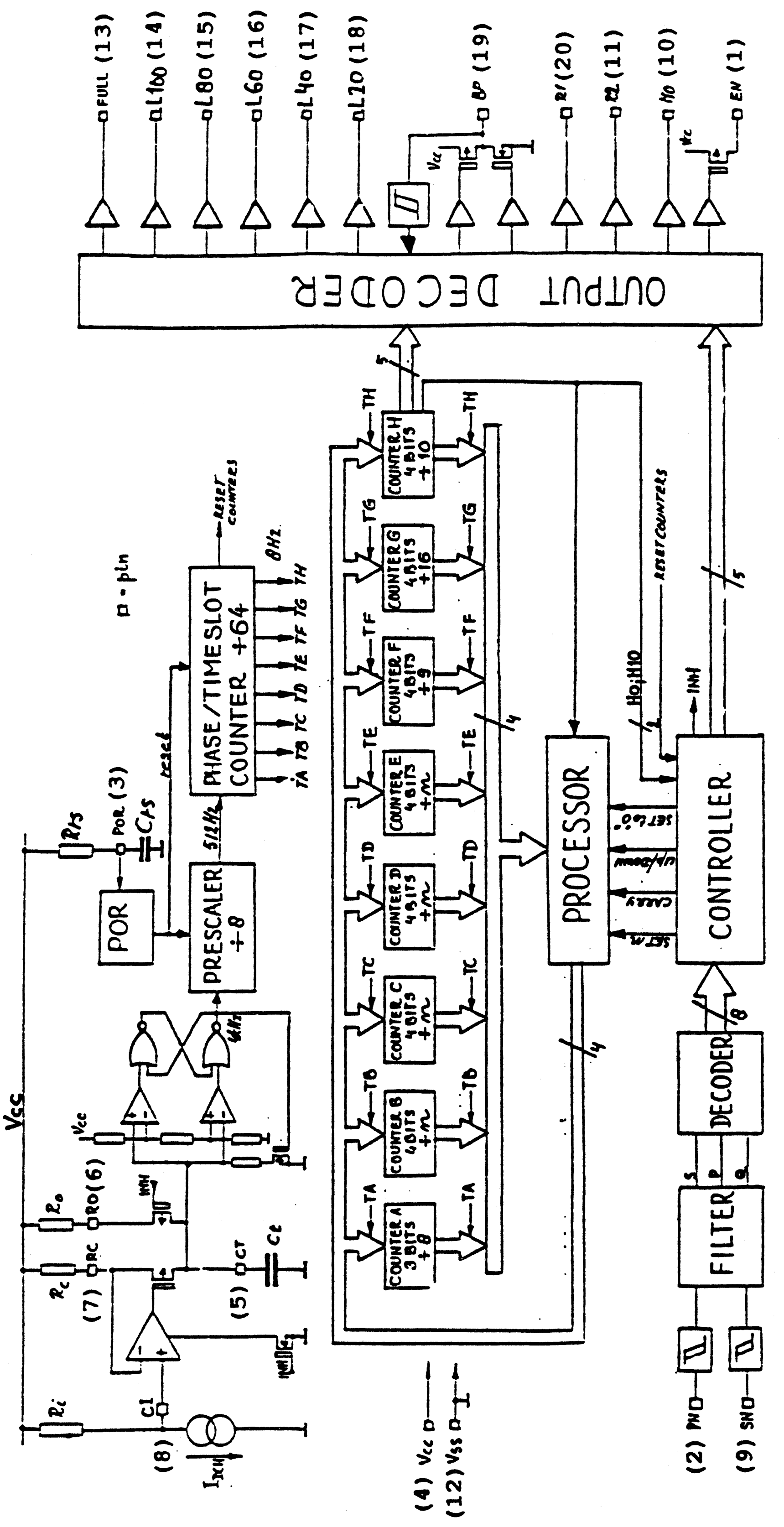


Fig 3 SAA1500T Block Diagram

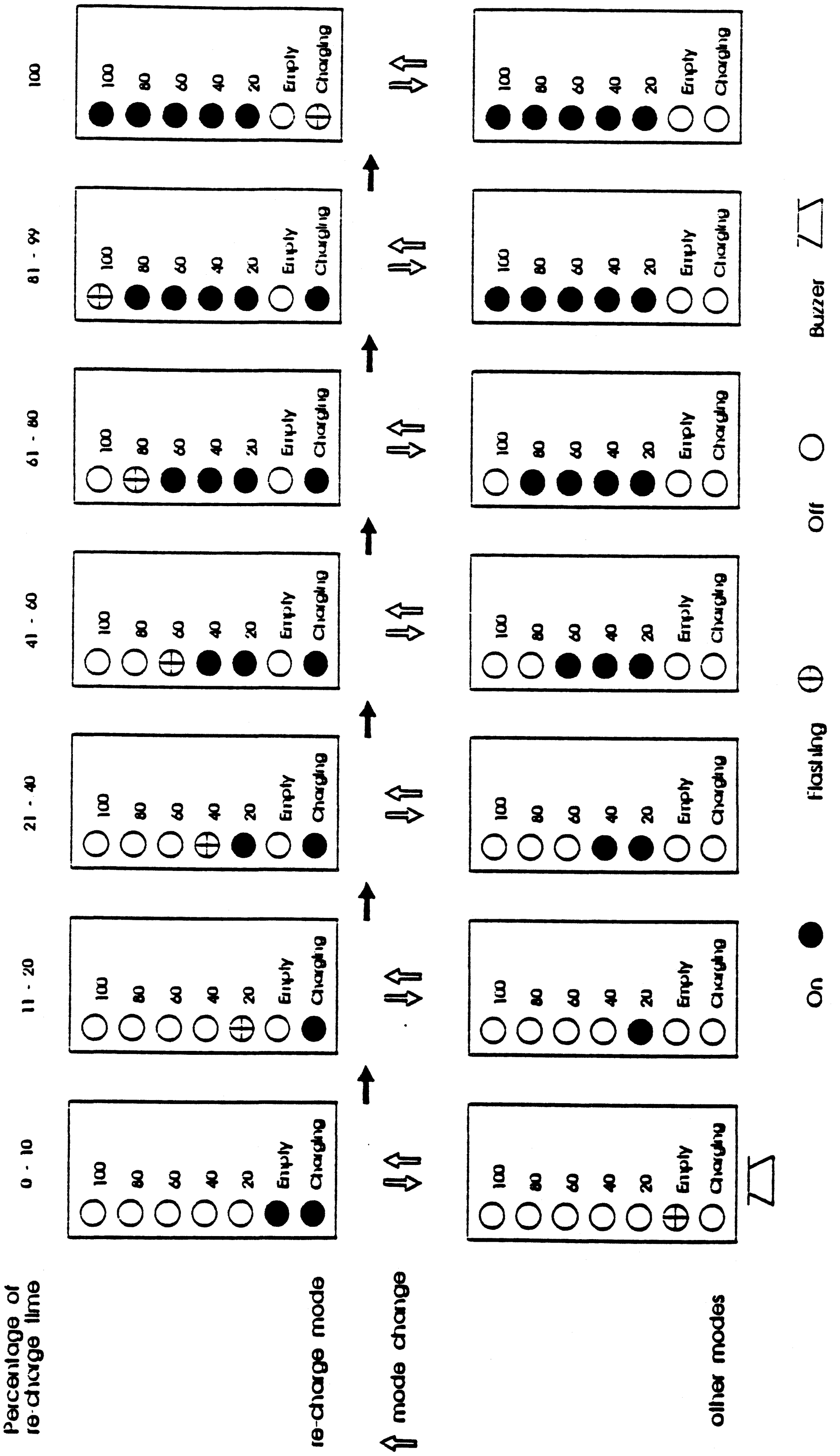


Fig 4 LED Display During Recharging

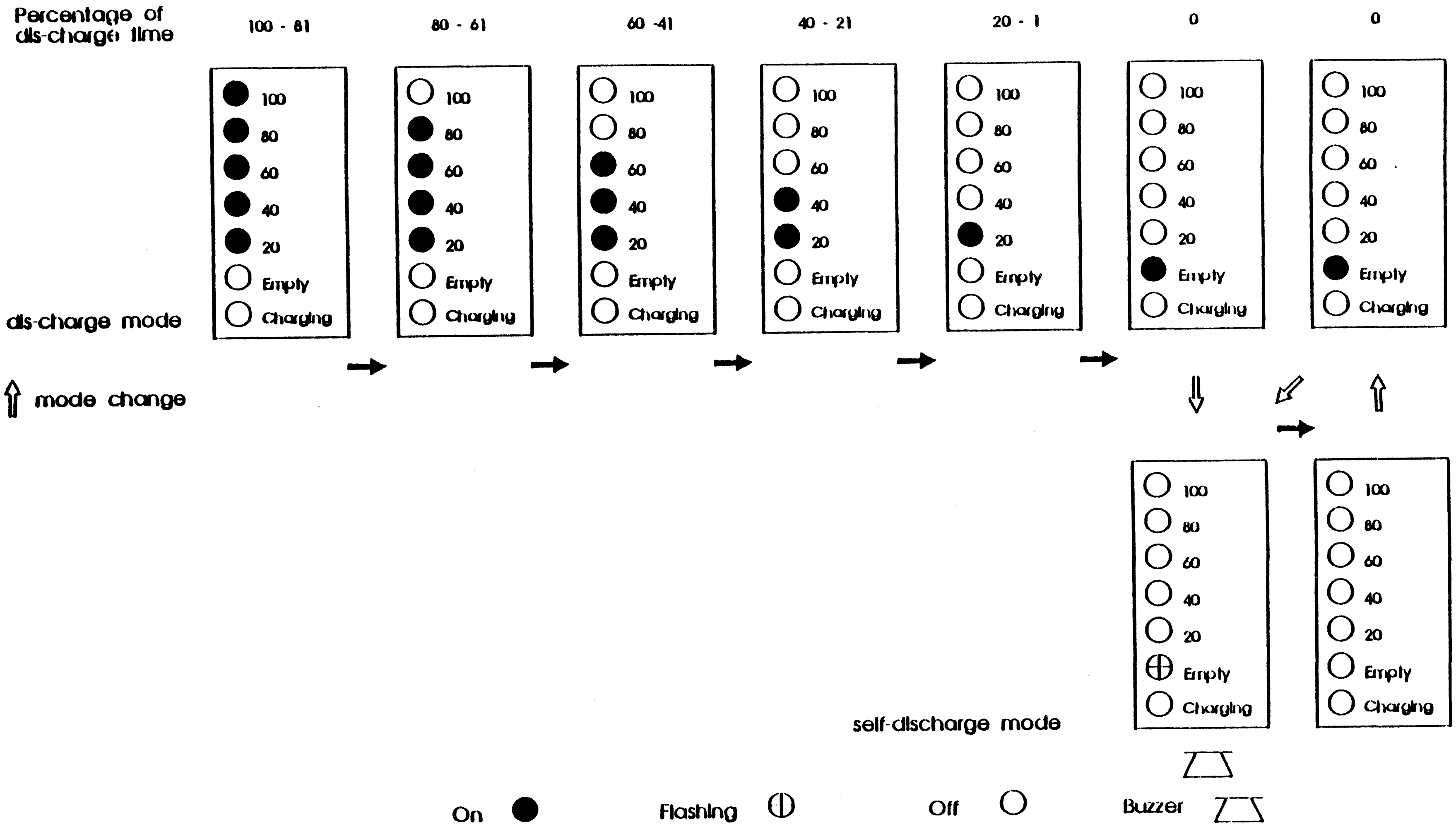


Fig 5 LED Display During Discharging

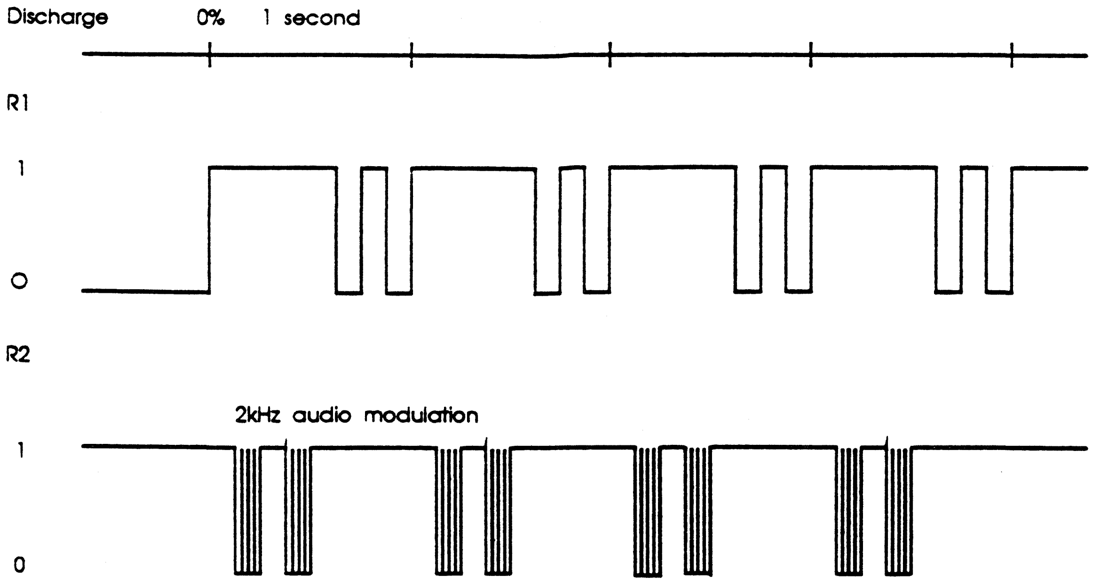
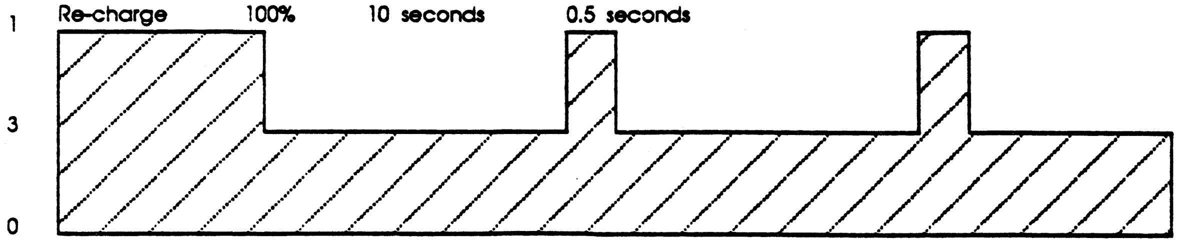


Fig 6 Waveforms of R1 and R2 Signalling Batteries Empty

0 = VSS
1 = VCC
3 = tristate - high impedance



EN signal with LED bargraph applied. With LCD bargragh the signal is continuously tristate after 100% indication.

Fig 7 Waveform of EN Signal with Full Batteries

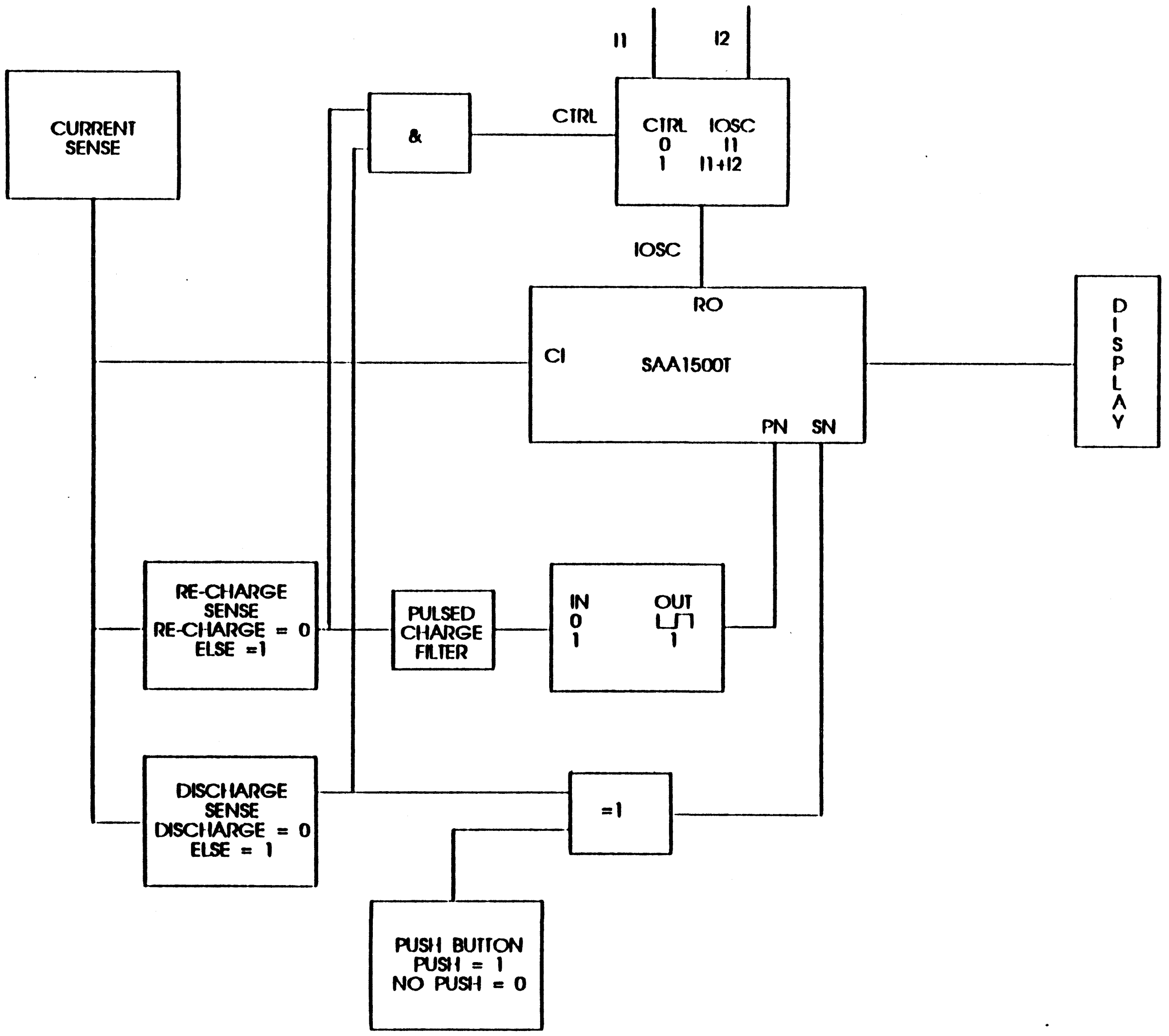


Fig 8 State Of Charge Indicator Block Diagram

5. CIRCUIT DESCRIPTION

In order to use SAA1500T as part of a battery pack with the usual two terminals, the counter control signals SN and PN must be generated inside the battery to ensure that the charge count progresses at the correct rate in the correct direction. The circuit block diagram of Fig 8 outlines how this is achieved. Figure 9 shows the complete schematic diagram.

5.1 Mode Determination

Only three of the possible eight modes are required; recharge, discharge, and self-discharge. To detect which of these modes is required, a bidirectional current sense circuit based on two op-amps is used. From the two op-amps emerge two signals which correspond to equipment not in use, and battery not being recharged. From these two lines the SN and PN signals are derived. The recharge line is buffered in such a way that pulsating trickle charge is counted in the same way as continuous charge current, but with the oscillator running at an increased rate.

5.2 Second Fixed Oscillator Frequency

To allow for a wide range of self-discharge times, a diode switch is used to control the resistance seen on the fixed oscillator input, giving two discrete frequencies of fixed oscillator. This is necessary since the numerical count for both recharge and self-discharge is fixed, and therefore the only way to vary self-discharge and fast charge time independently is to alter the fixed oscillator frequency.

5.3 Display Operation

To produce a display at the push of a button, the SN line is toggled. The user has to hold the button until a display is given (less than one second), and then release. Use of the button is not necessary during charging when the display is permanently lit. Additionally the nearly empty indicator comes on at appropriate times as a warning signal.

5.4 Supply Reduction

In order to keep within the specified operating supply range of the SAA1500T (2.0 V to 4.0 V), and yet keep at least 3.0 V applied to the peripheral circuitry, voltage regulation is required. Transistors V1 to V5 perform this function with minimal voltage drop or current drain.

6. STATE OF CHARGE INDICATOR PERFORMANCE CHARACTERISTICS

State of Charge Indicator Performance Characteristics	Min	Max
Battery voltage range.	3.0	18.0 (V)
Linear range of discharge current.	0.3	5 (C)
Quiescent current consumption.	---	600 (µA)

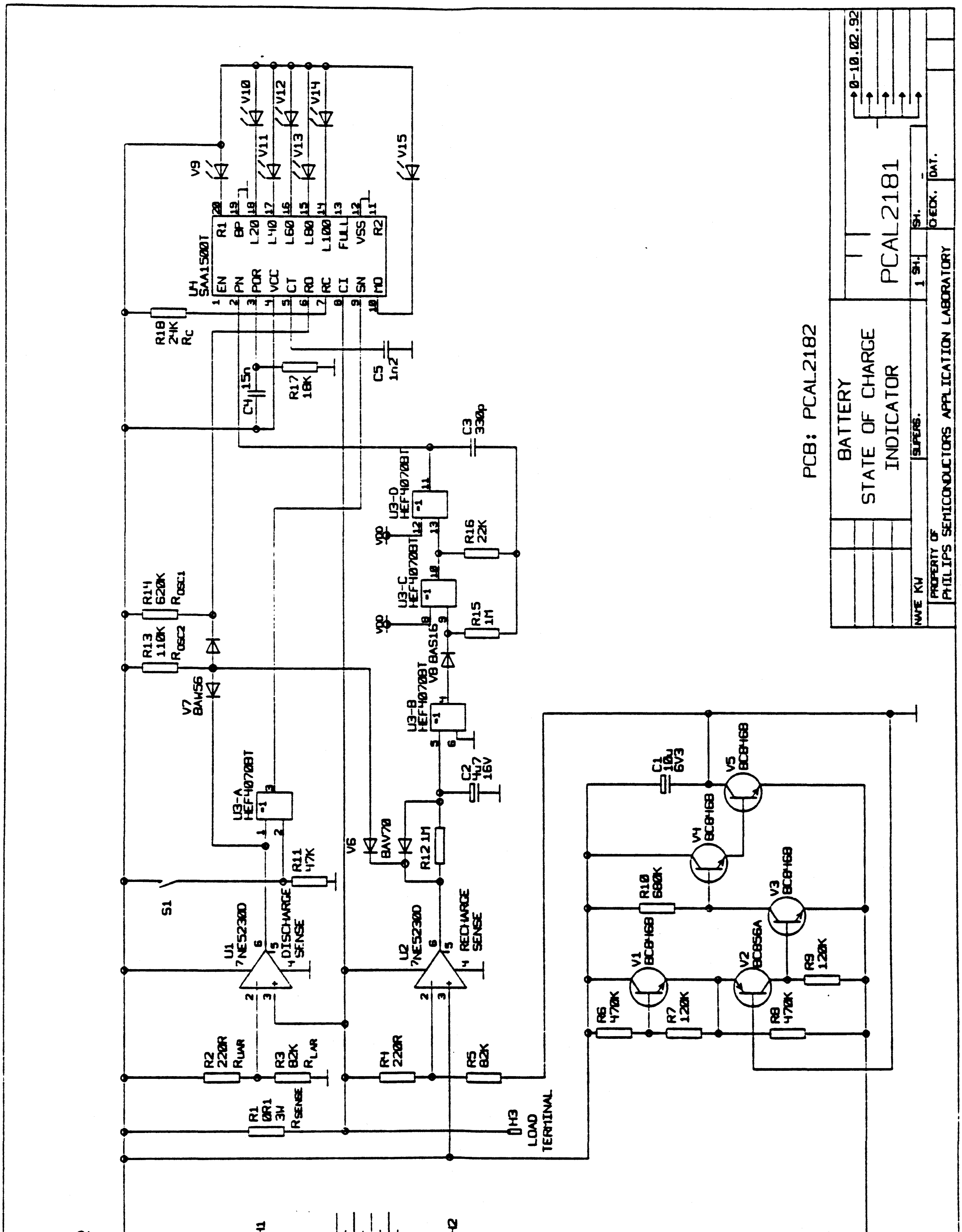


Fig 9 Battery State Of Charge Indicator Circuit Diagram

7. COMPONENT SELECTION

Selection of components and component values for various battery capacities, recharge time and self-discharge rate.

7.1 Op-Amp/Comparator Selection

The op-amps used as current detectors in this application have to work under rather awkward conditions.

- Input common mode range to include positive supply of op-amp.
- Very low quiescent current allowed (op-amps powered continuously).
- Low supply voltage.
- Logic compatible output.
- Input offset small to allow small current sense resistor and sensing of small currents.

The NE5230 is a suitable device for the system described in this report, however final selection will depend on the application.

7.2 Rsense

The value of Rsense is a compromise between a number of factors. A small value reduces power dissipation, minimises voltage drop, and retains the characteristically low impedance of NiCd batteries. Thus a low value is highly desirable. However too low a value causes a problem: Low recharge or discharge currents produce only a very small sensed voltage. This increases the cost of op-amp required to detect a given minimum current, (low offset voltage requirement) and potentially makes layout more critical.

The best value for Rsense then is the minimum consistent with being able to detect the minimum required current using the chosen op-amp.

7.3 Minimum Current Sense Setting Resistors

The two resistors Ruar (upper arm resistor) and Rlar (lower arm resistor) associated with each op-amp in the current sense circuit, set the minimum current that can be sensed in recharge (op-amp A1) and discharge (op-amp A2). When choosing these resistors the op-amps maximum offset voltage must be considered along with the minimum supply voltage and the minimum current that is required to be sensed.

The minimum sensed current Imin is:

$$Imin = \frac{Ruar\ Vsup}{(Ruar + Rlar)\ Rsense} \pm \frac{Vos}{Rsense}$$

Where Vos is the maximum op-amp offset.

Obviously the offset term must be less than the first term, or the circuit will behave very erratically!

7.4 Setting the Oscillator Frequency During Recharge - Roscl

Resistor Roscl determines the fixed oscillator frequency during recharging. This frequency should be set to give the maximum charge time that is possible with a given battery charger. Where the maximum charge time is not known, it can be calculated from the charge current as:

$$\frac{1.5 \times 60 \times Battery\ Capacity\ \ (Ah)}{Battery\ Charger\ Output\ Current\ \ (A)} \qquad (minutes)$$

$$Eqt\,1 \qquad (required)\ foscl\ (Hz) = \frac{122834}{Maximum\ Charge\ Time\ \ (minutes)}$$

$$Eqt\,2 \qquad Roscl = \frac{\left[\dfrac{1}{Foscl} - 17\ \mu s\right]}{0.7 \times Ct} \qquad (approximate\ equation)$$

7.5 Setting the Oscillator Frequency During Self-Discharge - Rosc2

Resistor Rosc2 is switched in parallel with Rosc1 to give an increased oscillator frequency during self-discharge mode. This is necessary when use of the fosc1 frequency would make the self-discharge time too slow.

$$Eqt\,3 \qquad (Required)\ Fosc2\ (Hz) = \frac{819444}{Self-discharge\ time\ Min\ (days)}$$

$$Eqt\,4 \qquad Rosc2 = \left[\frac{0.9 \times Ct}{\dfrac{1}{Fosc2} - 17\ \mu s} - \frac{1}{Rosc1} \right]^{-1}$$

approximate.

7.6 Setting the Discharge Count Down Rate - Rc

Resistor Rc determines the count rate during discharge.

$$(Required) \quad Fosc3 = \frac{3440 \times Idschg}{Ah\ capacity}$$

$$(In\ circuit) \quad Fosc3 = \frac{2.85\ Rsense \times Idschg}{Rc \times Vcc \times Ct}$$

Therefore:

$$Eqt\,5 \qquad Rc = \frac{Rsense \times Ah\ capacity}{1207 \times Vcc \times Ct}$$

8. DESIGN EXAMPLE - APPLICATION TO A 9.6 V 1500 mAh CAMCORDER BATTERY

Specification:

Battery:	Minimum capacity. Voltage. Self discharge time.	1300 mAh 9.6 V 100 days
Camcorder:	Maximum current in use. Minimum current in use. Minimum operating voltage.	3.0 A 150 mA 6.0 V
Battery Charger:	Recharge current. Maximum charge time.	2.25 A 60 mins

Step 1 Choose Rsense value

The minimum Rsense value is determined by the input voltage offset of the chosen op-amp, and the minimum current which must be sensed. If we choose NE5230 op-amps, then we have to accomodate a maximum offset voltage of 4 mV. Therefore, to sense the minimum charge current of 150 mA, the minimum possible Rsense value is:

$$\frac{2 \times 4\,mV}{150\,mA} = 54\,m\Omega$$

a value of 100 mΩ is chosen for a practical circuit to provide a further margin of safety.

Step 2 Calculate Rlar and Ruar.

To be able to sense a minimum of 150 mA, the nominal voltage reference at the op-amp input must be (allowing for offset voltage ± 4 mV and Rsense tolerance ± 10%):

$$150\,mA \times 90\,m\Omega - 4\,mV = 9.5\,mV$$

Which with a supply rail of 3.5 V requires a divider ratio of:

$$\frac{3.5\ V}{9.5\ mV} = 368$$

Thus resistor values for Ruar and Rlar of 220 Ω and 82 kΩ are appropriate. Keeping these values high reduces power consumption, but care needs to be taken with the op-amp input bias current.

Step 3 Determine Fosc1, Rosc1 (used in recharge)

$$Use\,Eqt\,1 \qquad Fosc1 = \frac{122834}{60} = 2048\,Hz$$

$$Use\,Eqt\,2 \qquad with\ Ct = 1n2$$

$$Rosc1 = \frac{\frac{1}{2048} - 17\ \mu s}{0.7 \times 1n2} = 561\ k\Omega$$

Which, choosing the next highest (conservatively slow charge rate) resistor value becomes 620 kΩ.

Step 4 Determine Fosc2, Rosc2 (used in self-discharge)

$$Use\,Eqt\,3 \qquad Fosc2 = \frac{819444}{100} = 8195\,Hz$$

$$Use\,Eqt\,4 \quad Rosc2 = \left[\frac{0.9 \times 1n2}{\frac{1}{8195} - 17\ \mu s} - \frac{1}{620\ k}\right]^{-1} = 115k3$$

Which choosing the next lowest resistor value (conservatively fast self-discharge time) becomes 110 k.

Step 5 Determine Rc (used in discharge).

$$Use\,Eqt\,5 \qquad Rc = \frac{0.1\,\Omega \times 1.3\,Ah}{1207 \times 3.5\,V \times 1n2} = 25k6$$

Which choosing the next smallest value (conservatively fast discharge time) becomes 24 kΩ.

All values are now determined, but since the oscillator equations are not exact, should be verified in circuit.

9. APPLICATIONS

The state of charge indicator presented here is designed to be part of a stand alone battery pack as used in Camcorders. However, the design is equally suited to a wide range of applications such as Laptop Computers, Personal Audio and Portable Tools, provided the SAA1500T is permanently connected to the battery pack. It should also be noted that in order to guarantee correct operation, only fully discharged batteries should be installed in the equipment.

APPENDIX A

Circuit Options

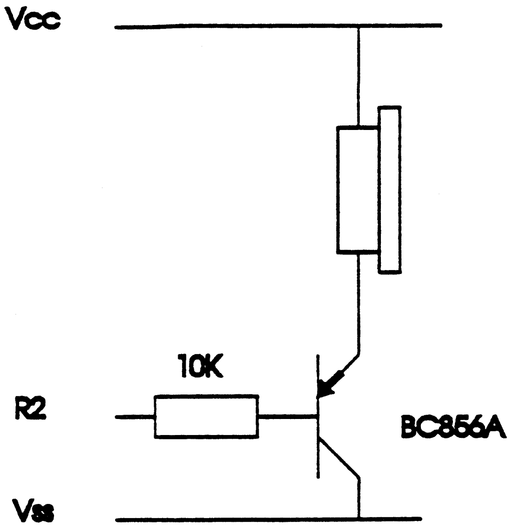


Fig 10 Audible Warning Output

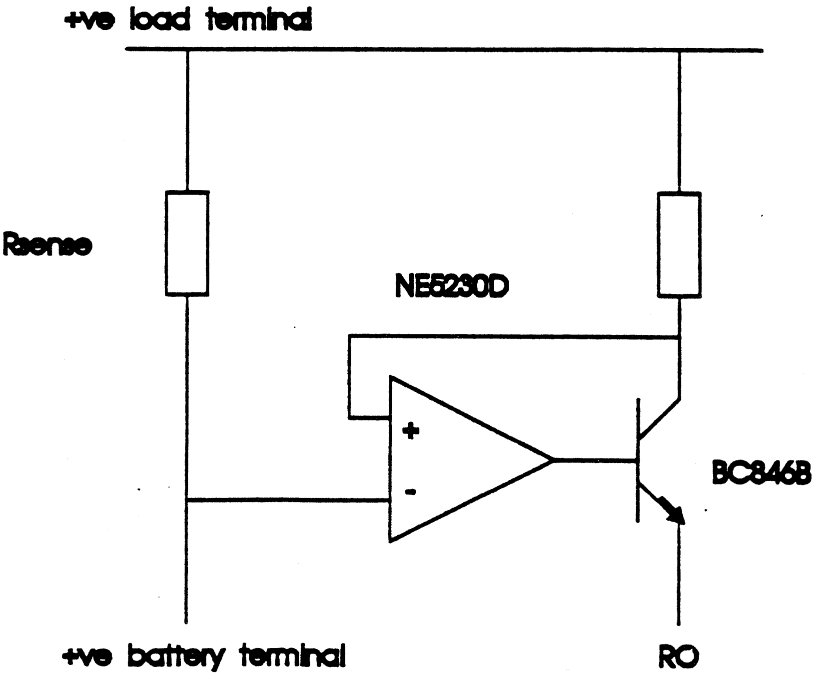


Fig 11 Recharge Current Measurement Circuit

APPENDIX B

SAA1500T Operating Characteristics

	Min	Max	Unit	Vcc	Condition
Operating voltage.	2.0	4.0	V		
Operating ambient temperature.	0	70	°C		
Output voltage HIGH outputs: BP, FULL.	2.0		V	2.4	Io = 650 µA
Output voltage HIGH outputs: R2, EN.	2.0		V	2.4	Io = 1 mA
Output voltage HIGH outputs: MO, R1, L20, L40, L60, L80, L100.	2.0		V	2.4	Io = 300 µA
Output voltage LOW outputs: BP, FULL.		0.4	V	2.4	Io = 650 µA
Output voltage HIGH outputs: R2, EN.		0.4	V	2.4	Io = 1 mA
Output voltage LOW outputs: MO, R1, L20, L40, L60, L80, L100.		0.55	V	2.4	Io = 5 mA
Schmitt trigger HIGH inputs: PN, SN, BP.	1.6 3.2		V V	2.0 4.0	
Schmitt trigger LOW inputs: PN, SN, BP.		0.4 0.8	V V	2.0 4.0	
Input voltage HIGH input POR.		1.55	V	1.8	
Input voltage LOW input: POR.	0.75		V	1.8	
Input current input: CT.	150	360	µA	2.4	VCT = 2.4 V
Supply current.		90	µA	2.6	VPOR = 0 VCI = 0 VCT = 0 VRC = 0

Absolute maximum storage temperature range: -65 to + 150 °C.

Package: S020L (SOT163.7).

APPLICATION NOTE

ICs for Battery Management

Intelligent Battery Packs
using SAA1501(T)

Report No: NPO/AN9401.

Guido B Banning

Customer & Product Support Group Consumer ICs Nijmegen, Power Conversion Section,
the Netherlands.

Keywords
State-Of-Charge-Indication
Intelligent Battery Packs
Coulomb Counting Principle
Charge/Discharge Compensation
LED and LCD read-out

Date : 06 April 1994
Pages: 39

Abstract:

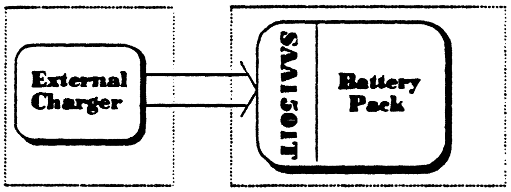
SAA1501T is a low-peripheral battery management circuit to be used in fast charging systems. The most important function of SAA1501T is monitoring the charge account in rechargeable batteries. During battery charging the charge current and charge time are registered in a coulomb counter, whereas during discharging the discharge current and time are recorded. The momentary charge account of the batteries can be displayed either on an LCDisplay or by an LED bargraph.

SAA1501T can be used for three application areas:
1) For intelligent battery pack applications. In order to get an intelligent battery pack, the state-of-charge indicator circuitry is incorporated into the battery pack. Charge control is carried out by the charger which is outside the battery pack (figure 1.a). Typical applications include Camcorder, tools and mobile telecommunications.

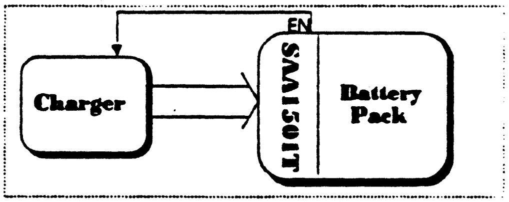
2) For applications with a built-in charger. Then SAA1501T carries out the charge control by means of a current control signal and thus effectively masters the charger unit (figure 1.b).

3) For applications with an external charger which is yet mastered by SAA1501T by means of an additional connection for the current control signal of SAA1501T to the charger unit.

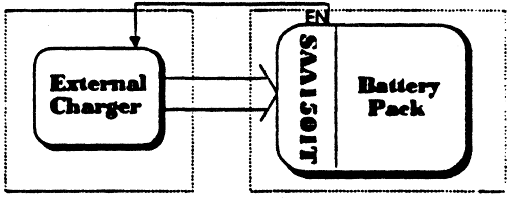
This application note deals with some basic applications for intelligent battery packs only (application area 1). A subsequent application note could be dedicated to the 'charge current regulation' function (application areas 2 and 3).



Intelligent Battery Pack Application
(Figure 1.a)



Application with built-in charger
mastered by SAA1501T
(Figure 1.b)



Application with external charger
mastered by SAA1501T
(Figure 1.c)

Figure 1: possible application areas with SAA1501T.

Table of Contents:

Appendices:

1. Introduction to state-of-charge indication

People tend to recharge batteries many times more often than actually necessary, simply because they are afraid of being caught out with empty batteries. This could result in early degradation of battery performance and lifetime. Besides, this can be a waste of time and effort. It can be even worse, when for instance you attempt to record a friend's first band-concert event with your camcorder of which - you think - the batteries are fully charged. However, after having been able to happily record the first part of the performance the battery goes flat just before your friend starts his solo.

Therefore an intelligent battery equipped with a battery state-of-charge-indicator function appears to be indispensible, whereby the function can be regarded as a fuel gauge in a car. The requirements of the indicator are that of an easy to read indication of the available battery energy and a battery low indication prior to when the batteries are really empty.

With SAA1501T we can offer the ideal compromise between prize and flexibility. The SAA1501T uses a highly accurate 'charge-sensing method' and enables charge/discharge efficiency compensation.
The application using SAA1501T only requires a few peripherals.

2. Battery measurement strategies

In order to determine the available battery energy two methods can be considered. Below, these two methods are described together with some elaboration on battery theory.

2.1 Absolute Battery voltage measurement

When considering the battery voltage as a function of charge time of NiCd and NiMH batteries (dashed line in figure 2) it may be concluded that it is possible to use battery voltage measurement to obtain an indication about the status of the battery capacity (state-of-charge indication). However, in case of discharging the batteries, with NiCd and NiMH batteries the flatness of the cell voltage during a single discharge cycle makes the 'absolute battery voltage measurement' method completely inappropriate.
Besides, this method would fail at all because of the dependence of the charge and discharge voltage with temperature, re-charging history and discharge rate.

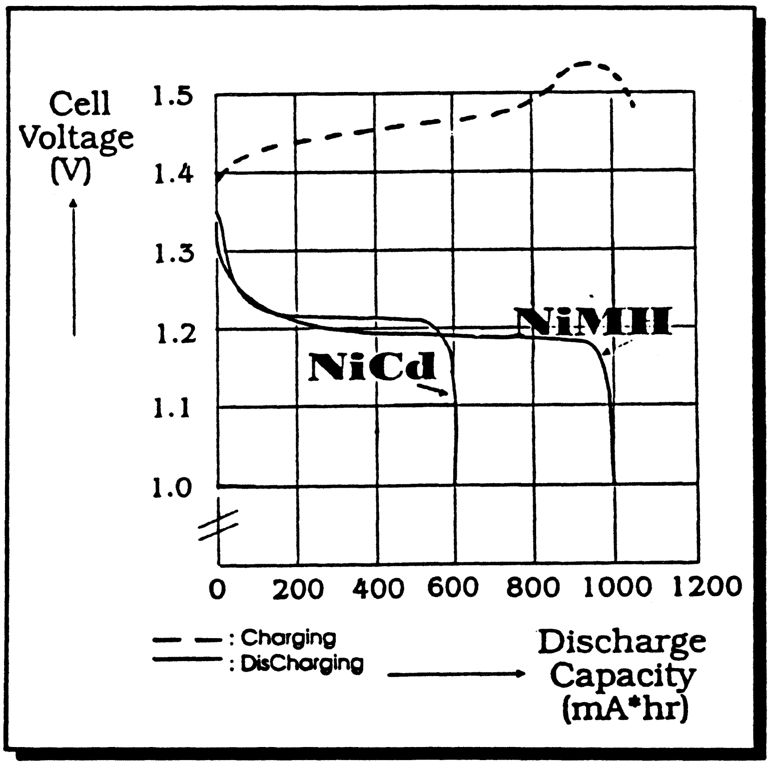


Figure 2: figure of nominal discharge characteristics.

2.2 Charge/discharge current measurement

Measuring the charge and discharge currents together with time gives us an accurate indication of how much energy is in the batteries. However, since the charge acceptance (charge efficiency) and available discharge capacity (discharge efficiency) decreases with charge and discharge rate respectively (there are no batteries with 100% efficiency), some allowance must be incalculated to prevent batteries being empty prior to battery low indication.
Further, some allowance must be incalculated due to battery temperature variation, which also influences charge and discharge efficiency.
SAA1501T uses Charge/DisCharge current measurement and enables charge and discharge efficiency to be set independently.

3. SAA1501T and its key features

3.1 Introduction SAA1501T

SAA1501T is a monolithic integrated circuit designed to provide state-of-charge-indication for rechargeable NiCd and NiMH batteries.

SAA1501T is highly appropriate for use in battery pack applications, as only a few peripherals are required and thus the overall application is extremely compact. A block diagram of a battery pack application using SAA1501T is shown in appendix B.

The circuit is processed in advanced BICMOS technology, wherein the benefits of mixed bipolar and CMOS components are fully utilized and due to this combination high-accuracy is guaranteed.
Since SAA1501T measures charge and discharge currents continuously, hence 'the charge sensing-method' of SAA1501T is suitable for detecting pulsating charge and discharge currents (GSM, camcorders and so on).

3.2 Key features of SAA1501T

- **High-accuracy charge/discharge current measurements.**
- **Large dynamic range of both charge and discharge currents.**
- **Independent settings of charge and discharge efficiency.**
- **Low power dissipation, 90μA quiescent current maximizes battery life.**
- **Few peripherals for easy integration in battery pack.**
- Appropriate for pulsating charge and discharge currents.
- Three charging levels, assuring batteries are fully charged (requires use of charge current regulation signal of SAA1501T (EN-signal) to be connected to the charger unit).
- Temperature-controlled battery self-discharge setting.
- Maximum battery temperature protection via absolute battery temperature measurement (adjustable Tmax).
- Two charge account display modes, six segment LCD or LED.
- LED and audio output for battery empty warning.
- Suitable for all NiMH and NiCd batteries and for each size of battery capacity.
- On-board regulator for controlling an external charger (EN-signal).

4 Functional description SAA1501T

This section explains the functional blocks of SAA1501T. The main system blocks of SAA1501T consist of the supply block, the U/I charge block, the U/I discharge block, the mode recognition block, the temperature block, the U/D Counter block, the display decoder driver block and regulation control block. A block diagram of the IC is shown in appendix C. Refer to the specification data of SAA1501T for the electrical characteristics.

For the terminology of the IC's input and output pins, table 1 can be referred to.

Pinnumber	Pin Name	Desciption
1	Vcc	Supply Voltage
2	EN	Enable
3	Cd	Duty cycle capacitor
4	Ccnt	Charge reservoir capacitor
5	Rmax	Maximum average charge
6	Rref	Current reference resistor
7	Rd	Discharge current conversion resistor
8	Rc	Charge current conversion resistor
9	GNDs	Charge sense input
10	Rsd	Discharge sense input
11	TEMP1	Temperature sensing resistor
12	TEMP2	Temperature setting resistor
13	Co	Oscillator capacitor
14	BUZ	Buzzer (audible BLI)
15	FULL	Full battery indication
16	100	100% segment indication
17	80	80% segment indication
18	60	60% segment indication
19	40	40% segment indication
20	20	20% segment indication
21	BP	LCD back plane drive
22	BLI	Battery Low Indication
23	PO	Power (Charge) On Indication
24	GNDp	Power ground

Table 1: SAA1501T I/Ps and O/Ps

4.1 The supply and bandgap generator block

The supply voltage of SAA1501T ranges from 2V to 5.5V. As the batteries are connected to SAA1501T (Vcc rises from 0 to supply voltage), the digital circuitry is reset (at 1.7V). As SAA1501T monitors 2 cells of the battery pack the system will reset to zero when each cell has a minimum voltage of 0.85V.

The bandgap generator block produces high-accuracy voltages for the oscillator and for the several reference currents which are being used in the U/D Counter block and the Charge current regulation block.

4.2 Charge/Discharge current sensing blocks

The charge and discharge currents are sensed across a sense resistor (Rsense) in series with the battery cells. The resulting sense voltages are applied to U/I convertors, which produce IC-acceptable down-converted charge and discharge currents. Refer to the figure below for the synthesis of the U/I convertors interfaced to the sense resistor.

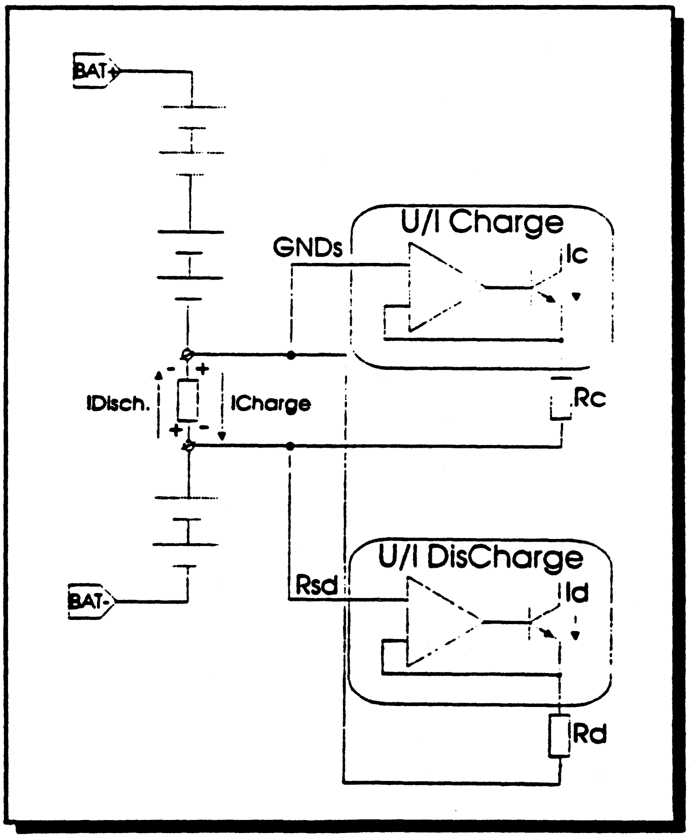


Figure 3: Charge and DisCharge U/I convertors.

4.3 Mode recognition block

For the intelligent battery pack application SAA1501T recognizes three modes of operation, viz. the Charge mode, the Discharge mode and StandBy mode (Self-discharge mode). The mode recognition block differentiates between the available modes of operation.
Refer to appendix D for an operation overview of SAA1501T (flow chart SAA1501T).

4.4 The U/D Counter Block

This block is a true coulomb counting circuit which processes the measured net charge and discharge currents and registrated charge and discharge times representing the available battery energy ($Q=I*t$). The output drives a two stage counter either up or down depending on whether charging or discharging takes place.

4.4.1 Charge account during the Charge Mode

In case a charge current is sensed across the sense resistor (positive voltage is sensed at input UGNDs) and the batteries are not loaded (no discharge current, so that voltage at Rsd=0V), the SAA1501T will set to Charge Mode. Charge Mode is recognized if the voltage at UGNDs > 3mV, which is called the Idle-level.
The sensed voltage is converted via Rconv to an IC acceptable charge current Iconv. A part of this converted current (1/6*Iconv) will be used for charging an external capacitor Ccnt. A fraction of Iconv is used, so that Ccnt can be small. The voltage on this capacitor will rise until a specified level is exceeded (Ucnt > Uh, refer to specification). On the first rising edge of the internal clock signal a fixed current Iref1 will be distracted from 1/6*Iconv (Icnt) during a defined time N*Tclk, in which N is the number of clock cycles Iref1 is subtracted and thus coulombs are counted. This will result in a voltage decrease at Ccnt in case 1/6*Iconv < Iref1, whereas if Icnt = Iref1, the voltage at Ccnt will not change. During the subtraction of Iref1 coulombs are counted ($Q = I*t$) during a defined time. Subsequently, after one clock period Iref1 is switched off, so that the voltage at Ccnt increases again, and so on.
Generally, the principle described above is called the 'coulomb counting principle', using a single ADC; refer to figure 4.

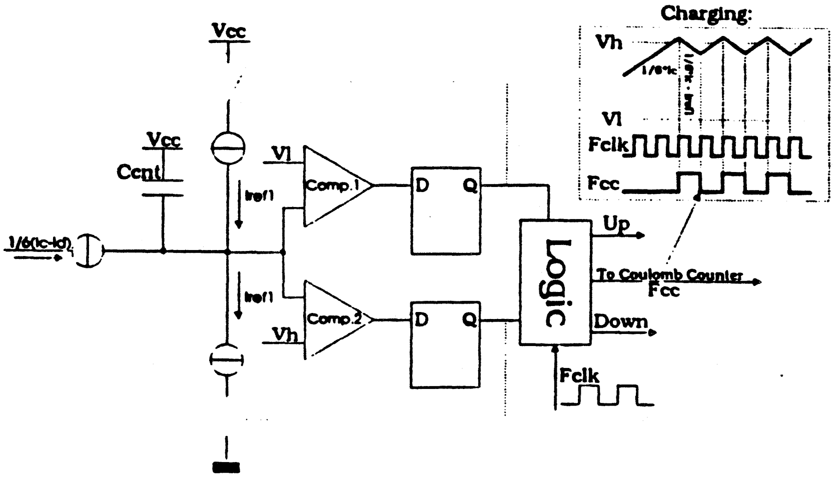


Figure 4: 'Coulomb Counting principle'.

4.4.2 Discharge account during the Discharge Mode

In case a discharge current (batteries are being loaded) is sensed across the sense resistor (positive voltage at Rsd), SAA1501T will set to Discharge Mode (Idle-level to detect discharge mode is also 3mV). The same principle of operation as described in the previous section applies for the Discharge Mode. However, in Discharge Mode Iref1 is added to -1/6Iconv in case Ucnt < Ul (refer to specification).

4.4.3 StandBy Mode

In case neither a charge current nor a discharge current is sensed, SAA1501T is in StandBy Mode. In this mode the counters will countdown in 200 days (Fosc=4kHz) to overcome the Self-discharge of the batteries. In addition, two absolute temperature measurements are carried out in StandBy mode, which enable setting of the countdown rate of the counters as a function of TBattery (refer to section 6.7). This function can be set optionally. The countdown rate is set as a function of TBattery in StandBy mode, otherwise large errors could be introduced.

The quiescent current consumption in StandBy mode is only 90μA. This is a negligible small current compared to the Self-discharge current of the batteries.

4.5 The display functions

With SAA1501T the available battery capacity can be output via six segment LEDs or on a six segment LC Display.

The energy gauge outputs consist of outputs FULL, L100 (100%), L80 (80%), L60 (60%), L40 (40%) and L20 (20%). Further, batteries nearly empty (battery low indication, indicated by flashing LED) and batteries being charged (Charge Mode) are indicated by two additional LEDs, B(attery)L(ow)I(ndication) and P(ower)O(n) respectively. Output BUZZ(er) produces a second battery nearly empty signal which can be used to drive a simple electroacoustic transducer via an external transistor (see figure 5).

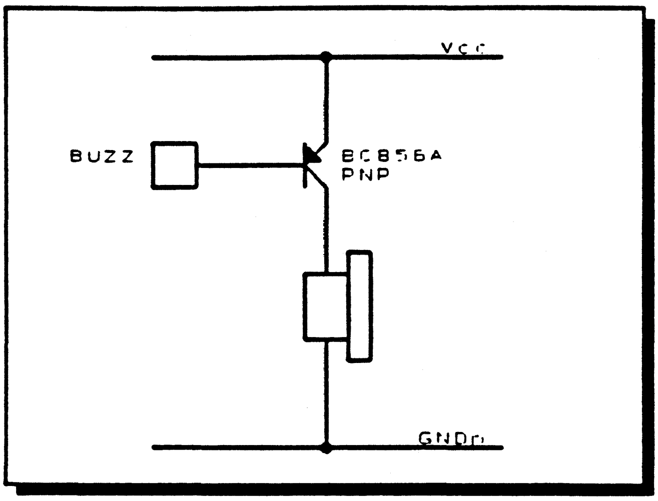


Figure 5: Audible 'Batteries Nearly Empty' Output

The B(ack)P(lane) pin is used to differentiate between LCD and LED modes. LED mode requires that BP is connected to ground, whereas LCD mode requires BP is connected to the backplane of the LCD bar.

In LED mode the energy gauge outputs are made visible by switching on constant current sources (10mA). In LCD mode BP generates a 64Hz block signal and the energy gauge outputs are made visible by producing block signals which are in anti-phase with the signal at BP.

When using LEDs for displaying the available battery capacity the LEDs are only continuously on in the Charge mode, whereas when using an LC Display the display is on in all modes (Charge mode, Discharge mode and StandBy mode). To be able to produce a state-of-charge indication when using LEDs in the StandBy mode, a mode change must be forced manually. In the battery pack application using LEDs this is carried out by pressing S1 (e.g. see appendix E), which thus forces a mode change from the StandBy mode to Discharge mode. Then. the LEDs will be on for 8 seconds.

The values of the voltage dividing resistors R1 and R2 have been chosen so that the input transistors (PNP) of the internal U/I convertors are not completely turned off. Resistor R1 of 10k introduces an offset voltage of 0.1mV only at the Rsd input of the discharge U/I convertor, since the bias currents of the inputs of the U/I convertors are 10nA.

4.6 Charge current regulation

This block produces a charge current regulation signal, to be used for controlling an external current power supply, such as the TEA1400. As this report is specifically meant for Battery Pack applications, the charge current regulation block is not utilized with the battery pack application. A follow-up application note on SAA1501T could be dedicated to applications using charge current regulation, e.g. desktop charger applications.

4.7 The oscillator and prescaler/controller

The oscillator is a relaxation type which drives the prescaler/controller. The prescaler/controller provides all system timings, which all are related to the oscillator frequency (Fosc). When running at a speed of 4kHz (Fosc), the prescaler/controller takes care of derivating signals of 2kHz for the buzzer (BUZZ), 64Hz for the LCD backplane, pulse trains for the temperature block and power sensing measurements, 2Hz signal for dynamic charge/discharge indication, Self-discharge timing-signals, and a signal with a duration of 8 seconds for read-out purpose of the LEDs as was already mentioned in section 4.5.

4.8 Temperature block

This block includes two functions. One function is to terminate fast charging when battery temperature exceeds a certain maximum temperature level. At this temperature TFC (Fast Charge Turnover point) temperature protection is activated which means that the external charger is requested to switch over to slow charging via a current regulation signal of SAA1501T. This is necessary since fast charging at high battery temperature (for instance 60°C) will damage the battery and thus degrade battery life and performance.

The other function is to adjust the Self-discharge countdown of SAA1501T as a function of temperature. This is necessary since the Self-discharge rate of the batteries is temperature dependent, i.e. the Self-discharge rate of the batteries increases with battery temperature.

The actual battery temperature is measured every 16 seconds (Fosc=4kHz) at pins Temp1 and Temp2 by use of an NTC and some setting resistors (all external). The number of resistors used is optional (2 up to 4). The NTC should be connected close to the batteries. During Charging (Charge Mode) one absolute temperature measurement (TFC) is carried out, whereas in StandBy Mode (Self-discharge Mode) two absolute temperature (TSB1 and TSB2) measurements are carried out. These three absolute temperature turnover points (TFC, TSB1 and TSB2) can be set independently.

It should be noted that TFC is not used in the battery pack application, as the charge intelligence is left to the external charger unit.
The StandBy turnover points TSB1 and TSB2 are directly related to TFC. Refer to section 6.7 how to set the three absolute temperature turnover points.

5 Regarding the SAA1501T system accuracies

5.1 Accuracy of battery characteristics

In the application examples given in this report, it is first tried to fit the batteries into a fixed model, at room temperature and nominal (0.1CA) charge/discharge rates. In here, charge and discharge efficiency of the batteries are taken into account by setting 'average efficiency factors'. These factors are determined by regarding the expected temperature operating range and Crate ranges of the batteries. Thus to assure that the batteries are not empty prior to battery low indication, e.g. the discharge efficiency is set to 70%, which means that 70% of the Standard battery capacity is available by indication during discharge (under-estimated indication). It should be noted that the 'efficiency factors' highly depend on the type of battery used.

The main error between the fixed model (above mentioned) and the perfectly compensated model for temperature and charge/discharge rates, determines the accuracy of the indicator.

Even with a perfectly compensated model the accuracy depends on the accuracies of the characteristics from the battery manufacturer. It is known from literature that the best accuracy with which the characteristics of the rechargeable battery can be attained for typical consumer use is about ±5%.

5.2 SAA1501T performance characteristics

The accuracy of the state-of-charge-indicator function of SAA1501T depends on the accuracy of the 'Coulomb counting' principle, i.e. the accuracy of the current conversion during an accurate time. These two 'settings' are defined by the accuracies of the oscillator frequency and the sense resistors together with the U/I convertor and the conversion resistors.

The error factor of the U/I convertor plays the greatest part in the overall accuracy of the 'coulomb counting principle'. SAA1501T features a wide dynamic range, though in order to achieve best accuracy it should be tried to operate SAA1501T at higher sense voltages (100mV and higher). The typical overall accuracy of the 'coulomb counting principle' is < 2%.

As an aside, it should be noted that the accuracy of the 'coulomb counting principle' is independent of the accuracy of the capacitor at Ccnt (charge counting capacitor). However, use of a low leakage capacitor is required.

6. External components selection

6.1 Rref and converted charge/discharge current range

With Rref the value of Iref is defined by 207mV/Rref wherein 207mV is a fixed (bandgap) voltage at Rref (pin 6). From Iref many bias currents are derived for use in most of the blocks in SAA1501T. Therefore the value of Rref cannot be freely chosen.
For maximum accuracy Iref must be chosen between 3.5μA and 8μA.

The sense voltages at the sense inputs GNDs (pin 9) and Rsd (pin 10) are converted to IC-acceptable currents (refer to section 4.3), i.e. during charging a converted current Ic is produced, whereas during discharging a converted current Id is produced.
The operating range of Ic(Id) as a function of Iref can be written as:

$$\frac{6*2.5*Iref}{100}\mu A \ (= Idle\ level) \leq Ic(Id) \leq 6*2.5*Iref\mu A \ [DC\ dynamic\ range=100] \qquad [1]$$

Formula [1]: the allowed DC range of operation for Ic/Id as function of Iref.

Thus for a nominal Iref=4μA (Rref=51kΩ) Ic(Id) should be in the range from 0.6μA to 60μA for best accuracy. In more detail, <u>then</u> Icmin(Idmin)=0.6μA since below 0.6μA StandBy mode (Self-discharge mode) is recognized, whereas <u>then</u> Icmax(Idmax)=60μA since for currents beyond this level the single bit AD convertor will fail (saturate). It should be noted that this applies for DC behaviour only.

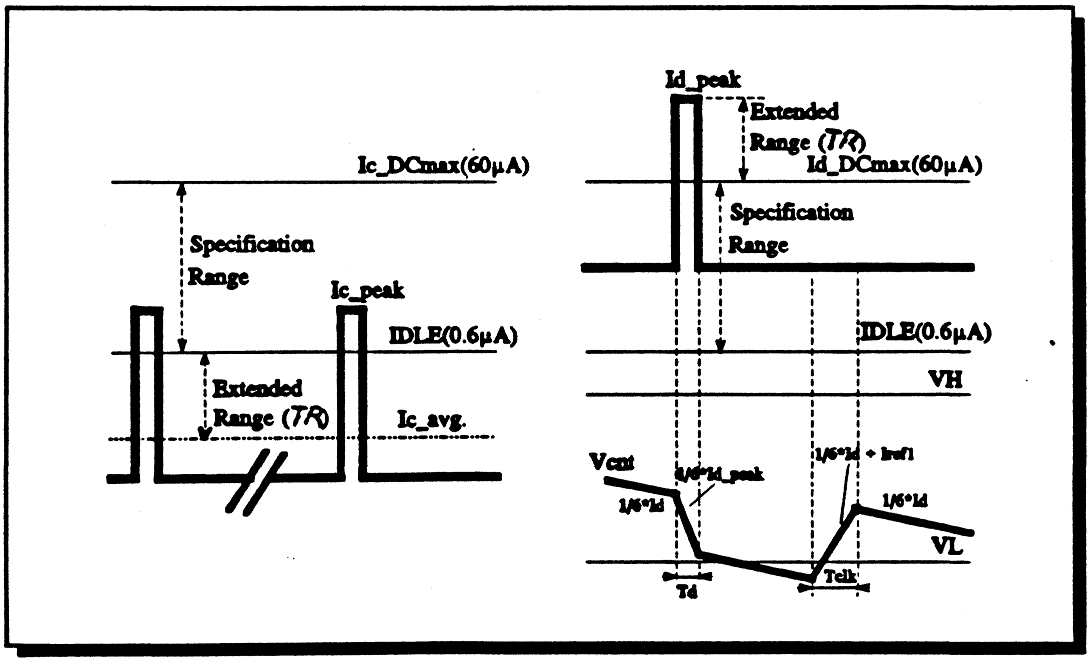


Figure 6: extended current range concerning TR behaviour.

In the previous paragraphs it is said that the dynamic range is a factor 100 (1:100). This applies for DC behaviour only. In case of a pulsated charge/discharge signal (see figure 5) with duty cycle δ and Icpeak(Idpeak) >= 0.6µA (Iref=4µA), the dynamic average charge/discharge range can be raised by $1/\delta$ (the lower range value). Also the upper range value can be raised because of the integration function of capacitor Ccnt. Id_{peak} must ensure that $Td*Id_{peak} <= Iref1*Tclk$ or $\Delta V < (Iref1*Tclk)/Ccnt$; refer to figure 6.
However, the higher the peak currents the more accuracy decreases and therefore it is recommended not to exceed the upper range value by a factor 2 (usually 120µA).
In short it can be said that concerning transient (TR) behaviour the dynamic range can be larger than 1000 times.

Note that the bandgap generator outputs Uref and Umax (pins Rref and Rmax) are sensitive to capacitive load. It is therefore recommended to connect the external resistors Rref and Rmax as close as possible to the IC pins.

6.2 Rsense

Three criteria determine how to choose Rsense:
1) Minimum sense voltage:
The minimum Rsense value must be chosen so that the voltage drop across Rsense does not fall below 4mV: the minimum sense voltage for which accuracy of the U/I convertors is still guaranteed (Rsense_min >= 4mV/Imin_charge(Imin_discharge)). In other words, in case it is tried to design-in for Rsense-values which lead to sense voltages smaller than 4mV at the specified charge/discharge currents, higher Rsense-values should be used.

2) Maximum sense voltage:
The maximum allowable sense voltage is limited by the IC supply (battery) voltage minus 1.6V (Ucc - 1.6V). During charging the supply (battery) voltage will be more than 2.5V with two battery cells. Therefore, sense voltages up to 900mV are feasible without any loss of accuracy.

During discharging however the minimum battery voltage of 2 cells can be as low as 2V. The maximum accuracy is now guaranteed for sense voltages up to 400mV.

3) The choice of Rsense also depends on the power rating ($Psense = Isense^2*Rsense$). It should be observed carefully, that Psense does not become unnecessarilly high.

6.3 The oscillator frequency

The basic intelligent battery pack application using SAA1501T is meant to operate at an internal clock frequency of 62.5Hz (Tclk=Fosc/64), which therefore means that the oscillator frequency at Co must be set to 4kHz (Fosc). Then the counter-synthesis of SAA1501T is able to track charge as well as discharge rates in the range from 0.05C up to 5C, provided that the converted charge currents (Ic) and discharge currents (Id) are in the range from 0.6μA (0.15*Iref) up to 60μA (15*Iref).

The following equation gives an idea on the range of charge and discharge rates as a function of the oscillator frequency:

$$Crate=\frac{Ic(Id)}{12*4*10^3*Tosc},\quad 0.6<=Ic(Id)<=60 \text{ when } Iref=4\mu A,\ Ic(Id)[\mu A] \qquad [2]$$

Formula [2]: Charge and discharge rates as function of Ic(Id) and Fosc.

Thus in order to be able to track higher charge and discharge rates, the oscillator frequency at Co must be increased by decreasing the capacitor at pin Co. For instance, an oscillator frequency of 40kHz would enable SAA1501T to track for charge and discharge rates from 0.5C to 50C. However, increasing the oscillator frequency will speed up the 100/200days countdown. Therefore some external circuitry is required to switch over to a lower oscillator frequency in StandBy (self-discharge) mode.

	Iref = Vref/Rref = 4μA (Rref = 51kΩ)							
	Fosc = 1/(5.6*Rref*Co)							
Fosc[kHz]	4		8		16		40	
Ic(Id)[μA]	12	60	6	60	6	60	0.6	60
Crate (CA)	1	5	1	10	2	20	0.5	50

Table 2: Ic(Id) values and the corresponding Crates as function of Fosc.

Fosc can also be varied by adjusting Iref, but this is not recommendable because the current swing is restricted to 8μA. Moreover, the TR expansion at the higher range would be zero because of accuracy reason as stated in section 6.2.

6.4 The counting capacitor Ccnt

Since this capacitor receives the down-converted charge and discharge currents, a low leakage type capacitor must be used.

The minimum value of the capacitor is determined by the switching levels Vh and Vl. Within one clock cycle, the voltage at Ccnt is not allowed to fall from Vh to Vl whilst charging, whereas the voltage at Ccnt is not allowed to rise from Vl to Vh whilst discharging. The formula for Ccnt (formula [15]) in the next chapter clarifies.

6.5 The conversion resistors Rc and Rd

By means of the resistors Rc and Rd the charge and discharge currents are converted to IC-acceptable levels. The Rc and Rd resistors are called charge and discharge converting resistors respectively.
The formulae below set out how the actual charge and discharge currents are converted to IC-acceptable current-levels. The Rc and Rd resistors define the charge and discharge rates respectively (refer to section 7.1).

$$Ic = \frac{Icharge * Rs}{Rc}$$

$$Id = \frac{Idischarge * Rs}{Rd} \qquad [3]$$

Formula [3]: conversion of actual charge/discharge currents to IC-acceptable levels.

From literature it is known that both the charge acceptance and the actual capacity of a cell charged at high temperatures are lower than those for a cell charged at room temperatures (23°C), i.e. the cell accepts less charge than that it would accept at room temperatures.
This cell characteristic can introduce an error in the state-of-charge indication. This error cannot be compensated since this would require the "converted" battery capacity in SAA1501T to be adjusted as a function of temperature and this function has not been implemented in SAA1501T, simply because of reasons of space and cost.
External compensation may be possible by using temperature dependent resistors for Rc and Rd so that the rate of the counters in SAA1501T (and thus Crate) is adjusted as a function of temperature. It is highly doubtful whether it is worthwile doing so. As the indication is carried out in rough steps of 20% it is allowed to have some mismatch between SAA1501T and the real battery capacity, provided that the charge/discharge rate ranges and temperature operating ranges stay within the recommended ranges as stated in the battery specification.

In case one is afraid prior mentioned mismatch could become unacceptably large, which then could lead to empty batteries prior to battery low indication, SAA1501T can be designed in so that average charge and discharge efficiency factors can be set independently.

With this effectively under-estimated indication the battery low indication will always take place prior to the batteries are empty.

Pins Rc and Rd are sensitive to capacitive load and it is therefore recommended to connect the external resistors for Rc and Rd as close as possible to pins Rc and Rd. This sensitivity (Cout*Rc(Rd) < 1μsec) is mainly due to the feedback from the outputs of the V/I sensing convertors to inputs Rc and Rd.

6.6 Setting the temperature block

The sections below set out how to set the absolute temperature turnover points. The principle used for the temperature measurement circuit is that of a resistor bridge configuration as shown in figure 7.

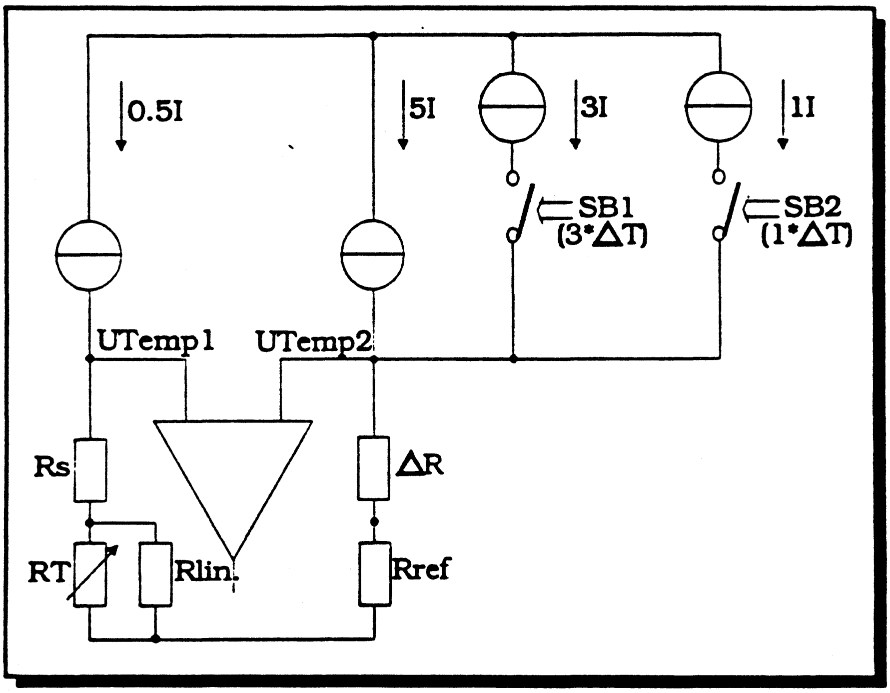


Figure 7: block diagram of temperature block.

6.6.1 Setting the Fast Charge turnover point

As can be seen from figure 7, absolute temperature takes place at UTemp1 = UTemp2 when TBattery = TFC (Fast Charge Turnover point). Then the following equation is valid:

$$0.5I*(Rs+Rp(Tfc))=5I*(Rref+\Delta R),\ \textit{so that} \quad [4]$$

$$Rp(Tfc)=10*(Rref+\Delta R)-Rs$$

, in which Rp (TFC) = RT(TFC)//Rlin.

In case 0.5I*Rs = 5I*ΔR, Rp(TFC) and thus TFC is only determined by Rref:

$$Rp(Tfc)=10*Rref \qquad [5]$$

Now, consider an example in which TFC is set to 60°C. For RT an NTC of type PHILIPS 2322 640 10104 (100kΩ @25°C) is used with characteristic as shown in figure 8. Further, in this example an Rlin. = 100kΩ is used for linearisation of the NTC and Rs and ΔR are chosen so that 0.5I*Rs = 5I*ΔR. Then for Rref a value of 1.87 kΩ (Rp(60°C)) is found (0.1*(100k/23k)).

It should be noted, that a linearisation resistor is only necessary when the NTC curve is not linear in the temperature operation range TSB2 to TFC. Further, the voltage at the Temp1 and Temp2 pins must be between 0 and 900mV.

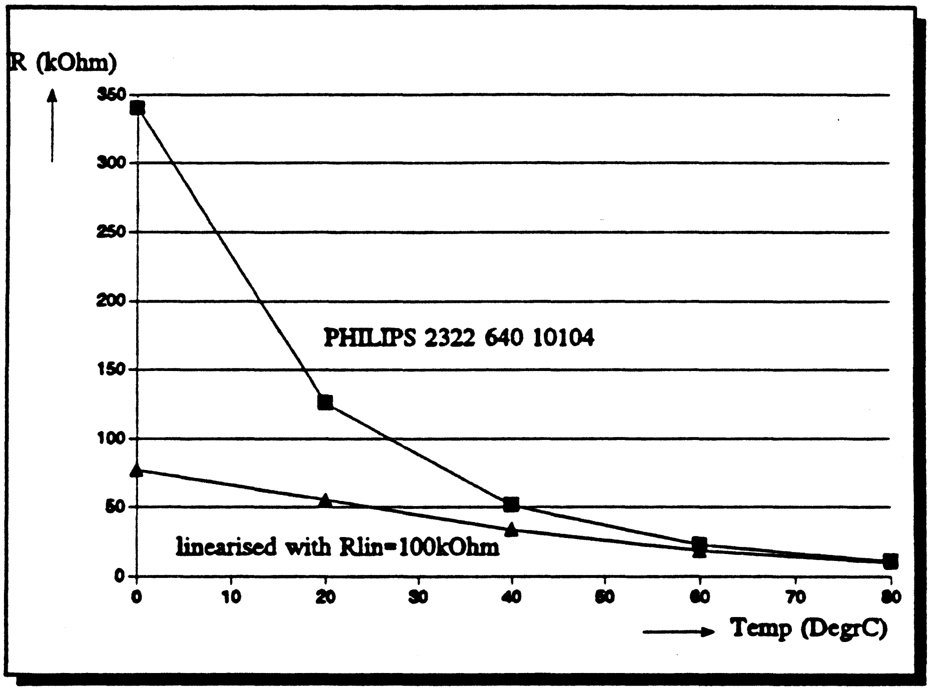


Figure 8: PHILIPS NTC 2322 640 10104 characteristic + linearisation.

6.7.2 Setting the Self-discharge turnover points

As mentioned above two absolute temperature measurements are carried out in StandBy Mode (Self-discharge Mode) and thus SAA1501T differentiates between three temperature ranges. One temperature range is that of battery temperatures beyond TSB1 which corresponds with a countdown rate of 33 days. The second temperature range is that of battery temperatures below TSB2 which corresponds with a countdown rate of 200 days. The remaining temperature range is between TSB1 and TSB2 and corresponds with a countdown rate of 100 days. Figure 9 clarifies.

As can be seen from figure 9, for the absolute temperature turnover points TSB1 and TSB2 can be written as:

$$
\begin{aligned}
TSB1 &= Tfc - 3\Delta T \\
TSB2 &= Tfc - 4\Delta T
\end{aligned}
\qquad [6]
$$

Further, for ΔT it can be found:

$$\Delta T = \frac{Rp(Tfc) + Rs}{5 * dR/dT} \qquad [7]$$

Now, consider an example in which TSB1 is set to 45°C and TSB2 to 40°C, so that ΔT must be set to 5°C.

Since it was assumed earlier on that 0.51*Rs = 51*ΔR, for Rs it can be written:

$$Rs = 5 * \Delta T * \frac{dRp}{dT} - 10 * Rref \qquad [8]$$

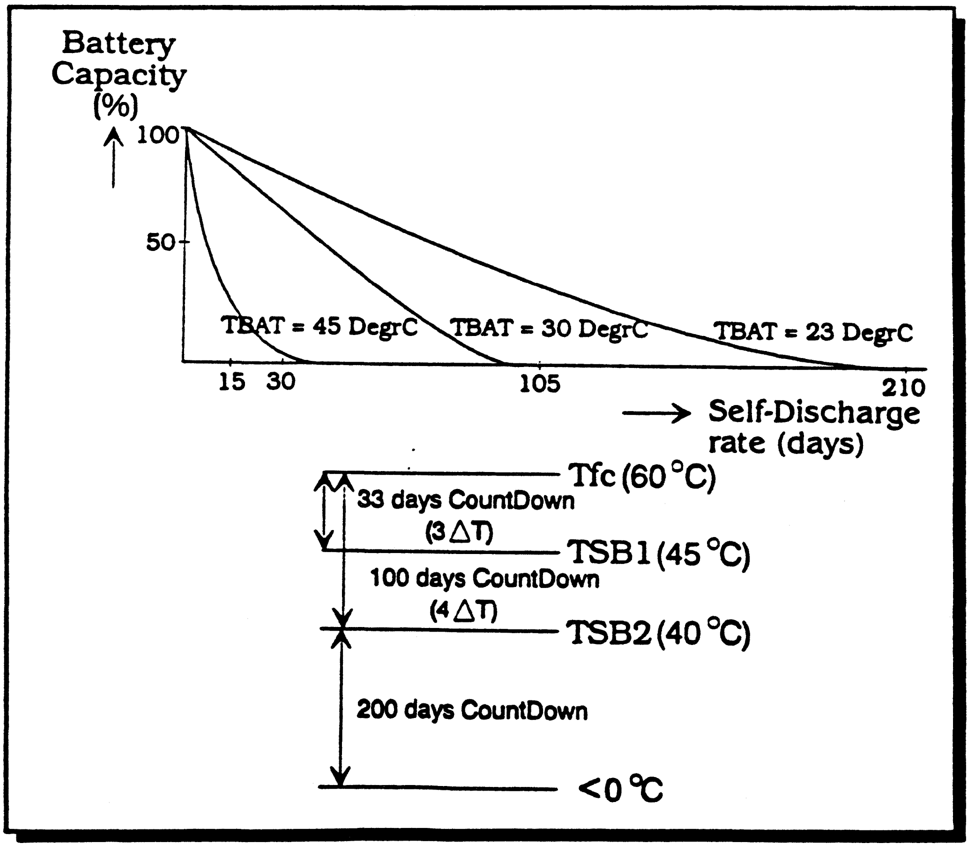


Figure 9: example of a battery with its Self-discharge characteristics as a function of battery temperature and derivation of the Self-discharge turnover points from these characteristics.

The differential value dR/dT follows from:

$$dRp/dT = abs(\frac{R(Tmin)-R(Tmax)}{Tmin-Tmax}) \qquad [9]$$

From figure 8 it follows that in the temperature area 40-60°C the dR/dT=762Ω/°C ((40k2-25k)/20) and thus from formula [5] it follows (using the value for Rref that was found in the previous section) that: Rs = 5*5*762 - 18k87 = 200Ω. Because it was assumed that 0.5I*Rs = 5I*ΔR, ΔR= 2kΩ is found.

In order to assure correct operation of the temperature block, it should be examined whether ΔU > Uoffset(=1.5mV) of the opamp at Temp1 and Temp2. ΔU is defined by formula:

$$\Delta U=0.5*2E-6*\Delta T*\frac{dRp(T)}{dT} \qquad [10]$$

In this example ΔU = 0.5*2E-6*5*762 = 3.81mV, which is well above 1.5mV.

7 Design-in examples

In this chapter first a general design-in strategy is given for the basic intelligent battery pack application.
Subsequently, an application example is set out by using this design-in strategy.
Further, an exceptional application example will be set out, which requires a non-standard design-in strategy.
In both applications only one sense resistor (Rs) is applied and Iref=4μA (Rref=51kΩ).

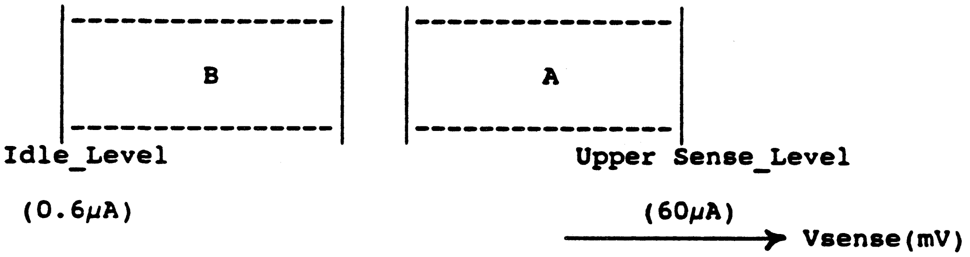
7.1 Design-in strategy for basic SCI application

Restrictions: Rref=51kOhm (Iref=4μA). It should be noted that the value of Iref will not affect the setting of the battery capacity.

1) Determine application charge and discharge rate ranges (in CA) by ICharge(IDischarge)/QBAT, in which QBAT is the Standard capacity value.

2) Examine whether the application Charge and Discharge ranges are within 0.05-5CA. If so, the oscillator frequency (Fosc) can operate at 4kHz (nominal frequency). Use formula [2].

3) Choose Rsense.
Refer to section 6.2 which requirements Rsense needs to satisfy. Further, ideally Rsense should be chosen so that the sense voltages are as high as possible as the relative accuracy of the U/I converters at high sense voltages is better than at low sense voltages.
Thus for the graph below it can be said, that it is best to have the U/I-converters operate in area A.



4) Make sure VIdle>4mV:
The current Idle-level is internally fixed to 0.6μA. This does not apply for the Idle sense voltage, as this depends on:

$$0.6\mu A = \frac{Rsense * ICharge\ (IDischarge)}{Rc(Rd)}$$

However, because accuracy of the U/I converters is guaranteed in the range from 4mV to 400mV, VIdle must be larger than 4mV.

Therefore from the afore mentioned formula it follows that Rc,Rd and Rsense must have minimum values so that VIdle>4mV.
Rsense_min follows from:
Rsense_min = 4mV/Iin_min, in which Iin_min = ICharge_min, if ICharge_min < IDisCharge_min; else Iin_min = IDisCharge_min, if IDisCharge_min < ICharge_min.
Rc_min and Rd_min (=4mV/0.6μA) do not have to be checked as this is implicitly accounted for in the formulae for determination of the values for Rc and Rd ([11] and [12]).

5) Determine values for the charge and discharge conversion resistors. Use the formulae as stated in [11]. These formulae do not take into account charge and discharge efficiency factors of the batteries.
To set average Charge and DisCharge efficiency factors the formulae in [12] should be used.

6) Choose value for reservoir capacitor Ccnt (Coulomb counting capacitor) using formula [13].

7.2 Camcorder battery pack

The application:
- 5 cells, 6Vnominal, 1.2A*hr batteries (Standard Capacity).
- Charge current range: 0.2A to 2.5A (0.167CA to 2.08CA).
- Discharge current range: 0.1A to 3.3A (0.083CA to 2.75CA).
- The application uses only one sense resistor.

1) First examine whether the operational charge and discharge current rates of the batteries are within 0.05CA and 5CA. If so, SAA1501T can operate at its nominal frequency of 4kHz. The charge and discharge rates are defined by: Ccharge_rate = Icharge/{Standard Capacity} and Cdischarge_rate = Idischarge/{Standard Capacity} respectively. Thus, a maximum charge rate of 2.5/1.2 (Icharge_max/QBAT) = 2.08CA and a maximum discharge rate of 3.3/1.2 = 2.75CA are found here. Further, a minimum charge rate of 0.167CA and a minimum discharge rate of 0.083CA are found. The given charge/discharge current rates are within range. This means that Fosc can easily operate at 4kHz.

   Co follows from the equation:

$$Co=\frac{1}{5.6*Rref*Fosc} \qquad [12]$$

   Formula [12]: Co as a function of Fosc.

   In this application example for Co a value of 820pF is found (Rref=51kΩ as Iref must be 4μA).

2) Further, it should be examined whether the sense voltages across Rsense stay within range of the U/I convertors. Preferrably in the range of 4mV up to 400mV.

   Regarding the specified charge current range, with a sense resistor of 20mΩ charge voltages of 4mV (20mΩ*0.2A) up to 50mV (20mΩ*2.5A) will be introduced at the inputs of the charge U/I convertor, which is just within range of the U/I converter. However, for the specified discharge current range, a sense resistor of 20mΩ will give discharge voltages of 2mV (20mΩ*0.1A) to 66mV (20mΩ*3.3A), which is not within range of the U/I convertor. The minimum Rsense value that must be used is 4mV/0.1A = 40mΩ. We choose for a commercially available resistor value of Rs=50mΩ, so that we find charge voltages in the range of 10mV-125mV and discharge voltages of 5mV-165mV, which is satisfactory. It should be noted that Rs may not be larger than 120mΩ, otherwise the allowed upper discharge voltage is exceeded (120mΩ*3.3A = 400mV).

   For an Rsense=50mΩ a maximum dissipation in Rsense of $P_{Rsense} = (3.3)^2*50E-3 = 0.5W$ is found.

3) The values for the charge and discharge converting resistors are found using the following formulae:

$$Rc = \frac{QBAT*Rs}{12*10^{-6}*4*10^{3}*Tosc} \quad , \textit{ with } Fosc[Hz]$$

$$Rd = \frac{QBAT*Rs}{12*10^{-6}*4*10^{3}*Tosc} \qquad [13]$$

,in which QBAT is the Standard Battery Capacity value.

As stated before, to prevent batteries being empty prior to battery low indication, we need to design-in so that SAA1501T operates under-estimated. Then, the charge and discharge efficiency of the batteries are taken into account by setting 'average efficiency factors'. These factors are determined by regarding the expected temperature operating range and Crate ranges of the batteries. From several battery characteristics of Actual Battery Capacity as a function of temperature and Crate, it is concluded that the 'worst case efficiency factor' is 70% when regarding a temperature range of 0-80°C and a Crate range up to 3CA. It should be noted that this factor is highly dependent on the battery type being used.

The afore given formulae for Rc and Rd do not take the charge and discharge efficiency into account. To take these into account the following formulae should be used:

$$Rc = \frac{QBAT * Rs}{12*10^{-6}*4*10^{3}*Tosc*(Charge\ efficiency)}$$

$$Rd = \frac{QBAT * Rs * Discharge\ efficiency}{12*10^{-6}*4*10^{3}*Tosc} \qquad [12]$$

In this basic application it is assumed that it takes 1.43 (charge efficiency 70%) times longer to fully charge the batteries; refer to table 3 for the underlying thought.
Thus for charge rates of 1CA it takes 86 minutes to charge the batteries from empty to full instead of 60 minutes nominally. The value of Rc = (1.2*50E-3)/(12E-6*0.7) = 7k2Ω.

Temp (DegrC) Crate (CA)	-20	0	20	40	60
C/10	70	90	100	102	105
C	50	65	80	82	95
3C	39	56	65	70	78
5C	30	43	58	65	70

Table 3: Actual battery capacity as a function of temperature and Crate.

In the basic application it is assumed for discharging that it will take 0.7 times (discharge efficiency 70%) shorter to discharge from full to empty. Thus for discharge rates of 1CA it takes 42 minutes to discharge batteries from full to empty instead of 60 minutes nominally. Therefore for Rd a value of Rd = (1.2*50E-3)/(12E-6*(1/0.7) = 3.5kΩ is found.

4) The minimum value for Ccnt follows from:

$$Ccnt > \frac{Irefl*64*Tosc}{0.66*(Vh-Vl)} > \frac{10*10^{-6}*64*Tosc}{0.66*0.4}, \text{ with } Tosc[Sec] \qquad [15]$$

,in which it is preferrable to set Irefl to 10μA (2.5*Iref), as stated in the Rref section. In this application example for Ccnt a value of 0.68μF is found.

To avoid continuous switching of the signal at pin CNT between Vh and Vl, Vh-Vl must be larger than ΔVmax = (Irefl*Tclk)/Ccnt. This is the case in this example: Vh-Vl=0.4V and ΔVmax = 0.24V.

7.3 A tools application

The application (an electrical drill-machine):
- 6 batteries, 9Vnominal, 2A*hr batteries (Standard Capacity).
- Charge current range: 0.1A - 14A (0.05CA to 7CA).
- Discharge current range: 1A - 100A (2CA to 50CA).
- Sense resistor: 5mΩ.
- The application uses only one sense resistor.

In the sections below two application-solutions are given. The first one enables SAA1501T to track the full charge/discharge ranges as given above. However, use of some additional circuitry is required. The second application-solution does not require any additional components, however the full discharge range can only be sensed accurately up to 20CA.

7.3.1 Full range solution

1) Usually, tools applications feature wide discharge currents as can be seen in this example. The current values specified here are mean current values. It is therefore feasible, that with a mean charge current of 0.1A a pulsated charge current with $\delta=10\%$ is sensed across the sense resistor (UGNDs=5mV during the on-time of the pulsated charge signal). And thus we are able to sense charge currents within a dynamic range of 1:140.

   In this example we need to be able to track discharge rates up to 50CA (maximum discharge rate) and minimum charge rates of 0.5CA ($Ic_{peak}=1A$), thus from formula [2] it is found that Fosc must run at 40kHz. From formula [12] it follows that Co=82pF.

2) Regarding the specified charge and discharge current ranges with an Rs=5mΩ, charge sense voltages of 5mV up to 70mV are introduced, whereas discharge sense voltages of 5mV up to 500mV are introduced. A discharge sense voltage of 500mV is out of range for Vsupply < 2.1V. However, it is feasible that this voltage occurs during circuit-shorts of the motor of the machine for short time, so that the inaccuracy being introduced during this short interval may be neglected.

   For an Rsense=5mΩ a maximum dissipation in Rsense of $P_{Rsense} = (100)^2*5E\text{-}3 = 50W$ is found, and thus use of a power resistor is required.

3) From formula [14] it is found that Rc = (2.0*5E-3)/(1.2E-6*0.7) = 11k9Ω (assumed is that the charge acceptance is 70%, refer to previous section).
   For Rd it is found that Rd = (2.0*5E-3*0.7)/(1.2E-6) = 5K8Ω (assumed is that the discharge efficiency is 70%, refer to previous section).

4) From formula [15] a value of 68nF for Ccnt is found.
   To avoid continuous switching of the signal at CNT between Vh and Vl, Vh - Vl must be larger than ΔVmax = (Irefl*Tclk)/Ccnt. This is the case in this example: Vh-Vl=0.4V and ΔVmax = 0.24V (Tclk = 1.6ms).

5) Since the countdown rate in self-discharge mode (SB mode) directly depends on the Fosc (2^{36} * Tosc), which thus means that with tenfold Fosc the countdown rate in Self-discharge mode is also tenfold (20/10 days instead of 200/100 days).
   In appendix F this problem has been solved by means of the circuitry among the opamps LM393A. In Standby mode this circuitry switches on C3, so that in this mode the countdown rate runs at its nominal speed of 62.5Hz (Fosc ~4kHz).

   Refer to appendix F for circuit diagram.

7.3.2 The straight-forward solution

In case it is satisfactory to have a concept which is not able to track for charge/discharge rates higher than 20CA with high accuracy, Fosc can run at lower frequency. Now, Fosc should run at a frequency so that it does not differ much from the nominal frequency (4kHz). As long as the countdown rate in self-discharge is kept above 50 days (nominally 200/100 days) the oscillator frequency does not have to be decreased in the Self-discharge mode.

1) In this example we choose for Fosc=16kHz which is still guaranteed by specification, so that the count-down rate is 50 days in the self-discharge mode.

   For Co a value of 180pF is found.

2) Refer to 7.2.1.

3) For Rc we find Rc = (2.0*5E-3)/(12E-6*0.25*0.7) = 4k8Ω (assumed is that the charge acceptance is 70%, refer to previous section).
   For Rd it is found that Rd = (2.0*5E-3*0.7)/(12E-6*0.25) = 2k3Ω (assumed is that the discharge efficiency is 70%, refer to previous section), whereby we realise that the discharge currents higher than 40A are taken into account as 20CA (maximum charge/discharge rate when Fosc runs at 16kHz).

4) From formula [15] a value of 220nF is found for Ccnt.

   To avoid continuous switching of the signal at CNT between Vh and Vl, Vh - Vl must be larger than ΔVmax = (Irefl*Tclk)/Ccnt. This is the case in this example: Vh-Vl=0.4V and ΔVmax = 0.24V (Tclk = 4ms).

5) With Fosc=16kHz the countdown rate in the Self-discharge mode is 50days.

   This concept does not require any additional components for correcting the oscillator frequency in the Self-discharge mode. However, it is not able to track the full discharge current range accurately; beyond 40A (20CA) additional inaccuracy is introduced.

   Appendix G shows the circuit diagram.

7.4 Typical design-in examples

The table below summarises some typical battery pack applications and their parameters and how to design-in the external components. As in the previous sections, it is assumed that the charge and discharge efficiency is 70%. In the following table this is accounted for in the calculations for Rc and Rd.

Application	Rsense (mΩ)	Batt. Cap. (mA*h)	Charge Curr. (A)	Disch. Curr. (A)	Fosc. (kHz)	Co (pF)	Ccnt (F)	Possible DC Ch./Disch. Rate Range (CA)	Rc (kΩ)	Actual Ch.Rate Range	Rd (kΩ)	Actual Disch. Rate Range (CA)
NoteBook (Colour Screen)	50	2300	1.2	0.8-1	4	820	0.68μ	0.05-5	13k7	0.366	6k7	0.497-0.62
NoteBook (Mono Screen)	50	2300	1.2	0.6-0.8	4	820	0.68μ	0.05-5	13k7	0.522	6k7	0.373 - 0.497
Camcorder	50	1200	0.2 - 2.5	0.1 - 3.3	4	820	0.68μ	0.05-5	7k2	0.116 - 1.45	3k5	0.12 - 3.92
Camcorder	50	2400	0.2 - 2.5	0.1 - 3.3	4	820	0.68μ	0.05-5	14k3	0.059 - 0.73	7k	0.060-1.96
Power tool	5	2000	1 - 14	1 - 70	40	68	68n	0.5-50	11k9	0.35 - 4.9	5k9	0.706 - 49.44
Battery Packs For Cellular telephone	100	1400	1.4	0.7 - 2	4	820	0.68μ	0.05-5	16k7	0.7	8k2	0.71 - 2.0

Table 4: typical SAA1501T application examples.

8 Application possibilities

In the first instance, SAA1501T has been designed for portable applications where battery energy indication is critical. Such applications include:

- Laptop, notebook, and handheld computers.
- Electronic organizers.
- Shavers.
- Camcorders.
- Cellular phones.
- Portable hand tools.
- Portable mobile radio.
- Portable instrumentation such as telecommunications, medical, and analytical.

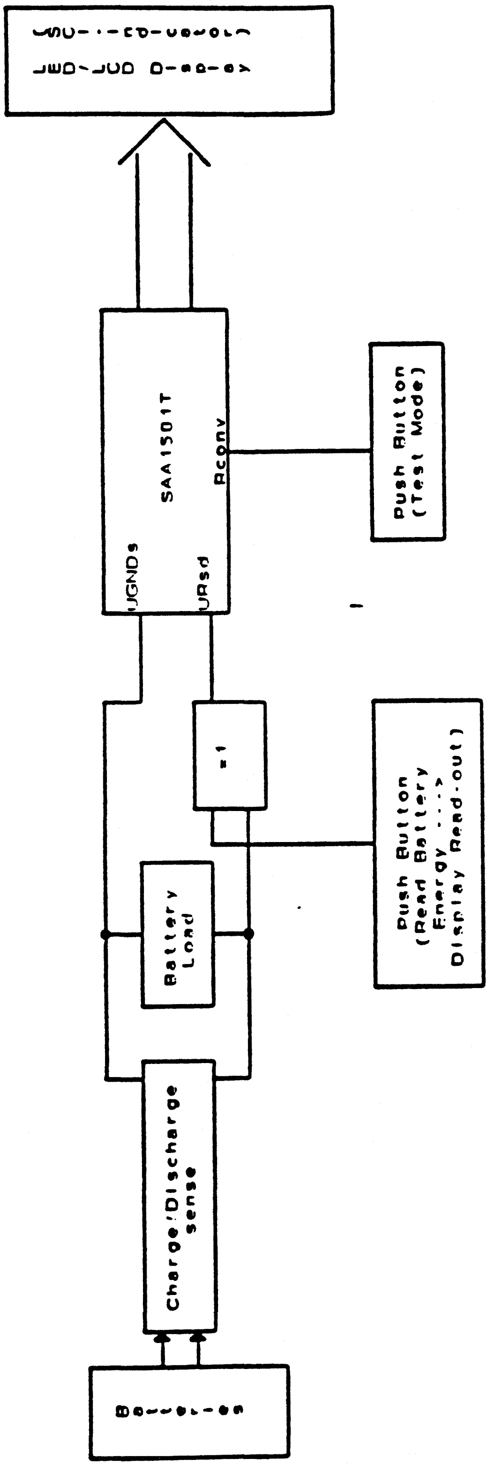
APPENDIX A

Glossary of Abbreviations and Terms

- Actual cell capacity	Measures the total cell capacity under all non-standard conditions that alter the amount of capacity which the fully charged cell is capable of delivering when full discharged.
- BICMOS	Combined Bipolar and Metal Oxide Semiconductor technology.
- BLI	Battery Low Indication output; for visible battery low indication.
- BP	Backplane pin; to differentiate between LED and LCD modes.
- BUZZ	Buzzer output; for audible battery low indication.
- CMOS	Complementary Metal Oxide Semiconductor technology.
- IC	Integrated Circuit.
- LCD	Liquid Crystal Display.
- LED	Light Emitting Diode.
- PO	Power On output.
- NiCd	Nickel Cadmium Battery.
- NiMH	Nickel Metal Hydride Battery.
- SB mode	StandBy mode; in this mode SAA1501T monitors the Self-discharge of the batteries.
- SCI	State-Of-Charge Indicator.
- TFC	Fast charge turnover point.
- TSB	StandBy turnover point.
- Standard cell capacity	Measures the total capacity that a relatively new production cell can store and discharge under a defined standard set of application conditions.

Appendix B

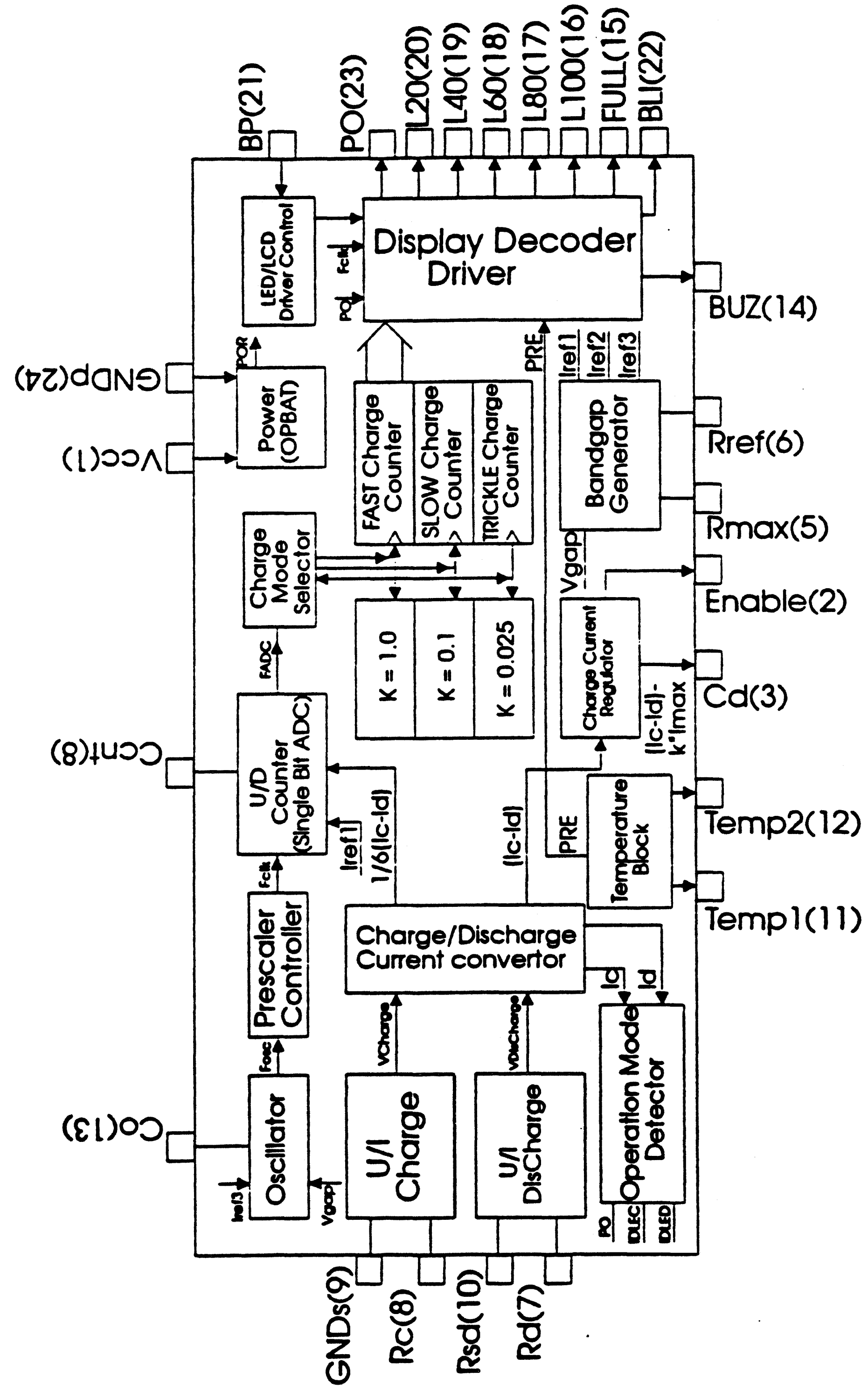
Block diagram of a battery pack application using SAA1501T



Intelligent Battery Pack Block Diagram
(State of Charge Indicator (SCI))

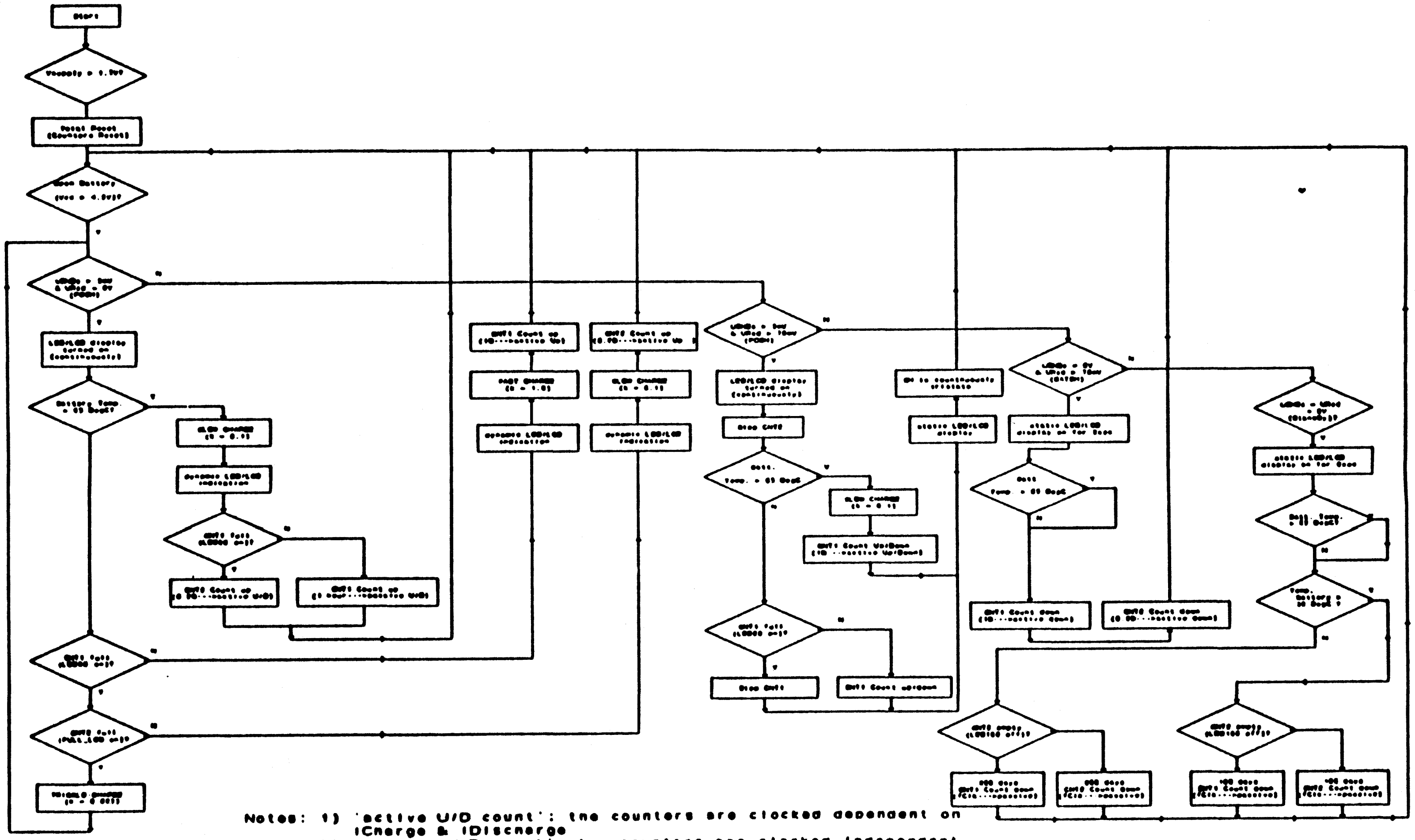
Appendix C

Block diagram of SAA1501T



Appendix D

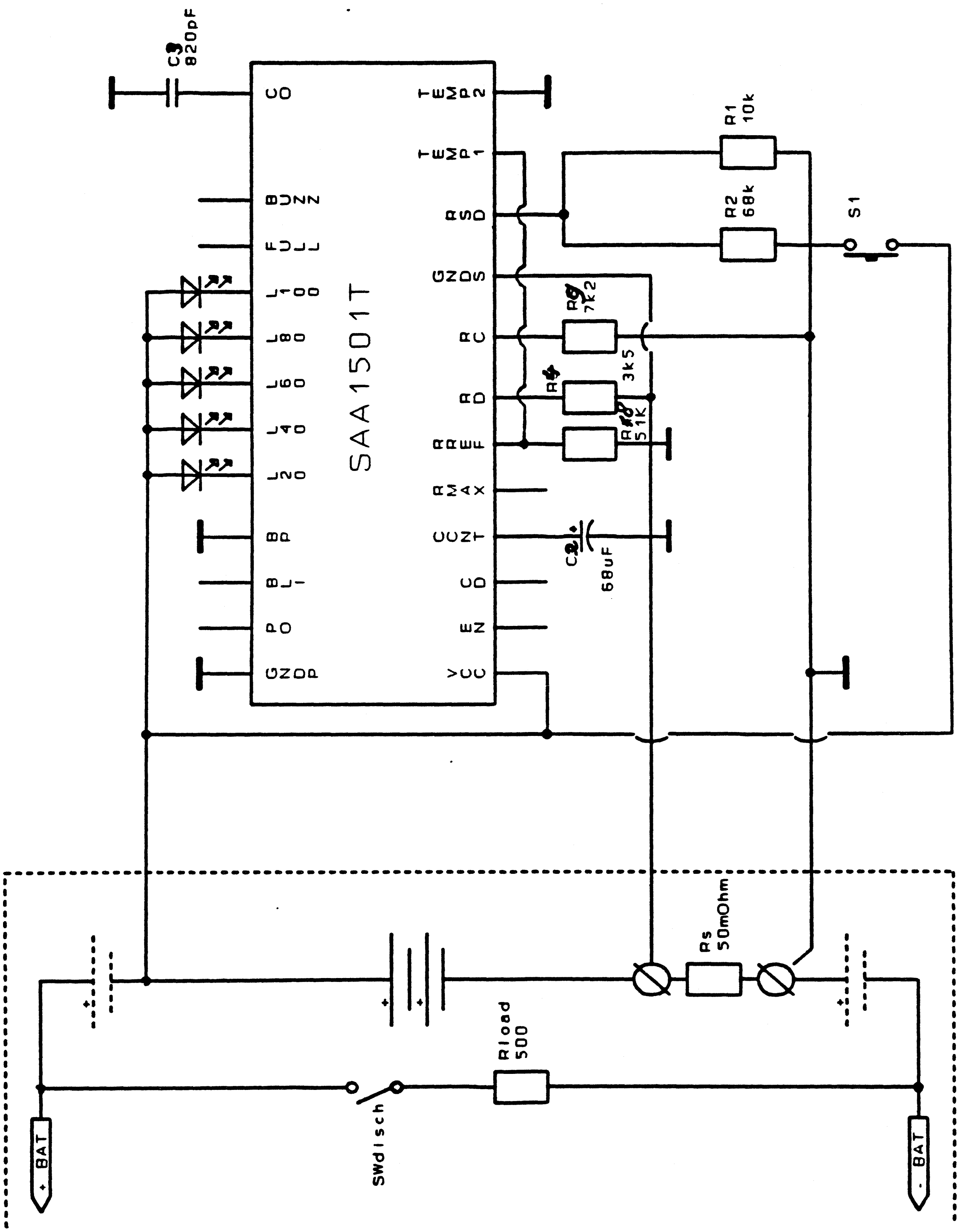
Flow chart of SAA1501T



Notes:
1) 'active U/D count': the counters are clocked dependent on ICharge & IDischarge
2) 'passive U/D count': the counters are clocked independent of ICharge & IDischarge (directly clocked by fClk)
3) 'dynamic charge': LEDs/LCDs are being read-out while current charge-segment LED/LCD is flashing
4) 'static charge': LEDs/LCDs are continuously on
5) This flow chart must not be regarded as sequential. Each operation mode is a separate continuous process.

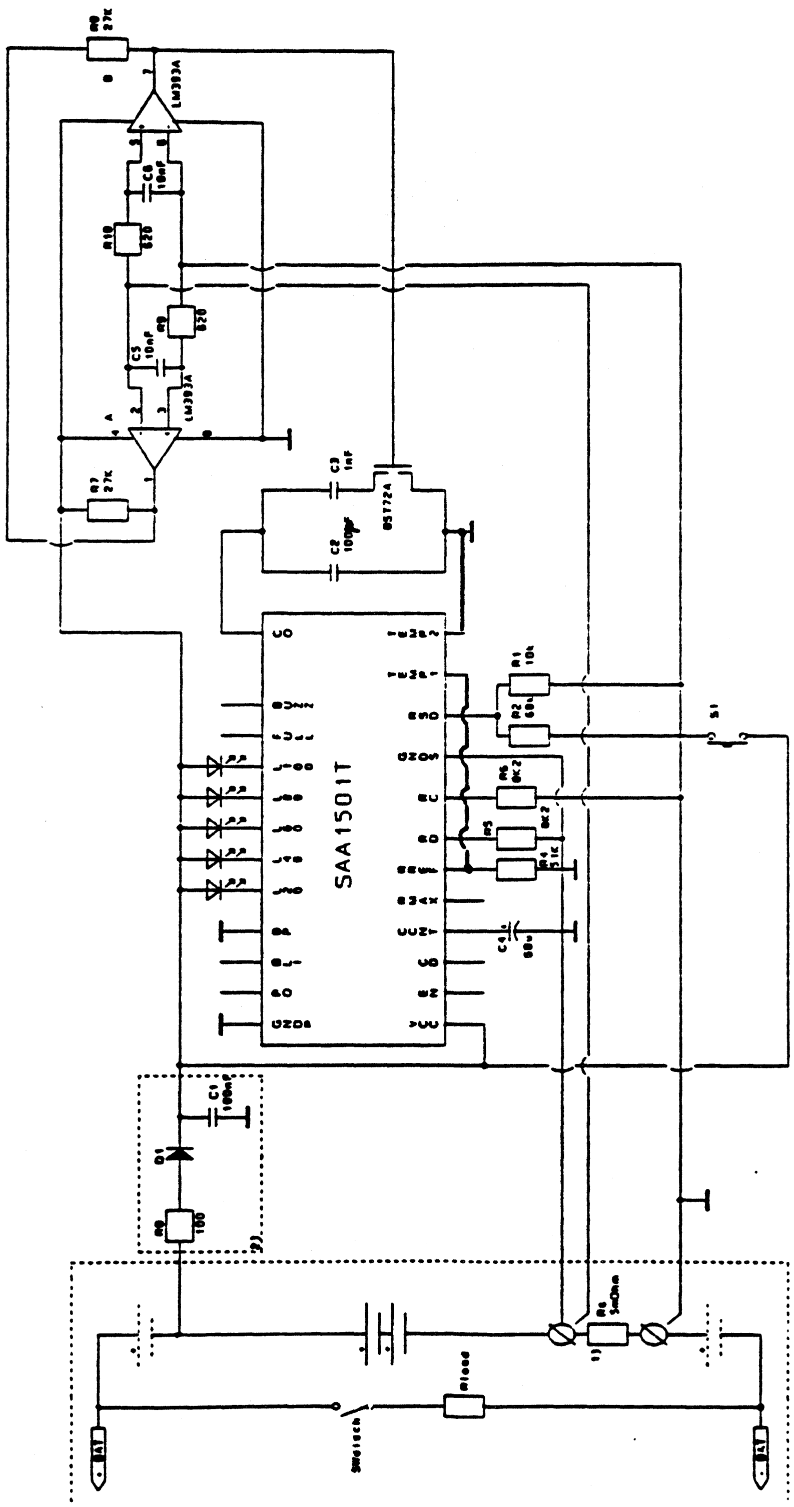
Appendix E

Camcorder Application circuit diagram (section 7.2)



Appendix F

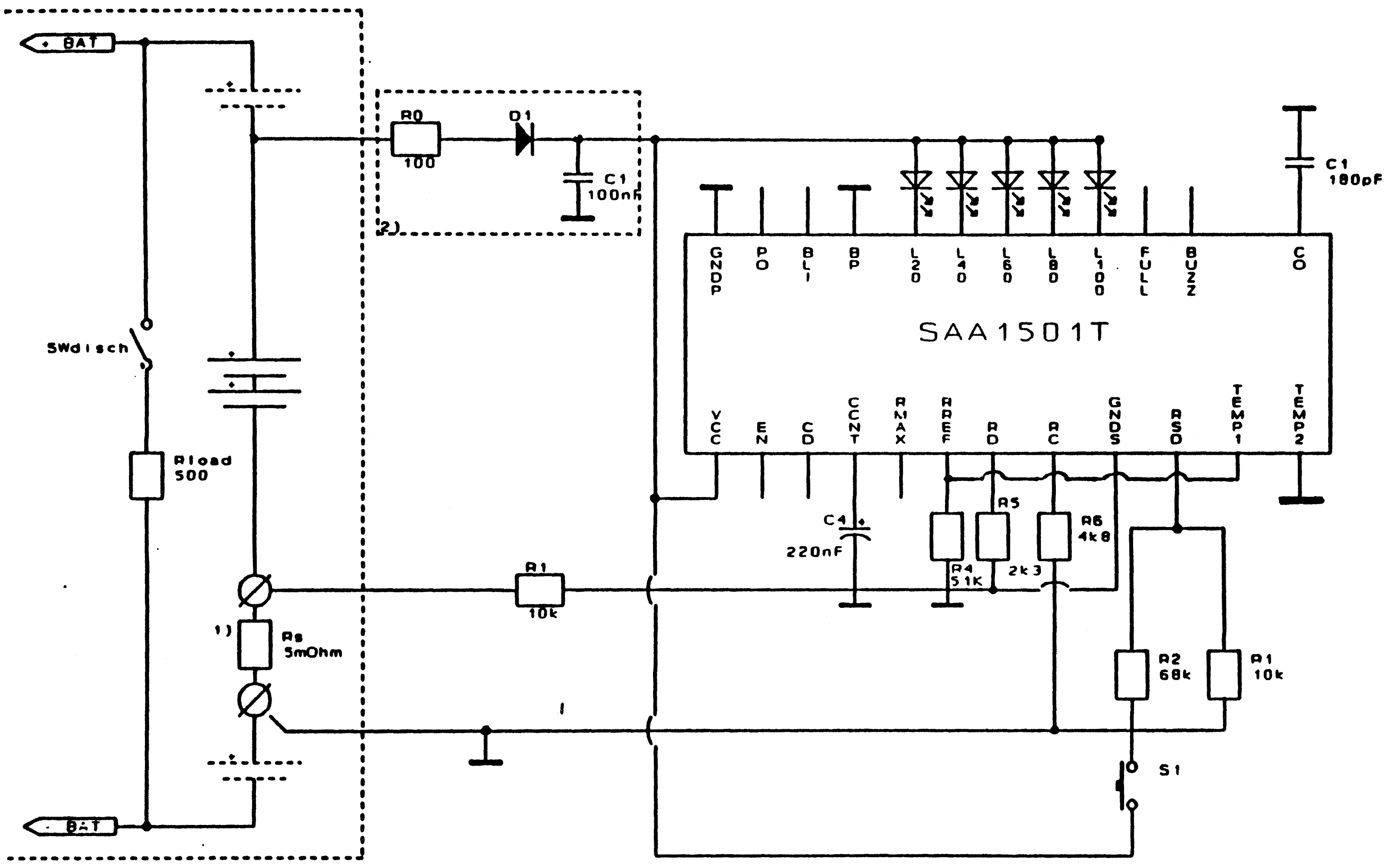
Non-basic tools Application circuit diagram (section 7.3.1)



1) The wire between two individual cells can be used as sense resistor.
2) Use of this 'voltage dip-protection' circuit is optional.
In case this dip-protection circuit is connected three cells - instead of two cells - need to be connected to SAA1501T, because of the voltage drop across the diode and series resistor.

Appendix G

Basic tools Application circuit diagram (section 7.3.2)



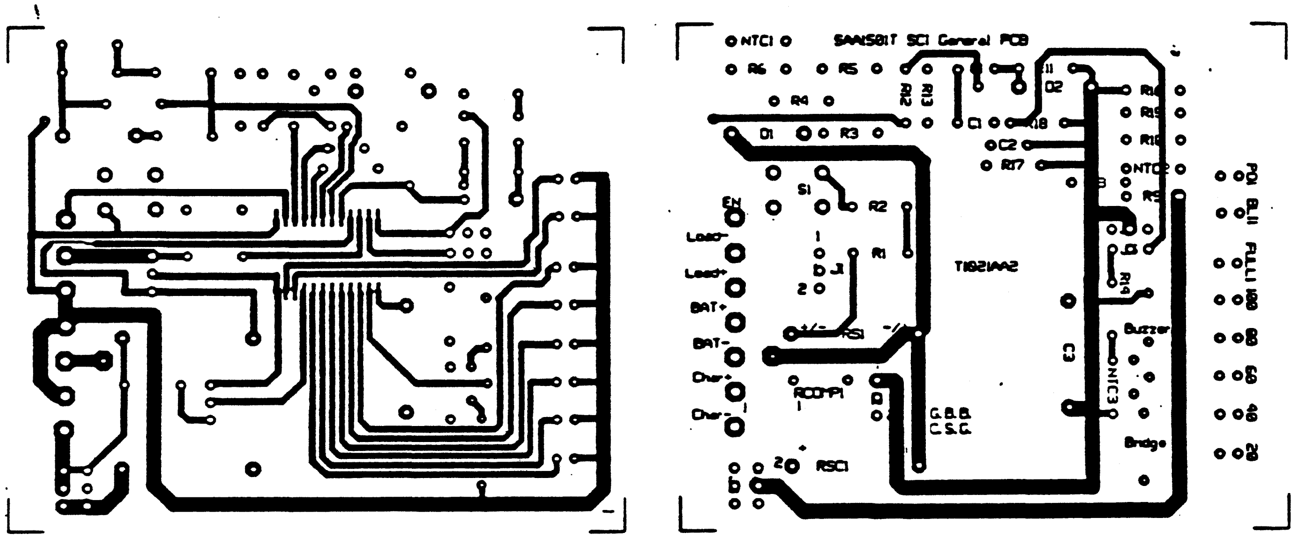
1) The wire between two individual cells can be used as sense resistor.
2) Use of this 'voltage dip-protection' circuit is optional.
In case this dip-protection circuit is connected three cells - instead of two cells - need to be connected to SAA1501T, because of the voltage drop across the diode and series resistor.

Appendix H

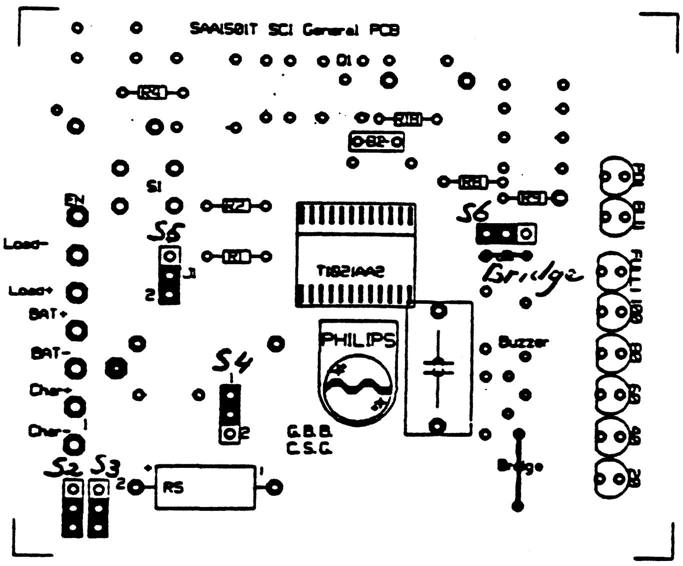
PCB Layout of the Camcorder SCI application

This PCB is for experimental purposes. Only the external components as shown in the circuit diagram in Appendix E have been connected. Other values for Rs, Rc and Rd have been used than those shown in Appendix E.

A) PCB Copper Side (Double Sided)



B) PCB Components Side



Rs	270mOhm
R1	10kOhm
R2	68kOhm
R4	20kOhm
R9	39kOhm
C2	680pF
C3	820pF

APPLICATION NOTE

ICs for Battery Management

Battery Pack Charge Gauge for Camcorders using SAA1501(T)

Report No: NPO/AN9406.

Guido B Banning

Customer & Product Support Group Consumer ICs Nijmegen, Power Conversion Section,
the Netherlands.

Keywords
State-Of-Charge-Indication
Intelligent Battery Packs
Coulomb Counting Principle
Charge/Discharge efficiency compensation

Date : 14 June 1994
Pages: 7

Abstract:

This note is a follow-up on note NPO/AN9401 and deals with a dedicated Battery Pack application for use with a camcorder. Included is a PCB layout for this application. PCBs are available for demonstration or experimental purpose.
For details on the SAA1501T the afore mentioned report should be referred to.

The SAA1501T is a low-cost battery management circuit to be used in fast charging systems. The most important function of SAA1501T is monitoring the charge account in rechargeable batteries. During battery charging the charge current and charge time are registered in a coulomb counter, whereas during discharging the discharge current and time are recorded. The momentary charge account of the batteries can be displayed either on an LC display or by an LED bargraph. In order to get an intelligent battery pack, the state-of-charge indicator function is incorporated into the battery pack.

Dedicated Intelligent Camcorder Battery Pack

The application:
- 5 cells, 6Vnominal, 1.2A*hr batteries (Standard Capacity).
- Charge current range: 1.1A to 1.4A (0.917CA to 1.167CA).
- Discharge current range: 700mA to 1.3A (0.583CA to 1.083CA).
- The application uses one sense resistor only: 70mΩ.
- Consider the battery pack of this application as a black box, i.e. it is not possible to access the single cell voltage with this application thus only the voltage from the complete battery pack unit (5 cells).

1) First examine whether the operational charge and discharge current rates of the batteries are within 0.05CA and 5CA. If so, SAA1501T can operate at its nominal frequency of 4kHz.
The charge and discharge rates are defined by:
Ccharge_rate = Icharge/{Standard Capacity} and Cdischarge_rate = Idischarge/{Standard Capacity} respectively.
Thus, a maximum charge rate of 1.4/1.2 (Icharge_max/QBAT) = 1.167CA and a maximum discharge rate of 1.3/1.2 = 1.083CA are found here. Further, a minimum charge rate of 0.917CA and a minimum discharge rate of 0.583CA are found. The given charge/discharge current rates are thus within range. This means that Fosc can easily operate at 4kHz.

   In this application example for Co a value of 820pF is found (Rref=51kΩ as an Iref of 4μA is preferred), which can be found using formula[12] as stated in note NPO/AN9401.

2) Further, it should be examined whether the sense voltages across Rsense stay within range of the U/I convertors. Preferrably in the range of 4mV up to 400mV.

   Regarding the specified charge current range, with a sense resistor of 70mΩ charge voltages of 77mV (70mΩ*1.1A) up to 98mV (20mΩ*1.4A) will be introduced at the inputs of the charge U/I convertor, which is easily within range of the U/I converter. The same applies for the discharge voltage range at the discharge convertor as with a Rsense=70mΩ a range of 49mV-91mV is found.

   It should be noted that Rsense may not be larger than 286mΩ, otherwise the allowed upper charge voltage is exceeded (286mΩ*1.4A = 400mV) and not smaller than 6mΩ otherwise the minimum discharge voltage will fall below the minimum allowed voltage at the U/I convertor (6mΩ*.7 = 4mV).

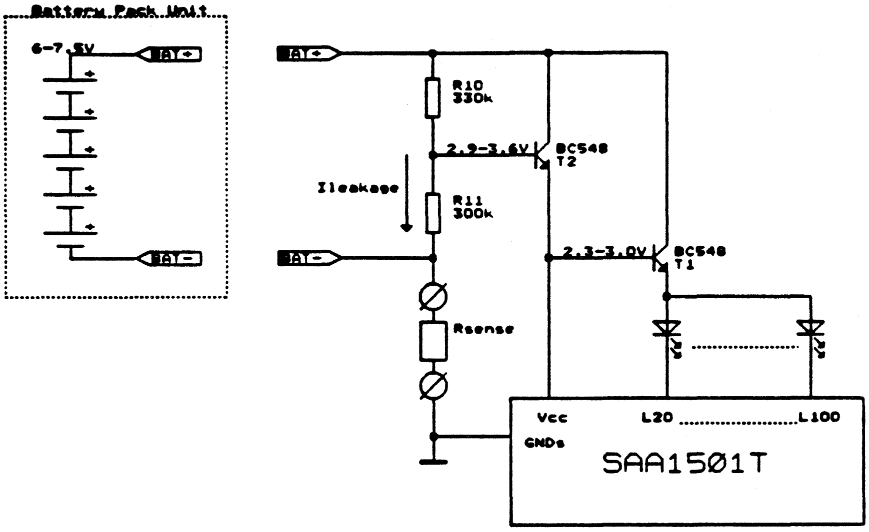
   For an Rsense=70mΩ a maximum dissipation in Rsense of $P_{Rsense} = (1.4)^2*70E-3 =$ 0.085W is found.

3) The values for the charge and discharge converting resistors are found using the formulae in [14] (report NPO/AN9401), in which for the Charge and Discharge efficiency a factor of 70% is used, so that the SAA1501T will operate under-estimated assuring that empty batteries will never occur prior to battery low indication.
Thus for Rc a value of Rc = (1.2*70E-3)/(12E-6*0.7) ~= 10kΩ is found, whereas for Rd a value of Rd = (1.2*70E-3)/(12E-6*(1/0.7) ~= 5.1kΩ is found.

4) The minimum value for Ccnt follows from formula[15] (NPO/AN9401), in which it is preferrable to set Iref1 to 10μA (2.5*Iref), as stated in section 6.1 of report NPO/AN9401.
In this application example for Ccnt a value of 0.68μF is found.

To avoid continuous switching of the signal at pin CNT between Vh and Vl, Vh-Vl must be larger than ΔVmax = (Iref1*Tclk)/Ccnt. This is the case in this example: Vh-Vl=0.4V and ΔVmax = 0.24V.

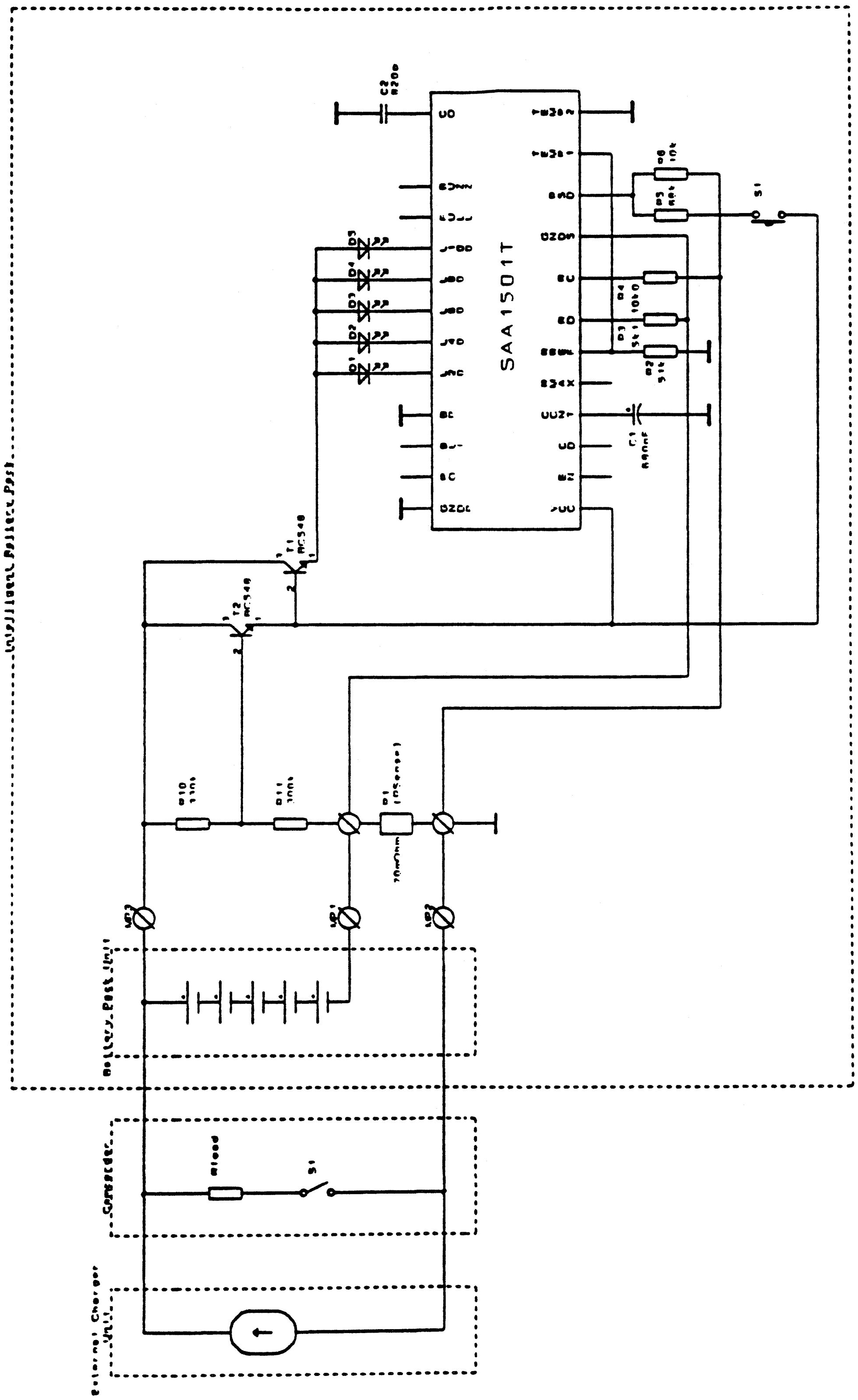
5) As only the voltage (6-7.5V) from the complete battery pack unit is accessible and the SAA1501T operates in the range 2.0-4.5V only, this means that the 5-cell voltage must be divided down. As can be seen in the diagram below the voltage division is carried out by series resistors R10 and R11 which are connected in parallel with the battery pack unit. Then these will cause leakage currents below 10μA, which are negilible compared to the self-discharge currents of the batteries. In order to deliver the supply current for the SAA1501T transistor T2 is used, which can be a simple low-power transistor as Icc< =1.3mA. Assuming the minimum battery pack unit voltage of 6V and maximum battery pack unit voltage of 7.5V the applied Vcc voltage range will be 2.3-3V ((300k/(300k+330k))*Vcc - 0.6V).



6) As the LEDs for the state-of-charge indication at full indication must be sourced with a current of 8*5> =40mA, use of an additional transistor (T1) is required.

Appendix A

Dedicated 5-cell Camcorder Intelligent Battery Pack Application



Appendix B

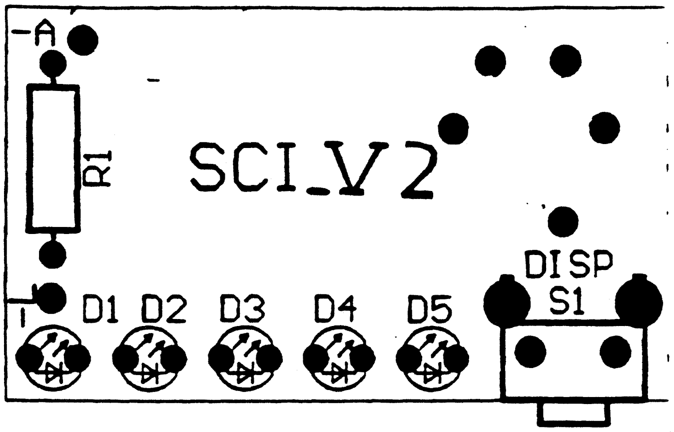
Part list of the application

Component Index	Part Nr.	Value	Case	Tol.	Mat.	Vol.
C1	D2022 029 00167	680nF	A	10%	Tantal	20V
C2	D2222 861 12681	820pF	0805	5%	NPO	63V
R1		70mΩ		5%		
R2	D2322 730 61513	51kΩ	0805	5%		
R3(Rd)	D2322 730 61512	5k1Ω	0805	5%		
R4(Rc)	D2322 730 50103	10kΩ	0805	5%		
R5	D2322 730 61623	62kΩ	0805	5%		
R6	D2322 730 50103	10kΩ	0805	5%		
R10	D2322 730 61304	300kΩ	0805	5%		
R11	D2322 730 61334	330kΩ	0805	5%		
T1,T2	D9335 896 40215	BC548C			SOT23	
IC1	SAA1501T				SO24	
5*LEDs						
1*Push Button						

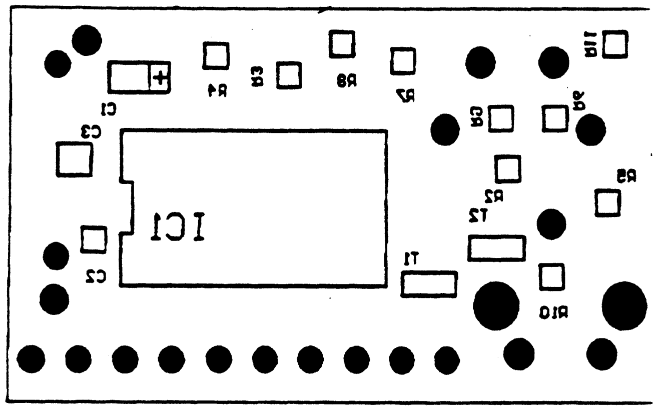
Appendix C

PCB Layout of the dedicated Camcorder Battery Pack Application

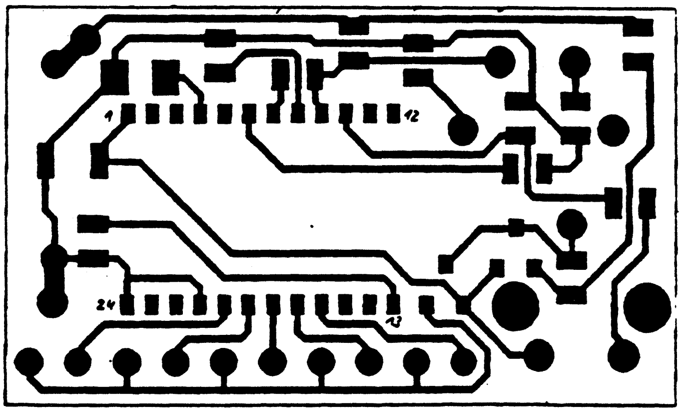
Top view (2.5:1):



Solder-side (2.5:1):



Solder-side: view from the top through the PCB (2.5:1):



APPLICATION NOTE

ICs for Battery Management

Battery Pack Charge Gauge with SAA1501(T) including charge current regulation

Report No: NPO/AN9409.

Guido B Banning

Customer & Product Support Group Consumer ICs Nijmegen, Power Conversion Section,
the Netherlands.

Keywords
State-Of-Charge-Indication
Charge current regulation
Intelligent Battery Packs
LED current compensation

Date : 15 September 1994
Pages: 7

Abstract:

Note NPO/AN9401 dealt with applications for intelligent battery packs, whereby the charge current control was carried out outside the intelligent battery pack (charge current control was incorporated into the charger unit).

As was mentioned in note NPO/AN9401 already the SAA1501T also includes charge current regulation circuitry to be used for mastering the charger unit, so that the charger unit can be very simple; a simple current source would satisfy. This application note deals with an application making use of this concept, i.e. the dedicated Battery Pack application for use with a camcorder as was described in NPO/AN9406 has been expanded with charge current regulation. Further, the application utilizes another feature of SAA1501T viz. 'LED current compensation' not having been dealt with in preceding application notes.
For details on the SAA1501T report NPO/AN9401 should be referred to.

The SAA1501T is a low-cost battery management circuit to be used in fast charging systems. The most important function of SAA1501T is monitoring the charge account in rechargeable batteries. During battery charging the charge current and charge time are registered in a coulomb counter, whereas during discharging the discharge current and time are recorded. The momentary charge account of the batteries can be displayed either on an LC display or by an LED bargraph. In order to get an intelligent battery pack, the state-of-charge indicator function is incorporated into the battery pack.

Table of Contents:

1 The remaining features of SAA1501T detailed

Two issues ('hiding' features) in the SAA1501T have not been discussed so far. The sections below describe these features in full detail.

1.1 The Charge Current Regulation Circuit

In this section the operation of this circuit is set out (refer to figure 1).
The capacitor Cd (at pin 3) is charged up to VHCd with current {Ic-k*Imax} during time interval t1, and subsequently discharged down to VLCd with current k*Imax during time interval t2 (Ic is the down-converted charge current and Imax is defined by Rmax at pin 5, viz. Imax = 210E-3/Rmax), and over and over again. The SAA1501T has three charge current levels (profiles) and the value of factor k depends on the charge level (status) of the SAA1501T. From the resulting integrating signal at pin Cd an ON/OFF signal at pin EN (pin 2) is derived which is on during t1 and off during t2. When solving the equation Q1 = Q2 by using formula Q = C*U and the currents and time intervals as given in the top part of figure 1, it is found that signal EN will regulate to δ*Ic - k*Imax = 0 (δ = t1/(t1+t2)). This signal is to be used for controlling an external charger or for instance a power transistor. The latter is done with the application in this note.

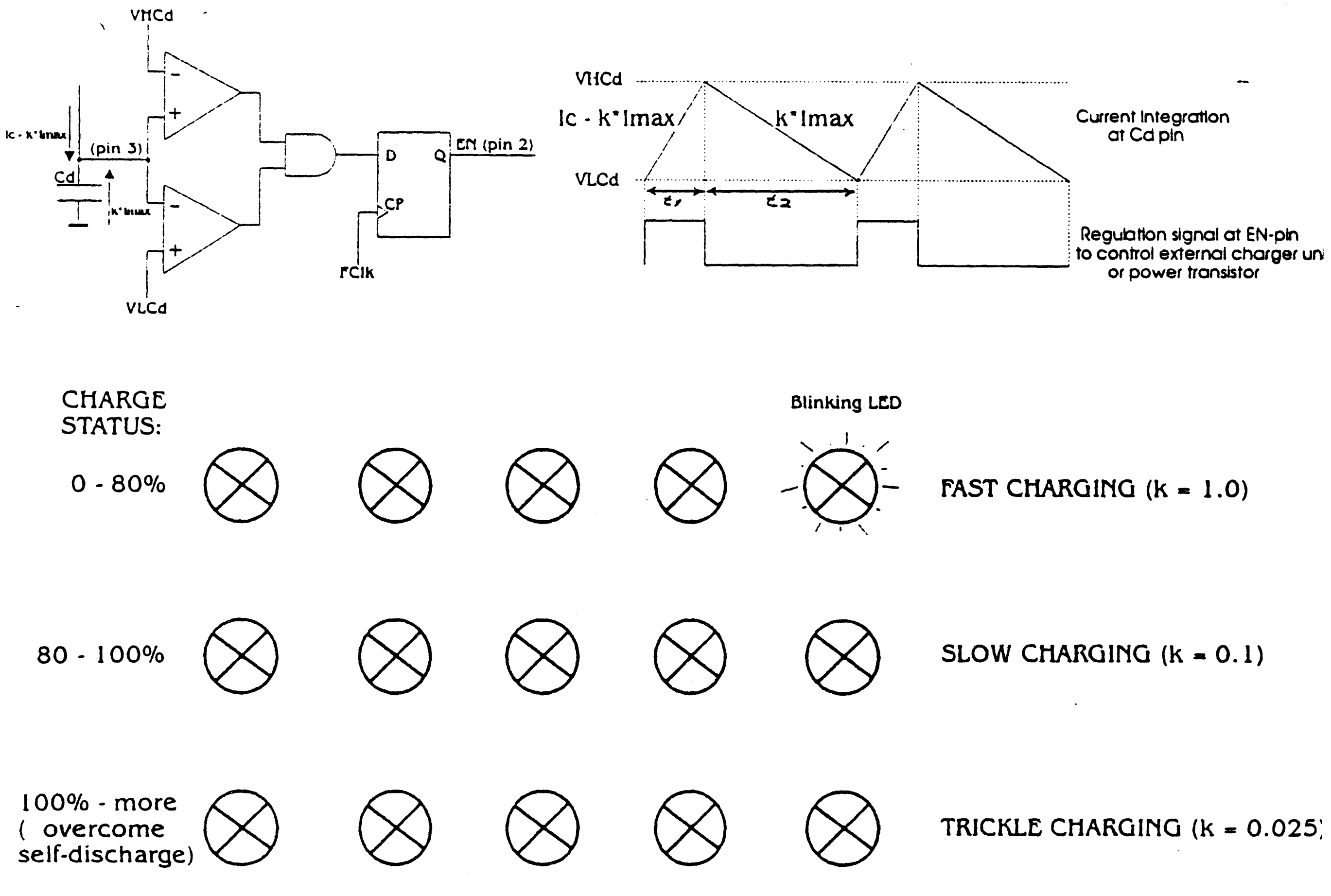


Figure 1: Charge Current Regulation Circuit + LED bargraph visualisation related to the available charge profiles.

1.2 'LED current compensation' by means of signal BP

In NPO/AN9401 it was explained that by means of pin BP the SAA1501T differentiates between LED and LCD mode. Additionally however, in the LED mode the BP produces a so called 'LED current compensation' signal whilst charging when connecting resistor Rcomp and the BP-pin as depicted in figure 2. This way this "additional" discharge current is used to compensate the current consumption used by the LEDs during charging. The mean current value of this signal equals $I_{BP} = N*(I_{LED}/125)$, in which N is a natural number between 1 and 5 depending on the charge status being indicated by L20 to L100. The factor 125 has been chosen so that the down-divided LED-current(s) at BP hardly contribute(s) to the IC's supply current. The 'LED current compensation' signal at BP is a 25% duty cycled signal in order to increase the measure level related to the offset voltage of the V/I convertor and in order to avoid that the SAA1501T will set to the discharge mode whilst charging, i.e. a continuous BP-signal would lead to erroneous influence on the display function.

For designing-in LED current compensation formula Rcomp = 125*Rsense must be used. In the application dealt with in this note therefore a value of 35Ω is found for Rcomp (refer to appendix A).

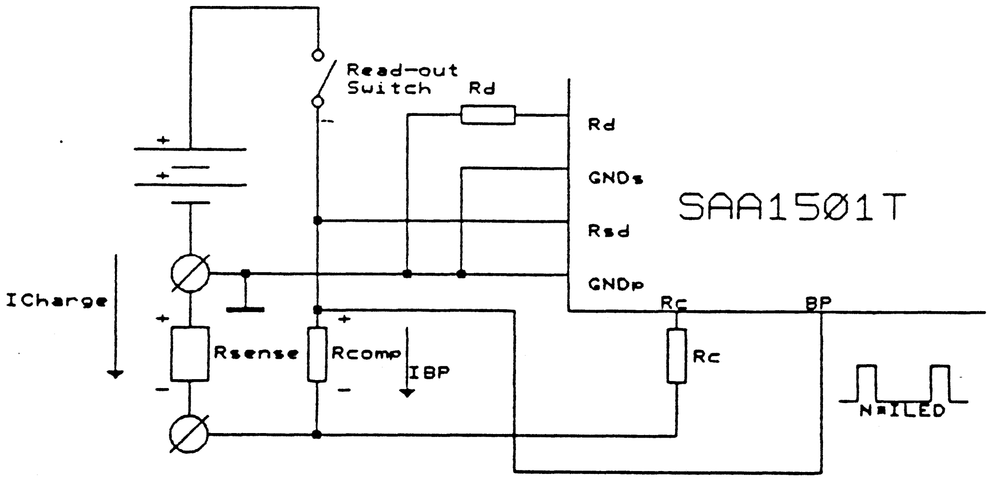


Figure 2: 'LED current compensation' utilized using the multi-functional BP-pin.

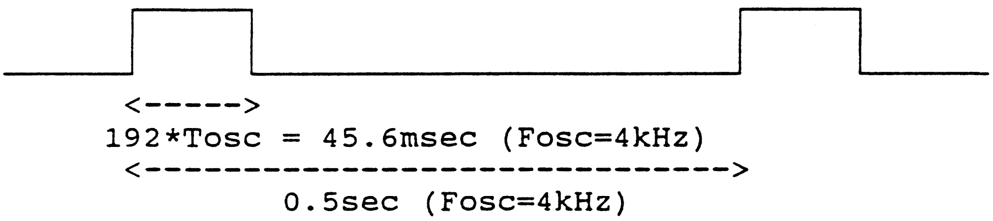
2 Design-in of the Charge Current Regulation function

In this section we will expand 'the dedicated Battery Pack application for use with a camcorder', as was described in NPO/AN9406, with the charge current regulation function, so that the SAA1501T will be in charge of setting the charge current level. This way the external charger can be very simple (no cut-off intelligence); a simple current source would satisfy.
To be able to carry out the design-in (section 2.2) for the charge current regulation function, first full notion of the 'Power Mode Detector'(PMDECT) must be conceived (section 2.1).

2.1 Functional description of the PMDECT

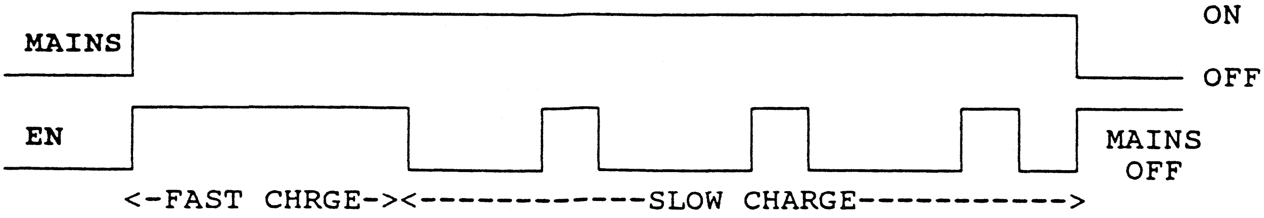
The PMDECT recognizes three modes of operation, viz. the Charge mode, the Discharge mode and the StandBy mode. To save supply current in StandBy mode some circuit blocks in the SAA1501T are switched off including the U/I convertors (for monitoring the charge and discharge currents). To be able to leave the StandBy mode and set to Charge or Discharge mode and thus (re-)activate monitoring of the charge and discharge currents, in StandBy mode the U/I convertors are "sensed" via the 'sense control signal' as depicted below.

'Sense control signal':



Let us now consider the Charge mode only. During the 45.6msec ON-interval of the 'sense control signal', the PMDECT needs to detect an Icharge (Ic) for a time interval > = 8msec (signal EN='1') to be able to set to Charge mode. In the first instance one could think that when choosing the value for capacitor Cd (at pin 3) one should meet the requirement that the ON-time of the EN-signal (which is being derived from the signal at Cd) must be > = 8msec. However, in case that the ON-time in the FAST CHARGE mode > = 8msec - and this is more often the case than not -, the ON-time in the subsequent SLOW and TRICKLE charge modes of the EN-signal may be less than 8msec. This is because then the PMDECT will be able to set to Charge mode during FAST CHARGE and will remain (EN='1') in Charge Mode when the SAA1501T moves to SLOW or TRICKLE CHARGE as t_{ON} < 8msec there is not enough time to detect StandBy mode. In case Power (external current source is switched off) is disconnected, signal EN will remain '1' and there will be enough time (t_{ON} > > 8msec) for the PMDECT to set to StandBy mode (which then should be).
It should be noted that in most cases the charge current regulation circuitry is designed-in so that the ON-time of the EN-signal is 100% during FAST CHARGE.

In short it can be concluded - assuming that the application meets the conditions as given above - that as far as the PMDECT circuit is concerned there is no restriction to the minimum value of Cd. However, as far as the oscillator period Tosc (250μA at Fosc=4kHz) is concerned the minimum ON-time of the EN-signal must be 250μs at minimum (Note that the response time (Slew rate) of the U/I convertors is negligible compared to Tosc). Further, there are no restrictions to the maximum t_{ON} time, however the higher the frequency of signal EN the faster the response time of the regulation loop.



2.2 Application using SAA1501T's charge current regulator

The application:
- Camcorder application as described in NPO/AN9406 (refer to this note for design-in of the 'coulomb counter circuitry' and the 'supply circuitry').
- ICharge = 1A (regulated charge current by SAA1501T in FAST CHARGE mode); ISource = 1.5A (current from external current source). Refer to appendix A for the circuit diagram.

1) Thus in this application example Isource > ICharge. We need to "tell" the charge current regulation circuitry to regulate at 1A (ICharge) so that δ_{EN} will be (1/1.5)*100 = 66.7%. This means that Imax must be equal to Ic during FAST CHARGE:

$$Imax = Ic = \frac{ICharge * Rsense}{Rc},$$

which follows from formula [3] in NPO/AN9401. With ICharge = 1A, Rc = 10kΩ (as was found in NPO/AN9406) and Rsense = 70mΩ, for Imax = Ic = 7μA is found. Thus for Rmax (pin 5) a value of 210E-3/7E-6 = 30kΩ must be used.

2) The value of Cd can be found using formula:

$$Cd = \frac{ICd_{mean} * tEN}{UHCd - ULCd} = \frac{ICd_{mean} * tEN}{0.45} = \frac{k * Imax * tEN}{0.45}$$

$$and\ tEN = \frac{ISource}{ICharge_{regulated} * k} * tON_{\min}$$

When assuming a minimum tONmin = 1mSec for this application example during TRICKLE CHARGE (k = 0.025), we will find a period time of the EN-signal of tEN = 60mSec.
For the minimum demand of Cd (Cd_{min}) it is found that:

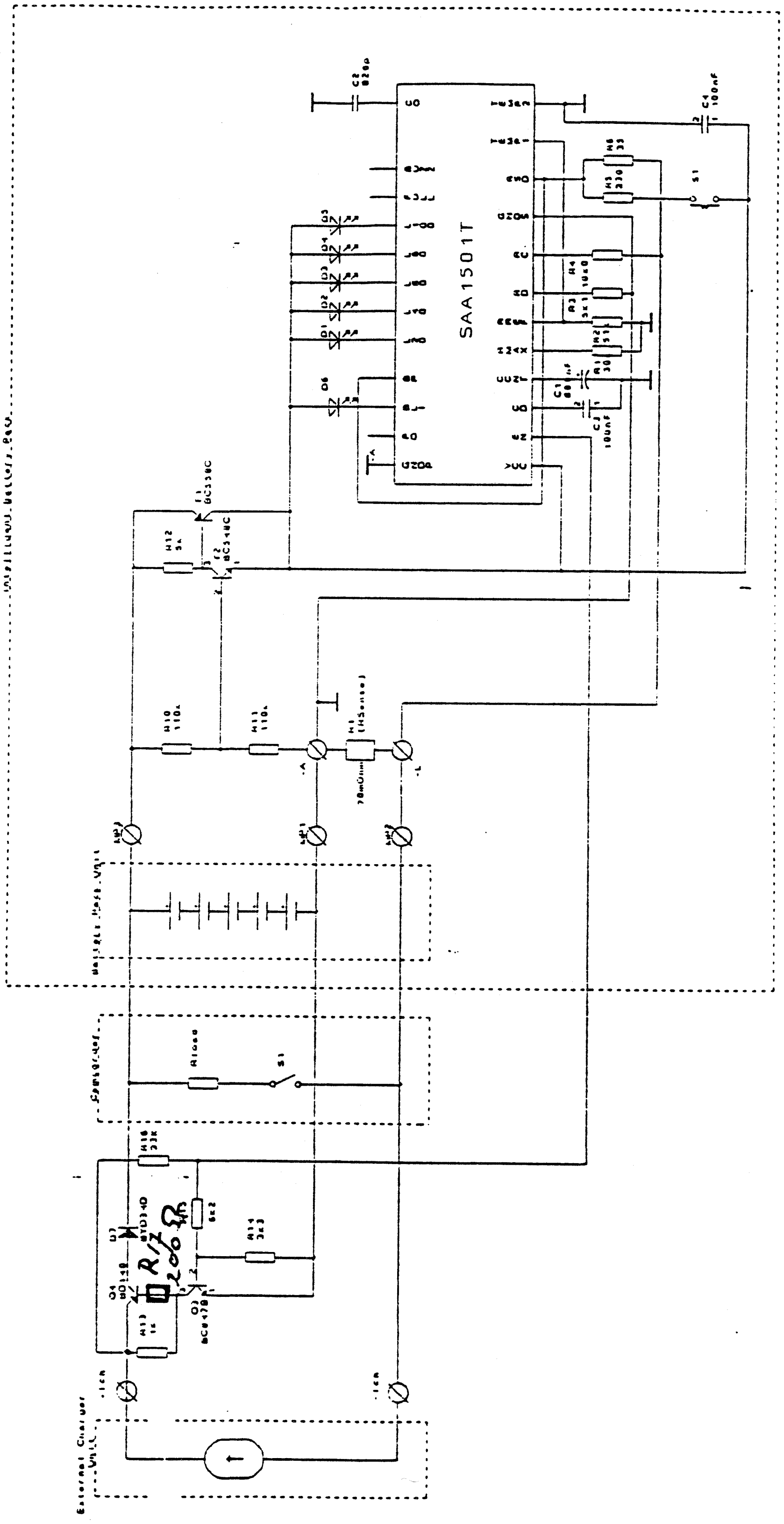
$$Cd_{\min} = \frac{Imax * Isource * tONmin_{EN}}{ICharge * 0.45}$$

so that for Cd_{min} a value of (7E-6*1.5*1E-3)/(1*0.45) = 23nF is found. Closest commercially available value for Cd is 27nF.
In SLOW CHARGE mode we then find f_{EN} = 57.5Hz (t_{ON} ~= 1.73mSec (0.1*1/57.5)), whereas in TRICKLE CHARGE mode we then find f_{EN} = 14.4Hz (t_{ON} ~= 1.73mSec (0.025*1/14.4)).

3) The circuitry connected up to pin EN (pin 2) is to control the charge current via power transistor BD140. Resistor R16 is necessary to pull the initially tristated EN-pin to '1'(Vcc) so that FAST CHARGE is activated. R17 is to restrict the current when 0V is applied to the base of the BC847B (transistor which controls the power transistor BD140).

Appendix A

Expanded 'Dedicated 5-cell Camcorder Intelligent Battery Pack' Application



Appendix B

Part list of the application

Component Index	Part Nr.	Value	Case	Tol.	Mat.	Vol.
C1	D2022 029 00167	680nF	A	10%	Tantal	20V
C2	D2222 861 12681	820pF	0805	5%	NPO	63V
R1		70mΩ		5%		
R2	D2322 730 61513	51kΩ	0805	5%		
R3(Rd)	D2322 730 61512	5k1Ω	0805	5%		
R4(Rc)	D2322 730 50103	10kΩ	0805	5%		
R5	D2322 730 61623	62kΩ	0805	5%		
R6	D2322 730 50103	10kΩ	0805	5%		
R10	D2322 730 61304	300kΩ	0805	5%		
R11	D2322 730 61334	330kΩ	0805	5%		
T1,T2	D9335 896 40215	BC548C			SOT23	
IC1	SAA1501T				SO24	
5*LEDs						
1*Push Button						

Philips Semiconductors – a worldwide company

Argentina IEROD, St. Juramento 1991 - 14 B, Buenos Aires.
Zip Code 1428, Phone/Fax: 54 1 7869367

Australia 34 Waterloo Road, NORTH RYDE, NSW 2113,
Tel, (02)8054455, Fax. (02)8054466

Austria Triester Str. 64, 1101 WIEN,
Tel. (0222)60 101-0, Fax. (0222)60 101-1975

Belgium 80 Rue Des Deux Gares, B-1070 BRUXELLES,
Tel. (02)5256111, Fax. (02)5257246

Brazil Rua do Rocia 220, SAO PAULO-SP, CEP 4552,
P.O. Box 7383, CEP 01051, Tel. (011)829-1166,
Fax. (011)829-1849

Canada DISCRETE SEMICONDUCTORS, 601 Milner Ave,
SCARBOROUGH, ONTARIO, M1B 1M8,
Tel. (416)292-5161
INTEGRATED CIRCUITS, 1 Eva Road, Suite 411,
ETOBICOKE, Ontario, M9C 4Z5, Tel. (416)626-6676

Chile Av. Santa Maria 0760, SANTIAGO,
Tel. (02)773816

Colombia Carrera 21 No. 56-17, BOGOTA, D.E.,
P.O. Box 77621, Tel. (01)2497624

Denmark Prags Boulevard 80, PB 1919, DK-2300
COPENHAGEN S, Tel. 32-883333, Fax. 32-960125

Finland Sinikalliontie 3, SF-2630 ESPOO,
Tel. 358-0-50261, Fax. 358-0-520039

France 117 Quai du Président Roosevelt, 92134 ISSY-LES-MOULINEAUX Cedex, Tel. (01)40938000,
Fax. (01)40938127

Germany Burchardstrasse 19, D-2 HAMBURG 1,
Tel. (040)3296-0, Fax. (040)3296213

Greece No. 15, 25th March Street, GR 17778 TAVROS,
Tel. (01)4894339/4894911

Hong Kong 15/F Philips Ind. Bldg., 24-28 Kun Yip St.,
KWAI CHUNG, Tel. (0)-4245121, Fax. (0)4806960

India Shivsagar Estate 'A' Block, P.O. Box 6598, 254-D
Dr. Annie Besant Rd., BOMBAY-40018,
Tel. (022)4921500-4921515, Fax. (022)49419063

Indonesia Setiabudi 11 Building, 6th Fl., Jalan H.R. Rasuna
Said, P.O. Box 223/KBY, Kuningan, JAKARTA 12910,
Tel. (021)517995

Ireland Newstead, Clonskeagh, DUBLIN 14,
Tel. (01)693355, Fax. (01)697856

Italy Piazza IV Novembre 3, 1-20124 MILANO,
Tel. (02)67522642, Fax. (02)67522648

Japan Philips Bldg 13-37, Kohnan 2-chome, Minato-ku,
TOKIO 108, Tel. (03)813-3740-5101,
Fax. (03)81337400570

Korea (Republic of) Philips House, 260-199 Itaewon dong,
Yongsan-ku, SEOUL, Tel. (02)794-5011, Fax. (02)798-8022

Malaysia 3 Jalan SS15/2A SUBANG, 47500 PETALIN JAYA,
Tel. (03)7345511, Fax. (03)7345494

Mexico Paseo Triunfo de la Republica, No. 215 Local 5,
Cd Juarez CHI HUA HUA 32340, Tel. (16)18-67-01/02

Netherlands Postbus 90050, 5600 PB EINDHOVEN,
Tel. (040)783749, Fax. (040)788399

New Zealand 2 Wagener Place, C.P.O. Box 1041, AUCKLAND,
Tel. (09)894-160, Fax. (09)897-811

Norway Box 1, Manglerud 0612, OSLO,
Tel. (02)748000, Fax. (02)748341

Pakistan Philips Markaz, M.A. Jinnah Rd., KARACHI-3,
Tel. (021)725772

Peru Carretera Central 6.500, LIMA 3, Apartado 5612,
Tel. 51-14-350059

Philippines PHILIPS SEMICONDUCTORS PHILIPPINES Inc,
106 Valero St. Salcedo Village, P.O. Box 911, MAKATI,
Metro MANILA, Tel. (63-2)810-0161, Fax. (63-2)8173474

Portugal Av. Eng. Duarte Pacheco 6, 1009 LISBOA Codex,
Tel. (019)683121, Fax. (019)658013

Singapore Lorong 1, Toa Payoh, SINGAPORE 1231,
Tel. 3502000, Fax. 2516500

South Africa 195-215 Main Road, JOHANNESBURG 2000,
P.O. Box 7430, Tel. (011)8893911, Fax. (011)8893191

Spain Balmes 22, 08007 BARCELONA,
Tel. (03)3016312, Fax. (03)3014243

Sweden Tegeluddsvägen 1, S-11584 STOCKHOLM,
Tel. (0)8-7821000, Fax. (0)8-6603201

Switzerland Allmendstrasse 140-142, CH-8027 ZÜRICH,
Tel. (01)4882211, FAX. (01)4828595

Taiwan 581 Min Sheng East Road, P.O. Box 22978,
TAIPEI 10446, Tel. 886-2-5097666, Fax. 88625005899

Thailand 283 Silom Road, P.O. Box 961, BANGKOK,
Tel. (02)233-6330-9

Turkey Talatpasa Cad. No. 5, 80640 LEVENT/ISTANBUL,
Tel. (01)1792770, Fax. (01)1693094

United Kingdom Philips Semiconductors Limited, P.O. Box
65, Philips House, Torrington Place, LONDON,
WC1E 7HD, Tel. (071)4364144, Fax. (071)3230342

United States INTEGRATED CIRCUITS, 811 East Arques
Avenue, SUNNYVALE, CA 94088-3409,
Tel. (800)227-1817, Ext. 900, Fax. (408)991-3581
DISCRETE SEMICONDUCTORS, 2001 West Blue Heron
Blvd., P.O. Box 10330, RIVIERA BEACH, FLORIDA 33404,
Tel. (407)881-3200, Fax. (407)881-3300

Uruguay Coronel Mora 433, MONTEVIDEO,
Tel. (02)70-4044

Venezuela Calle 6, Ed. Las Tres Jotas, CARACAS, 1074A,
App. Post. 78117, Tel. (02)2417509

Zimbabwe 62 Mutare Road, HARARE, P.O. Box 994,
Tel. 47211

**For all other countries apply to: Philips Semiconductors,
International Marketing and Sales, Building BAF-1,
P.O. Box 218, 5600 MD, EINDHOVEN, The Netherlands,
Telex 35000 phtcnl, Fax. +31-40-724825**

SCB6

The information presented in this document does not form part of any quotation or contract, is believed to be accurate and reliable and may be changed without notice. No liability will be accepted by the publisher for any consequence of its use. Publication thereof does not convey nor imply any license under patent- or other industrial or intellectual property rights.

Printed in the Netherlands **Date of release: 6-'92**

Philips Semiconductors

PHILIPS

PHILIPS