INTEGRATED CIRCUITS

Semiconductors for Wired Telecom Systems

1997

Application Handbook IC03b

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PREFACE

Thank you for your interest in wired telecom products from Philips Semiconductors. As a leading supplier to the telecom market, we offer a wide range of discrete and integrated semiconductor components.

This Application Handbook is largely a compilation of already-published laboratory reports from several Philips Application Laboratories. It contains extensive practical information for those designing-in Philips Semiconductors ICs into a variety of telecom sets - from the simple to the most advanced featurephone. It supplements the latest information given in product data sheets and in the Philips Semiconductors data handbook 'Semiconductors for Wired Telecom Systems' (IC03a).

Whilst the individual reports have been grouped and ordered logically, no attempt has been made to make a continuous story, nor to edit individual reports to avoid repetition of subject matter. In addition, though the principles described remain valid today, some of the components solutions have been superseded; no attempt has been made to replace old type numbers.

More information is available from Philips System Laboratories world-wide.

July 1996

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Chapter 8 REFERENCE MATERIAL

APPLICATION NOTE SUMMARIES

Below are summaries of application notes available for wired telecom systems. Application notes marked with an * are published in this Application Handbook for Wired Telecom Systems (IC03b). For all other application notes, please contact Philips Semiconductors.

Note: The stated references refer to the actual application report.

* AN94016

Basics of PCA1070: a programmable analogue CMOS transmission IC (PACT)

Details: 102 pages, March 1994

The PCA1070 opens the possibility to manufacture a programmable transmission part for use in the telephone line interface of electronic devices such as electronic telephone sets and feature phones, cordless phones, answering machines, modems and fax equipment.

The PCA1070 is a CMOS integrated circuit performing speech and line interface functions. It needs a minimum number of external components. The transmission parameters are programmable via the I^2C-bus. This makes the IC adaptable to nearly all country requirements in the world, and to a various range of speech transducers, without changing the (few) external components of the basic application. With the PCA1070 the number of printed circuit board versions and the number of components is minimised. This reduces the costs of logistics and manufacturing operation extensively. For some countries a 12 kHz or 16 kHz tax pulse filter has to complete the design.

This report gives a detailed description of the PCA1070 and its basic application in electronic telephone sets. Also EMC aspects, protection and tax pulse filtering are discussed. Furthermore an application example of PCA1070 together with a pre-programmed µC PCD3353A/008 is given and some measurement results of this application are shown.

AN94017

Tax pulse filters at 12 and 16 kHz for the PCA1070

Details: 13 pages, March 1994

Tax pulse filters for 12 and 16 kHz metering systems were designed for the PCA1070. With these filters, the BRL of all West-European countries can be met by adaptation of the programmable set impedance, if needed. The range of the programmable sidetone impedance was adequate for the test method used. For the filter design an exchange is possible between receive gain of the PCA1070, the filter Q-factor and spread of filter components. For Switzerland with 12 kHz/10 Vrms tax pulse the maximum receive gain is +6 dB for a filter with a Q of 15,3% inductors and 5% capacitors. For Germany with 16 kHz/9.8 Vrms the maximum receive gain is +10 dB for a Q-factor of 20,3% inductors and 5% capacitors.

AN94021

French current-regulation circuit for the PCA1070

Details: 17 pages, March 1994

This report describes the application of the programmable transmission circuit PCA1070 in the French current-regulation system. A detailed circuit is given with measurements of the V-I characteristic, BRL, side tone, send and receive gain and line current during dialling and start-up. The measurements show that the influence of the regulation circuit on the original transmission parameters is small and that the combination fulfils the requirements.

*** AN94069 (SUPERSEDES ETT/AN94002)**

High-end telephones with PCA1070, TEA1093 and PCD3755A

Details: 104 pages, September 1994

The PCA1070 programmable transmission circuit, the TEA1093 handsfree circuit and the PCD3335x μC can be used to design a high-end handsfree telephone set. With PCA1070 the number of printed circuit board versions to suit the various country transmission requirements is minimised because many country and control parameters can be programmed via the I^2C-bus. The number of components that are needed for the application is reduced when compared to an application using a TEA106x family member. This reduces the cost of logistics and manufacturing operation extensively. For some countries a 12 kHz or 16 kHz tax pulse filter has to complete the design. Additionally the PCA1070/PCD335x combination offers the possibility to realise extra features by means of software (e.g. hold function, NSA or DMO, multiple anti sidetone).

This report describes hardware and software design considerations for a high-end telephone set application with PCA1070, TEA1093 and the PCD3755A. The PCD3755A is a One Time Programmable (OTP) μC which supports the PCD3351A, 52A and 53A μC. In the report also EMC aspects and protection are discussed. Furthermore some measurement results of the described application with a pre-programmed μC PCD3755A are given. A complete functional description of the SW used for the evaluation is given in Appendix D.

This report is not meant to give a turn-key application solution but merely gives first application ideas and test results that may be helpful to start a design of a high-end telephone set with PCA1070, TEA1093 and PCD335x. The printed circuit board which is mentioned in this report is intended for evaluation purposes only and is not meant to be a demonstration board. The application is intended for countries which require voltage regulation for DC termination of the telephone line.

*** AN95023**

Application of the UBA1702/A line-interrupter driver and ringer circuit

Details: 57 pages, November 1995

The UBA1702/A performs the high voltage interface and ringer functions for corded telephone sets. The IC itself, its typical behaviour and the combination of this circuit with a transmission IC (a member of the TEA106x family) and a micro-controller are described in detail. Information on adjustments to fulfil country requirements and several application proposals are given.

*** AN95050**

Application of the TEA1112 and TEA1112A transmission circuits

Details: 67 pages, November 1995

The TEA1112 and the TEA1112A are bipolar transmission circuits for use in electronic telephone sets. They are added to the range of well-known transmission circuits of the TEA1060-family.

This report contains a detailed description of the circuit blocks of the TEA1112 and the TEA1112A. Two application examples of the TEA1112 are given. The report handles the consecutive steps to design or to adjust the basic application with these ICs The EMC behaviour of an evaluation board with the TEA1112 or the TEA1112A is included.

The General notation in this report for both ICs, TEA1112 and TEA1112A, is: TEA1112/A.

*** CTT/AN95083**

TEA1095 voice-switched speaker-phone IC

Details: 42 pages, September 1995

The TEA1095 is a high-performance, low-power consumption voice switched speakerphone IC designed for integration of a handsfree function in terminal environments.

A detailed description of the IC, advises on the adjustments and a worked-out application example are contained in this report.

EIE/AN91001

Workbench EMC evaluation method

Details: 18 pages, December 1991

With the proposed EMC workbench evaluation method small application boards, printed wiring boards containing one or more ICs can be verified on basic EMC parameters such as Radio Frequency (RF-) emission and immunity over the frequency range from 150 kHz to 230 MHz, or even 1 Ghz.

This EMC evaluation method can be used in a qualification procedure to select between new electronic circuit (systems) implementations from various supplies or as a design tool when developing new products based on modular circuit blocks.

EIE94006

A comparison of test signals for testing the RF immunity with burst-mode disturbances

Details: 18 pages, November 1994

It is expected that RF immunity standards for all electronic equipment will soon be extended with requirements on GSM and DECT frequencies. As the existing test methods use sine wave amplitude modulated RF signals, it is interesting to know the relation to the newly proposed block pulse modulation.

For the judgement of self-pollution, e.g. in a DECT base station, it is also required to look at the psophometrically weighted noise that is added as noise to the line. This report presents a relation between the results as obtained with various test signals.

ESG8801

An evaluation method to characterize the EMC performance of PCBs containing ICs

Details: 44 pages, July 1988

The evaluation of the ElectroMagnetic Compatibility (EMC) performance of Printed Circuit Boards (PCBs) is the most essential part to obtain proper EMC performance of the entire product from the very start of development. These methods can even be applied to evaluate the performance of Integrated Circuits (ICs) either as single device or implemented in a typical application.

It is tried to establish an evaluation method without ambiguity, which is comprehensive and economic in three senses:

- The initial costs of the verification tools
- The ease to perform the measurements
- The benefits: reduction of development throughput-time and re-design.

From the theory and practical experience, based on this EMC-performance evaluation procedure, it can be concluded that only a small number of tests are necessary to determine whether or not a PCB fulfils certain requirements.

Application notes summaries

ESG89001

Electromagnetic compatibility and printed circuit board (PCB) constraints

Details: 30 pages, April 1989

Today's Printed Circuit Board designs often fail to comply with Electro Magnetic Emission and immunity requirements applicable for complete products. Herewith, guidelines are given which will enable the designer to take those precautions necessary to tailor PCBs, that they will fulfil those requirements.

*** ETT/AN8903**

Galvanic separation of the I^2C-bus

Details: 11 pages, March 1989

In some telephone applications, there is a need for galvanic separation of the I^2C-bus, for example if I^2C devices coupled to the bus are not fed from the same internal power supply. In this note, a technical solution is given where the I^2C-bus specification can be met. As galvanic separation device, an optocoupler is used.

*** ETT/AN89008**

Application of the speech-transmission circuit TEA1062

Details: 24 pages, October 1989

The TEA106x family consists of a range of bipolar integrated circuits performing all speech and line interface functions required in fully electronic telephone sets. The TEA1060 family designers guide (lit.1) provides information on most of the members of the TEA106x. The TEA1062 is not described in this guide. Detailed information about this circuit can be found in the data sheet (lit.2).

The TEA1062 is a low-voltage speech-transmission IC able to operate down to a dc line voltage of 1.6 V to facilitate the use of more telephone sets in parallel. The TEA1062 has a modified performance and less features compared to the TEA1067.

In this report differences between the TEA1062 and the TEA1067 as described in lit.1 are elucidated. When applying the TEA1062 this report should be used in combination with the designers guide.

The figures given in this report all refer to the TEA1062. The TEA1062 is described in this report with respect to the basic application circuit shown in fig.1 (lit.3)

*** ETT/AN90005**

The TEA1064A with complex set impedance and complex line termination

Details: 8 pages, February 1990

More and more PTTs require a complex set impedance for telephone sets. It can be expected that the sending frequency curves will be measured with a complex line termination instead of 600 Ω. The German PTT is the first one that actually changed over from 600 Ω to a complex termination. Furthermore they require a better accuracy of the complex set impedance (balance return loss up to 18 dB). To prevent high frequency roll-off in sending direction with the TEA1064A in such a case, the application hint mentioned in this report should be considered.

*** ETT/AN90017**

Software controlled ringer for German market

Details: 32 pages, July 1990

In sets with the listening-in feature, it is possible to generate a ringing signal via the loudspeaker instead of using a separate mechanical bell or PXE. By using the Philips components TEA1064A (speech-transmission), TEA1085 (listening-in), PCD3312 (DTMF-generator), PCD3346 (micro-controller plus EEPROM) and a power converter it is possible to do this software controlled such that both the German TEL02 ringer requirements and the speech-transmission requirements are fulfilled.

*** ETT/AN91010**

User's manual PR4535x DEMO board with TEA1083A-TEA1064 call progress monitoring application.

Details: 25 pages, June 1991

The printed circuit board PR4535x demonstrates the TEA1083A call progress monitoring circuit in combination with the TEA1064A transmission IC. The TEA1083A may be replaced by the TEA1083.

This report describes the printed circuit board and gives some application information. However, it is not intended to be an application report. Refer to the literature given in chapter 7 for application details.

*** ETT/AN91012**

User's Manual - PR4516X; demo board with the TEA1085A-TEA1064A listening-in application

Details: 29 pages, July 1991

The printed circuit board PR4516x demonstrates the TEA1085A listening-in IC in combination with the TEA1064A as transmission IC. This report describes the printed circuit board and gives some application information. However it is not intended to be an application report. Refer to the literature given in chapter 7 for application details.

First version: ETT/AN90006, May 1990. Updates: ETT/AN91012, July 1991: TEA1085 replaced by the TEA1085A. Demonstration board PR4516x is delivered with the TEA1085A. The TEA1085A may be replaced by the TEA1085.

ETT/AN91013

Micro-controller PCF84C12/xxx - for PCD4440 scrambler/descrambler applications

Details: 18 pages, May 1991

An analogue voice scrambler device, the PCD4440 needs to be driven by a controller circuit via the serial inputs SDA (Serial Data), and SCL (Serial Clock), to switch the device to the correct split frequencies in the time domain. Also other features like crystal oscillator frequency generation, initialisation, transparent mode and mute mode switching can be done via the serial bus, the I^2C-bus.

In a system where no controller function is needed for other tasks, this micro-controller can provide these functions. It includes different feature setting via keyboard inputs, while the controller can be switched to the AUTOMATIC or BCD modes so that it can operate without a keyboard as well.

ETT/AN91014

Demonstration board PR45284 for PCD4440 scrambler and NE577 compander

Details: 16 pages, May 1991

The PCD4440 scrambler/descrambler IC has been developed for use in cordless telephones for the purpose of scrambling and descrambling the analogue speech signals which are transmitted by a radio link between handsets and base units. This report describes a printed circuit board designed to exercise the PCD4440, which in combination with the NE577 compander IC provides an excellent vehicle for the purposes of demonstration and measurement. A micro-controller is used to control the PCD4440 via an I^2C-Bus.

*** ETT/AN91016**

TEA1085A/TEA1085 - a listening-in facility for electronic telephone sets

Details: 64 pages, September 1991

The TEA1085A and the TEA1085 are designed for use in line powered telephone sets. Besides the listening-in function, they incorporate an effective dynamic limiter and a Larsen Level Limiter to reduce annoying howling effects.

They have to be combined with a transmission IC of the TEA1060 family. Nearly all line current can be utilised for powering the loudspeaker without affecting the transmission characteristics. This report describes the TEA1085A/TEA1085 and an application with the TEA1064A.

*** ETT/AN92010**

User's manual OM4723 demo board: PCD3330-1/TEA1067/TEA1083A feature-phone application

Details: 37 pages, July 1992

This feature-phone board OM4723 is designed to demonstrate the PCD3330-1: a mixed-mode multi-standard repertory dialler/ringer with EEPROM. The on-hook dialling feature is realised with the TEA1083A call progress monitoring IC.

The main components on the board are:

- Multi-standard repertory dialler/ringer IC-PCD3330-1
- Versatile Speech/Transmission IC-TEA1067
- Call progress monitoring IC-TEA1083A
- Line current interrupter DMOST-BSP254A.

Application notes summaries

ETT/AN92011

Software specification to control the PCA1070 PACT-IC (PCD3353A/008)

Details: 38 pages, January 1993

This report gives the objective target specification of demonstration and test software to control the PCA1070 which is a Programmable Analogue CMOS Transmission (PACT) circuit, with the PCD3353A micro-computer.

The PCD3353A/008 is a member of the PCD3353A micro-computer family, special developed for telephone applications. This PCD3353A micro-controller family has on-board DTMF generator, EEPROM and a special ringer output and is fully able to control the PCA1070 via its software I^2C-bus.

The PCA1070/PCD3353A circuits combination performs the following functions:

- Pulse and DTMF dialling
- Redial and repertory dial stored in EEPROM
- PCA1070 control
- PCA1070 variables stored in EEPROM
- PCA1070 variable programmable via I^2C-bus and Key-board
- Dialling options programmable via keyboard stored in EEPROM
- Ringer functions incorporated 3-tone, 4-speeds and 4 volumes
- Display.

The program will firstly be used for evaluation of the first samples of this PCA1070 in application with the PCD3353A micro-computer.

To use it for demonstration purposes at customers and to complete the dynamic evaluation a mask programmed version (PCD3353A/008) is made.

ETT/AN93010

User's Manual of the OM4737 evaluation kit for the PCA1070

Details: 8 pages, July 1993

This report describes the OM4737 evaluation kit. The kit is intended to be used for laboratory evaluation of the PCA1070 Multi-standard Programmable Analogue CMOS Transmission IC (PACT).

The kit consists of:

- PCA1070 evaluation board
- I^2C-bus interface board for use with an IBM-compatible PC
- Evaluation software with the I^2C-bus control program (on 3.5" diskette).

*** ETT/AN93015 (SUPERSEDES REPORT NOVEMBER 1993 WITH SAME NUMBER)**

Application of the TEA1093 hands-free circuit

Details: 67 pages, 29 November 1993 (revised 30 November 1995)

The TEA1093 is a hands-free telephone IC, that can be used in combination with a transmission IC of the TEA106x, TEA111x family or the PCA1070, in line powered telephone sets and in mains supplied sets, to extend the function of the transmission IC with a handsfree function. It provides a half duplex connection. The decision logic eliminates back ground noise from speech signals. If both sides are quiet the TEA1093 goes into idle mode to avoid the 'line dead' phenomenon.

This report contains a number of application examples that can be taken as starting point for new designs and application hints that can be used during design phase.

*** ETT/AN93017**

User's manual for the OM4736 demo kit: a multi-standard telephone set with PCA1070 and PCD3353A/008

Details: 34 pages, December 1993

The feature-phone kit OM4736 is designed to demonstrate the programmable transmission IC PCA1070 with the help of the PCD3353A/008 pre-programmed micro-computer.

The PCA1070 is a Multi-standard Programmable Analogue CMOS Transmission (PACT) integrated circuit performing all speech and line interface functions required in fully electronic telephone sets. It needs a minimum number of external components. The transmission parameters are programmable via the I^2C-bus, which makes the IC suitable for nearly all countries in the world without changing the (few) external components.

The PCD3353A/008 has three functions:

- Control of the normal feature phone functions such as pulse/tone dialling, redial/repertory dialling and software controlled ringer function
- Setting of the transmission parameters of the PCA1070, which are stored in the on-chip EEPROM, via the I^2C-bus
- Setting of the dialler and the ringer parameters which are also stored in the EEPROM
- Changing of all the programmable parameters via keyboard to show the flexibility of the total application.

All dialled or programmed numbers, programmed or stored parameters for dialling ringer and PCA1070 control, are shown on an LCD-Display. This report gives a brief description of the application and gives a guide-line for operation of the DEMO kit.

*** ETT/AN94001**

User manual for OM4750: Demonstration board TEA1093 and TEA1094

Details: 16 pages, April 1996

This report describes the TEA1093/TEA1094 demonstration board. The board demonstrates the performance of the TEA1093 and TEA1094 handsfree circuit in conjunction with a TEA1062 or TEA1067 speech transmission circuit. It provides handset, handsfree and listening-in operation. The difference between the TEA1093 and TEA1094 is that the TEA1093 is line powered while the TEA1094 is externally powered. An external power supply is useful, for instance, in answering machines. It has the advantage that a high loudspeaker output power can be realised.

*** ETT/AN94004**

Application of the TEA1094 hands-free circuit

Details: 49 pages, March 1994

A detailed description of the TEA1094 is given. In conjunction with a member of the TEA106x family transmission circuits, it offers a hands-free function. It incorporates a microphone amplifier, a loudspeaker amplifier and a duplex controller with signal and noise monitors on the transmit and the receive channel. A cookbook gives the general application steps and also several application examples are given including listening-in, cordless-base and answering machine.

*** ETT/UM95004.0**

OM4775 user manual - basic phone demonstration board OM4775. UBA1702/A, TEA1062/1064B and PCD3755A

Details: 25 pages, November 1995

The OM4775 is a demonstration board of a basic phone realised with the UBA1702/A line interrupter driver and ringer circuit, the TEA1062/1064B transmission circuit and the PCD3755A micro-controller.

The basic phone is in fact a general application with different demonstration possibilities with respect to interrupter/ringer and transmission ICs. It is not made to fulfil specific country requirements, however adaptation is possible to a certain extent and is made easy by mounting concerning components on sockets or solder pins. These components are underlined in the text of chapter 2 and 4.

The OM4775 contains a handset with microphone and earpiece and a base with the hardware including a piezo-buzzer for ringing and a keyboard. The OM4775 is delivered with the UBA1702, BSP254A, TEA1064B and PCD3755A. The UBA1702 and BSP254A can be replaced by the UBA1702A and MPSA92, the TEA1064B by the TEA1062.

ETT/UM95011.0

OM4776 Evaluation board for TEA1112(A) and TEA1113

Details: 17 pages, October 1995

This document describes the evaluation kit OM4776. This kit is intended to be used for laboratory evaluation of the TEA111x.

The TEA1112 and the TEA1112A are speech transmission ICs for application in electronic telephone sets. They are added to the range of transmission circuits of the TEA106x-family. Besides the required basic interface functions between microphone capsule, earpiece, telephone line and dialler circuit (DTMF and pulse dialling) they offer the user a Hook-Status indicator function (LED output) and a transmit mute function (MMUTE). The EMC behaviour is improved with respect to the TEA106x Ics.

The TEA112A differs from the TEA1112 by the inverted MUTE and the MMUTE input (Ref 1). The TEA1113 differs from the TEA1112 by the inverted MUTE and by the use of a dynamic limiter in combination with MMUTE (Ref. 2).

ETT/UM95013.0

OM4766 evaluation kit for TEA1069N speech/dialler/ringer

Details: 25 pages, February 1996

This document describes the evaluation kit OM4766. This kit is intended to be used for laboratory evaluation of Speech/Dialler/Ringer IC TEA1069N.

The purpose of the evaluation kit is to evaluate and demonstrate the basic features of the TEA1069N one-chip telephone IC. The board behaves as a normal telephone. Using the handset and the keyboard telephone calls can be made, and incoming calls can be accepted. In order to accommodate easy evaluation of the additional features of the TEA1069N it is possible to access the different pins of this chip.

ETT/UM96003.0 (ETT/AN89006 REVISION)

OM4729 basic application board for the TEA1062 and TEA1062A

Details: 25 pages, February 1996

This document describes the evaluation kit OM4729 (formerly known as CAB3422) This kit is intended to be used for laboratory evaluation of the TEA1062(A).

The TEA1062 and TEA1062A are speech transmission ICs for application in electronic telephone sets. They are part of the range of transmission circuits of the TEA106x-family.

The TEA1062A differs from the TEA1062 by the inverted MUTE input (Ref. 1).

*** ETT/UM96006.0**

OM4784 user manual - System board of TEA1069N, TEA1093 and UBA1702/A

Details: 33 pages, March 1996

This document describes the system board OM4784. This board is intended to be used for laboratory evaluation of the Speech/Dialler/Ringer IC TEA1069N in combination with the TEA1093 and the UBA1702/A.

The purpose of the system board is to evaluate and demonstrate the features of the TEA1069N one-chip telephone IC in combination with the TEA1093 (handsfree) and the UBA1702A (line-interface). The board is a feature phone with Music On Hold, keytone, parallel set detection and handsfree.

ETT8503

The coupling network between the PCD3311/12 DTMF generator and the TEA1060/61 transmission circuit

Details: 17 pages, March 1985

The DTMF generator PCD3311/12 needs a transmission circuit as an interface to the telephone line. The report describes the design considerations for a fixed passive coupling network between the generator and a TEA1060/61. The worst case variation of the DTMF voltage on the line will, according to the specifications of both devices, exceed the CEPT recommendation. However, by using statistical methods it can be proved that 99% of the PCD3311/12-TEA1060/61 combinations will meet the recommendation.

*** ETT8612**

Application of the PCD3310 bilingual dialler in electronic telephone sets

Details: 27 pages, October 1986

The PCD3310 is a CMOS integrated dialling circuit with dual-standard dialling for either pulse dialling (PD) or dual tone multi frequency (DTMF) dialling with last-number redial.

This report gives a description of the circuit, the dialling procedures and the subscriber set architectures of the PCD3310. Also the Electro Magnetic immunity compared with other dialling circuits is given.

Application notes summaries

*** ETT8707**

Application of the TEA1081 - a supply IC for peripheral circuits in electronic telephone sets

Details: 45 pages, 15-9-87

TheTEA1081 is intended for application in electronic telephone sets which are powered from the telephone line. This circuit performs the interface function between telephone line and supply line for peripheral circuits. In combination with a transmission circuit of the TEA1060 family it improves the power supply capabilities of the set depending on the available line current.

The TEA1081 is a successor of the TEA1080, major differences with the TEA1080 are the improved performance and an extension of the circuit with a power down function.

Described are the comparison of the TEA1081 with the TEA1080, applications of the TEA1081 with the TEA1060 including its effect on the transmission characteristics, and application of the TEA1081 in a pulse dial telephone set.

*** ETT8710**

Specification of quartz and ceramic resonators for PCD33xx and PCF84cxx CMOS ICs

Details: 34 pages, September 1987

The oscillator of the PCD33xx and PCF84cxx CMOS integrated circuits was originally designed to operate with a quartz resonator. Ceramic (or PXE) resonators form an economically attractive alternative but can not always replace the quartz. A procedure is described which explains how to decide whether or not a ceramic resonator is suited for the oscillator. The report gives also general and more specific information on oscillators, on start-up time and on parasitic effects in quartz or ceramic resonators.

ETT8711

Bridge: A software tool for optimising the TEA1060 anti-sidetone bridge

Details: 11 pages, September 1987

Bridge is a software tool that helps you to find the optimal anti-sidetone balance network for your TEA1060 application. It is designed for use with the IBM PC or compatibles.

Starting with your initial guess BRIDGE can calculate a better balance network resulting in a lower average sidetone level. It can also present you the performance of balance networks graphically (if your computer is equipped with a colour/graphics adapter).

BRIDGE allows you full editing of all relevant parameters and components. Besides sets of parameters/components can be stored on disk for later retrieval.

There are two versions available on the diskette:

- BRIDGE which will run on every IBM PC or compatible
- BRIDGE87 which runs 4 times faster, but only on PCs equipped with a 8087 mathematical co-processor.

ETT8805

Controller (PCD3344/006) programmed for telephone sets with on-hook dialling and up to 20 numbers repertory dial facilities.

Details: 24 pages, July 1988

The PCD3344/006 is a programmed version of the standard telephony micro-controller PCD3344 which has an on-chip DTMF/melody generator. With the software described in this report the /006 behaves as a 20 number repertory pulse/DTMF dialler controlling the listening-in part of the telephone set realised by the TEA1081 supply IC and the low frequency amplifier TDA7050.

*** ETT89008**

Listening-in with the TEA1081, TDA7050 and TEA1064

Details: 27 pages, May 1989

This report describes a listening-in application with the TEA1081 supply IC, the TDA7050 loudspeaker amplifier and the TEA1064 transmission circuit for use in electronic telephone sets.

*** ETT89009**

Application of the versatile speech/transmission circuit TEA1064 in full electronic telephone sets

Details: 84 pages, August 1989

The TEA1064 is a speech/transmission integrated circuit for use in analogue electronic telephone sets. It is added to the range of well-known transmission circuits of the TEA1060-family. The circuit incorporates a relatively powerful supply for peripheral circuits (such as microcomputers and diallers) with two options: a) stabilised supply or b) regulated line voltage. Furthermore a dynamic limiter in the sending channel has been included that on the one hand limits the maximum level (and prevents distortion) of the transmitted line signal and on the other hand improves sidetone performance by reducing the maximum sidetone level and the distortion. This report gives a complete description of the circuit and provides the necessary information to the set-design-engineer in order to enable him to take full advantage of the many application possibilities.

*** ETT89016 (SUPERSEDES ETT8613)**

Measures to meet EMC requirements for TEA1060-family speech/transmission circuits

Details: 19 pages, 1-10-89

Recently the Electro Magnetic Compatibility (EMC) of the TEA1060-family, speech/transmission circuits, have been investigated again. If a few components are placed at the right spots the TEA1060 circuits reach a good performance. This report describes the measures which can be taken without influencing the basic functioning of the IC.

*** ETT94001**

Application of the TEA1096 transmission and listening-in circuit

Details: 60 pages, January 1994

The TEA1096 as well as the TEA1096A are bipolar telephony IC's for use in line powered telephone sets. They offer a transmission function and a group listening-in (or monitoring) facility of the received line signal via a loudspeaker. The TEA1096 and TEA1096A incorporate a line interface with active output stage, a stabilised supply, send and receive amplifiers, a double anti side tone circuit, line loss compensation, a loudspeaker amplifier and dynamic limiters for transmit and loudspeaker signal.

This report gives a detailed description of the TEA1096 and the TEA1096A and an application example. The description is given by means of block diagrams and discussion of the details of the sub-blocks. For product details is referred to the Device Specification ref.1.

ETT94008 (SUPERSEDES ETT/AN92011)

Functional specification of the demo PCD3353A/008 to control the PCA1070 (PACT)

Details: 39 pages, May 1995

This report gives a description of demonstration and test software to control the PCA1070 which is a Programmable Analogue CMOS Transmission (PACT) circuit, with the PCD3353A microcomputer.

The PCD3353A/008, a member of the PCD335xA family of micro-computer special developed for telecom applications. This PCD3353A micro-controller has an on-board DTMF generator, 6K ROM, 128 bytes RAM, 128 bytes EEPROM, 20 I/O and a special ringer output. And is fully able to control the PCA1070 via it's software I^2C-bus.

The PCA1070/PCD3353A circuits combination performs the following functions:

- Pulse and DTMF dialling
- Redial and repertory dial stored in EEPROM
- PCA1070 control
- PCA1070 variable stored in EEPROM
- PCA1070 variable programmable via I^2C-bus and Keyboard
- Dialling options programmable via keyboard stored in EEPROM
- Ringer functions incorporated in 3-tone, 4-speeds and 4 volumes
- Display.

*** ETT95007.0.0**

OM4757 demonstration board with the PCD3332-3, TEA1064B/1062 and TEA1093/1094

Details: 37 pages, April 1995

This OM4757 is a demonstration model of a feature phone realised with the PCD3332-3 dialler/ringer, the TEA1064B or TEA1062 transmission IC, the TEA1093 or TEA1094 handsfree circuit and a ringer stage.

The feature phone contains a general application with different demonstration possibilities with respect to transmission and handsfree ICs. It is not made to fulfil specific country requirements.

The OM4757 contains a handset with microphone and earpiece and a base with the hardware including buzzer, microphone and loudspeaker. The OM4757 is delivered with the TEA1064B, TEA1093 and PCD3332-3. The TEA1064B can be replaced by the TEA1062 and the TEA1093 by the TEA1094.

The TEA1093 is line powered. Although the TEA1094 is intended for supply by an external source, the demonstration model is provided with the TEA1094-supply from the line as well.

*** 9398 341 10011**

TEA1060 family designer's guide - versatile speech/transmission ICs for electronic telephone sets

Details: 72 pages, July 1987

This designer's guide gives an overview, subscriber set architectures, a detailed functional description, hints for PCB layout and anti-sidetone bridge calculations for the TEA1060 family. Powered by the telephone line, these ICs can be used with virtually any kind of microphone and the set's impedance can be complex or real. Besides their application in basic and feature phones, they can be used in automatic answering machines, facsimile equipment and cordless phones. The choice of TEA1060 family IC depends upon the microphone used and local PTT requirements.

*** 9398 061 30011**

An introduction to the basics of electronic telephone sets

Details: 53 pages, January 1994

The Telecom Applications Group of Philips Components Division has organised field application engineer (FAE) training sessions on several occasions. These training sessions cover both theoretical and practical work. This course note summarises much of the basic theory presented at the training sessions. It's aim is to give engineers working in the area of components sales, application, and design, a thorough introduction to the fundamentals of analogue telephony circuitry. This consists of the analogue telephone set and other subscriber equipment connected to the analogue local subscriber lines (also referred to as central office lines, PABX lines, a/b wires or ring/tip wires. In later course notes subjects covered will include: switching, cordless telephony, mobile telephony, ISDN, transmission, radio paging, data communication.

The introduction in section 1 gives a brief description of the fundamentals of classical telephone sets (with rotary dial and carbon microphone). Section 2 covers the DC requirements of a telephone set in relation to the line current and operating voltage. Section 3 describes the speech circuitry in more detail. Theory as well as practical electronic replacement for the coil hybrid are described. New features not available with carbon microphone sets are also included. In section 4 the dialling circuitry is described. Electronic pulse diallers, the modern DTMF dialling system, and PTT requirements are also covered in section 4. The ringer is described in section 5, and section 6 summarises all the user-friendly enhanced features of an advanced electronic telephone set.

THE I²C-BUS CONCEPT

Details: 27 pages

This document gives an overview of the design of I²C-bus. The complete specification is given in publication "The I²C-bus and how to use it (including specifications), ordering code 9398 393 40011, April 1995.

*IEEE JOURNAL OF SOLID-STATE CIRCUITS, VOL. SC-19, NO. 3, JUNE 1984

Dual tone and modem frequency generator with on-chip filters and voltage reference

Details: 10 pages, June 1984, (Reprint from IEEE journal of solid-state circuits)

A single-chip CMOS circuit is described that contains a dual tone multi-frequency (DTMF) and modern frequency generator. For optimum performance and economy, switched capacitor techniques are used for the on-chip bandgap reference voltage, D-A converters, and filter. CEPT recommendations on output level stability and distortion are met without recourse to external filtering and without a stabilised supply or external reference voltage. A self aligned contact (SAC) CMOS process with 4 μm design rules and with 50 nm thick gate oxide is used to manufacture the circuit.

TTE87132

Documentation for printed circuit board CAB3467

Details: 6 pages, June 1987

This printed circuit board incorporates a full-wave rectifier especially suited for use as polarity guard in electronic pulse-dialling/flash telephone sets. Optionally also an electronic interrupter is presented on the board.

The bridge is realised with DMOS-field effect transistors, combining a very low forward voltage drop (total voltage drop is typ. 0.18 V at 10 mA) with a very high maximum voltage rating (200 V).

The (optional) interrupter is equipped with a DMOS P-channel FET which combines the advantages of conventional N-channel FETs (virtually no driving current required) with those of PNP-transistor interrupters (very simple interface between pulse dialler and interrupter).

The complete board is realised in SMD technology.

TTE87168

Documentation for TEA1064 printed circuit board CAB3458

Details: 21 pages, November 1987

This report describes the printed circuit board CAB3458. This printed circuit board is meant to demonstrate the speech transmission IC TEA1064. This report gives some brief application information, but is not meant as an application report. Please refer to the literature given in chapter 4 for application details.

1 INTRODUCTION TO BASICS

CONTENTS

APPLICATION NOTE Nr	9398 061 30011
TITLE	An Introduction to the Basics of Electronic Telephone Sets
AUTHOR	J. van Tiggelen, J. van der Hof
DATE	December 1988

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1. CLASSICAL TELEPHONE SET

1.1 Introduction

Almost unchanged for many decades, the rotary dial carbon microphone set is used all over the world. However, this position seems to be changing rapidly in favour of electronic telephone sets. This section covers the basic principles of classical telephone sets. It consists of three main parts:

- speech part with microphone, earpiece and the connection to the line
- rotary dial for dialling the required telephone number
- ringer or alert which tells the subscriber that a call is on the line

1.2 Speech transmission function

1.2.1 Carbon microphone

Speech (acoustic pressure waves) can be transformed into electrical signals by a carbon microphone consisting of a container filled with carbon particles. At one end, the container is closed by a flexible membrane. When acoustic pressure waves impinge on the membrane, the carbon particles are compressed resulting in small changes in resistance.

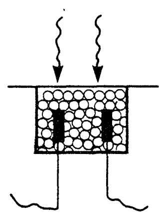


Fig.1.1 - Carbon microphone

The variation in resistance is used to modulate a current (DC) which flows through the microphone (see Fig. 1.2).

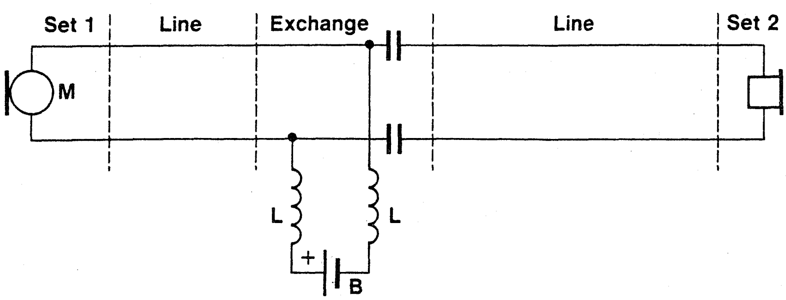


Fig.1.2 - One-way speech transmission

The current (10 to 100 mA DC) which flows through the carbon microphone M is supplied by battery B (usually 36 to 60 V). The current is modulated by the microphone in the rhythm of the speech (Fig.1.3).

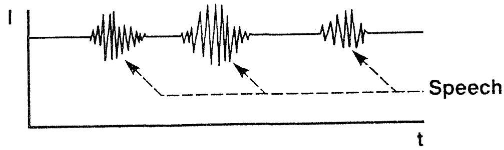


Fig.1.3 - Microphone current as a function of time

The resulting AC can be heard in the earpiece of the receiving telephone set (the DC is blocked by capacitors). The earpiece is normally an electromagnetic transducer (Fig.1.4).

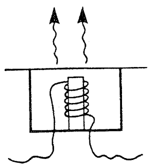


Fig.1.4 - Electromagnetic transducer

The coils (L, usually about 10 H) prevent short-circuiting of the AC signals via the battery. With a mirror-image system (Fig.1.2) a full-duplex telephone system can be realized with four-wires connecting each telephone set to the exchange.

1.2.2 2-wire full-duplex operation

To save two wires, another system is used which makes it possible to use the same pair of wires for both transmission and reception of speech signals. This is called a 2-to-4 wire conversion or hybrid function. The principles of the hybrid circuit are quite simple. In a Wheatstone bridge (Fig.1.5) if $R1 \times R_L = R2 \times R3$:

$$\text{then } V3 = \frac{R_L}{R1 + R_L} \times V1, \text{ and } V2 = 0$$

i.e. if the bridge is balanced, the generator signal V1 does not cause any signal V2. Maximum transfer will occur if $R1 = R2 = R3 = R_L$ results in a transmit signal $V3 = {}^1/_2\, V1$.

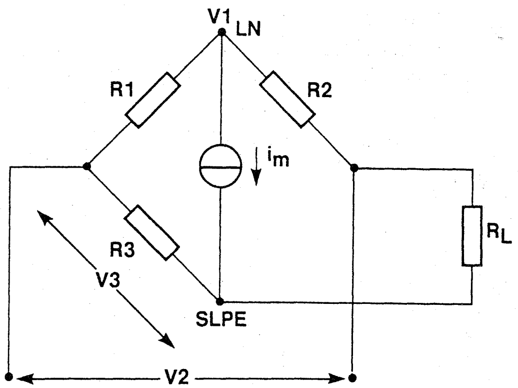


Fig.1.5 Wheatstone bridge principle

If V1 comes from a microphone and V2 from an earpiece, then the speech signal from the microphone is only available on the telephone line R_L (V3) and not to the earpiece. This 2-to-4 wire principle is one of the most basic circuits in the telephone system. It exists even in the most modern digital version called ISDN, but then it's called an echo canceller. In the electro-mechanical telephone set it's realized with a transformer (Fig 1.6). In this Wheatstone bridge configuration the transformer replaces R1 and R2, the balance network Z replaces R3, and R_L represents the combined impedances of the telephone line and the other telephone set.

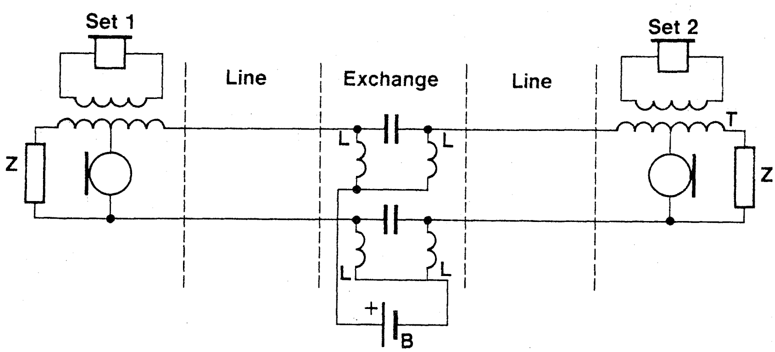


Fig.1.6 - Full duplex speech transmission system

The carbon microphone modulates a direct current i. Because it's connected to the centre tap of the primary winding of transformer T (the hybrid transformer) the AC splits into current i' to the left and a current i" to the right (Fig.1.7).

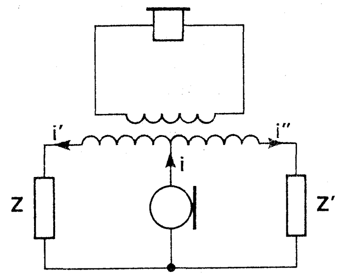


Fig. 1.7 - Function of hybrid transformer (sending condition)

If current i' equals current i" (Z = Z') there will be no resulting AC signal in the secondary winding, and the audio generated in the microphone won't be audible in the earpiece.

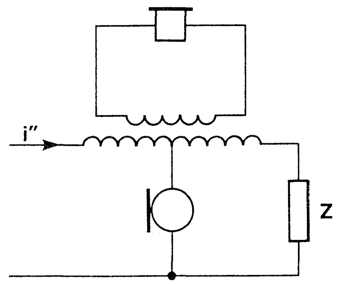


Fig. 1.8 - Function of hybrid transformer (receiving condition)

At the other end of the line, current i" induces (neglecting a small dissipation in the carbon microphone and in impedance Z) a voltage in the secondary winding of transformer T which results in audio signals in the earpiece (Fig.1.8).

This situation is the ideal one, but in practice impedance Z' is dependent on both the impedance of the telephone set connected to the other end of the line (usually 600 Ω) and on the impedance of the telephone line itself. Therefore, complete cancellation of the microphone signals can only be achieved for one particular telephone line length. For all other line lengths, part of the signal generated in the microphone will be audible in the earpiece (sidetone). A small amount of sidetone is not a drawback and is, in fact, required to give the user audio feedback of his/her voice to avoid the temptation to shout during a call. However, too much imbalance in the bridge circuit makes line loss compensation more difficult as maximum amplifier gain will be limited by the sidetone. This can be a real problem with long distance connections. These transmission problems will be described in detail in later course notes.

1.3 Signalling functions

1.3.1 On-hook/off-hook detection

The telephone exchange must know the status of the telephone sets connected to it. The DC line current is used for this purpose. Each set has a hook switch which interrupts the line current (Fig.9) if the handset is placed on the cradle (on-hook). On-hook/off-hook detection is realized by a relay R connected in series with the DC path in the exchange.

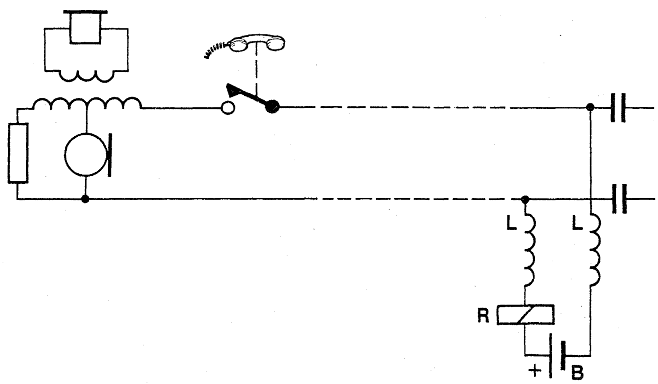


Fig.1.9 - On-hook/off-hook signalling

This system enables the same pair of wires to be used for both speech transmission and detection of the hook status. The resulting current waveforms are given in Fig.1.10.

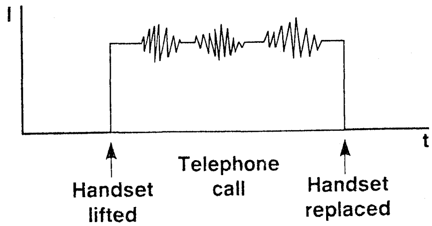


Fig.1.10 - Telephone set current as function of time

1.3.2 Subscriber number dialling

In automatic telephone systems, subscribers must be able to indicate the number of the subscriber they want to call to the exchange. This is done by turning a rotary dialler which interrupts the DC supply current (by means of S1 in Fig.1.11) in a predetermined rhythm (usually 10 Hz). The number of interruptions corresponds to the digit dialled (1 interruption = '1', 2 interruptions = '2', etc., 10 interruptions = '0'). To avoid loud clicks in the earpiece, the speech part of the telephone set is muted during dialling by an extra switch in the rotary dial (S2).

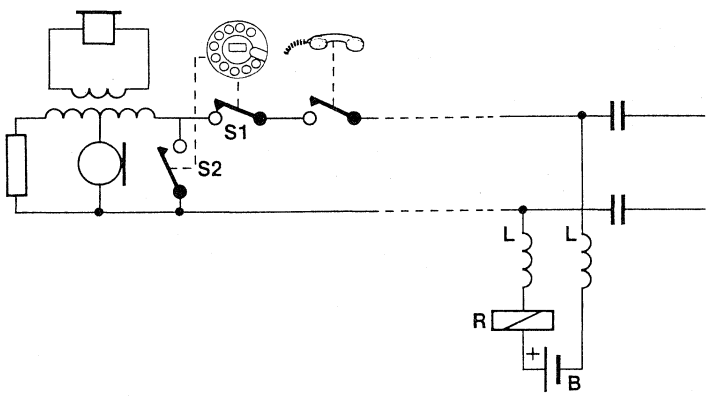


Fig.1.11 - On-hook/off-hook detection and dial pulse generation and detection

The digits dialled can be detected by relay (R) in the exchange (on-hook/off-hook detection). The resulting current waveforms are given in Fig.1.12.

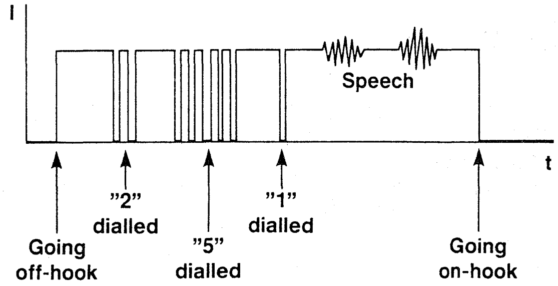


Fig.1.12 - Telephone set current as function of time

1.4 Alert function

Every telephone set incorporates a ringer (Fig.1.13) which has to alert the subscriber to incoming calls. Usually this ringer is an electro-mechanical bell which is driven by an AC supply of 40 to 70 V at about 25 Hz. A capacitor in series with the ringer blocks DC.

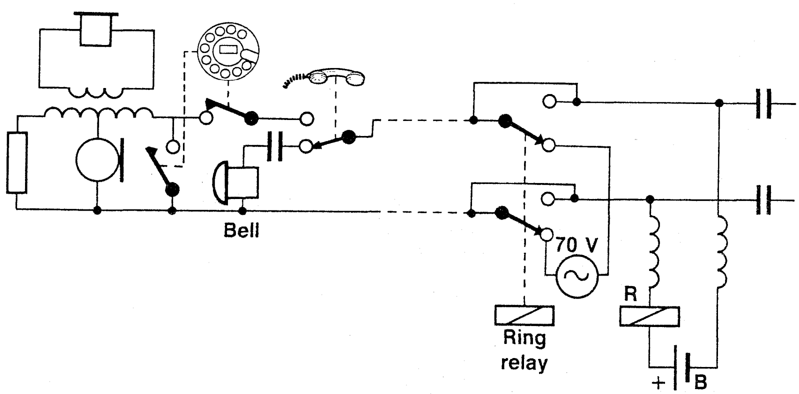


Fig.1.13 - Circuit diagram of telephone set including ringer

The ringing voltage is applied to the telephone set only when the telephone set is on-hook. As soon as the handset is lifted (detected in the exchange by the flow of DC) the ringing voltage is switched off to avoid unpleasant sounds in the earpiece. The resulting current waveforms are given in Fig.1.14.

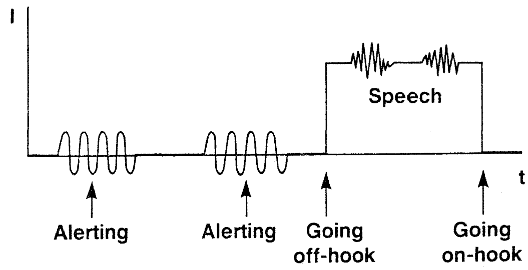


Fig.1.14 - Telephone set current as function of time

2. DC REQUIREMENTS

2.1 Polarity guard

A telephone set must function properly independent of the polarity of the line voltage applied to it. This is no problem for carbon microphone types with a rotary dial since there are no polarity-sensitive components in such telephone sets. It becomes a problem when electronic components are introduced into a telephone set. Transistors and integrated circuits cannot operate if the supply polarity is reversed and can even be damaged by a reversal. A diode bridge is therefore always connected between the telephone line and the electronic circuits in the telephone set. The circuit diagram of such a bridge is given in Fig.2.1.

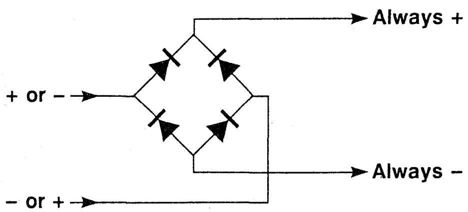


Fig.2.1 - Circuit diagram of a polarity guard diode bridge

Notice that this bridge has no rectifier function; the telephone line already has a direct current supply. A drawback of this polarity guard is the extra voltage drop of two forward diode voltages (1.4 V with typical diodes or 0.6 V for more expensive Schottky diodes).

Sometimes the diode bridge has an additional lightning protection function. Usually two of the diodes are then replaced by zener diodes or transient suppressor diodes (e.g.BZW14).

2.2 DC voltage

A telephone set has to guarantee a minimum operating line current to ensure that all relays in the exchange function properly.

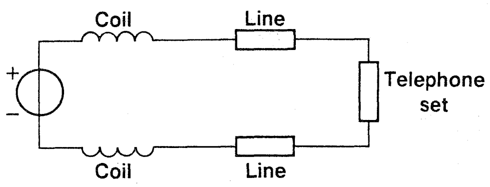


Fig.2.2 - Direct current path in subscriber loop

As shown in Fig.2.2 the direct current in the subscriber loop is determined by; the supply voltage of the exchange, the resistance of the coils in the exchange, the resistance of the telephone line between subscriber and exchange, and by the resistance of the subscriber set.

The battery voltage of most exchanges is normally 48 V, but it can vary between 24 and 100 V; the coil resistance is normally 2 x 200 Ω or 2 x 400 Ω, the line resistance (for 0.5 mm diameter line) is about 2 x 90 Ω/km, and the set impedance of a (carbon microphone) telephone set varies between 100 and 300 Ω. An example of a worst case loop current calculation is:

coils resistance (2 x 400 Ω)	:	800 Ω
10 km telephone line	:	1800 Ω
telephone set resistance	:	300 Ω
TOTAL	:	2900 Ω
Minimum battery voltage	:	42 V (48 V nominally)
Minimum loop current	:	42 V/2900 Ω = 14.5 mA

Figure 2.3 gives the minimum line current for some European countries.

Country	Minimum line current
Belgium	20 mA
Denmark	15.3 mA
W.Germany	17 mA
France	12 mA
Netherlands	16 mA
Norway	17 mA
United Kingdom	25 mA
Sweden	11 mA

Fig.2.3 - Table of minimum line currents in Europe

In the past, most PTTs have defined a maximum DC resistance for telephone sets to guarantee the minimum loop current. However, for electronic telephone sets it is not very practical to define a maximum DC resistance since they have a non-linear voltage-current characteristic with (due to the polarity guard bridge) a very high resistance at low currents. Nowadays, most PTTs specify voltage-current masks. Figure 2.4 gives the voltage-current mask for the USA.

Some PTTs allow a higher line voltage during DTMF dialling because this system works without dial pulse detection relays at the exchange. The USA normally allows a line voltage of 6 V at 20 mA, but during DTMF dialling it may be 8 V at 20 mA. It is therefore easier to power the DTMF generator in countries with this type of specification.

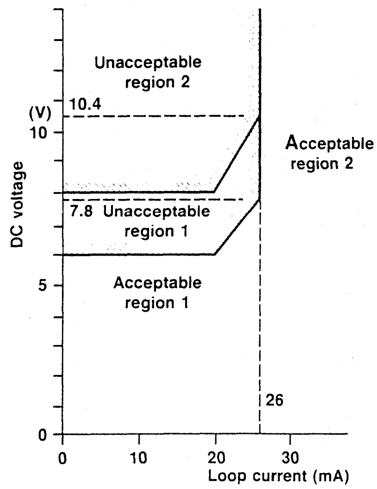


Fig.2.4 - USA voltage-current mask for telephone sets

Some PTTs require a lower line voltage during pulse dialling to make it easier for the exchange to detect line current interruptions with relays. In Italy, normally the maximum line voltage is almost 8 V (at 12 mA), but in the make periods of pulse dialling it can be less than 4 V. These PTT requirements can significantly influence the cost price of a telephone set.

2.3 Parallel operation of telephone sets

Many subscribers connect more than one telephone set to their telephone line. However, most PTTs don't allow more than one off-hook telephone set to be connected to the telephone line at the same time. Extra hook-switch contacts or relays must therefore be used to prevent other sets going off-hook in parallel with a set that is already off-hook. To remove the need for the extra components, there is a trend (in USA and in Japan) to allow subscribers to simply connect their telephone sets in parallel. If more than one telephone set is off-hook during switch over for a short time, they have to share the total line current available. This can result in lower performance. Real problems can arise when a carbon microphone type of telephone set is connected in parallel with an electronic type, especially when the subscriber is a long way from the exchange (long lines).

Take for example a carbon microphone set with a DC resistance of 200 Ω connected to a 20 mA line (Fig.2.5). The line voltage then will be 4 V. Now connect an electronic telephone set in parallel which needs at least 4 mA to have acceptable performance. This leaves 16 mA for the carbon microphone set and gives a line voltage of 3.2 V. This means that the electronic telephone set only has a supply voltage of 1.8 V! (3.2 V minus the polarity guard voltage drop). A specially designed electronic speech circuit is therefore required to achieve acceptable performance at such a low supply voltage.

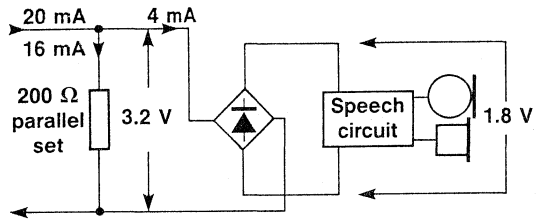


Fig.2.5 - Parallel operation of classical and electronic telephone set

Of course, no guarantee can be given for correct dialling when two telephone sets are connected to one telephone line at the same time. Even pulse dialling won't be possible because the current through the parallel set can't be interrupted; DTMF dialling will be difficult because DTMF diallers need a minimum supply voltage (usually 2.5 V) which can't be guaranteed under these circumstances. In general, parallel operation is only required during conversation.

3. THE SPEECH PART OF A TELEPHONE SET

3.1 Newton, Pascal and (deci)Bell

All sounds consist of acoustic pressure waves. A sound source modulates the air pressure and this modulation propagates at the speed of sound (about 330 m/s). Air pressure can be measured in Newtons per square meter (N/m^2) or in Pascal (1 Pa = 1 N/m^2). The nominal ambient air pressure at sea level is 100 kPa. We can hear air pressure variations between roughly 16 Hz to about 15 kHz. The minimum sound level we hear is (frequency dependent) about 20 μPa; the maximum level (threshold of pain) is about 200 Pa.

20 µPa	0 dB SPL	→ Threshold of hearing
	10 dB SPL	
		→ Forest
200 µPa	20 dB SPL	
		→ Bedroom
	30 dB SPL	
		→ Library
2 mPa	40 dB SPL	→ Living room
	50 dB SPL	
20 mPa	60 dB SPL	→ Conversational speech
		→ Business office
	70 dB SPL	
200 mPa	80 dB SPL	
		→ Average street traffic
	90 dB SPL	→ Heavy truck
	94 dB SPL	**→ Typical telephone microphone**
2 Pa	100 dB SPL	→ Pneumatic drill
	110 dB	→ Pop group
20 Pa	120 dB SPL	
		→ Jet take-off (at 100 m)
	130 dB SPL	
200 Pa	140 dB SPL	→ Threshold of pain

Fig.3.1 - Real-life examples of Sound Pressure Levels (SPL)

Because our ears have a very large dynamic range (about 1 to 10^8) it is easier to express sound pressures on a logarithmic scale (in decibel):

$$\text{sound pressure (dB)} = 20 \log \frac{\text{sound pressure (Pa)}}{\text{reference sound pressure (Pa)}}$$

A commonly used reference level is the threshold of hearing, i.e. 20 µPa. This reference level is indicated by the abbreviation 'SPL' placed directly after the 'dB'. In Fig.3.1 a list of absolute and relative sound pressure levels is given with a real-life example of where the level occurs. For telephony applications during normal speech, the average sound pressure level at the microphone of the handset is about 94 dB SPL. The peak level which is important for the maximum drive level of the amplifiers is about 20 dB higher. The human ear is not equally sensitive at all frequencies - a 70 dB SPL, 100 Hz signal seems less loud than a 70 dB SPL 1 kHz signal. For this reason loudness measurements are used which are weighted according to the different sensitivities of the human ear.

The best known weighting curve is the A-curve with dB(A) values. Another weighting curve is the psophometric CCITT P53 curve which is used for noise loudness measurements in telephony. Measurements according to the P53 curve are often characterized by a suffix 'p'. The A- and the p-curves are shown in Fig.3.2.

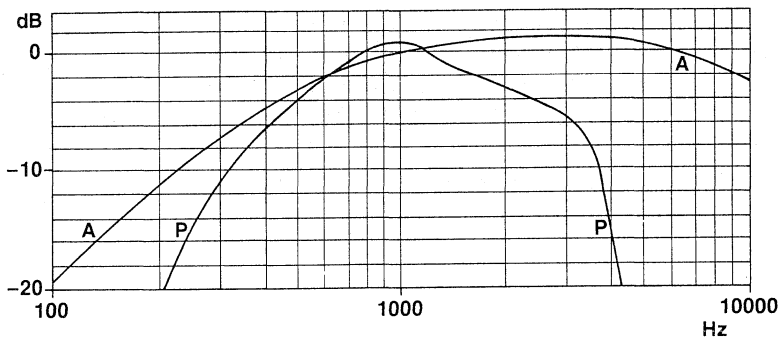


Fig.3.2 - The A- and the p-weighting curves

Some other 'decibel type' symbols are frequently used in the world of telephony.

dBm - The reference level of 0 dBm is defined as the voltage resulting from 1 mW of power dissipation in (usually) 600 Ω, i.e. a voltage of 0.775 V. Therefore, a voltage of 100 mV results in 20 log (0.1/0.775) = -17.8 dBm.

dBmp - This describes a psophometrically weighted dBm level.

dBV - This has a reference level of 1 V.

dBmV - This has a reference level of 1 mV. Thus 0 dBV corresponds to 1 V. and 0 dBmV corresponds to 1 mV.

3.2 Transducers

Transducers in telephone sets are used to convert sound into electrical signals (microphones) and back again (earpieces and loudspeakers). An important parameter of transducers is their sensitivity. This is expressed in V/Pa for microphones and Pa/V or Pa/W for earpieces and loudspeakers.

3.2.1 Carbon microphones

The carbon microphone is the most frequently used type of microphone in the world. The reason for this popularity is its sensitivity of 100 mV/Pa or more, makes it the only type of microphone that doesn't need additional amplification. This point was of crucial importance in the pre-transistor era.

The carbon microphone is a passive microphone; it can't generate any signals without an external power source because sound signals only result in changes in the microphone resistance. By passing a current through it, the resistance variations are transformed (linearly) into voltage variations.

The linear relationship between AC output voltage and direct current is a weak point of the carbon microphone. High direct currents usually occur in telephone sets connected to the exchange via short telephone lines (low resistance) and low currents occur in long telephone lines (high resistance). Speech signals of long-line subscribers are attenuated more than those of short-line subscribers, and the sensitivity of the microphone is better at high currents which is the opposite of what is required.

Other weak points of the carbon microphone include its instability (the characteristics of the microphone, especially the resistance, are position dependent) and its poor reliability (carbon particles can 'weld' together resulting in a defect). The carbon microphone also produces a lot of non-linear and frequency distortion. It's also subject to the effects of aging.

A positive point of the carbon microphone is its 'squelch' effect. The relationship between sound pressure and output voltage is not linear (Fig.3.3) and it has a very low sensitivity to weak signals; background sounds are therefore automatically attenuated, which gives better audibility.

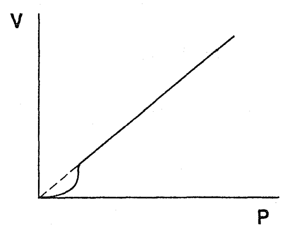


Fig.3.3 - Output voltage versus sound pressure showing squelch effect

3.2.2 Dynamic transducers

Dynamic transducers can be used both as microphones and earpieces. Two versions are available; the electro-dynamic or moving coil transducer (Fig.3.4a) and the magneto-dynamic or moving-magnet transducer (Fig.3.4b).

Fig.3.4 - Dynamic transducers
a) electro-dynamic
b) magneto-dynamic

These have a coil and a magnet which can move with respect to each other. In the electro-dynamic type the magnet is fixed and the coil, which is suspended in the magnetic field of the magnet by a membrane, can move. Most loudspeakers work like this. In the magneto-dynamic type, the coil is fixed and the magnet can move. This principle is used in some record player cartridges.

If the device is used as a microphone, sounds generate an electromagnetic force in the coil which in turn results in an output voltage. Used as an ear-piece, an AC signal will generate an alternating magnetic field and this results in movement of the membrane.

Dynamic transducers have a low impedance (a few hundred ohms) and can produce very good sound quality, but they are not very sensitive (used as a microphone they produce only about 1 mV/Pa).

3.2.3 Magnetic transducers

Magnetic or rocking armature transducers have a magnet with an air gap in which a piece of iron, mounted on a flexible membrane, can move (Fig.3.5).

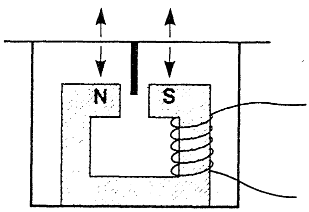


Fig.3.5 - Magnetic or rocking armature transducer

Used as a microphone, the movements of the piece of iron affect the magnetic field and this generates an electromagnetic force in the coil; this in turn results in an output voltage. Used as an earpiece, an AC signal in the coil will generate a magnetic field resulting in magnetic forces on the piece of iron. Movements of the piece of iron are transformed into sound by the membrane.

Magnetic transducers don't produce good sound quality; the frequency characteristic is not flat and a lot of non-linear distortion is produced. The sensitivity, however, is fairly high and the impedance is low (a few hundred ohms).

3.2.4 Piezo-electric transducers

Some materials (e.g. BaTiO) generate a voltage when they are mechanically deformed. This process can also be inverted; applying a voltage to the material will deform it. This effect is called the piezo-electric effect and can be used to make microphones and earpieces which can be very inexpensive. Electrically, they behave as a capacitor of several tenths of a nano-Farad. An amplifier driving an piezo-electric transducer must be carefully designed because this capacitive characteristic can easily cause instability.

The sensitivity of piezo-electric transducers is fairly high when used as a microphone (about 10 mV/Pa) but low when used as an earpiece. The frequency characteristic is bad; there are a lot of peaks in it caused by resonance effects which can be used in ceramic resonators for oscillators. However, a specially shaped housing which damps these resonant peaks, can greatly improve its frequency characteristic (Fig.3.6).

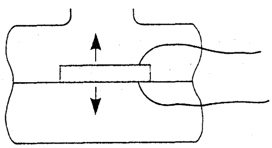


Fig.3.6 - Piezo-electric transducer

A very popular application of piezo-electric devices is as an electronic ringer. The resonance effect can be used to create very high volume chimes very efficiently.

3.2.5 Electret microphones

An electret microphone has a pre-charged capacitor with a flexible plate that can move under the influence of sound signals. Such movements result in changes of the capacitance and therefore in changes of the voltage across the microphone (according to the law $V = Q/C$) because the charge across the capacitor remains constant (Fig.3.7).

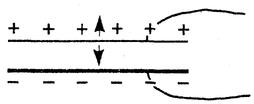


Fig.3.7 - Electret microphone

Electret microphones are sensitive (about 10 mV/Pa) but need an extremely high resistance termination to avoid leakage of the charge. An FET is normally incorporated in the microphone as a preamplifier to bring down the impedance to a more practical value (several kΩ).

Figure 3.8a is the circuit diagram of an electret microphone with an FET buffer and its three connections. The FET is used as a source follower. Pins 1 and 3 are connected to the supply voltage (1.5 to 5 V) and the AC signal is available across pins 2 and 3.

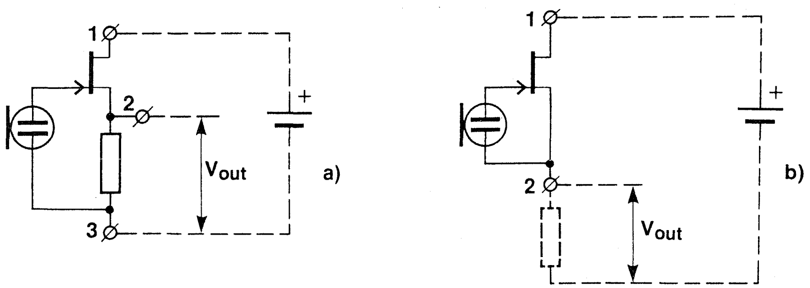


Fig.3.8 - Circuit diagrams of electret microphone with FET buffer

In Fig.3.8b, a type with only 2 connections is shown. The FET is used here as a current source modulated by the electret microphone. Pin 1 is connected to the positive supply voltage and pin 2 is connected to the negative side of the supply via a resistor (usually several kohms). The sound signals are available across this resistor. The advantage of type b is that only 2 wires (instead of 3 in type a) are required.

3.3 AC requirements of telephone sets

The most important AC requirements that electronic telephone sets have to fulfil, and how they can be implemented, are described in the next few sections.

3.3.1 Maximum output voltage; output stage configurations

The sending amplifier of a telephone set must be capable of modulating the direct current to obtain a certain output voltage. The output impedance of the sending amplifier is usually 600 Ω, and the load impedance for measurements is also usually 600 Ω. The maximum required output voltage is usually about 3 dBm but it can be as high as 9 dBm depending on the individual PTT. There are two ways to make such a sending amplifier. The first is to use an output amplifier (voltage source) with an (internally determined) impedance of 600 Ω (Fig.3.9a). The second is to use a very high resistance output amplifier (current source) and to adjust the impedance to that required using parallel passive components (a 600 Ω resistor in Fig.3.9b). This is the method used with the TEA1060 family of speech/transmission* circuits. Both methods have advantages and disadvantages.

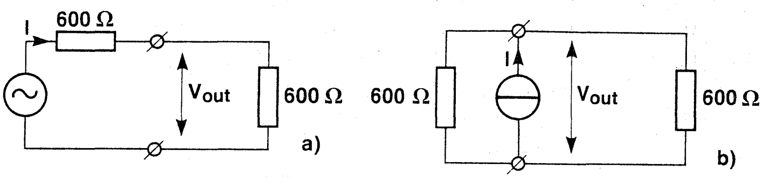


Fig.3.9 - Two methods of realizing an output stage:
a) active 600 Ω output stage
b) high resistance output stage with parallel passive 600 Ω termination.

In the TEA1060 family, the method in Fig.3.9b is combined with a very simple way of making a DC supply point for the internal amplifiers and external peripherals. A capacitor is connected in series with the 600 Ω resistor (Fig.3.10). If the value of the capacitor is high enough, it will have no influence on the set impedance.

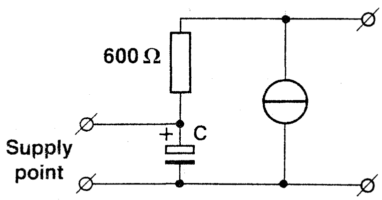


Fig.3.10 - Simple supply point with current source output stage

* Note: *The electronic circuit which interfaces the microphone and earpiece with the telephone line is called a speech/transmission circuit. Philips produces a range of integrated speech/transmission circuits called the TEA1060 family.*

3.3.2 Impedances; balance return loss

The most popular telephone set impedance is still 600 Ω. This value originates from the use of air-spaced lines with a characteristic impedance of 600 Ω. Underground lines used more often nowadays can be better characterized by a complex impedance. The trend is therefore to switch over to complex set impedances (e.g. in United Kingdom, West Germany and France). Normally, the required set impedance is not expressed as an impedance but as a balance return loss (BRL) figure. The balance return loss describes how accurately a telephone set impedance approximates the nominal impedance specified by the PTT:

$$BRL = 20 \log \frac{|Z + Z_{nom}|}{|Z - Z_{nom}|}$$

A high BRL indicates good matching, a low BRL indicates poor matching. In Fig.3.11, circles of constant BRL are drawn in a complex impedance plane using 600 Ω as the nominal reference.

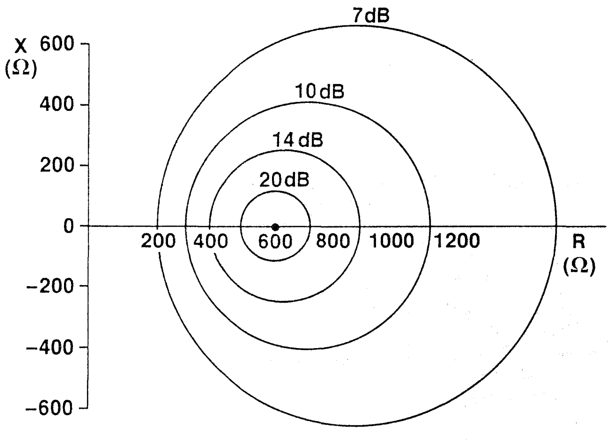


Fig.3.11 - BRL with respect to 600 Ω in complex impedance plane

If the characteristic impedance of the telephone lines is complex, then the termination of the set also has to be a complex impedance. A network used in the United Kingdom is given in Fig.3.12.

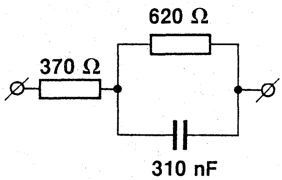


Fig.3.12 - Complex network used in the UK to terminate telephone lines

The use of such a complex network doesn't make any difference to the definition of BRL. The formula is valid for both real and complex line terminations. However, it's not possible to use Fig.3.11 to read the BRL for a particular telephone set impedance, because the ideal set impedance is frequency dependent. A set of circles of constant BRL must be therefore be calculated for each frequency. Circles of constant BRL for the network shown in Fig.3.12 are given in Fig.3.13 for frequencies of 300, 1000 and 3400 Hz.

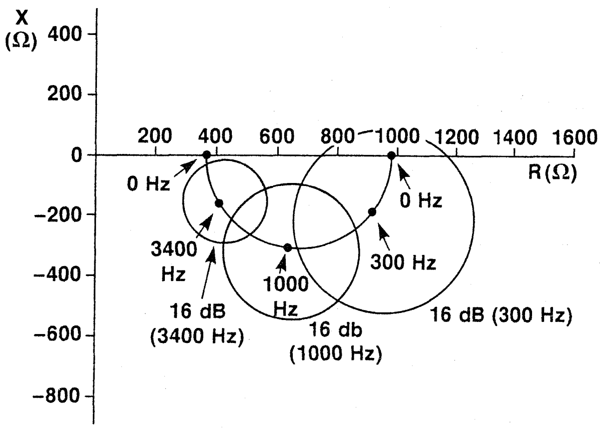


Fig.3.13 - Circles of 16 dB BRL with respect to network of Fig.3.12

3.3.3 Line loss compensation

Speech signals are attenuated by the telephone line between subscriber and exchange. The average attenuation of a 0.5 mm^2 telephone line is about 1.2 dB/km. Assuming that telephone lines have a length between 0 and 10 km, speech signals arriving in the exchange can vary by more than 10 dB. CCITT recommend a maximum loss on the subscriber line of 12 dB. The total attenuation including the other subscriber's telephone set can be twice that (up to 24 dB) (Fig.3.14).

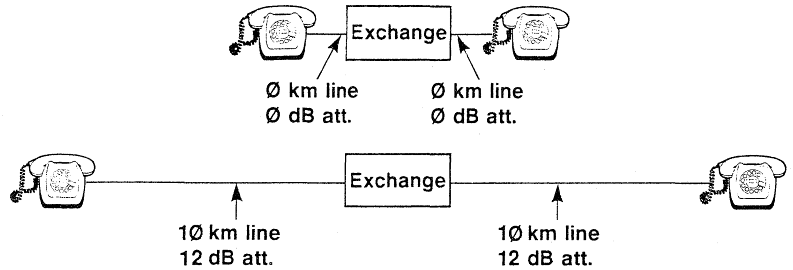


Fig.3.14 - Attenuation caused by the telephone line

3.3.3.1 Line loss compensation by adapting microphone and earpiece

Some PTTs have taken measures to equalize the line loss variation. One way is to use different types of microphone with different sensitivity. For long-line subscribers, a more sensitive type can be installed to compensate for the extra line loss. In electronic telephone sets this system can be replaced by a system in which the gain of the amplifiers can be easily adjusted (e.g. by jumpers). This is done in Austria with two or three gain levels in the sending and receiving amplifiers.

3.3.3.2 Line loss compensation by line current dependent AGC

Because telephone lines have a fixed resistance per unit length, and exchanges a fixed supply voltage and coil resistance, the line current in a particular subscriber line gives a indication of the line length according to the following formula:

$$\text{line current} = \frac{V_{exchange} - V_{set}}{R_{exchange} + R_{set} + R_{line}}$$

The line current can therefore be used to adjust the gain of the amplifiers in the telephone set automatically in such a way that (part of) the line loss is compensated.

This is an option with the TEA1060 family. Here, the gain of the sending and receiving amplifiers is changed for line currents which correspond to telephone lines of 0 km (6 dB gain reduction) to 5 km (0 dB gain reduction) (assuming a loss of 1.2 dB/km). The gain control facility must be adapted to the supply voltage and coil resistance of the exchange by a resistor. Figure 3.15 illustrates this principle with the TEA1060 for a variety of exchanges.

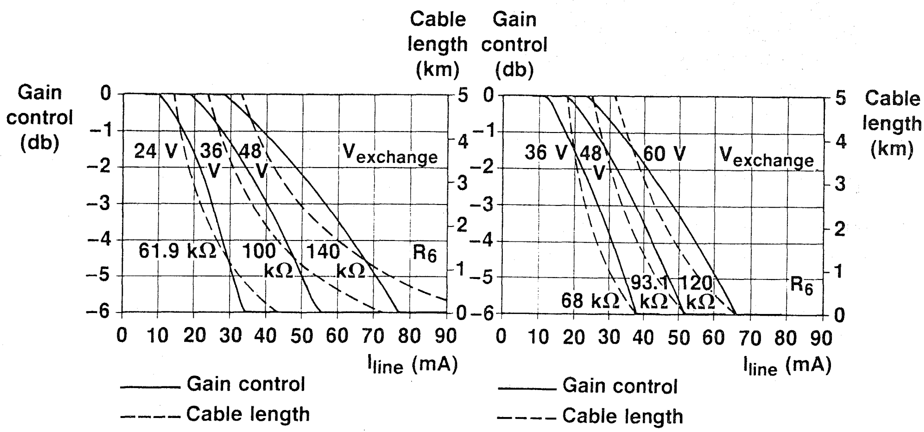


Fig.3.15 - Automatic line loss compensation of the TEA1060 for 400 Ω (left) and 800 Ω (right) bridges

In France, a different system of automatic gain control is used. The line current in France must be limited to a maximum of 50 mA by a current limiter in the telephone set. As a result, the line voltage will increase automatically if the current limiter comes into action. The line voltage therefore gives an indication of the line length and can be used to realize an automatic gain control system.

3.3.3.3 Line loss compensation by dynamic gain control

This method of dynamic gain control assumes that the average sound level of an individual's speech is constant. The gain control of the sending and receiving amplifiers is based on the average amplitude of the AC signals. People who speak softly into the microphone, therefore, get a high gain, and people who speak loudly, a low gain. One method is dynamic limiting of the microphone stage (horizontal part of the curve). Limiting avoids clipping the sending signal with very high level microphone signals; *this facility is built into the TEA1064 speech circuitry*. A further improvement can be realized if electronic squelch is incorporated in the speech circuitry; i.e. reducing the amplifier gain for low amplitude inputs to eliminate background noise (Fig.3.16 dotted line).

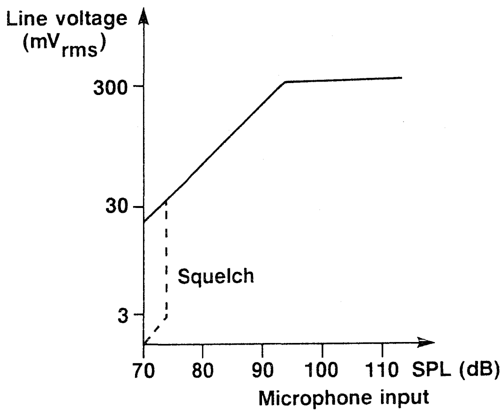


Fig.3.16 Dynamic gain control in the transmitter channel

3.3.4 Anti-sidetone circuits

As previously explained, a balance transformer can be used to prevent the microphone signal being heard in the earpiece of the same telephone set (Fig.3.17). The microphone is connected to the centre tap of the transformer, the telephone line (characterized by its impedance Z) to the left connection, and the balance impedance (impedance Z') to the right connection. If impedances Z and Z' are equal, no microphone signals will be induced in the secondary winding and be heard in the earpiece.

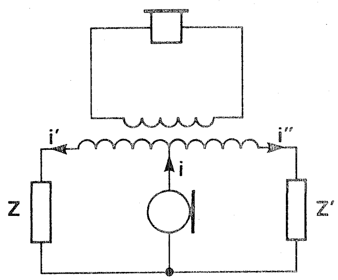


Fig.3.17 - Anti-sidetone circuit with transformer

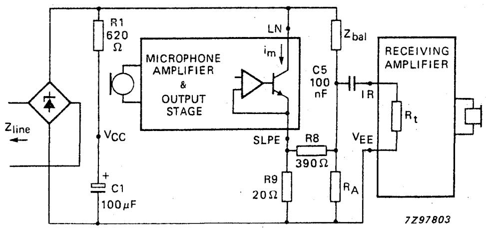
Each telephone set has its own telephone line characterized by its own length and therefore by its own impedance. In practice, telephone lines between the subscriber and the exchange can be from 0 km to more than 10 km. Ideally, the balance network Z' should be optimized for each subscriber, but this is not practical. Therefore, most PTTs optimize the balance network Z' for an average telephone line (e.g. a line of 3 km for systems without, and 5 km for systems with, automatic gain control; for all other line lengths a certain amount of sidetone is accepted). This level can be reduced further by using two anti-sidetone networks: one for short lines, the other for long lines. These networks can be automatically selected by measuring the line current (for an indication of the line length).

In electronic telephone sets, the function of the transformer is performed by a transformer-less bridge circuit. This can be a simple Wheatstone bridge (section 1.2.2 and 3.3.4.1) or the TEA1060 anti-sidetone bridge (section 3.3.4.2).

3.3.4.1 Wheatstone bridge

This well-known bridge circuit (Fig.3.18) is shown in an application with a speech/transmission circuit. Figure 3.18a is drawn in a normal way; Fig.3.18b is the same circuit but redrawn in such a way that the bridge components can be recognized more easily.

(a)



(b)

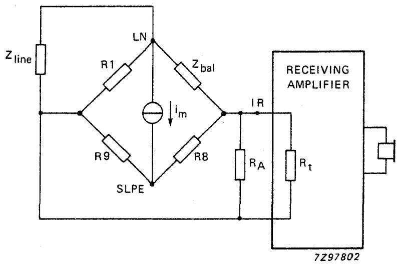


Fig.3.18 - Wheatstone anti-sidetone bridge with a speech/transmission circuit

The balance condition for this bridge is:

$$Z_{bal} \times R9 = R8 \times R1 \,//\, Z_{line}$$

or $$Z_{bal} = \frac{R8}{R9} \times \frac{R1 \times Z_{line}}{R1 + Z_{line}}$$

These calculations show that it would be difficult to have complete sidetone cancellation for every length of subscriber line. This would require a different Z_{bal} for each line length.

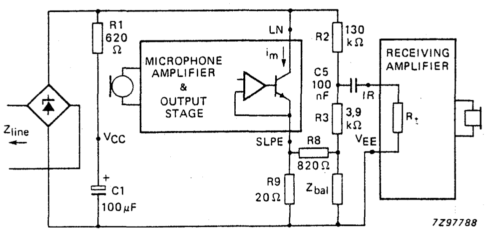
An advantage of this bridge is its simplicity. Furthermore if R8 >> R9 then Z_{bal} is a high impedance which results in a smaller value of capacitor in Z_{bal}.

The main disadvantage can be seen in Fig.3.18; the balance network is part of the voltage divider for the received signals, so this bridge will introduce a frequency dependency for received signals (normally 3 to 4 dB over the 300 to 3400 Hz band).

3.3.4.2 The TEA1060 anti-sidetone bridge

The TEA1060 can be used in the Wheatstone bridge configuration of Fig.3.18. However, moving Z_{bal} to another point in the bridge avoids frequency dependency problems with the received signal. This is illustrated by Fig.3.19 and the new bridge is called the TEA1060 anti-sidetone bridge. This bridge is shown in Fig.3.19a drawn in the normal way, and Fig.3.19b shows the bridge drawn in such a form that the bridge components can be more easily recognized.

(a)



(b)

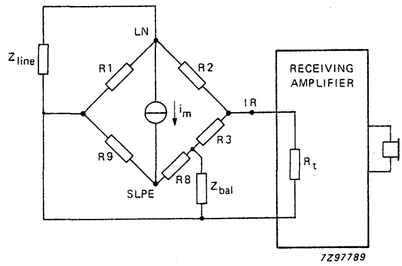


Fig.3.19 - TEA1060 anti-sidetone bridge

The balance condition for this bridge is more complicated to calculate; two equations must be fulfilled simultaneously:

a) $R9 \times R2 = R1\ (R3 + (R8//Z_{bal}))$

b) $Z_{bal} = \dfrac{R8}{R1} \times Z_{line}$

This bridge configuration only introduces some tenths of a dB of frequency dependency for the amplitude of received signals. More details on bridge configurations are given in Philips publication 9398 341 10011; "The TEA1060 family designers' guide".

3.4 Loudspeaking facilities

There are two main loudspeaking features:
- listening-in (also used as on-hook monitor)
- hands-free or speaker-phone facility

3.4.1 Listening-in/on-hook dialling monitor

This facility is the amplification of the receiving signal for a loudspeaker so that the telephone conversation can be followed by more than one person in the room. This loudspeaker function can also be used during on-hook dialling to monitor the dialling and ringing tone, and to hear when the called party answers the call. Figure 3.20 shows an application circuit where a loudspeaker amplifier is connected to the output of the speech/transmission circuit (QR-) and a potentiometer is used to control volume.

There are two problems with listening-in circuitry; the limited telephone line supply, and feedback from the loudspeaker to the handset microphone. To deliver sufficient energy to the speaker, an electronic "inductor" (TEA1081 in Fig.3.20) or a bulky coil can be used to increase the current available from the telephone line. A mains supply can be used but this will make the system significantly more expensive and extra mains isolation (section 6.4) will also be necessary.

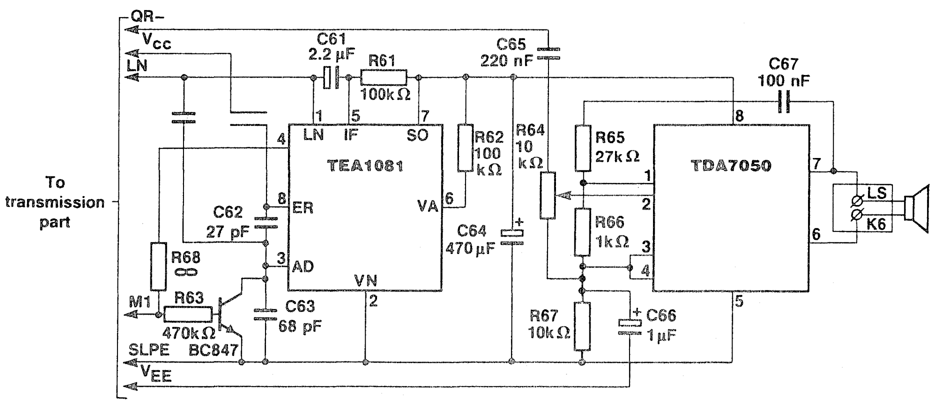


Fig.3.20 - Schematic of the sound part of listening-in telephone set with TDA7050 listening-in amplifier and TEA1081 supply circuit

The schematic for the audio path is shown in Fig.3.21. Depending on the distance between the loudspeaker in the base of the telephone set and the handset microphone, part of the received signal is fed back through the audio path by the microphone to the telephone line. It's attenuated by the anti-sidetone network to the receiver input (the total attenuation depends on the line length and the optimization of the sidetone circuit). Amplified to the earpiece output (QR) and again amplified by the loudspeaker amplifier, the same signal will be output by the loudspeaker. Howling (singing) will occur if the total gain in signal path is greater than or equal to 1 (the Larsen effect). This howling occurs only when the handset is close to the base; mostly when the handset is taken off-hook or replaced at the beginning or end of a conversation. Sophisticated telephone sets implement special precautions which avoid or limit this howling effect. The solution shown in Fig.3.20 does not have any anti-Larsen precautions.

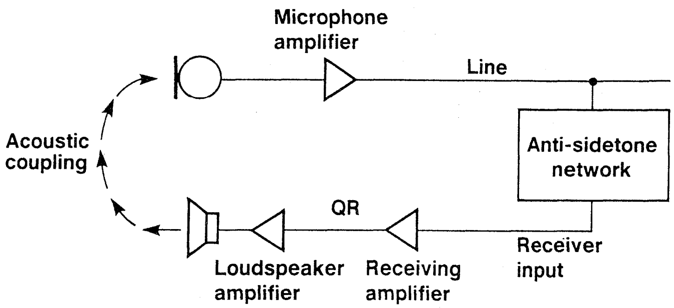


Fig.3.21 - The sound path with a listening-in amplifier telephone set and the howlround problem

3.4.2 Hands-free telephones

In 3.4.1 the received line signal had additional amplification to drive a loudspeaker and the listener did not need to listen via the earpiece. If the microphone sensitivity is increased by means of extra amplification the "user" can also talk to the telephone set from some distance. Usually, a separate microphone is built-in to the base of the telephone set, and the user can converse without having to hold the handset. A telephone with this facility is called a hands-free telephone. With listening-in, the loop from loudspeaker, to line, to receiver input, to loudspeaker can generate howling. This loop is still present with hands-free sets; the base microphone (with high sensitivity) is close to the loudspeaker and the loop can easily oscillate if no special precautions are taken by means of voice-controlled switches.

Figure 3.22 is a block diagram of a hands-free/loudspeaking telephone set. The function of voice switches VS1 and VS2 is to switch off the channel that is not active. If the set is in the sending mode because sound is picked up by the base microphone, VS1 is closed and VS2 is opened to avoid feedback via the hybrid coupling network (TEA1060 family speech/transmission circuit). If the set is in the receiving mode (audio from line to the loudspeaker) then switch VS1 is opened and VS2 is closed so feedback via the base microphone is blocked.

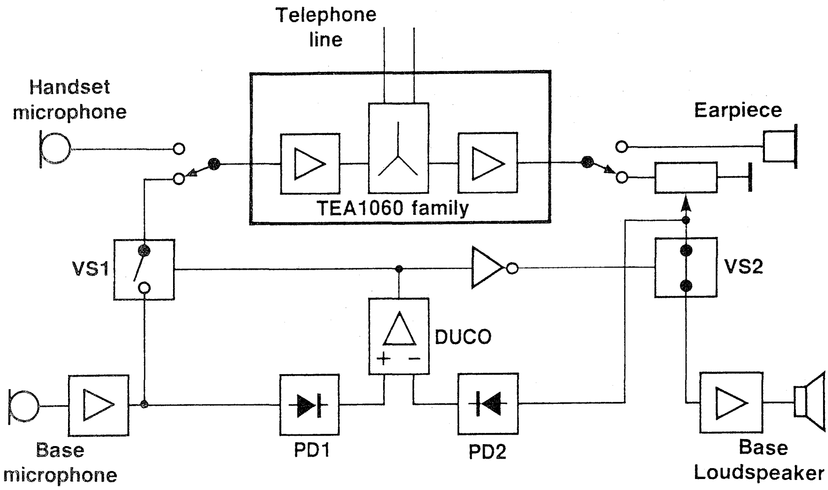


Fig.3.22 - Hands-free/loudspeaking telephone set

Superior performance can be achieved by substituting switchable attenuators in both channels (instead of switching the channels completely on or off). Both voice switches are controlled by a duplex controller (DUCO) which is essential to the operation of this type of set. The DUCO detects which channel the strongest signal is present. It then switches off the attenuator in that channel, and inserts the attenuator in the channel with the weaker signal. An application with the TEA1060 family in a line powered hands-free telephone set is described in laboratory report ETT8508. The attenuation between loudspeaker and microphone is heavily dependant on the acoustics of the telephone casing and the room in which the telephone is located. Also background noise and line signal conditions have a strong influence on the correct operation of this voice-controlled system. In other words the hands-free loudspeaking feature is a very complex problem for the designer of a telephone set.

4. THE DIALLING PART OF A TELEPHONE SET

4.1 Electronic pulse dialling

The pulse dialling system (section 1.3.2) uses line current interruptions to signal the digits dialled to the exchange. The number of line current interruptions corresponds with the digit dialled except for the digit '0' which is characterized by 10 interruptions. The interruption rate normally is 10 Hz ($t_m + t_b = 100$ ms in Fig.4.1) although faster dialling rates are also used (16 Hz in Colombia and, 20 Hz in Japan and in some private exchanges (PABX)). Digits are separated by the time required for the dial to rewind and be wound up again. This is called the interdigit pause (t_{id} in Fig.4.1). The ratio of the off-time (t_b) and the on-time (t_m) of the line current switch is called the break-make-ratio (BMR) or mark-space-ratio. It is typically 3:2 (60 ms break and 40 ms make) or 2:1 (67 ms break and 33 ms make) in a 10 Hz system.

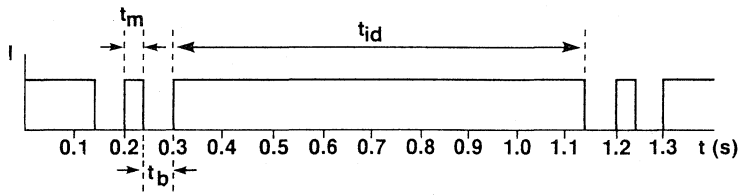


Fig.4.1 - Telephone set current during pulse dialling
(in this case dialling ..-2-2-..)

4.1.1 Electronic pulse diallers

Keypad-operated electronic pulse diallers have been developed to offer subscribers connected to older exchanges with only pulse dialling, similar advantages to those connected to exchanges compatible with keypad operated DTMF sets. These pulse diallers convert push-button inputs into a stream of binary codes that can drive an electronic line current interrupter (relay or transistor) and a mute switch for the speech part. Together they emulate the rotary dialler function.

Since digits can be entered on the keypad much faster than they can be entered with a dial, a buffer memory is required. This buffer memory can be adapted very easily to function as a last number redial memory. This allows redialling of the last number dialled simply by pressing one button (usually the 'R' or '#' button). Most electronic pulse diallers have this redial function.

Usually an inaccuracy of ±10% is allowed in pulse dialling systems. Some types of pulse diallers use a stable and reliable quartz or ceramic resonator but an RC controlled oscillator can be used if the ±10% accuracy can be met.

The logic output signals generated by electronic pulse diallers have "standardized" names:

DP: Dial pulse signal controlling the line current interrupter

$\overline{DP}$: Inverted DP signal

M1: Mute signal indicating when the speech part has to be muted. Active during dialling including interdigit pauses,

$\overline{M1}$: Inverted M1 signal

M2: Digit mute signal. Unlike the M1 signal this signal is only active during breaks and makes but not during the inter digit pauses

M3: Logic combination of the M1 and $\overline{DP}$ signal ($M3 = M1.\overline{DP}$) needed in the parallel architecture system (see section 4.1.2).

These logic signals are illustrated in Fig.4.2.

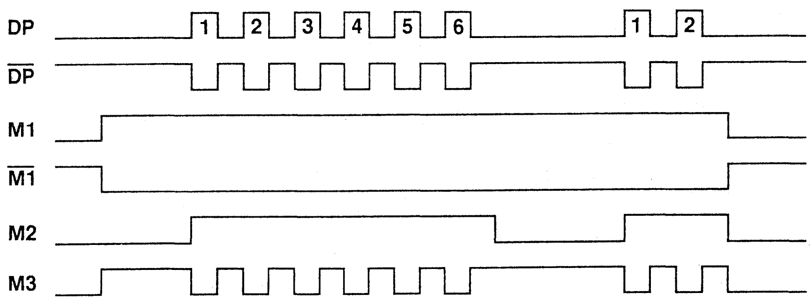


Fig.4.2 - Logic signals of a pulse dialler if '62' is dialled

4.1.2 Architecture of a telephone set

Principally, there are two ways to connect a pulse dialler interrupter in a telephone set:

- In the series pulsing system, the interrupter is connected in series with the speech part (Fig.4.3a). The interrupter must be driven by the DP signal, while the speech part must be muted (or short circuited) by M1.
- In the parallel pulsing system, the interrupter is connected in parallel with the speech part (Fig.4.3b). In the speech mode, the interrupter is opened. As soon as the set enters the dialling mode, the speech branch is discon-nected by $\overline{M1}$ while the current is taken over by the inter-rupter switch in the dial branch (M3). After dialling (M3), the inter-rupter switch is opened and the speech branch is connected to the line again.

With the series pulsing system it is easy to construct an interface (supply and logic signals) between the speech and dialler circuitry. The advantages are:

- only one line interface - fewer external components
- click-free muting because the speech circuit stays switched on
- easy addition of extra dialling facilities.

The parallel pulsing system results in at least two line interfaces. Switch-over is controlled by M1 and pulse dialling by M3. The advantages are:

- its compatible with a conventional set
- the speech circuit can be in the handset where it will be less sensitive to Radio Frequency Interference (RFI) because it's connected by shorter wires.

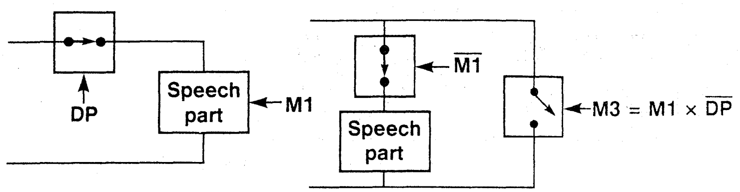


Fig.4.3 - Pulsing systems
a) series b) parallel

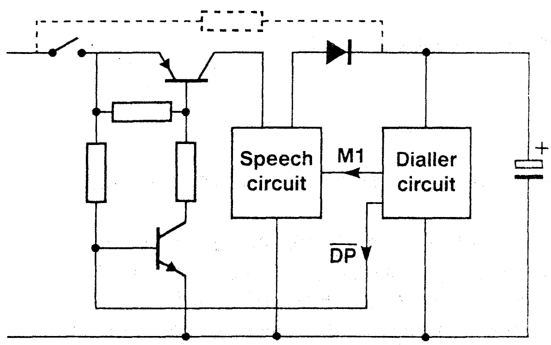


Fig.4.4 - Popular interrupter circuit for series pulsing

Figure 4.4 shows a practical series pulsing circuit with two transistors. Notice that to drive this circuit correctly, the pulse dialler must generate a $\overline{DP}$ signal at an open drain output. The redial capacitor provides a back-up voltage for the redial memory of the pulse dialler while the handset is on-hook. The diode prevents discharging of the redial capacitor via the switched-off speech/transmission circuit.

The duration that the contents of the redial memory can be retained in this way varies from several minutes to half an hour (depending on the initial voltage, the retention current, and the capacitance and leakage current of the redial capacitor). The time can be extended by taking the retention current (some micro amperes) from the telephone line via a resistor of several MΩ (dashed in Fig.4.4). This method is not allowed by all PTTs, but another way to extend the redial time is by using a battery to provide the back-up voltage (section 6.3).

4.1.3 Alternative pulse dialling systems

There are other pulse dialling systems e.g. that used in the city of Oslo and in Sweden. In Oslo the digits are (except for the'0' which is normal) distributed in a complementary way by the rotary dialler. In Sweden, the number of pulses is equal to the digit dialled plus 1, and '0' equals 1 pulse. The differences are listed in the Fig.4.5.

number of interruptions	digit dialled		
	normal	Oslo	Sweden
1	1	9	0
2	2	8	1
3	3	7	2
4	4	6	3
5	5	5	4
6	6	4	5
7	7	3	6
8	8	2	7
9	9	1	8
10	0	0	9

Fig.4.5 - Oslo and Swedish system of pulse dialling

These differences don't cause problems in telephone sets equipped for pulse dialling only, because adapting normal pulse dialling circuits to one of these systems is simply a matter of renaming the push-buttons. For telephone sets incorporating both the pulse and DTMF function, it is more of a problem because the push-buttons don't have the same meaning for both pulse dialling and DTMF dialling. The problem can only be solved by using a special dialler.

4.2 DTMF dialling

4.2.1 Principles

To speed up the dialling procedure and to make it more reliable, a new dialling system was introduced in 1970. In this system digits are transmitted as two tones simultaneously. This explains the name Dual-Tone Multi-Frequency (It's also known as DTMF, tone dialling, and MF dialling). The tone frequencies are selected to avoid harmonic interference from speech signals. There are eight frequencies defined in the DTMF system; four in a low frequency group (697-941 Hz) and four in a high frequency group (1209-1633 Hz). A valid digit is defined as one tone out of the low frequency group together with one tone out of the high frequency group. In total there are sixteen combinations possible as indicated in Fig.4.6.

		high frequency group			
		1209 Hz	1336 Hz	1477 Hz	1633 Hz
low	697 Hz	1	2	3	A
frequency	770 Hz	4	5	6	B
group	852 Hz	7	8	9	C
	941 Hz	*	0	#	D

Fig.4.6 - DTMF tone assignments

Normally, only the digits 0,1,...,9 are used, but some systems also support the * and # signals or even all 16 signals for special functions.

The maximum dialling speed with a DTMF system (limited by the detection time required by the DTMF receiver in the exchange) is typically 7 digits/sec. (assuming a tone burst of 70 ms). With the pulse dialling system, the dialling speed (assuming a dialling rate of 10 Hz and interdigit pauses of 0.8 s) varies between 1.1 (for '1's) to 0.56 (for '0's) digits per second (average rate 0.8 digits per second). The DTMF system is therefore almost 10 times faster!

As well as the dialling function, another application for DTMF is low-speed data transfer e.g. home banking, credit card verification, domestic remote control and exchange facilities.

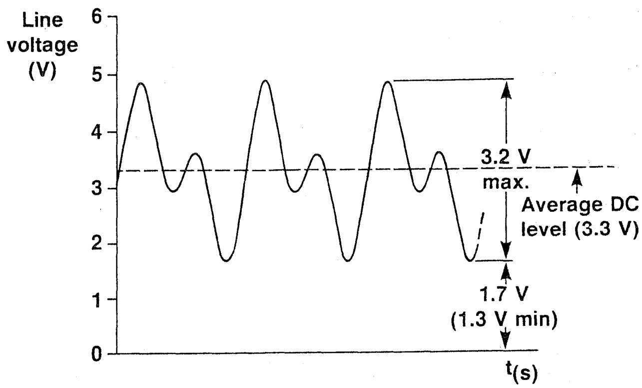


Fig.4.7 - DTMF signal on telephone line

4.2.2 DTMF generation

Two tunable oscillators (one for the low frequency group and one for the high frequency group) can be used to generate DTMF tones (Fig.4.7). However, due to the rather accurate frequency demands (usually ±1.5% spread), LC oscillators incorporating a heavy, expensive coil are required. When IC techniques became available (around 1974) ICs were put onto the market with a crystal oscillator and two synthesizers which generate the DTMF tones digitally.

Although it can't synthesize the exact DTMF frequencies, an inexpensive crystal (originally intended for NTSC colour television) has turned out to be by far the most popular type of DTMF synthesizer clock in Europe. It generates a frequency of 3 579 545 Hz and this can be divided down to the DTMF fre-quencies with only a small error (Fig.4.8). Low cost 3.58 MHz ceramic resonators can also be used in various countries.

frequency required	frequency generated	error in Hz	error in %
697	697.90	+0.90	0.13
770	770.46	+0.46	0.06
852	850.45	-1.55	0.18
941	943.23	+2.23	0.24
1209	1206.45	-2.55	0.21
1336	1341.66	+5.66	0.42
1477	1482.21	+5.21	0.35
1633	1638.24	+5.24	0.32

Fig.4.8 - Division errors in PCD3311/12 DTMF generator

The resulting dividing errors are well within the tolerances allowed (±1.5%) even when the inaccuracy of the quartz crystal itself, is included.

4.2.3 Amplitude and distortion requirements

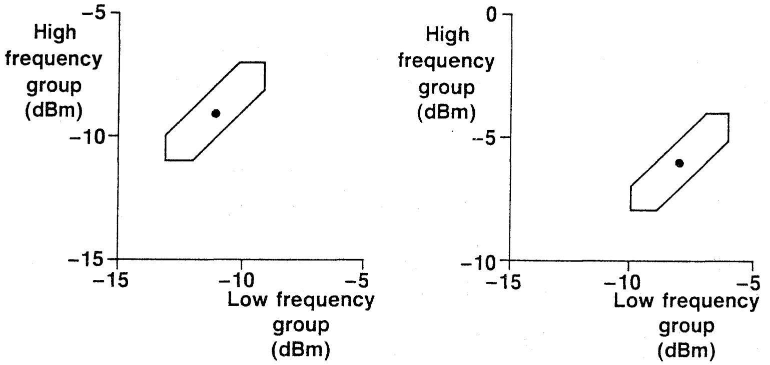
4.2.3.1 European requirements (CEPT recommendations)

The CEPT (Conference des Administrations des Postes et des Telecommunicatins) specify two options for the amplitude requirements:

Option 1: amplitude low frequency group tones: -11 dBm ± 2 dB
amplitude high frequency group tones: -9 dBm ± 2 dB
pre-emphasis: 2 dB ± 1 dB

Option 2: amplitude low frequency group tones: -8 dBm ± 2 dB
amplitude high frequency group tones: -6 dBm ± 2 dB
pre-emphasis: 2 dB ± 1 dB

Figures 4.9a and 4.9b show these two options graphically. The cross-point of the low and high frequency group amplitudes must fall within the boxed areas.



(a) option 1 (b) option 2

Fig.4.9 - CEPT DTMF-amplitude requirements

The difference between the amplitudes of the high and low frequency groups is called pre-emphasis; this compensates for the line losses which increase with frequency. In Europe the amplitude requirements must be fulfilled over the whole line current and ambient temperature range. Often an extra stabilized supply voltage is required to achieve this. All Philips CMOS diallers offer voltage stabilization on-chip.

The CEPT specify two requirements for DTMF tone distortion:

1) The total power level of all unwanted frequency components must be at least 20 dB below the low frequency group component of the signal.
2) The level of any unwanted frequency component must be below the following limits:

300 Hz - 4300 Hz	:	-33 dBm
4300 Hz - 28 kHz	:	-37 dBm at 4300 Hz falling 12 dB/octave
28 kHz - 70 kHz	:	-70 dBm
70 kHz - 200 kHz	:	-80 dBm
above 200 kHz	:	-70 dBm

See Fig.4.10.

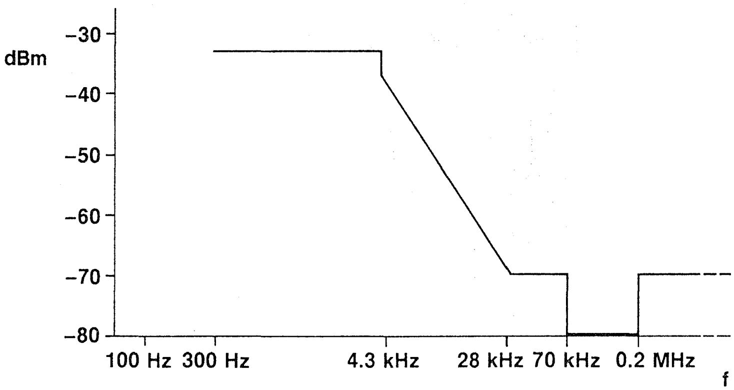


Fig.4.10 - Distortion requirement 2 according to CEPT

These distortion requirements are not so much a problem for the LC or RC DTMF generators, but more for the synthesizers which generate sinewaves approximated from square waves.

In the PCD3310/11/12, the frequencies are composed of 23 square wave steps per full cycle of the sine wave. This results in strong 22^{nd}, 24^{th}, 45^{th}, 47^{th} harmonics. Figure 4.11 shows a synthesized sine wave. The number of time slots is the same for all tones (23) and the amplitudes are quantized to 5-bit accuracy.

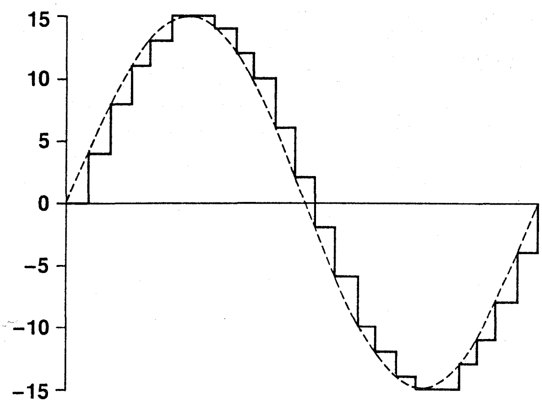


Fig.4.11 - The synthesized sine wave.

To bring this harmonic distortion within the CEPT requirements, extra filter components are required. Note that such a filter can influence the amplitude of the DTMF signals, especially over the high frequency group, and this may affect the pre-emphasis.

All Philips CMOS DTMF diallers, generators, or microcontrollers with DTMF generators on-chip, also have the complete filter on-chip e.g. the PCD3310 which has an on-chip switched-capacitor filter. The spectrum for the digit '1' generated by the PCD3310 is given in Fig.4.12.

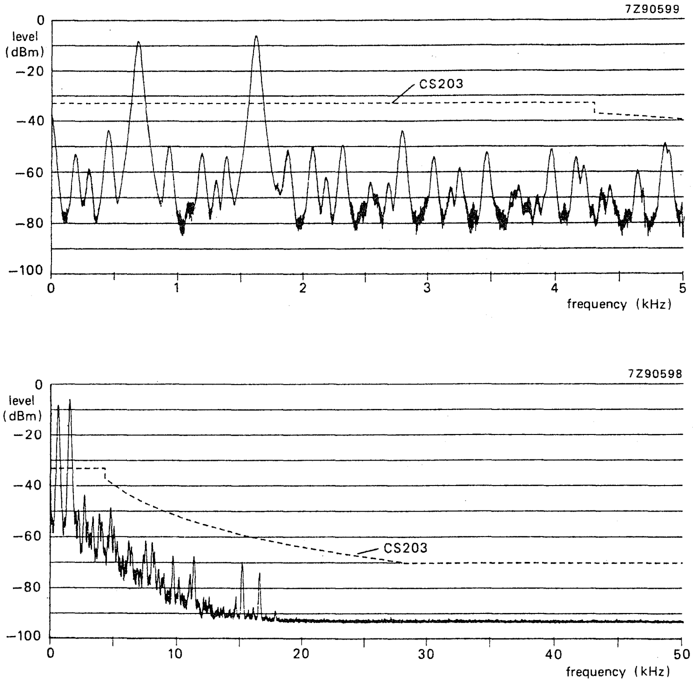


Fig.4.12 - Typical frequency spectra of a dual-tone signal after flat-band amplification of 6 dB

4.2.3.2 USA requirements (RS470)

The requirements for the amplitude and distortion for the DTMF tones in the USA are considerably easier to fulfill than those for Europe. The amplitude requirements are shown in Fig.4.13. There are two limits given; one for a line current of 25 mA, and one for a line current of 100 mA. For intermediate line currents, the maximum and minimum limits can be read from the figure.

Automatic gain control is permitted but not required. If no automatic gain control is used, the amplitudes of the DTMF tones must remain within the cross-hatched area of the figure. Note that the maximum level is given for maximum addition of the high and low frequency group levels.

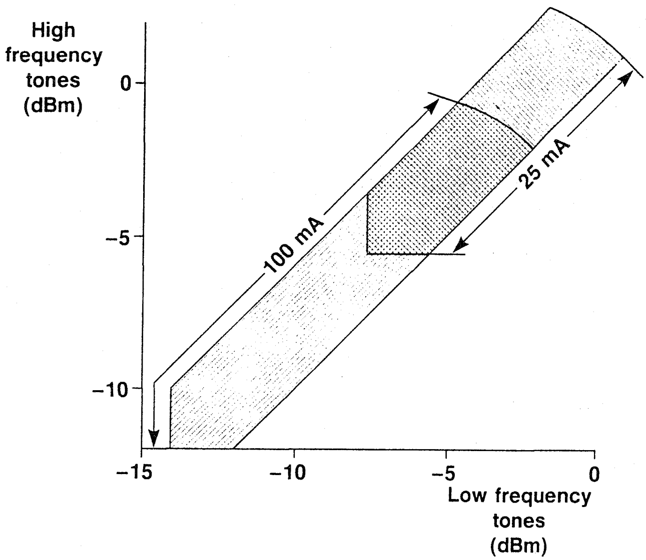


Fig.4.13 - RS470 DTMF amplitude requirements

The total power of unwanted frequency components within the band 500 Hz to 3400 Hz must remain at least 20 dB below the power of the DTMF frequency pair.

4.2.4 Interface to the telephone line

The DTMF tones generated by the DTMF dialler must be applied to the telephone line respecting the AC and DC requirements of the PTT (i.e. BRL and voltage drop as a function of line current). Most bipolar DTMF diallers incorporate an on-chip line interface making it possible to connect them (via a polarity guard bridge) directly to the telephone line (Fig.4.14).

CMOS DTMF diallers are designed to operate together with an electronic speech transmission IC with a DTMF interface. They don't incorporate an on-chip line interface. This approach results in very simple and efficient circuit designs. The DTMF dialler is powered from the speech-circuit peripheral supply point. The DTMF tones are transmitted to the telephone line via the speech-circuit line interface. The mute signal generated by the DTMF dialler, controls the speech circuit and determines when to transmit speech and DTMF signals. The switch-over from speech mode to dialling mode (and back again) can be realized without noticeable audible clicks.

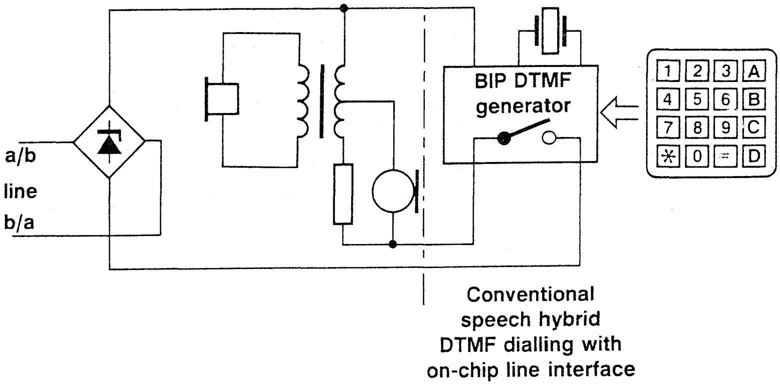


Fig.4.14 - Interfacing a bipolar DTMF dialler to the telephone line

If the speech circuit passes part of the signals on its DTMF input to the earpiece output, a confidence tone will be introduced. This approach is called the common line interface architecture because both the speech and dialling parts of the telephone set are connected to the line by the same interface (see Fig.4.15 and section 4.1.2).

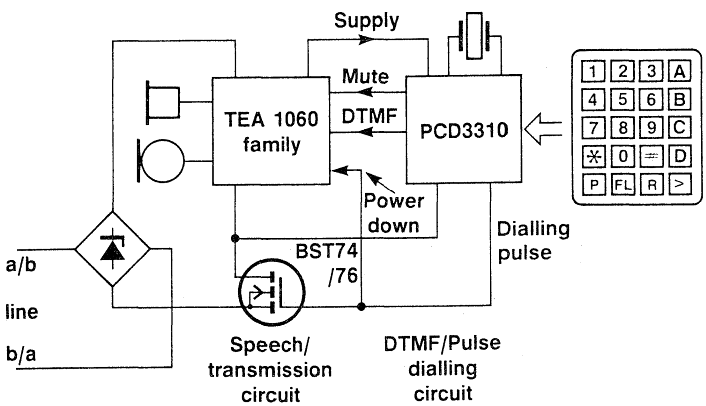


Fig.4.15 - Example of common line interface architecture

If an appropriate speech circuit is not available for interfacing the CMOS DTMF dialler to the telephone line, a separate line interface for the dialler must be used. This usually requires a fairly large number of discrete components.

5. THE RINGER

5.1 Supply

The principles of the telephone set supply are explained in section 2; the DC line current enters the telephone ringer via a feeding bridge which includes some resistance and high inductance. In the exchange, several AC signals are superimposed on this DC; dialling tone, busy tone, tariff pulse, and a ringer voltage. A rather high voltage is used to alert a subscriber. This may be superimposed on the DC signal (depending on the exchange).

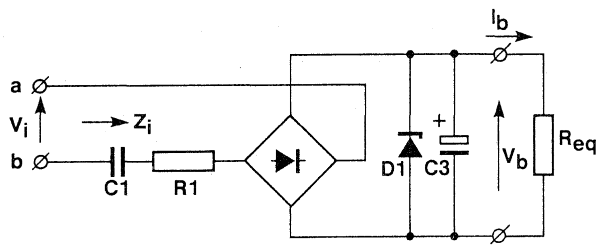


Fig.5.1 - AC-to-DC conversion

On-hook, the ringer circuit is connected to the telephone line via a series-connected capacitor C1 and resistor R1. The capacitor blocks DC and the resistor defines the minimum impedance, affords protection against mains spikes and lightning induced surges, and sufficiently reduces the power requirements to allow IC implementation of an electronic ringer. The AC line voltage (e.g. 50 V RMS/25 Hz) is converted to a DC supply voltage V_b that delivers power to the ringer circuit (equivalent resistance R_{eq} in Fig.5.1.). This DC supply is converted by the ringer, via an output driver to AC for the transducer which converts it to audible sound. The AC-to-DC conversion depends on the ringer frequency, and on the values of C1, R1, and R_{eq}. The required minimum input impedance Z_i varies according to PTT requirements, and on whether series or parallel connection of ringers is permitted (Z_i is the RMS input voltage divided by the RMS input current). Figure 5.2 shows an application diagram of a simple 2-tone ringer circuit with a piezo-electric (PXE) transducer.

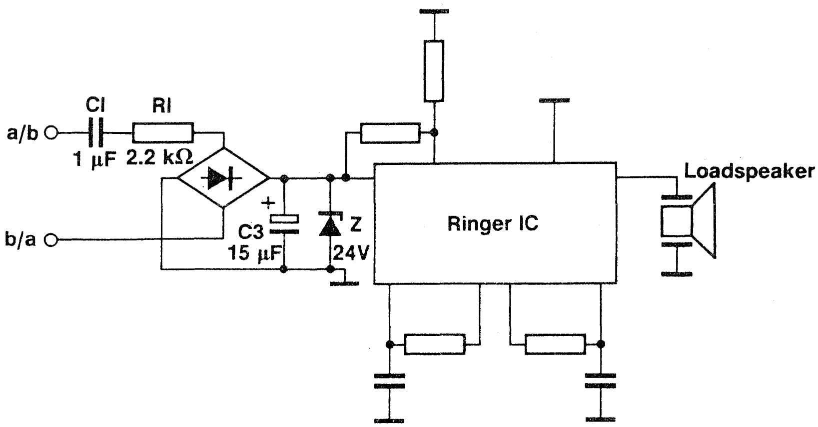


Fig.5.2 - Ringer circuit

5.2 Characteristics

The AC ringer voltage varies from 32 to 90 V RMS, and its frequency from 25 to 50 Hz (minimum of 16 Hz e.g. Japan, maximum 60 Hz). The ringer operates when a burst (or series) of alternating voltages are applied. Timing of the AC voltage varies considerably for different countries and exchange types. Another very important characteristic is the impedance of the ringer circuit. On one hand, the ringer energy is limited, but on the other, an alerting signal is required. The signal level is not only influenced by the line length, but also by the number of ringer circuits connected in parallel with the same line. To achieve a sufficiently loud ring, large and expensive bells or sirens are needed. The introduction of electronic ringers allows a much higher frequency (500 - 1500 Hz) alert signal with increased audibility because the sensitivity of the human ear in this frequency region is greater. Elderly and disabled people can also hear these frequencies much better.

5.3 Multiple-tone ringers

Electronic tone ringers generate tone bursts during the AC bursts. Tone generation may be single-, dual-, triple-, or multi-tone sequences depending on the ringer type. The more advanced multi-tone types are more pleasant. These have a ringer tone generator which generates the tone frequency during the on-sequence; the frequency is modulated in an LF rhythm. This produces a warble tone in sync. with the AC burst. The 2- and 3-tone ringers reproduce their fixed multiple frequencies continuously.

Yet more advanced is the multi-tone ringer which can generate complete melodies to customize telephone ringers in an office environment. The four standard PCD3360 melodies are shown in Fig.5.3.

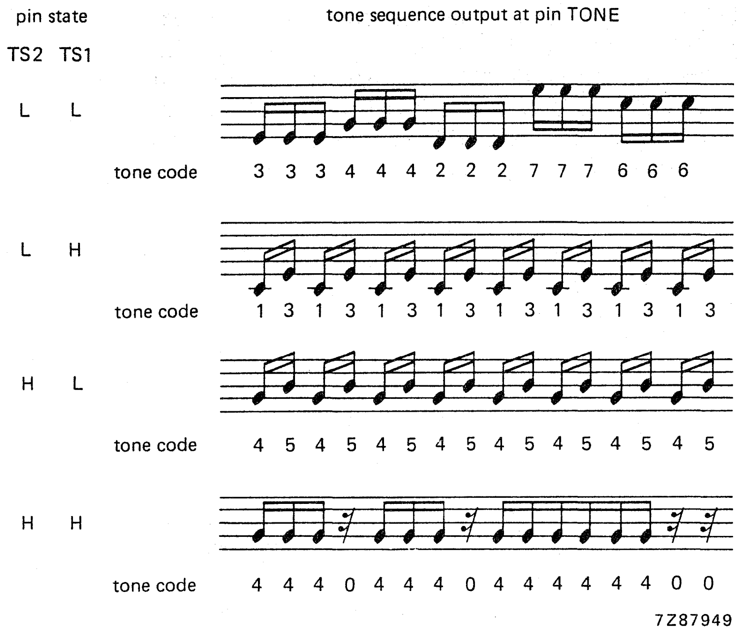


Fig.5.3 - PCD3360 Standard melodies

The ringer can be a PXE transducer or a small loudspeaker. The PXE transducer is very attractive because it can be connected directly to a high impedance square wave output and, if properly designed, can be very efficient at the higher frequencies. The difficulty with a small loudspeaker is that is must have a high impedance (a few kΩ) and is therefore rather expensive. An impedance transformer or a class-D output stage must be used to drive a low impedance loudspeaker (e.g. 50 Ω). When set in the loudspeaker mode (DM = LOW), the PCD3360 has a class-D output (Fig.5.4).

5.4 Features

Extra features are possible with electronic ringers. Anti-tinkle protection avoids ringer activation during pulse dialling (10 Hz) on a parallel connected set. Another feature is automatic swell. Here the output level progressively increases in steps after the ringer voltage has been applied for the first time. This gently increases the alert volume.

The output from the PCD3360 is a delta-modulated signal that approximates a sine wave sampled at 32 kHz. The duty cycle of this signal can be controlled thereby also controlling the DC resistance, the input impedance and the alert volume (Fig.5.4). Automatic swell can be selected whereby the duty cycle, and therefore the alert volume, increases in three steps so that the maximum level is reached after the third AC ringing burst.

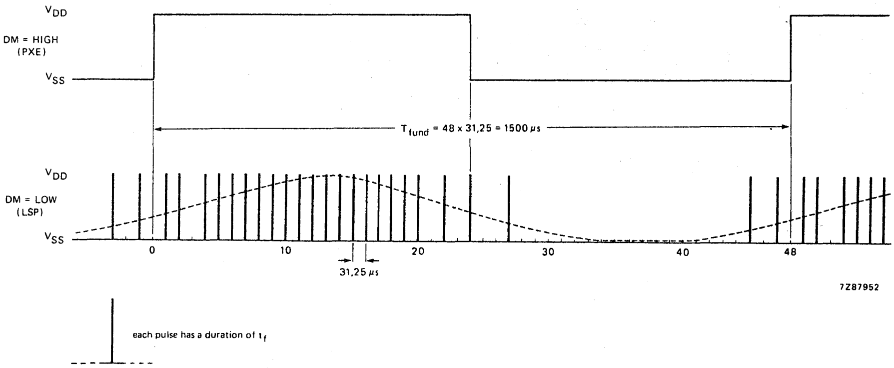


Fig 5.4 Output waveforms of the PCD3360 in PXE mode and loudspeaker mode

6. ADDITIONAL FACILITIES WITH FEATURE-PHONES

Telephone sets with enhanced user facilities are called feature-phones. These facilities enhance the three basic operations; speech processing, dialling, and alerting:

Speech processing enhancements:

- receiver volume control
- listening-in
- hands-free.

Dialling enhancements:
- redial
- repertory dial
- extended redial
- notepad
- flash
- PABX digits
- direct dialling
- on-hook dialling
- music on hold.

Ringer enhancements:
- multiple tone and dedicated melody generation
- automatic swell
- built-in anti-tinkling.

6.1 Architecture

For a basic telephone set, there may be a requirement for one IC for each of the three basic functions (dialling, speech and alert). With feature-phones, additional functions may require several more ICs or a higher degree of integration. For basic telephone sets, the availability of sufficient current to supply the ICs presents the problems described in section 2. With feature-phones, these supply problems are even more severe because of the current requirements of the additional ICs. In some circumstances, it may be necessary to consider a mains power supply. For cost/production reasons the future trend will be in the direction of integrating the three basic functions on one chip. Of course, in future, BiCMOS or analogue CMOS will facilitate effective integration of the three basic functions on one IC. However, for now, it is still far more effective to use bipolar technology for the speech and ringer parts, and CMOS technology for the dialling part. The architectural approaches for the three basic parts must therefore be analyzed separately.

6.1.1 Speech/transmission architecture

In the speech/transmission part, two architectural approaches are possible:
- single-chip approach - the basic speech/transmission functions and facilities for listening-in and hands-free are built-in. The drawback is that it requires at least three separate solutions; one for basic telephones, one for basic telephones with listening-in facilities, and one for feature-phones with listening-in and hands-free facilities.
- building block approach - for the basic speech/transmission functions, one common solution is used for all telephone sets (the TEA1060 approach). An additional block for telephone sets with a listening-in facility (TEA1081 and TDA7050 solution) and another block for the hands-free facility can simply be added. This gives manufactures planning production of comprehensive families of telephone sets, a more flexible and economic solution than the single-chip approach.

6.1.2 Dialler architecture

The dialler has three main architectures:

1. **a one-chip architecture** - including microcontroller, memory, LCD drivers, DTMF and clock/calender on one chip. It may offer a low-cost solution, but the design is less flexible.
2. **building block architecture by means of I^2C-bus* connected ICs** - This is a multi-chip, highly flexible solution, with the advantage of short development times. This is an economic approach (see PCD3315 and PCD3343 I^2C-bus feature-phone concept Figs 6.1 and 6.2). The PCD3315/43 feature great flexibility with a low pin-count; all information is handled by the I^2C-bus via the serial SDA and SCL lines, and no extra pins are required for decoding, addressing, or data conversion. The 100 kbit/s I^2C-bus is fast enough for most telephone set applications. The ICs should either have the I^2C-bus interface on-chip, or a software-emulated equivalent can be implemented using two of the micro-controller's I/O pins.
3. **one-chip basic telephone set architecture with add-on module for LCD and clock calender function** - The controller includes the dialler, memory, and DTMF generator. An additional LCD module,including a clock/calender function, can be added for the feature-phone if required. This has economic advantages over the two other architectures. It offers one solution for all electronic telephone sets with the flexibility of the add-on LCD display and clock/calender function (with an additional module) offering increased layout flexibility (examples: PCD3344/002 and PCD3344/004; Fig.6.3).

* *Purchase of Philips I^2C components conveys a license under the Philips' I^2C patent to use the components in the I^2C-system provided the system conforms to the I^2C specification defined by Philips. More information can be obtained from the brochure "The I^2C-bus ICs" publication no. 9398 342 9011, and the "I^2C-bus specification" publication no. 9398 336 70011.*

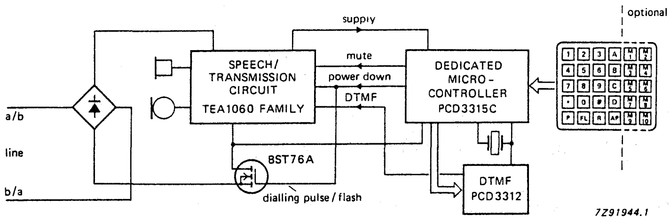


Fig.6.1 - Feature-phone using dedicated microcontroller PCD3315C

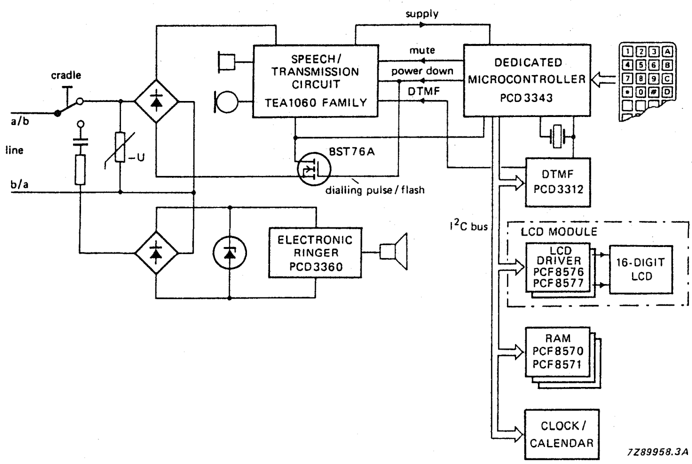


Fig.6.2 - Feature-phone using dedicated microcontroller PCD3343

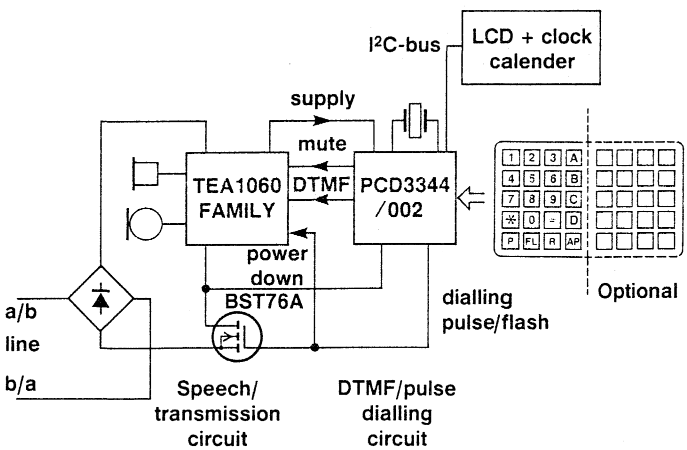


Fig.6.3 - The PCD3344 with a microcontroller and a DTMF generator on-chip with an add-on LCD module

6.2 Feature-phone facilities

6.2.1 Speech part facilities

- **Speech/transmission circuit** - This is the main line-interface for microphone and earpiece signals, and also for the DTMF transmission and peripheral supply. It can be adjusted to meet the DC and AC requirements of a telephone set. The speech part is limited to 'soft speaking' via the handset.
- **Listening-in** - This facility is used to amplify the received signal for a loudspeaker so that more than one listener can follow the conversation. Problems to solve include loudspeaker feedback through the microphone and the ability to power the peripheral circuitry from the telephone line (power for the loudspeaker). See section 3.4.1.
- **Hands-free** - With this feature the handset can be used for 'soft speaking' and an additional microphone and loudspeaker for conference-type communication. As well as the listening-in supply problems, another problem is how to avoid feedback from the loudspeaker to the microphone. One method is the use of a voice-controlled switch or duplex controller (DUCO). This controls switches which connect either the receiving or the sending channel to the line. See section 3.4.2.

6.2.2 Enhanced dialling facilities

- **Redial** - The last dialled telephone number is stored in a dedicated memory and the previous one is overwritten. Repeated dialling is possible with a single key press (redial button) if the line is engaged, or there is no answer.
- **Extended redial** - Telephone numbers are stored in another memory for a longer time and not overwritten after a new call. A special key procedure is required to store and recall this extended memory number.
- **Emergency call** - This feature is also called direct call or hot line. In an emergency, this number will be transmitted if the set is put in a special state (e.g. via a key switch) and any key pressed.
- **Note pad** - This is a dial facility to program a telephone number during a conversation with directory enquiries. After on-hook/off-hook this number can be transmitted. Sometimes the memory allocated for this, is common to redial and/or extended redial memory. This means for recall, the user has to perform a similar key procedure.
- **Repertory dial** - This facility stores a catalogue of regularly dialled telephone numbers. Two approaches are possible. Using a standard keypad (0 - 9), a maximum of 10 repertory numbers can be redialled with 2 keys (e.g. P then 0 - 9), or a maximum of 100 numbers can be recalled with 3 keys (e.g. P then 0 - 9 and 0 - 9). The second solution is an extended keypad with special buttons for repertory numbers; each button represents one (sometimes two) repertory numbers, so the user may attach a name card to it. This is also called one-touch memory.

- **Chain dialling** - This is a dialling procedure to dial several repertory numbers within one call. This facility is useful if a long telephone number consists of several parts, stored in different locations of the repertory memory.
- **PABX pre-digits** - If the set is connected to a private automatic branch exchange (PABX), the dialling procedure for external calls to the public exchange (also known as central office) usually has to stop momentarily (during access pause) to allow switch-through by the PABX. Because these pre-digits are determined by the PABX, they're fixed and the user always has to dial the same code for external calls (e.g. 0). A circuit can be built-in to the dialler to detect this pre-digit(s) for automatic insertion with an access pause of a fixed duration e.g. 1 to 3 seconds.
- **Flash** - This facility is also called register recall or white button function. It is used to warn the local exchange that a call has to be transferred. After the dialling tone has been heard, the caller can continue dialling in the normal way. Two methods are used to warn the exchange that a recall is required:
  - connection of the a/b wires to the earth terminal of the telephone set
  - a timed break of the DC line current.

  Both can be detected in the exchange (PABX) and offer the user transfer of a call from one set to another within the same exchange. The duration of such a flash is standardized to one of several values: 100 ms, 270 ms (France), 650 ms (USA).
- **Floating RAM** - This is not a user facility but a method of using RAM more efficiently. The telephone number for an international call may be very long, but the average numbers may be much shorter than the maximum length. Instead of allocating a full memory block for each number, RAM capacity can be used more efficiently by filling up the RAM space completely without spaces. Usually an indication of a full memory is required.
- **Music on hold** - This facility allows a melody to be generated which can be heard on the other subscribers' phone to indicate that the subscriber is on hold and not disconnected.
- **Confidence tone** - Soft tones can be generated in the earpiece to indicate when keys are pressed.
- **On-hook dialling** - This is a dialling procedure that allows the user to dial without lifting the handset. After the connection has been made, the handset is lifted and conversation can start. To monitor the dialling procedure by the caller, a listening-in (or on-hook dialling) loudspeaker facility may be built-in.

- **Mix-mode dialling** - There are two dialling methods; pulse and DTMF. In most countries, there is a mixture of pulse and DTMF exchanges. For economic reasons the PTT uses telephone sets with both dialling options in the same set. These sets can also be used for mixed-mode dialling. Here, the first part of the telephone number must be dialled by pulse dialling and the second part (usually for long distance calls) by DTMF dialling. It is also possible to use these sets for facilities which require low-speed data transmission e.g. home banking, credit card verification, remote control etc. The use of pulse and DTMF dialling during one call is called mix-mode dialling.

6.3 Memory retention

Memory retention of the stored telephone numbers is required even when the handset is on-hook. There are several ways to do this:

- **Mains power** - the set and RAM supply is from the mains. Problems occur when the set is disconnected from the mains or if there is a mains failure. It's a very unreliable way to ensure memory retention (Fig.6.4).
- **On-hook current** - some administrations allow current to be drawn from the telephone line while the handset is on-hook. However, only a very limited standby current of a few microamps (via. a bleeder resistor) is permitted.
- **Battery** - The disadvantages of battery back-up are the limited lifetime, extra space, cost and separation of the two supplies. This requires two extra diodes to safeguard supply to the IC either from the telephone line (during off-hook) or from the battery (during on-hook). See Fig.6.4. Another disadvantage is that one forward diode voltage drop is lost. The left-hand diode is therefore usually a Schottky type to limit the loss to 0.3 V instead of 0.7 V.

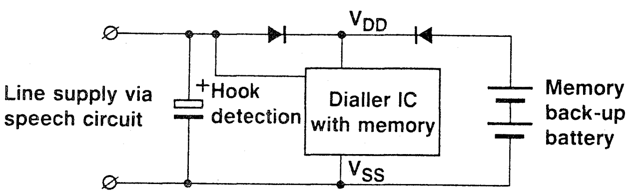


Fig.6.4 - Feature-phone with memory supplied by battery

- **Capacitor** - In Fig.6.4, the battery can be replaced by a large capacitor if only the redial memory is on-chip, if storage of the number last dialled is only required for 5 to 30 minutes. The diode on the right-hand side of Fig.6.4 can then be omitted.
- **Non-volatile RAM (e.g. EEPROM)** - It is possible to design RAM which can be written and erased electrically so that the RAM contents are retained when the phone is on-hook. Battery or capacitor supplies are then unnecessary. Access times for writing and erasing can be rather long (some 10 to 30 ms) but, for telephone number storage, this is still acceptable. The Philips PCB8582 is a serial EEPROM with a capacity of 256 x 8-bit. Microcontrollers with on-board EEPROM are becoming more and more common.

Where information has to be retained by the memory during stand-by, certain procedures have to be built-in to prevent data loss. The Philips concept has 2 procedures:

- **Power-on reset circuit on-chip** - As soon as the supply voltage for the memory falls below the minimum retention voltage, a reset is automatically generated which can be used to clear part, or all of the RAM; this avoids corruption of telephone number data and dialling of a wrong number.
- **RAM check procedure** - This can be built-in to the microcontroller software. Just prior to going on-hook, all RAM addresses with stored telephone numbers are added and the 8 least significant bits are stored in a special RAM address. After the handset is lifted again, the above procedures are repeated and the new summation is compared with the previous one. Inequality causes the memory to clear, and prevents recall of incorrect telephone numbers.

6.4 Mains power supply

There are a few things to be considered with mains supplies:

- **Isolation** - PTTs require the isolation between the telephone line and the mains to withstand a test voltage of 2 to 4 kV. Digital and analogue circuitry connected between the telephone line and the mains must also meet these safety standards; transformers and optocouplers can be used for this purpose. Transformers use no secondary power but can only interface AC signals. Optocouplers can interface AC and DC signals, but also need supply current on the secondary side. With optocouplers, non-linear behaviour with analogue signals causes problems with microphone, earpiece and DTMF-dialling signals. Laboratory report ETT8709 describes how optocouplers can be used for analogue signal communication with little distortion.
- **POTS** - This abbreviation stands for Plain Old Telephone Service. A feature-phone must always be able to perform simple manual dialling during a mains failure. Usually, this requires extra dialling and speech components at additional cost.

2 HIGH VOLTAGE LINE INTERFACES

CONTENTS

APPLICATION NOTE Nr	AN95023
TITLE	Application of the UBA1702/A Line Interrupter Driver and Ringer Circuit
AUTHOR	H. Derks
DATE	November 1995

CONTENTS

LIST OF FIGURES

1. INTRODUCTION

The considerations which played an important role during development of the UBA1702/A Line Interrupter Driver and Ringer Circuit were the following:

- the circuit must be complementary to the existing range transmission circuits (TEA106X/TEA111X family) and microcontrollers (PCD33XX) in such a way that these ICs in combination with the UBA1702/A offer a complete system solution. Therefore the software controlled ringer concept is supported. In this concept the ringer-melody is generated under control of (microcontroller-)software. This flexible approach makes adaptation of the melody to the individual preference of the user possible.
- a certain degree of versatility must be built in to meet requirements for various countries.
- the number of required external (discrete) components must be minimized.

The proposed system concept will be illustrated with the help of figure 1.

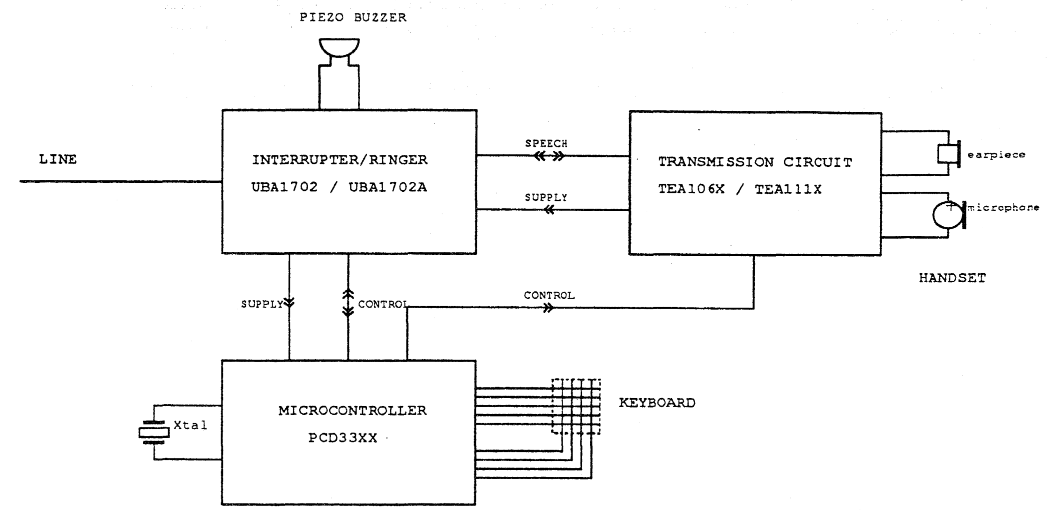


Fig.1 System concept corded telephone.

Two different operation modes can be determined:

the ringer mode. This mode occurs during 'on hook', the set is in quiescent condition. After applying an AC ringer signal to the set a square wave with twice the ringer signal frequency is generated by the UBA1702/A. This square wave signal is checked by the microcontroller for the correct frequency (frequency discrimination). In case of a correct ringer frequency a melody is generated by the microcontroller which is fed back to the UBA1702/A in combination with a specific output volume setting. After a check for the correct amplitude of the ringer signal the generated melody is amplified by the UBA1702/A as well to the desired level.
A piezo transducer makes the melody audible. The microcontroller is supplied from the telephone line via the UBA1702/A.

the conversation / dialling mode. This mode is related with 'off hook', the set is in dialling or loop condition. In this mode the UBA1702/A serves as an overvoltage- and overcurrent protection circuit for the transmission circuit. The IC also contains a line current detection circuit to inform the microcontroller about the status of the line and the hookswitch. In pulse dialling mode the UBA1702/A takes care of interrupting the loop current during the breakperiod and for proper set resistance during the make period. The microcontroller is supplied via the UBA1702/A directly from the line or indirectly by the supply point of the transmission circuit.

To meet specific market demands two different versions are developed:
- the UBA1702, intended to drive an external PMOST interrupter and
- the UBA1702A which is adapted to drive an external bipolar PNP interrupter.

The UBA1702/A is fabricated in Philips' proprietary High Voltage Bipolar Complementary and Double-diffused Metal Oxyde Semiconductor (HV-BCDMOS) process. The circuit is capable of withstanding up to 400 V on pins connected directly or indirectly to the telephone line.

The detailed description in this report is given by means of the blockdiagram of the UBA1702/A (ch. 2) and by discussing the details of each block (ch. 3). Notes on the overall system performance can be found in ch. 4. The application is treated by giving guidelines for application also in relation with specific requirements (ch. 5) and by giving a number of worked-out examples including electronic hookswitch application (ch. 6). EMC aspects are discussed (ch. 7) and a list with references is given (ch.8). The appendices contain quick reference data for the TEA106X family (A), a list of abbreviations (B) and application diagrams of the UBA1702/A printed on A3 size paper (C).

Note: by the addition /A in UBA1702/A two different circuits are meant at the same time: UBA1702 as well as UBA1702A.

2. BLOCKDIAGRAM

The blockdiagram of the UBA1702/A is depicted in figure 2 while the pin configuration is given in figure 3 with the belonging table.

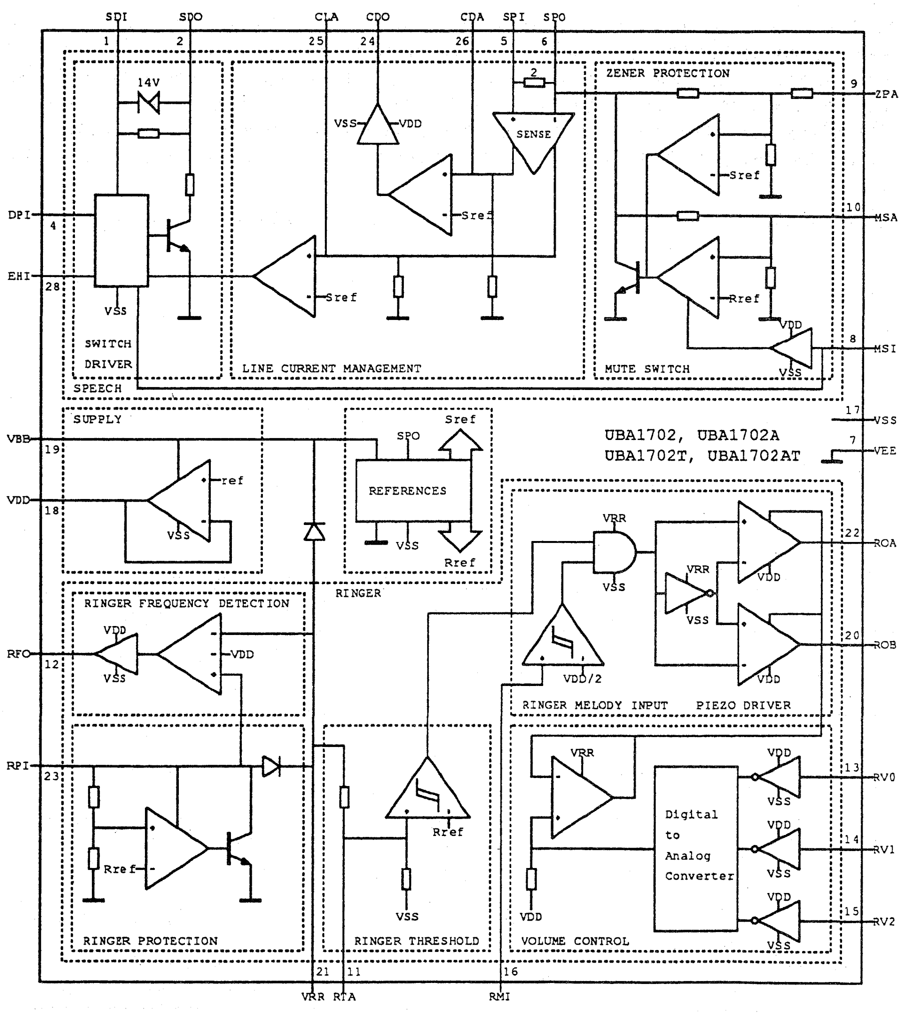


Fig.2 Blockdiagram

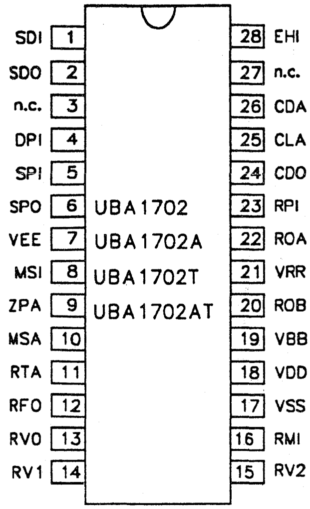


Fig.3 Pinconfiguration

SYMBOL	PIN	DESCRIPTION	SYMBOL	PIN	DESCRIPTION
SDI	1	Switch Driver Input.	RV2	15	Ringer Volume; bit 2 (MSB).
SDO	2	Switch Driver Output.	RMI	16	Ringer Melody Input.
nc	3	not connected.	VSS	17	Ground ref. micro and ringer.
DPI	4	Dialling Pulse Input.	VDD	18	Microcontroller supply.
SPI	5	Speech Part Input.	VBB	19	Supply from transm. circuit.
SPO	6	Speech Part Output.	ROB	20	Ringer Output B.
VEE	7	Ground ref. transm. circuit.	VRR	21	Ringer supply.
MSI	8	Mute Switch Input.	ROA	22	Ringer Output A.
ZPA	9	Zener Protection Adjustment.	RPI	23	Ringer Part Input.
MSA	10	Mute Switch Adjustment.	CDO	24	Current Detection Output.
RTA	11	Ringer Threshold Adjustment.	CLA	25	Current Limitation Adjustment.
RFO	12	Ringer Frequency Output.	CDA	26	Current Detection Adjustment.
RV0	13	Ringer Volume; bit 0 (LSB).	nc	27	not connected.
RV1	14	Ringer Volume; bit 1.	EHI	28	Electronic Hookswitch Input.

In the blockdiagram of figure 2 three main parts can be distinguished:

- speech interface part
- ringer part
- supply

The subparts / blocks which correspond with these parts and the related pins are shortly described below.

Speech interface part

This part can be subdivided into:

switchdriver. This block controls via pins SDI and SDO the external PMOST (in case of UBA1702) or external PNP (in case of UBA1702A) interrupter switch. By making DPI high the interrupter is switched off, which corresponds with a line break. For basic applications pin SDI and EHI are shorted, but by applying a control signal to EHI the interrupter switch can be used as hookswitch.

line current management. This block measures the line current flowing through pins SPI and SPO. This line current information is used for two purposes:

- protection: the line current has to be limited in case of over current. The limiting value is adjustable via pin CLA.
- detection: a minimum value has to be detected to determine line- and hookswitch status. This minimum value is adjustable via pin CDA, the status information (logic signal) is available at pin CDO.

mute switch / zener protection. This block influences the voltage present on pin SPO in the following cases:

- overvoltage: for protection of the transmission circuit the voltage is clamped to a value adjustable by pin ZPA.
- DMO operation: to assure a specific set resistance during the make period of pulse dialling the voltage is clamped at a low value (adjustable via pin MSA) under control of pin MSI. This function is also designated as NSA or Mute2.

Ringer part

This part includes:

ringer protection. This block shows similarities with a zener diode. It limits the ringer voltage present on pin RPI.

ringer frequency detection. The AC ringer signal is converted by this block to a logic (square wave) signal with twice the frequency. This signal is available at pin RFO.

ringer threshold. The ringer signal is checked for the required minimum level by this block. The level is adjustable by means of pin RTA.

volume control. To control the ringer output volume three control inputs: RV0, RV1 and RV2 are available.

melody input / piezo driver. The ringer melody is applied to pin RMI. The output stage is enabled by the ringer threshold block if the voltage on pin VRR crosses a certain threshold. The piezo transducer can be connected between pins ROA and ROB.

Supply

A stabilized voltage is available to supply the microcontroller via pin VDD. Input of this supply is VRR in ringer mode and VBB in conversation / dialling mode.

VEE is the ground reference for speech and VSS for the ringer and the microcontroller.

3. BLOCKDESCRIPTION OF THE UBA1702/A

This chapter describes in detail the three main parts of the IC: the speechpart (ch. 3.1), the ringer part (ch. 3.2) and the supply (ch. 3.3). For each part the operation principle is described and its adjustments and performance are discussed.

All values and graphs given in this chapter are typical and at room temperature unless otherwise stated.

3.1 Speech interface part

The speech interface part consists of the switchdriver (ch. 3.1.1), the line current management (ch. 3.1.2) and the zener protection / mute switch (ch. 3.1.3).

3.1.1 Switchdriver of the UBA1702

Principle of operation

The blockdiagram of the switchdriver with external components is given in figure 4.

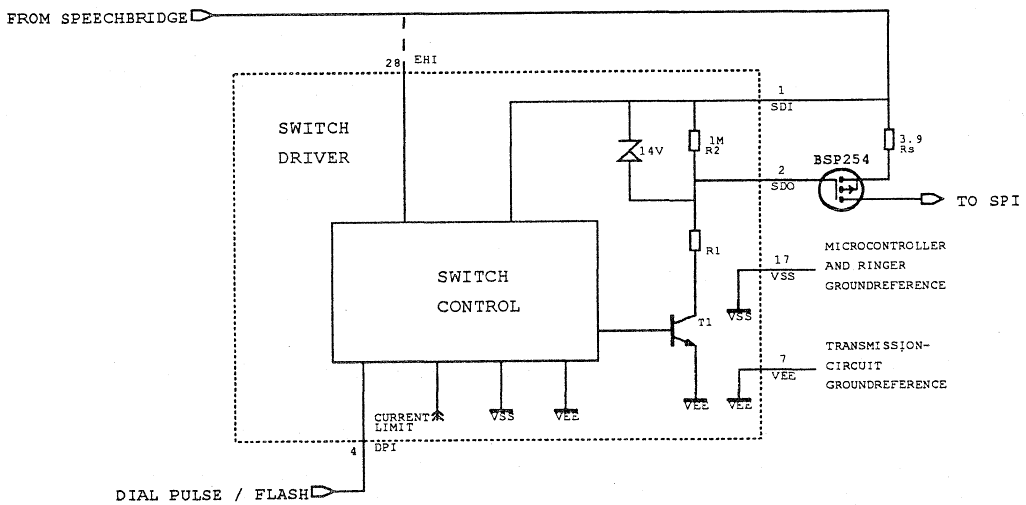


Fig.4 Switchdriver of the UBA1702

Start-up of the circuit is accomplished by applying a voltage greater than one junction voltage (about 0.6 V) to pin EHI. In the basic application with mechanical hookswitch this voltage is derived from the telephone line: EHI has to be connected to SDI. In case the interrupter switch is also used as an Electronic Hook Switch (EHS) the output of a microcontroller may be connected to pin EHI. This pin has a high voltage capability (400 V_{peak}). The same holds for SDI and SDO.

The input voltage on EHI is converted to a drive current which finally controls the external PMOST by creating a voltage drop across R2. By making DPI high the drive current is interrupted and the PMOST is switched off. The reference of DPI is VSS, which may be different from the one of the transmission circuit (shifted reference), which is VEE. This solution makes it possible to use all members of the TEA106X family:

- in case of TEA1064 peripherals (e.g. microcontroller) are supplied between VCC and SLPE. The VSS pin of the peripheral has to be connected to SLPE and the VEE pin of the UBA1702/A with the VEE pin of the TEA1064.
- in all other cases peripherals are supplied between VCC and VEE. The VSS and VEE pins of the UBA1702/A have to be interconnected.

The PMOST is switched on by the voltage across R2, which is caused by the drive current. The voltage is limited by an internal zenerdiode to 14 V. A current source, controlled by the DMO signal (see ch. 3.1.3), has been added to supply drive current during low voltage DMO operation. To put it in another way: if MSI is high (= VDD) it is not possible to create a linebreak by making EHI low, this is only accomplished by making DPI high (= VDD).

The active state of the DPI input has been chosen in such a way that it has the same parity as the power down input of the transmission circuit, viz low is make and high (> VSS + 1.5 V) is break (corresponds with power down). In case very fast transients may occur (e.g. lightning) a series resistor Rs (value 3.9 Ω) in the source connection of the PMOST is recommended.

Performance

The performance of the switchdriver / PMOST combination depends mainly on the characteristics of the PMOST: the outputvoltage on pin SDO (reference VEE, DPI low, voltage on SDI < 12 V) is lower than 0.2 V. Consequently almost the entire voltage across the transmission circuit is available to minimize the ON-resistance of the PMOST and maximize the output swing of the transmission circuit. Delay and switching times are determined by the value of the drive current, R2 and capacitances and voltages across the PMOST. The drive current is about

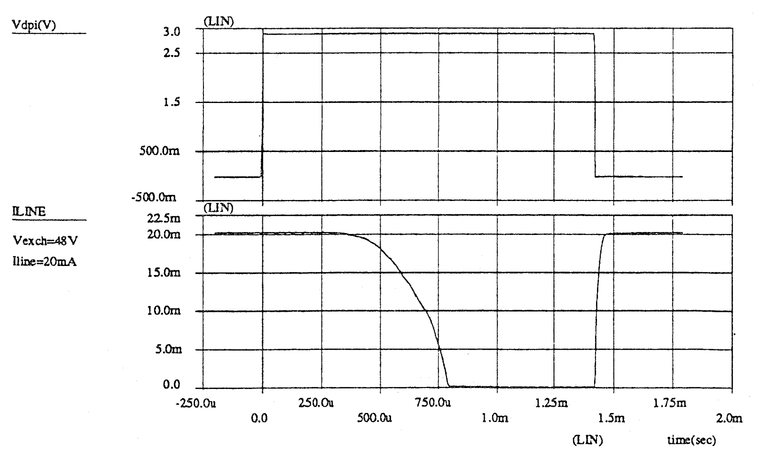


Fig.5 Typical switching times of UBA1702 combined with BSP254

20-30 μA (if voltage on EHI > 1.5 V). If a BSP254 PMOST is used the typical turn-off delay time is 700 μs, the switching-off time 250 μs, the turn-on delay time 50 μs and the switching-on time 25 μs, see figure 5. For testcircuit the basic application (see figure C1 in the appendix) has been used.
It may be clear that if EHI is low, no drive current is generated by T1 and no current is drawn from SDI. This is important for EHS application while SDI is permanently connected to the line. Therefore during 'on hook', while the set is in quiescent condition, an insulation resistance of greater than 5 MΩ can be guaranteed. Insulation resistance is not always equal to break resistance. Break resistance is the loop resistance during the break period of pulse dialling:

- in basic application (interconnection between EHI and SDI) this resistance is solely determined by the input resistance of pin EHI and greater than 100 kΩ. If the voltage at EHI exceeds a certain value (typical 35 V) the current through this pin remains constant and therefore the resistance increases.
- in EHS application (EHI driven from external source) the break resistance is equal to the insulation resistance (> 5 MΩ).

The difference between break and insulation resistance is elucidated by means of figure 6. The given resistance values are for indication only.

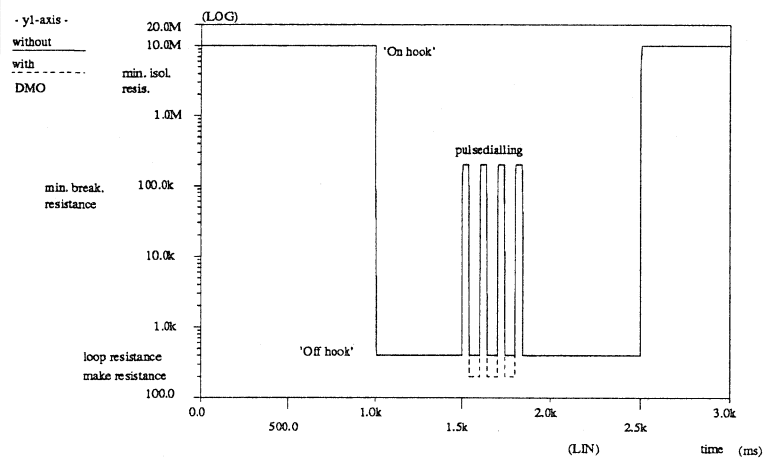


Fig.6 Loopresistance as function of operation mode (basic set)

3.1.2 Switchdriver of the UBA1702A

Principle of operation

A bipolar interrupter switch requires a considerable base drive current, consequently T1 in figure 4 is replaced by a darlington resistor combination with T1 and T2. Remaining circuitry has not been altered, see figure 7.

Due to the small voltage difference between SDI and SDO (one junction voltage) a series resistance (R1) is necessary to limit the basecurrent. Its value (2 kΩ) is a compromise:

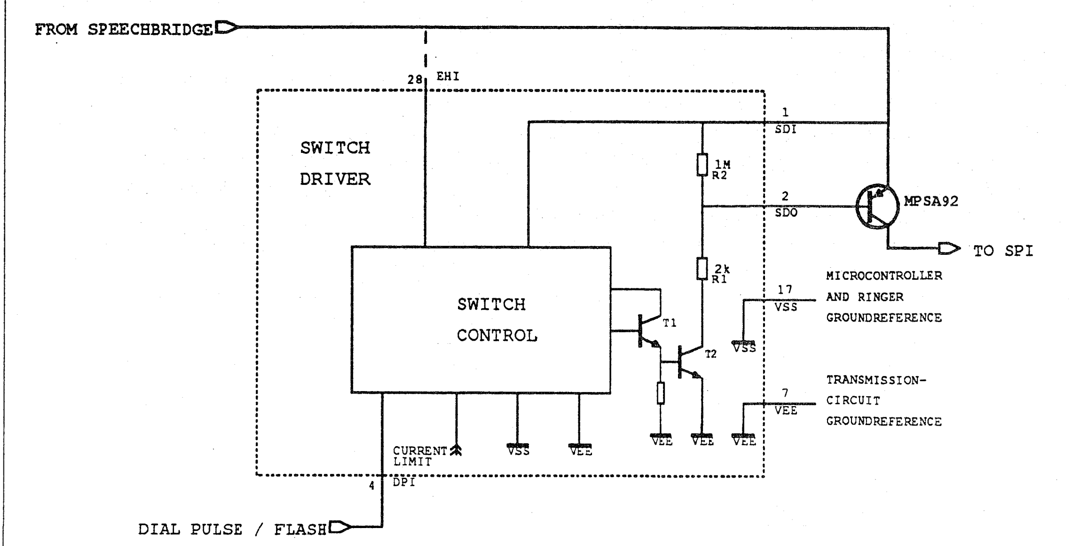


Fig.7 Switchdriver of the UBA1702A

- a low value is necessary for sufficient base current to keep the bipolar transistor saturated at high line currents. In case of non-saturation the impedance and set voltage would increase considerably.
- this resistance is connected in parallel with the set and to not deteriorate the set characteristics too much a high value is necessary.

It may be clear that the H_{FE} in saturation in the required line current range is an important parameter of the external PNP interrupter, it may be necessary to select for this parameter.

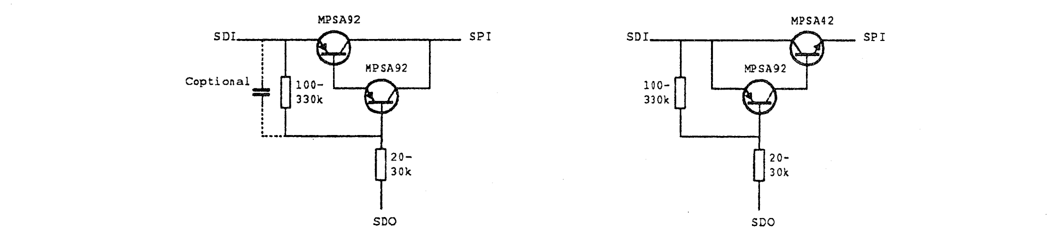


Fig.8 Various two transistor combinations to replace single PNP

An alternative is to add a transistor to make a PNP / PNP or PNP / NPN combination, see figure 8. Disadvantages of a two transistor solution are:

- increased number of required components (four instead of one)
- increased voltage drop across the interrupter switch (one junction voltage extra).

Advantages are:

- the possibility to increase the series resistance and hence the parallel set resistance by adding a resistor of a few times 10 kΩ
- the interrupter transistor remains saturated even at high line currents.

In case very fast voltage transients may occur (e.g. lightning) it is recommended to connect a capacitor Coptional in parallel with the be-resistor in the first configuration. This is not required for the second configuration.

Performance

Performance of driver / PNP combination is to a large extent dependent on the characteristics of the PNP. Because of the fact that this bipolar transistor is current driven instead of voltage driven like a PMOST, delay times (turn-on: 5 µs, turn-off: 80 µs) are shorter and switching (on: 5 µs, off: 70 µs) is faster, see figure 9. For testcircuit the basic application given in figure C2 has been used.

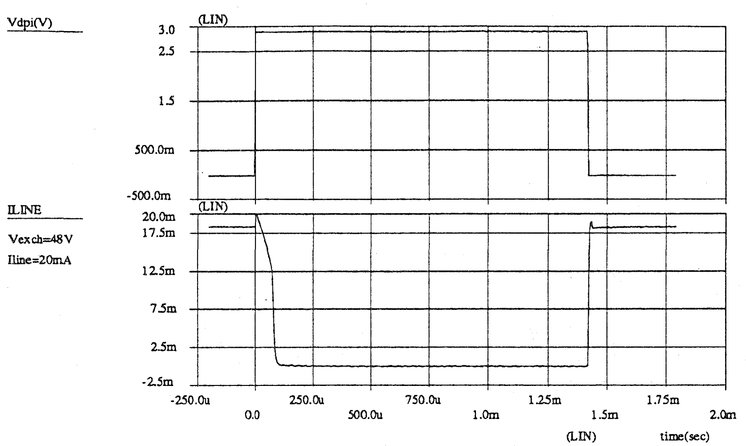


Fig.9 Typical switching times of UBA1702A in combination with MPSA92

Insulation and break resistance are equal to UBA1702.

3.1.3 Line current management circuit

Principle of operation

The blockdiagram is depicted in figure 10.

The line current is flowing through SPI, the sense resistor R1 (typical value 2 Ω) and SPO. The voltage drop across the resistor is converted to a small sense current which is used for two purposes:

- current detection (minimum level). For this purpose the sense current is reconverted to a voltage by R2 and compared to a reference voltage. A voltage across R2 greater than the reference results in the logic output CDO high (= VDD).
- current limitation (maximum level). Operation is comparable to the way line current is detected. The output of the comparator however is not used for driving a pin, but it directly controls the switch driver.

Reference voltages are generated by an internal bandgap voltage source.

Current detection is intended for use by microcontrollers to give information about hookswitch- and line status.

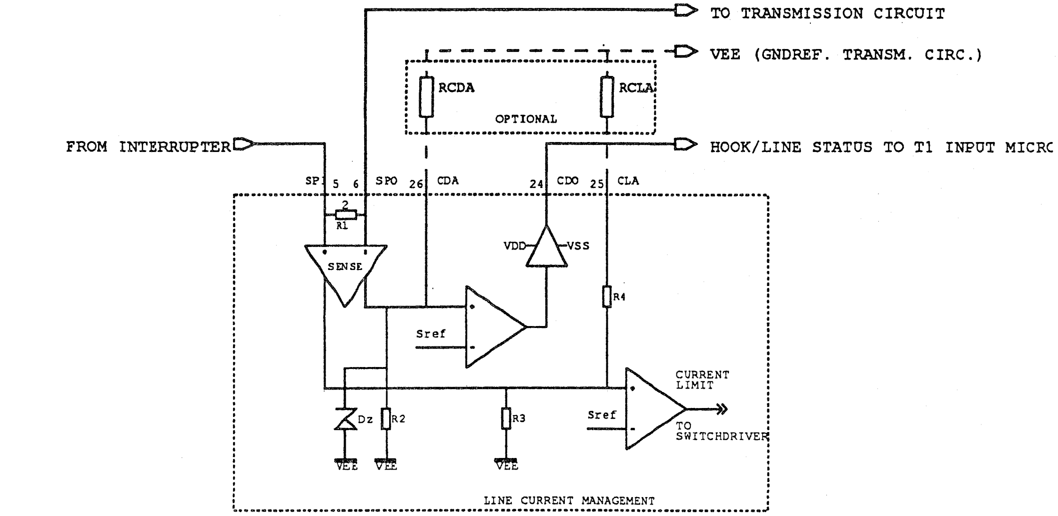


Fig.10 Linecurrent management circuit

Therefore CDO has to be connected with the 'test 1' (T1, CSI, hook) input of the microcontroller. In case of a line current interruption the microcontroller can force the transmission circuit into a powerdown condition during which the charge on the supply capacitors is maintained. A large AC line voltage swing can also cause the detection threshold to be passed. To avoid an unwanted powerdown two measures can be taken:

- apply a sufficiently large time window in software. This is usually the case.
- introduce delay by connecting a capacitor to pin CDA

Both measures have influence on the current detection response time.

Adjustments and performance

The current detection threshold and the current limit level can be adjusted by connecting resistors to respective pins CDA and CLA. The voltages at CDA and CLA are proportional to the line current. However, the voltage at CDA is limited by Dz, which represents an internal zener diode. The influence of R_{CDA} (the resistance between pin CDA and VEE) is given in figure 11.

The graph for the current limiting action is given in figure 12. R_{CLA} is the resistance between pin CLA and VEE.

Note that the maximum current is fixed at approx. 120 mA. This is achieved by resistor R4 in series with pin CLA.

Stability of the currentlimiter for low line currents (approx. 45 mA) can be improved by connecting a capacitor of 10 nF between pin SPO and VEE. However in most cases this capacitor is already present for EMC purposes.

If necessary it is possible to decrease the minimum level for current detection (3 mA) and current limiting (45 mA) even further by connecting the respective pins instead of VEE to a constant voltage by means of an appropriate resistor.

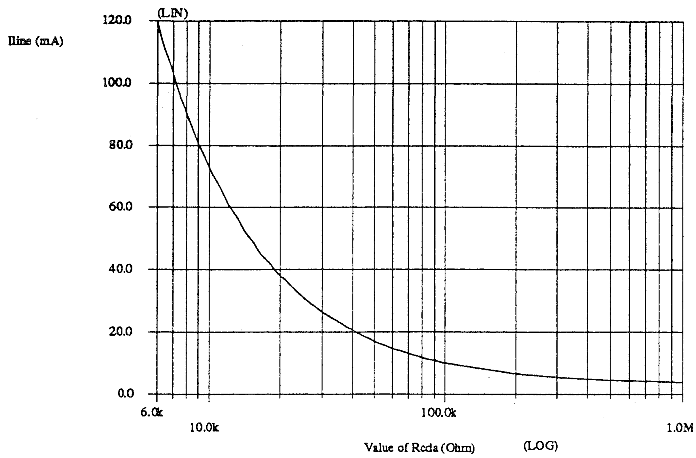


Fig.11 Current detection threshold as function of R_{CDA}

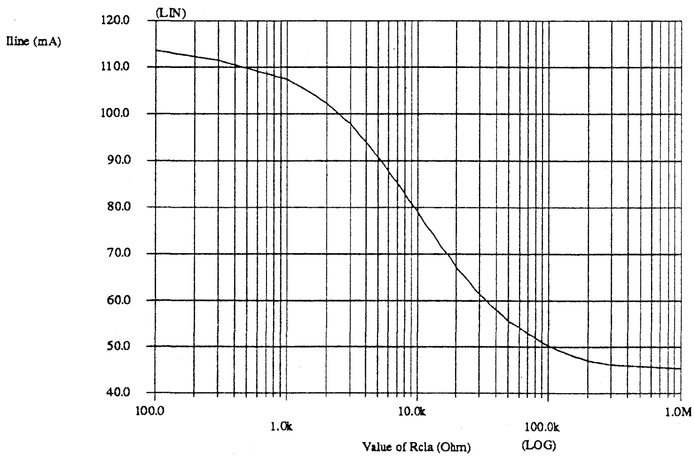


Fig.12 Current limiting level as function of R_{CLA}

3.1.4 The mute switch / zener protection circuit

Principle of operation

The blockdiagram is given in figure 13.

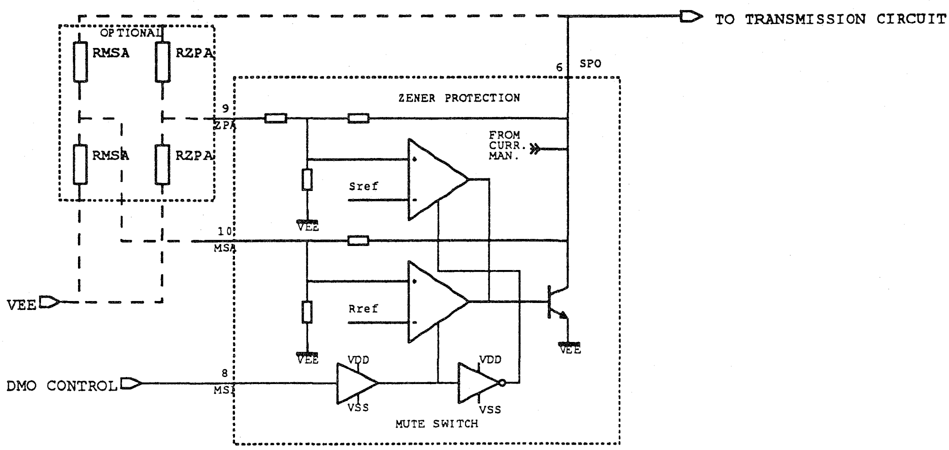


Fig.13 Mute switch and zener protection

This part of the UBA1702/A controls the maximum voltage on pin SPO. Dependent on the level on pin MSI two different modes exist, these two modes are mutual exclusive:

- DMO-mode (MSI = high). This mode guarantees a decreased loop resistance of the set during the make period of pulse dialling. Therefore a logic control signal is generated by the microcontroller, which is applied to pin MSI. A buffer has been added to convert the logic levels to appropriate values. In case of a high MSI input level the voltage on SPO is compared with a reference and dependent on the difference more or less current flows through the shunt transistor. The DMO voltage can be adjusted by connecting a resistor between pins MSA and SPO. A provision has been made for proper operation of the switch driver even if the voltage at SDI drops to a low value due to low voltage DMO operation.

Remarks:

* to achieve low DMO voltages (< 2.5 V, R_{MSA} between MSA and SPO) it is required that VDD and therefore VBB does not drop below a certain minimum value (about 2 - 2.5 V).
* the differential resistance of the shunt at low DMO voltage (1.6 V) has increased from < 0.5 Ω to 5 - 6 Ω.
* the control voltage on pin MSI may not exceed VDD.
* it is recommended not to increase the DMO voltage to values greater than 4 V.

- protection mode (MSI = low). This is the default operation mode which is intended to protect the transmission-circuit against overvoltage. Its operation principle is similar to the DMO mode, with this exception that a different voltagedivider and comparator has been used. For the zener voltage can be chosen between three values: by connecting pin ZPA to pin SPO, pin ZPA to pin VEE or by leaving pin ZPA open. Intermediate values can be realized by replacing the shorts by a resistor.

Adjustments and performance

The influence of a resistor R_{MSA} between pins MSA and SPO / VEE is depicted in figure 14 and 15 and of R_{ZPA} between pins ZPA and SPO / VEE in figure 16 and 17.

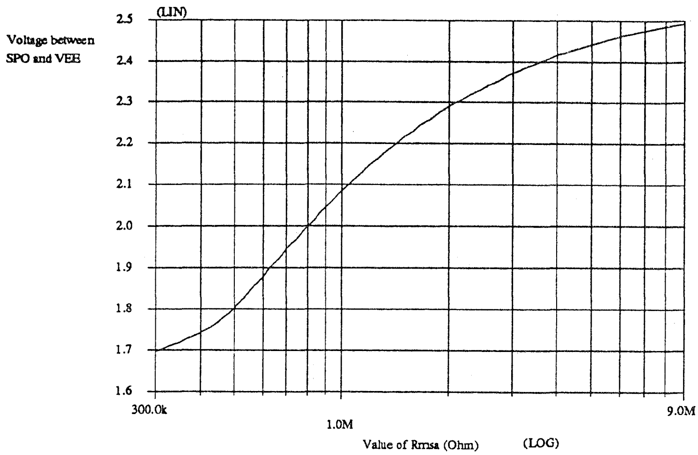


Fig.14 DMO voltage as function of R_{MSA} (connection to SPO)

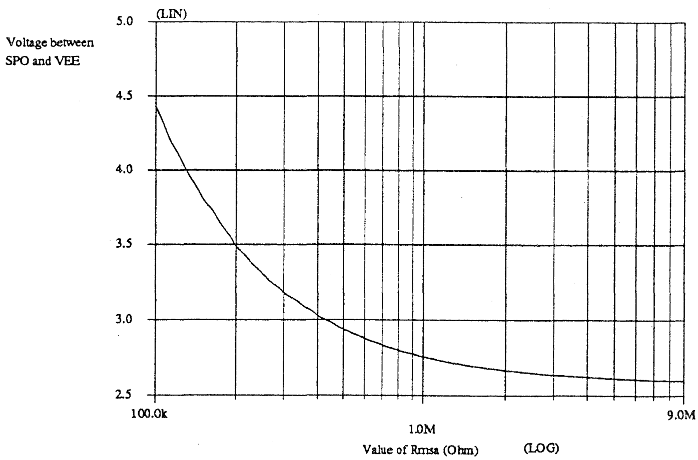


Fig.15 DMO voltage as function of R_{MSA} (connection to VEE)

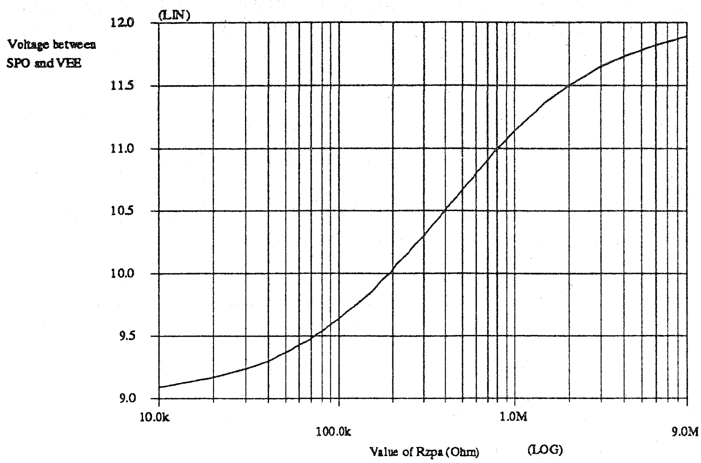


Fig.16 Zenerprotection voltage as function of R_{ZPA} (connection to SPO)

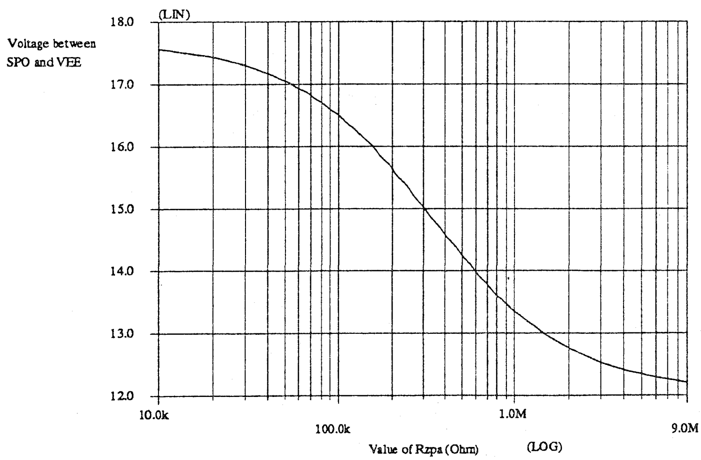


Fig.17 Zenerprotection voltage as function of R_{ZPA} (connection to VEE)

Note that the minimum and maximum zener protection voltages (respectively 9 and 18 V) have finite values. This is accomplished by the resistor in series with pin ZPA. In this way the minimum and maximum voltage settings do not require external components and have a reasonable accuracy. This accuracy can not be achieved with external resistors because they do not match with on chip resistors .

3.2 Ringer part

The ringer part consists of the protection circuit (ch. 3.2.1), the frequency detection (ch. 3.2.2), the threshold comparator (ch. 3.2.3), the volume control (ch. 3.2.4) and the melody input / piezo driver (ch. 3.2.5).

3.2.1 Protection circuit

Principle of operation

The voltage between pins RPI and VEE is limited to a fixed level (typ. 20.4 V) by a comparator / shunttransistor combination which represents an electronic zener diode, see figure 18.

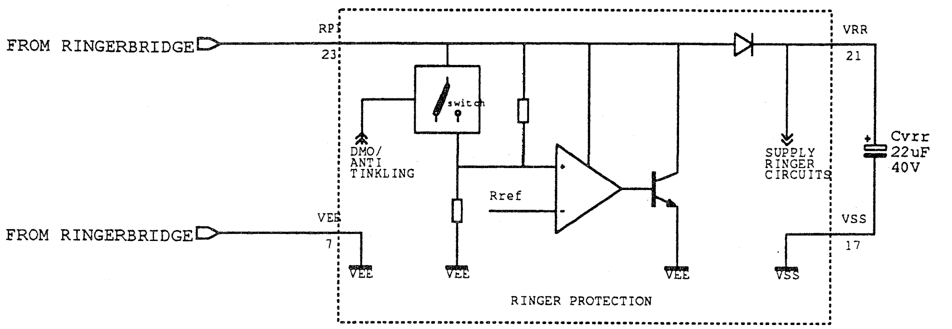


Fig.18 Ringer protection and limiting circuit

This input circuit has to be preceded by a ringer bridge, consisting of a series RC-combination and four diodes to rectify the AC-ringer signal. The capacitor blocks the DC current and determines in combination with the series resistor the input impedance for voltages higher than the limiting voltage. The RC network has also a current limiting function.

A provision has been built in to avoid ringers connected in parallel with the ringer bridge and which share the same RC-network from making noise during pulse dialling, the so called anti-tinkling function. Condition is that during dialling a high level (= VDD) is applied to pin MSI. This signal causes a switch across the voltage divider to close and in this way the input voltage across RPI and VEE is limited to a value insufficient for ringers to make any noise (typ. 2 V + voltage drop ringer bridge). In case the anti-tinkling function is wanted but the DMO function not, the latter can be made inoperative by connecting an appropriate resistor between pin MSA and VEE with such a value that the voltage in zenerprotection mode and DMO mode are equal, see figure 15. This is necessary because the speech zenerprotection is inoperative during MSI = high.
The internal diode between pins RPI and VRR has two functions:

- the voltage at RPI must return to zero to make frequency detection possible
- this diode only conducts if the input voltage is above a certain level. In this way a high series input impedance for speech signals is created, which is required for EHS application.

Performance

The voltage divider which determines the limiting voltage is not external accessible and this voltage is therefore not adjustable.
The ringer supply voltage is smoothed by buffercapacitor C_{VRR}. The recommended value for this capacitor is 22 µF. A larger value will result in a longer ringer start-up time, while a smaller value will cause increased ripple amplitude, especially at low ringersignal frequencies.

3.2.2 Frequency detector

Principle of operation

This block consists of a comparator with hysteresis and an output buffer for level shifting, see figure 19.

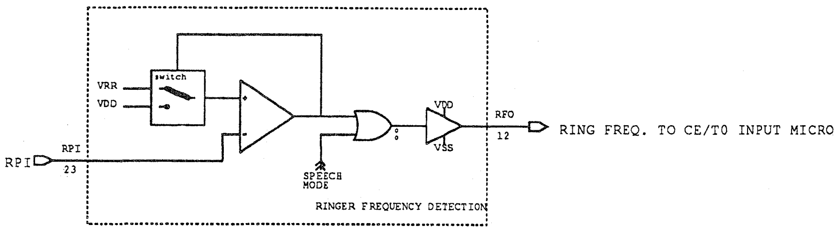


Fig.19 Frequency detector.

The non-inverting input of the comparator is connected to VDD or VRR, dependent on the most recently passed threshold, see figure 20.

The full wave rectified AC ringer signal is applied to pin RPI. This signal must cross a high threshold level before RFO is high and the output on RFO only returns to a low output level if a low threshold level is crossed. The selection of these threshold levels is rather arbitrary under restriction of the following conditions:

- the levels must lay within the amplitude range of V_{RPI}
- the voltage difference between the levels have to be as large as possible.

For the high level VRR and for the low level VDD have been chosen. These voltages satisfy the conditions and are already available.
Advantage of the applied hysteresis is the insensitivity to superimposed signals, this insensitivity is required for some countries.
RFO is a logical output (levels VSS and VDD) intended for processing by a microcontroller. Therefore this pin must be connected to the 'chip enable / not test zero' (CE/notT0, FDI, RF) input of the microcontroller. Note: because the AC ringer signal is full wave rectified the frequency of the RFO signal is twice the frequency of the original ringer signal!
In speech mode the CE/notT0 input has to be high. Hence the speechmode infosignal is combined with the ringer signal by an OR-gate.

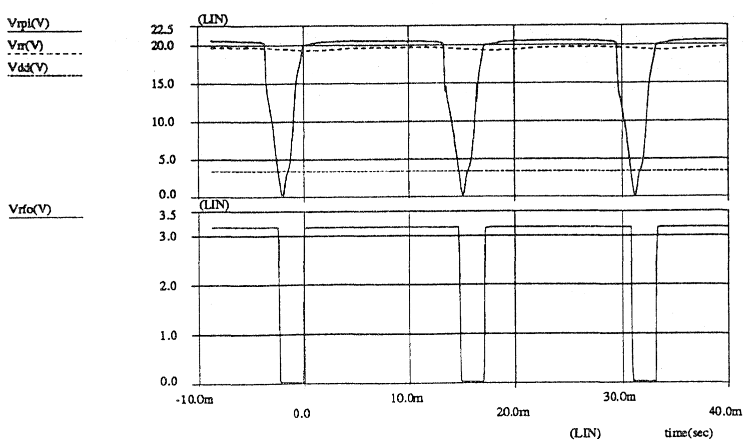


Fig.20 Switching thresholds frequency detector

Performance

In the previous paragraph it is pointed out that the threshold levels (VRR and VDD) for the ring frequency detector are fixed.

The insensitivity to superimposed signals of the circuit is made visible in figure 21. A 6 V_{RMS}, 1 kHz sinewave is

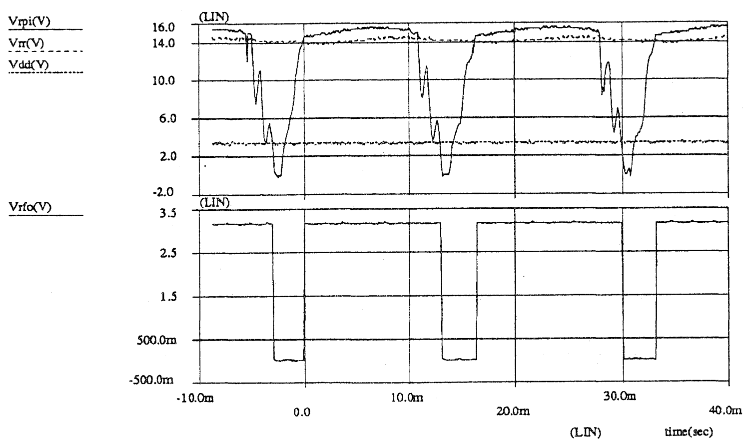


Fig.21 Response of frequency detection circuit to superimposed signals

superimposed to a 45 V_{RMS}, 20 Hz ringer signal. Despite various single threshold crossings within one period of the inputsignal the interference of the superimposed signal can not found back in the outputsignal available at pin RFO.

Care must be taken in case EMC capacitors are connected between the lineterminals and ground (VEE). During the zero crossings of the AC ringer signal these capacitors are not discharged completely which can result in not crossing the lower threshold and hence not generating a 2F ringer frequency signal. This can be solved by connecting a high ohmic resistor (in the order of hundred kilo Ohm) between RPI and VEE.

3.2.3 Threshold comparator

Principle of operation

This block checks the amplitude of the ringer signal. If this signal has the required level this block enables the input of the piezo output stage, see figure 22.

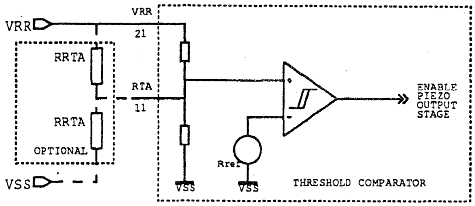


Fig.22 Threshold comparator

After applying a ringer signal to the protection circuit (ch. 3.2.1) capacitor C_{VRR} is charged. The voltage across the capacitor is divided by a resistor network and compared with a reference voltage. If VRR crosses an adjustable threshold the output of the comparator is switched to a high level.
A hysteresis has been built in to avoid faltering. In case the output stage is enabled and a melody is applied the current consumption of the circuit will increase. Consequently VRR will drop and without hysteresis the output stage would be disabled. The current consumption would decrease, VRR would rise and the cycle would repeat itself.

Adjustment and performance

The threshold level can be adjusted by connecting a resistor R_{RTA} between pins RTA and VSS (to increase the level) and between RTA and VRR (to decrease the level). In figure 23 and 24 the upper threshold level and hysteresis as function of R_{RTA} is given.
The hysteresis is not independently adjustable but correlated with the threshold level: it is about 60 % of the upper threshold level.

If pin RTA is not connected (default threshold values) the UBA1702/A is suitable for the German FTZ specification as described in 1TR2 2.4.3.1/2 (June 1990).

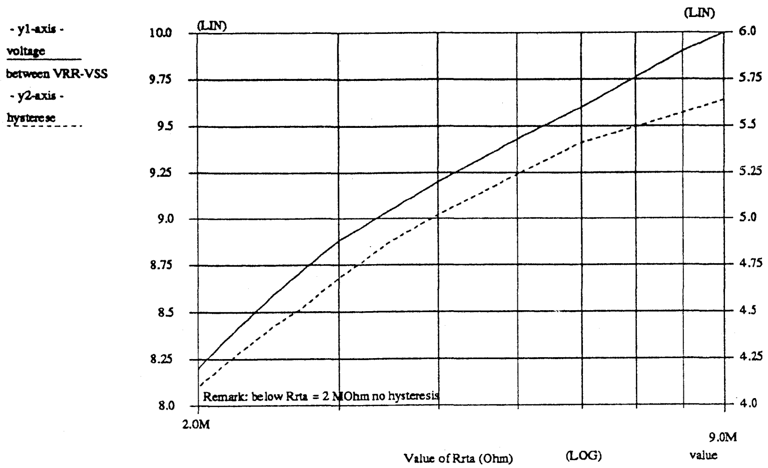


Fig.23 Upper threshold level and hysteresis as function of R_{RTA} (connection to VRR)

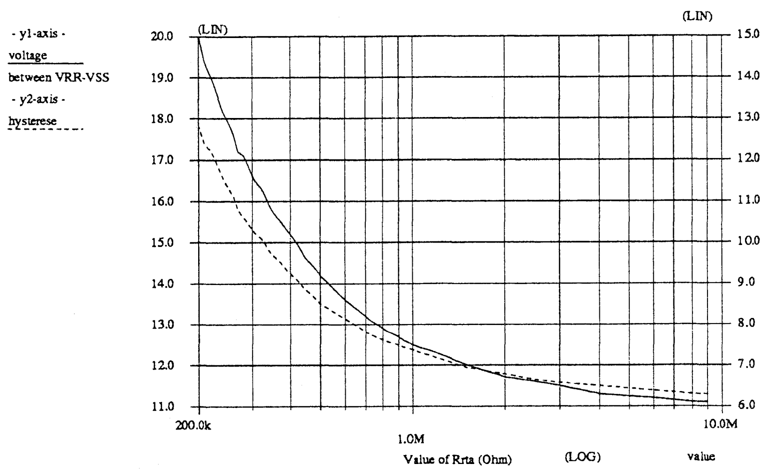


Fig.24 Upper threshold level and hysteresis as function of R_{RTA} (connection to VSS)

3.2.4 Volume control

Principle of operation

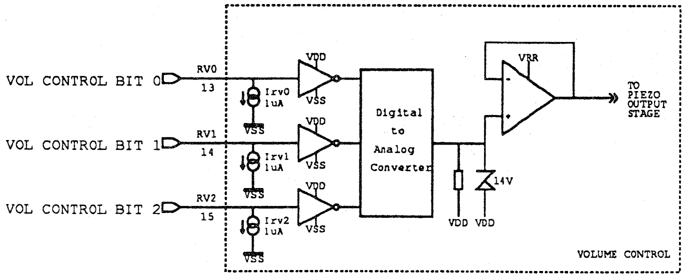


Fig.25 Volume control

Volume control is implemented rather straightforward with a D/A-converter, three inputbuffers and an output-buffer, see figure 25.
The three logic inputs RV0 (least significant bit), RV1 and RV2 (most significant bit) are provided with pull-down current sources, which sink about 1 μA. In case of floating (high Z) microcontrolleroutputs connected to RV0-2 (for instance during start-up) these sources assure a minimum volume setting. The output of the three bits D/A-converter is a controlcurrent which is converted to a voltage with reference VDD by resistor R and limited by a zenerdiode Dz. This voltage is buffered and determines the positive peaklevel of the piezo output signal. The negative peaklevel is fixed and close to VDD. Note: the voltage on pins RV0, RV1 and RV2 may not exceed VDD.

Performance

Output voltage across ROA and ROB for minimum volume setting (bits RV2-0 all low) is 150 mV_{pp}. This level is independent of ringersupply conditions. For the following six steps (RV2-0 (0,0,1) to (1,1,0)) each step doubles the output voltage (6 dB steps). The last but one **level** (RV2-0 (1,1,0)) is $2^6 = 64$ times the first level and therefore 64 * 0.15 = 9.6 V_{pp}, on condition that the ringersupply voltage suffices. The last step is the largest and consequently the positive peaklevel of the piezo output signal is almost equal to the supply voltage VRR. However to protect the piezo element the output amplitude is internally limited by an internal zenerdiode to about 2 * 14 = 28 V_{pp} (unloaded).
The level of the outputsignal (available between pins ROA and ROB) as function of the volume setting is given in figure 26.

3.2.5 Melody input and piezo driver

Principle of operation

This (sub)block amplifies the ringermelody generated by the microcontroller to the desired level for the piezo. It contains an input comparator, an AND-gate, an inverter and two outputbuffers for BTL operation, see figure 27.

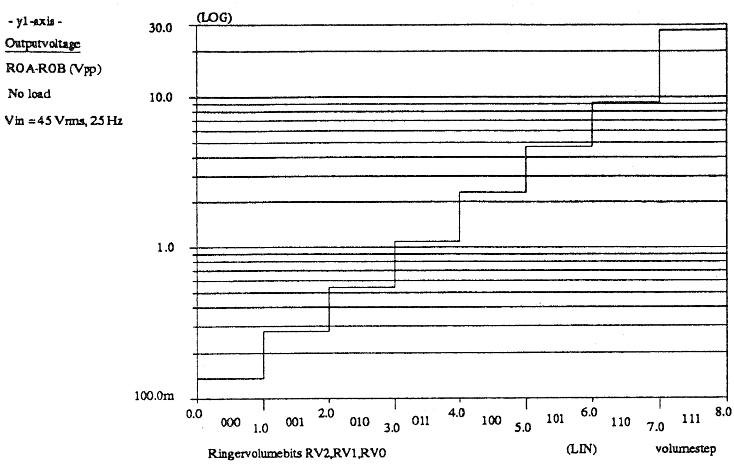


Fig.26 Output level as function of volumesetting

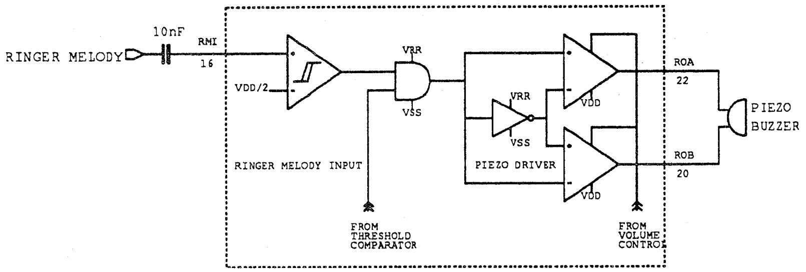


Fig.27 Melody input and piezo driver

The input comparator has a threshold at ½ VDD. It is possible to apply square wave signals (min. / max. level VSS and VDD) as well as sinewave signals superimposed on a DC level of ½ VDD. However if the DTMF output is used (sinewave signals) it is advised to use a series capacitor of 10 nF to avoid problems caused by DC shift as a result of e.g. resistor mismatch. A small hysteresis of about 20 mV has been built in for schmitt-trigger action to assure jitterfree operation. This hysteresis has no influence on the duty cycle of 50 %. Because of the high inputimpedance of the RMI input (in the order of hundred kilo Ohm) the attenuation for DTMF signals is low: less than 0.05 dB.

The square wave output of the comparator is combined in an AND-gate with the output of ring threshold comparator (ch. 3.2.3). If VRR has reached the required level the ring melody is routed to the output buffers. These buffers switch their outputs (ROA / ROB) between a DC level coming from the volume control and VDD. The resulting signal is a squarewave which is applied to the piezo. The slew rate of the outputs is rather high and the outputsignal contains therefore a lot of higher harmonics which contribute considerably to the produced sound-pressure. Because the piezo shows equivalence with a capacitor there flows no DC current through it and therefore the output stage may operate in Single Ended Load (SEL) configuration with the piezo connected between ROA or ROB and VDD.

Performance

It is already mentioned that the piezo shows a capacitive behaviour, the outputcurrent reaches its peak value during switching.
Current capability is an important specification parameter for sound pressure: if limiting at a low current level occurs there will be less harmonics and the produced sound will be dull. The UBA1702/A can deliver more than 100 mA and piezo transducers with a capacitance up to at least 80 nF can be used.
In stead of a piezo transducer a high ohmic (> 25 Ω) loudspeaker can be used to make the ringer melody audible. A capacitor in the range 100 - 470 nF must be connected in series.

3.3 Supply part

Principle of operation

The reference block generates all the bias currents and reference voltages for the IC. All reference voltages are derived from bandgaps and converted to biascurrents by internal resistors. The supply can be split up in a series and a shunt regulator, see figure 28.

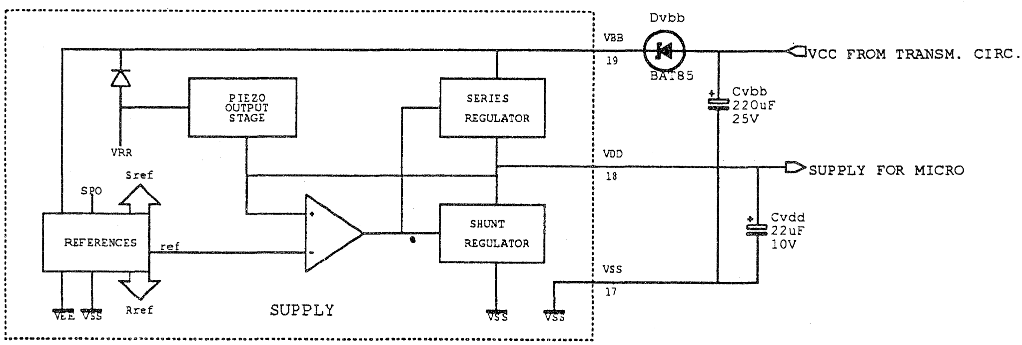


Fig.28 Supply

Operation is different for speech and ringer mode:

- speech mode. The supply is powered via pin VBB by VCC of the transmission circuit. The internal diode between VRR and VBB avoids current flowing to ringcapacitor C_{VRR} and other ringer parts. There is no current flowing through the ringer output stage, consequently all current for VDD flows through the series regulator. VDD is stabilized by comparing it with a reference voltage and by feeding the errorsignal back to the regulator.
- ringer mode. In ringer mode the supply point is VRR. To avoid current flowing through VBB to the transmission circuit an external seriesdiode has to be added. To minimize the voltage drop (important in case of low line currents) it is recommended to use a schottky type (e.g. BAT85).

Current to VDD can be delivered by the series regulator as well as the piezo outputstage which is connected in parallel. It is likely that at high output volume the output stage consumes (sources) more current than the micro-controller sinks from VDD. As a consequence VDD would increase.
To keep VDD constant a shunt regulator has been added to sink the superfluous current. This series concept (outputstage and microcontroller supply in series) has been chosen because it has advantages in case of low ringer power conditions: the ring current is used more efficiently. VDD is also used to supply some internal blocks.

Performance

The output of the supply is a fixed 3.3 V stabilized voltage. Advantage of a rather low supply voltage is a decrease of current consumption of the microcontroller. This can be attractive during line current interruptions. The output voltage VDD is smoothed by C_{VDD}. Besides a smoothing function this capacitor is also essential for stability of the supply. The recommended value is 22 µF. A larger value will result in increased start-up times in speech and ringer mode. For supplying the microcontroller during long line interruptions it is advised to choose a larger value for C_{VBB}.

In quiescent condition (supply point voltages VBB and VRR low) the internal current consumption from VDD is minimized. With this provision memory retention of the microcontroller is possible with a resistor (> 5 MΩ) connected in front of the hook switch to the line (trickle current), see figure C2.
Because of the rather low current consumption of microcontrollers the source capability of the supply is accordingly low: max. 2 mA. The currentsink capability (necessary in ringer mode at high volume settings) is much higher. The voltage drop between VBB and VDD at low values for VBB is minimized (100 mV @ IDD = 1 mA) to make operation with low line currents possible. The output voltage VDD as function of IDD for two values of VBB is given in figure 29. IDD is positive in case pin VDD sources current and negative if current is sinked.

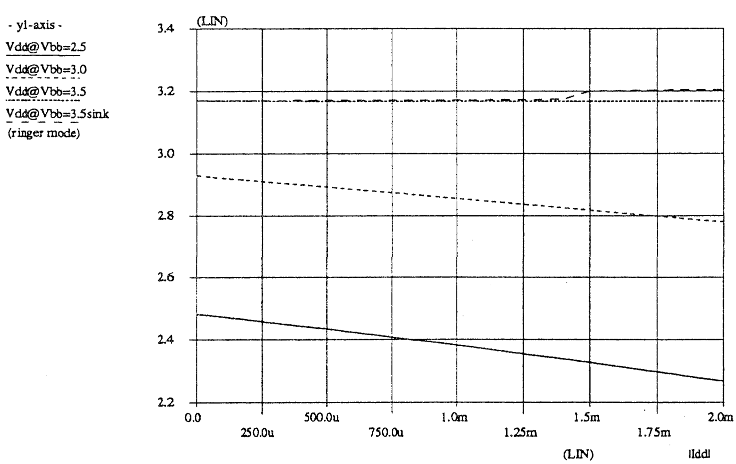


Fig.29 VDD as function of +/- IDD

4. DESCRIPTION OF THE COMPLETE SYSTEM FEATURING THE UBA1702/A

Apart from the UBA1702/A a transmission circuit and a microcontroller (dialler / ringer) are required to form a (basic) telephone set. Typical set parameters are not solely dependent on the individual circuit characteristics but also on their interaction. This interaction will be described in this chapter with the help of a typical application diagram which is given in the appendix, see figure C1.
Two different modes can be distinguished: the 'off hook' situation (ch. 4.1) and the 'on hook' situation (ch. 4.2).

4.1 'Off hook' situation

In this mode a conversation mode (ch. 4.1.1) and pulse dial / flash operation (ch. 4.1.2) can be distinguished.

4.1.1 Conversation mode

Principle of operation (See also appendix figure C1)

Start-up is accomplished by putting the hookswitch S1 in an upwards position. The ringer series RC-network is shortcircuited. The current through the switch and C_{rng} is limited by R_{rng}. A DC line current starts flowing and after the voltage on EHI has crossed the threshold value interrupter T1 starts conducting under condition that DPI is low. As soon as the line current is larger than 3 mA (default) the microcontroller receives a chip enable (CE). Because of the electronic coil function of the transmission circuit it takes some time before the line current has reached its endvalue. In the meantime the voltage across the transmission circuit (and the set) is limited by the electronic zener of the UBA1702/A through which most of the line current is flowing. C1 is charged by R1, the transmission circuit is supplied via VCC, starts operating and takes over the current from the zener.
Dependent on the used transmission circuit there are two ways to feed the supply point VBB of the UBA1702/A:

- ground reference is VEE. This is the case for all transmissioncircuits except the TEA1064A. VSS and VEE of the UBA1702/A are interconnected and VBB to the supplypoint of the transmission circuit VCC. A series diode is necessary to separate the ringersupply from the speech part. Because of the rather large series resistance the supply current capability is small, but sufficient for the UBA1702/A in combination with a microcontroller.
- ground reference is SLPE. This structure is possible if the transmissioncircuit has a powerdown / mute input which is referenced to SLPE: TEA1064A or TEA1064B. VSS of the UBA1702/A is connected to SLPE and VEE with VEE. In this case the supply point is not VCC but directly the line with a (small) series resistor. Due to the reference of this supply point (SLPE), this resistor has no influence on the set impedance, see [Ref. 5]. This results in a better supply capability. To bridge large supply interruptions the value of C_{VBB} can be increased. This has almost no influence on the start-up time of the transmission circuit.

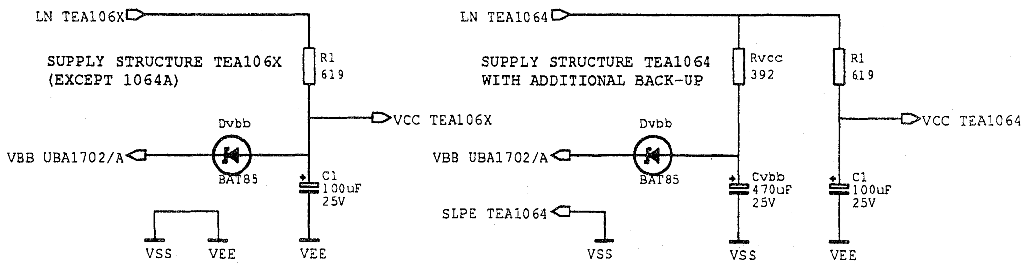


Fig.30 Supply structures for combination with TEA106X

After supplying the UBA1702/A via VBB, C_{VDD} is charged, VDD rises and when the reset reference voltage of the microcontroller has been reached start-up is accomplished under condition the CE/notT0 input is high.

This CE/notT0 (FDI,RF) input of the microcontroller has two functions:

- as chip enable input it is used to initialize the system and in combination with the test1 input of the micro-controller it determines the operation mode (speech, ringer or stand by).
- as ringer frequency discriminator input the time between two low-to-high transients (reciprocal to the ringer-frequency) of RFO is measured, see table.

TABLE 1 Operation modes of microcontroller

	CE/notT0 = 0	CE/notT0 = sq. wave	CE/notT0 = 1
T1 = 0	standby	ringer	standby *)
T1 = 1	X	X	speech

*) if T1 = 0 longer than reset delay time

Performance

In conversation mode the UBA1702/A and its peripheral components has a slight influence on the set characteristics. This influence is caused by the (small) current consumption and (large) parallel impedance of the UBA1702/A. However, in case the UBA1702A with a single bipolar interrupter is applied the parallel resistance is considerably low, see also ch. 3.1.2. The influenced set characteristics are:

Balance Return Loss. The BRL is a figure that gives the accuracy a set impedance approximates the nominal impedance required by the local PTT. Because of the high parallel impedance (> 20 kΩ) of the UBA1702 the deterioration of the BRL is almost negligible with respect to the situation without UBA1702. For the UBA1702A the effect is considerable because of the low baseresistor value. To a certain extent the effect can be compensated by increasing the value of R1, although the supply capability (voltage drop) is deteriorated. The same holds for the transmit / receive gain.

Sidetone Suppression. A certain amount of attenuation (suppression) of transmitted signal to the earpiece is required. The parallel impedance of the UBA1702 has only a minor effect on the anti sidetone network. In case of the UBA1702A the sidetone suppression will be worse.

Maximum Sending Level. Sending level can be limited by voltage or by current. Maximum voltage swing is determined by the voltage difference between LN and SLPE of the transmission circuit. The UBA1702A has no influence on this parameter, in case of the UBA1702 the minimum peak may be limited by the threshold voltage V_{th} of the applied PMOST (for BSP254 V_{th} can vary between 0.8 and 2.8 V). A capacitor between source and gate can give improvement (bootstrapping), though switching times are getting worse.
Limitation by current occurs if there is not enough current available to be modulated by the outputstage of the transmissioncircuit. Because of the low current consumption the influence is negligible.

Automatic Gain Control. A provision can be present to compensate transmit / receive gain for long line lengths. This gain control is determined by the line current flowing through the transmission circuit. Current consumption of the UBA1702/A (UBA1702: typ. 350 μA, UBA1702A: typ. 500 μA PNP interrupter base current excluded) reduces this current, which results in a very small gain increase in the relevant range. If really necessary this gain increase can be compensated by decreasing the value of R6.

Low Voltage / Parallel Set Operation. In case of bad supply conditions e.g. operation with long lines or with a classical set (impedance about 200 Ω) connected in parallel the available supplyvoltage may be insufficient for proper operation of transmission circuit and microcontroller. The influence of the UBA1702/A is small, see figure 31.

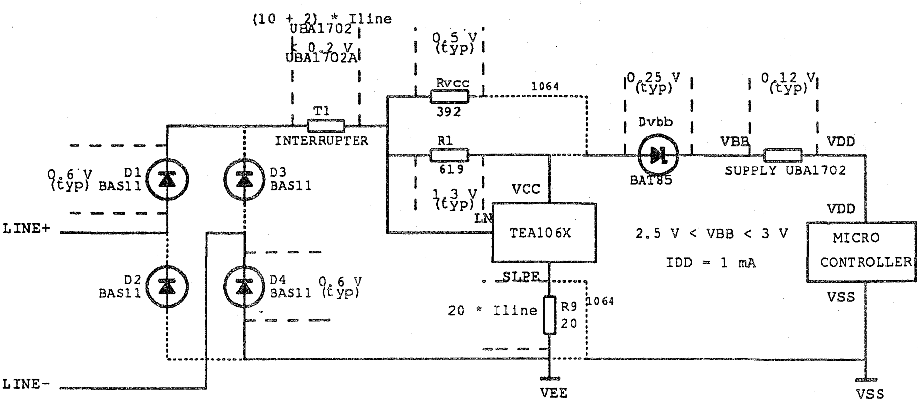


Fig.31 Supply voltage drop

4.1.2 Pulse dial / flash operation

Principle of operation

The dial- and flash pulses are generated by the microcontroller under control of the keyboard. Interrupting the line current has consequences for the supply and several measures may be taken to guarantee a non-interrupted supply of the microcontroller. The kind of measure depends on the type of transmission circuit used.

Circuits without power down function (e.g. TEA1062/A). During the breakperiod of pulse dialling and eventually flash operation there is still flowing current because of the electronic coil function. C3 (C_{reg}) and C1 are discharged and VCC drops to unacceptable level to supply VBB and consequently the microcontroller. Increasing the value of C1 is not recommendable because the discharge current requires unrealistic capacitance values. Workable alternatives are the following two solutions:
Solution 1 requires an additional series low drop diode (Schottky e.g. BAT85) and capacitor connected between the diodes and VEE. Disadvantage is the additional voltage drop caused by the diode.
Solution 2 comprises only an increase of the value of C_{VDD}. Disadvantage is an increase in microcontroller start-up time.

Circuits with power down and VEE as logic reference (most of the circuits of the TEA106X family). Because of the power down function the current consumption is reduced almost a factor 20 and increasing the value of C1 could be an alternative to increase of C_{VDD}.
Note that the power down function can be used for external generated interrupts (e.g. due to switching of a connected PABX) as well. The CDO output of the UBA1702/A controls the power down mode of transmission circuit and microcontroller. Restriction is that the power down control and the pulse dial / flash output can not be combined.

Circuits with power down and SLPE as logic reference (e.g. TEA1064A and TEA1064B). C_{VBB} (see fig. 30) can be chosen in accordance with the required interruption time.

Performance

The pulse dialling waveform is mainly determined by the transmission circuit if DMO is not used. In case C3 (C_{reg}) is discharged during the break period of pulse dialling due to a missing power down mode or because of a resistor R17 between REG and SLPE, it takes some time before C3 is charged again and the line current / voltage across the set has reached its steady state. Considerable pulse distortion occurs. In stead of using R17 to increase the voltage between LN and SLPE silicon diodes in series with pin LN may be used, however in practice the exchange feeding bridge may be highly inductive and in this case the inductors determine the current waveform.
If DMO is used the set voltage during the pulsing period is determined by the UBA1702/A and not by the transmission circuit. Pulse distortion occurs not until the transmission circuit has taken over the line current after the DMO input (MSI) has become low, hence in the post-pulsing period. Dialling waveforms are depicted in figure 32.

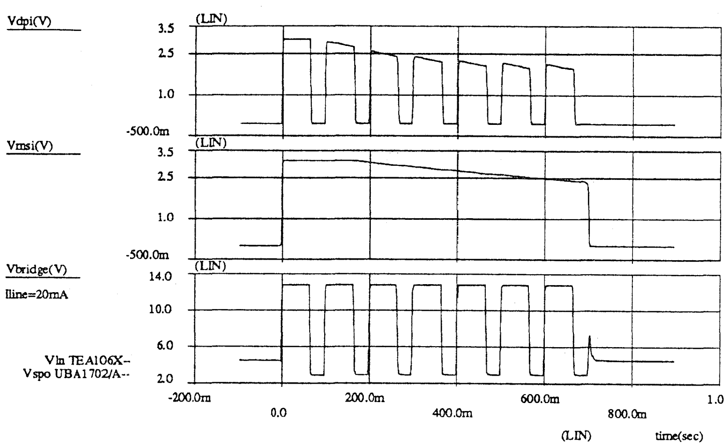


Fig.32 Pulse dialling waveforms with DMO.

4.2 'On hook' situation

In this mode the DC set resistance must be very high. After applying an AC voltage the set can enter the ringer mode (ch. 4.2.2), otherwise the memory retention / stand-by mode (ch. 4.2.1) can be maintained.

4.2.1 Memory retention / stand-by mode

Principle of operation

The PTT in some countries allows the subscriber to draw a very small DC current (in the order of 10 µA) from the line for e.g. memory retention. For this purpose the microcontroller has a stand-by mode which is entered by making the CE/notT0 input zero for a longer time interval. In this mode the VDD supply voltage is allowed to drop to a low value (1 V) and the current is likewise decreased to a very low value (about 1 µA for the PCD3349).

The UBA1702/A has a stand-by mode too, entered under condition that the voltages on supply points VBB and VRR are low. In this mode the current sink from the VDD supply point is minimized by disconnecting some internal circuit parts.

In the 'on hook' situation DC line current is blocked by the ringer capacitor C_{rng} and the interrupter. These have to be bridged by high ohmic resistors (for most countries a total minimum resistance of 5 MΩ is required) for direct supply of VDD from the line.

In case of an electronic hookswitch (the interrupter may incorporate the hookswitch function as well) it also possible to achieve set start up from this mode. For this purpose a key out of the keyboard matrix can be used. By depressing this key and at the same time making the CE/notT0 input high the microcontroller can enter the conversation / dialling mode and force the interrupter via EHI in the conducting state. An application proposal is given in which this function has been worked out further.

Performance

In stand-by mode the UBA1702/A has a resistance between VDD and VSS of more than 300 kΩ at 1 V. If the trickle current is 9 µA, more than 5 µA remains for the microcontroller.
It may be clear that after connecting a set with a discharged VDD capacitor it will take some time before this stand-by mode is operational. This is caused by the large RC time constant.

4.2.2 Ringer mode

The functions required for ringer operation have been partitioned between UBA1702/A and microcontroller:

TABLE 2 Function partitioning between UBA1702/A en microcontroller in ringer mode

UBA1702/A	microcontroller
Line termination ringer signal	ringer signal frequency check
Overvoltage protection	ringer melody generation
Supply	volume control
Superimposed signal discrimination	
Ringer signal amplitude check	
Volume controlled piezo output stage	

Note: ringer signal: low frequency AC signal generated by the exchange
ringer melody: audio frequency melody (warble) generated by microcontroller to be made audible by piezotransducer.

Principle of operation

A concise description of the operation principle of the software controlled ringer concept has already been given in the introduction: the AC ringer signal is rectified and if necessary limited and used for supply. A square wave with twice the frequency of the ringer signal is generated and available at pin RFO. Frequency check is done by the microcontroller which measures the time interval between low to high transients. In most cases the required frequency interval can be set by internal bits for versions with EEPROM or by diode option with versions without EEEPROM. If the measured time interval is according the setting a melody is generated. After the voltage at VRR (below the limiting voltage proportional to the ringer voltage) has crossed a minimum threshold the melody is routed to the piezo output stage. This is the amplitude check. Advantage of applying the result of the amplitude check at this point instead of gating the ringer frequency output is gain in start up time: charging of the VRR capacitor is done in parallel with the frequency check. This is particularly advantageous when the ringer signal has a low amplitude and frequency. The ringer melody is amplified to a level according the volume setting. If only two volume control bits are available it is recommended to interconnect the two least significant bits (RV0 and RV1) of the UBA1702/A. The steps are larger (12 dB instead of 6), but with exception of the last one the levels are still equidistant.

Performance

Ring Input Impedance. The input impedance is dependent on the ringer signal amplitude and the value of the series RC-network (R_{rng} and C_{rng}). Because of the limiting function it is also not linear. Above a certain amplitude the impedance is only influenced by the RC network. Decreasing the resistance and / or increasing the capacitance to increase output power has then little use because most of the additional available power is dissipated in the limiting / protection circuit. Minimum ringimpedance is in most cases fixed by local PTT requirements.
The V / I curve for a typical application (figure C1) is given in figure 33.

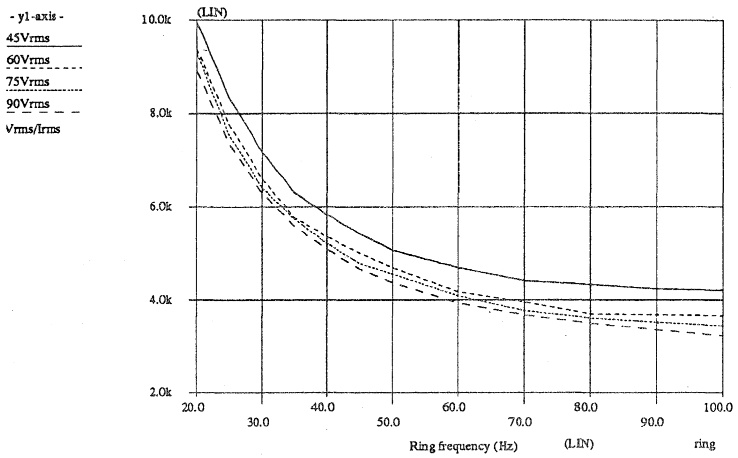


Fig.33 Ring input impedance as function of ringervoltage

Impedance is defined as the RMS value of the ringer voltage across the line terminals divided by the RMS value of the ringer current. The impedance for speech signals (up to 1 V_{rms}) is larger than 200 kΩ.

Start up time. Ringer start up time can be defined as time between applying a ringer signal the moment the melody is audible. This time has been minimized by paralleling supply capacitor charge and frequency check and is determined by various parameters like capacitance values for C_{VRR} and C_{VDD}, frequency and amplitude of ringer signal, reset level of microcontroller and number of required low to high transients for the frequency check. In a typical application with a ringer signal of 45 V_{rms} and 25 Hz this time is about 80 ms. See figure 34.
The frequency check can only be carried out after the microcontroller has been reset. In case this reset is generated externally by means of an RC-network the time constant of this network adds up to the total start time. Between ringer bursts VDD can drop below the reset level because of the VDD current consumption of the UBA1702/A. So every burst a start-up delay will occur. However, an external reset is not necessary under condition that the XTAL minimum operating voltagelevel (for f_{XTAL} = 3.579545 MHz this voltagelevel is 1.8 V) is available on VDD before the clock generated by this XTAL is applied. This condition is generally fulfilled with quartz resonators since oscillator start-up takes several milliseconds.

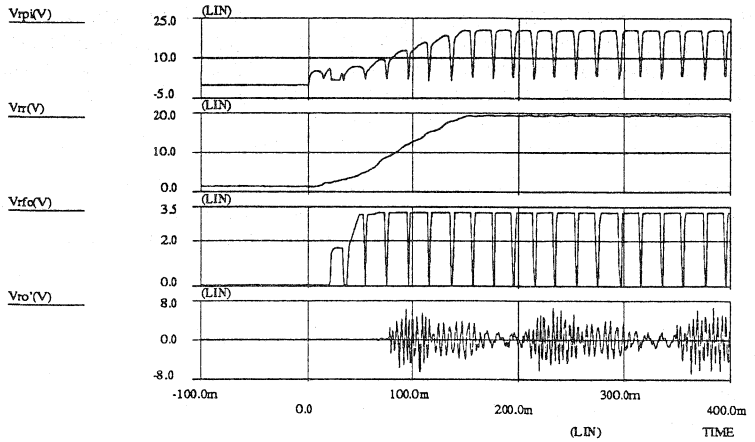


Fig.34 Ringer start up

Output sound pressure. The output sound pressure is dependent on the available output voltage, the applied piezo transducer and the acoustical properties of the cabinet. The output voltage of the UBA1702/A is limited to 28 V_{pp}, a value most piezo's can handle without degradation. The frequency characteristic of the piezo is very irregular: output levels within the specified frequency range can vary 20 dB.

5. APPLICATION GUIDELINE

In this chapter the procedure for making a basic application with a transmission circuit of the TEA106X-family and the UBA1702/A will be given. By means of figure C1 (basic application, see appendix) the design flow is given as a number of consecutive steps which should be taken. As far as possible for every step also the components involved and their influence on every step are given, the preferred value is given between brackets: [...]. For the UBA1702/A also a reference to the relevant graph in chapter 3 or 4 is made. Two adjustment resistors, R_{ZPA} and R_{RTA}, can be connected in different ways, only one is given in the application diagram.
For more information on the settings of the TEA106X see datahandbook (IC03), appendix A and [Ref. 4-6].

The basic application given in figure C1 comprises a TEA106X transmission circuit, a PCD3349A microcontroller and the UBA1702/A. As can be seen only few components have a fixed value. These components are R5, setting the reference current for the transmission circuit and various EMC components, which are indicated by a '★' in the diagram.

STEP	ADJUSTMENT
1 OFF HOOK CONDITION	
1.1 Conversation mode	
1.1.1 DC settings. First adjust the DC settings of the UBA1702/A and TEA106X to the local PTT requirements and maximum ratings of the transmission circuit.	
a) Voltage limit between LN and VEE	R_{ZPA} [open], see fig. 16/17
b) Line current limit	R_{CLA} [short to VEE], see fig. 12
c) Line current detection sensitivity	R_{CDA} [open], see fig. 11
d) Voltage between LN and SLPE	R17 or silicon diodes in series with pin LN
e) DC slope	R9 [20Ω], combination with $R_{onT1}+R_{SPI\text{-}SPO}$ (= 12 + 2 Ω)
f) Supply point VCC	C1 [100 μF]
g) Artificial inductor	C3 [4.7 μF]
h) Reset time microcontroller	C_{rst} [open], R_{rst} [short]
i) Supply point VDD	C_{VDD} [22 μF]
1.1.2 Impedance and sidetone. After setting the setimpedance, the sidetone has to be optimized for mean linelength and linetype. Also AGC can be chosen.	
a) Set impedance	Z1 (R_{onT1} and $R_{SPI\text{-}SPO}$ are in series)
b) Sidetone	R2, R3, R8, R11, R12 and C12
c) AGC	R6
1.1.3 TEA106X microphone and earpiece amplifiers, see also appendix A. After adjusting the microphone sensitivity, the gain and frequency curve can be set for the desired value. The same holds for the ear piece.	
a) Sensitivity microphone	R20, R21
b) Microphone gain	R7 (TEA106X dependent), R_{mic}, R_{mic-}, R_{mic+}
c) Low pass	C6 (combination with R7 + 3.5 kΩ)

d) Stability	C20 (= 10 * C6) (TEA1062 only)
e) High pass	R_{mic}, C8, C9 (combination with inputimpedance)
f) Earpiece gain	R4 (TEA106X dependent)
g) Low pass	C4 (combination with R4)
h) Stability	C7 (= 10 * C4)
i) High pass	C11 (combination with IR inputimpedance), C2 (combination with earpiece impedance)
1.2 Dialling mode	
1.2.1 DTMF dialling	
a) DTMF attenuation	R_{dtmf1}, R_{dtmf2}
b) High pass	C_{dtmf1}, C_{dtmf2} (combination with R_{dtmf1}, R_{dtmf2} and impedance DTMF input)
1.2.2 Pulse dialling	
a) Make resistance	DMO not used: R9 DMO used: R_{MSA} [open], see fig. 14/15

STEP	ADJUSTMENT
2	ON HOOK CONDITION
2.1 Ringer mode. First select the value of C_{rng} (often a certain value is prescribed by local PTT requirements). R_{rng} will follow as a result of minimum allowed impedance.	
a) AC input impedance	R_{rng} [2.2 kΩ], C_{rng} [1 µF], see fig. 33
b) Ring threshold sensitivity	R_{RTA} [open], see fig. 23/24
c) Start-up time	C_{VRR} [22 µF], C_{rst} [open], R_{rst} [short]
d) Input coupling (DC block)	C_{RMI} [10 nF]
2.2 Standby mode. This mode is optional and not required if microcontroller has EEPROM or memory retention is not necessary.	
a) Standby current	R_{sb1} in series with R_{sb2} [2 * 2.7 MΩ]

6. APPLICATION EXAMPLES

As a follow up of the preceding guideline some practical examples of applications with the UBA1702/A are given in this chapter. The proposed examples include all functions required for a complete functional set like ringer, interrupter, 4 to 2 wire conversion, protection etc. Therefore the PCD3349A/53A single chip 8-bit telecom micro-controllers are used which can be programmed as multistandard repertory dialler/ringer: PCD3332-2 and PCD3330-1, see [Ref. 7] and [Ref. 8].

Figure C1 gives the basic application which has already been discussed in chapter 5.
Pulse or DTMF dialling can be selected by means of switch S2 (diode option). Because the power down function is not present, flash operation can be critical: the supply of the microcontroller (VDD) can drop below the reset level during flash and the microcontroller will reset itself. The same holds for DMO operation: the setvoltage during the make period can be too low to 'survive' the break period.

Figure C2 is more or less an extension of figure C1 for which the guideline is followed. Instead of a PMOST interrupter a bipolar transistor (MPSA92) is used.

DC settings:

• Voltage limit LN - VEE:	12 V.
• Line current limit:	120 mA.
• Line current detection sensitivity:	3 V.
• Voltage LN - SLPE:	4.5 V.
• DC slope:	25 Ω (at low line currents).
• On-hook loopresistance:	> 5 MΩ.

AC settings:

• Voltage gain mic.:	52 dB.
• Voltage gain DTMF:	25.5 dB.
• Transmit cut-off frequency:	23.4 kHz.
• Voltage gain IR - QR:	31 dB.
• Receive cut-off frequency:	15.9 kHz.
• AGC startvalue:	25 mA.
• AGC stopvalue (gain -6 dB):	60 mA.

Ringer:

• Ring threshold sensitivity:	10.5 V.
• Ringer melody high pass cut-off:	65 Hz.

Figure C3 depicts an application with the TEA1064B and the PCD3353A microcontroller. The latter contains EEPROM. This kind of memory makes number storage without the need for trickle current (flowing through resistors R_{sb1} and R_{sb2} in fig. C2) possible.
The TEA1064B features a power down function, selectable reference for logical inputs and a dynamic limiter. This dynamic limiter is combined with a mic.mute (switch Smmute). The reference input (VEE2) is in this application connected to SLPE. This results in combination with R_{VBB} and C_{VBB} in improved supply capabilities.
EMC measures are extended by the addition of C31 and C32. R_{RPI} is then required for correct 2F ringer detection. Compared to the previous example the following settings have been changed:

DC settings:

• Voltage between LN and SLPE:	4.2 V

- DC slope: 35 Ω.
- On hook loop resistance: > 5 MΩ.

AC settings:

- AGC start value: 40 mA.
- AGC stop value (gain -6 dB): 80 mA.

The last proposal is given in figure C4 and features on-hook dialling and call progress monitoring (TEA1083A). The on-hook dialling facility requires a separate ringerbridge (D7-10) capacitively coupled by C_{rng1} and C_{rng2}. Two capacitors are needed because of the common reference of ringer and speech part: VEE. The voltage across both capacitors has always the same polarity and therefore the capacitors can be unipolar. Going off-hook can be done in two ways:

- by closing the cradle switch (corresponds with picking up the handset). The EHI pin of the UBA1702 is connected to the line via R_{ehs} and T1 starts conducting. The microcontroller is informed whether the cradle has been operated by its hookinput. Going on-hook is achieved by opening the cradle switch. Consequently the hookinput of the micro is pulled down by R_{hook1}.
- by operation of the pushbutton 'HOOK'. Start up is comparable to the previous case, however a takeover signal is necessary to keep pin EHI high. This signal is generated by the microcontroller and available on pin LSE (LoudSpeaker Enable). At the same time T_{hook} conducts to simulate a switch in the matrix.

In case the handset is lifted (cradle switch closed) the hook push button can be used to switch the LS amplifier on and off. A diode (D14) has been added to protect the microcontroller input for overvoltage. The TEA1083A is supplied from the line, therefore the line current is split up in a constant part flowing through R22 (about 3 mA) and a line current dependent part flowing through the TEA1083A to SLPE. The output volume can be adjusted by means of R22. To save a microcontroller pin two ringer volumebits are combined. The stepsize is doubled and as a consequence there are 4 levels instead of 8.

7. ELECTROMAGNETIC COMPATIBILITY

With respect to electromagnetic compatibility (EMC) no common European or international specification yet exists. Also the measurement methods differ and are not always reproducible. At the application laboratory in Eindhoven (PCALE) the German current injection method is used (VDE 0878 part 200). It is a reliable method of measuring and is highly reproducible. The method is described in [Ref. 3]. The hints for EMC of the TEA106X transmission circuit given in the second paragraph of this chapter are based on this method. The same counts for the hints of the printed circuit board design given in the first paragraph.

7.1 Printed circuit board

In the current injection method, radio frequency (RF) signal currents enter the telephone set at the a/b wires and leave the set via any capacitive coupling to ground. Normally, in a telephone set the handset has the largest capacitance to ground and thus the main part of the RF signal current flows to ground via the handset. However, a proper PCB layout is essential for good EMC.

The first measure to be taken is to create a groundplane on the PCB. The RF signals entering the PCB should be decoupled immediately to this groundplane. Preferably this groundplane is homogeneous and is not cut into parts by interconnection wires. To reach this a double layered PCB, with interconnection wires on one side of the board and a groundplane on the other side, is a minimum. When interconnection wires within the groundplane are inevitable, the continuity of the groundplane should be restored. This is done by cross coupling these interconnection wires by jumpers and wires at the interconnect side of the PCB. In this way RF signal currents can flow freely over the groundplane.

Another measure is to keep the length of the wires between the different components as short as possible. Of course this measure is especially important for those wires which interconnect one or more RF signal sensitive parts, in particular the wire connected to SLPE which can also be a reference for the UBA1702/A. As any current, RF signal currents prefer to flow to ground via the lowest ohmic path. Therefore, it should be noted that a wire of 10 mm corresponds to an inductor of 10 nH.

7.2 TEA106X

In this paragraph the standard measures for the TEA106X are given. When, after the proposed measures are taken, the EMC behaviour has to be optimized, it is advised to start with the transmit direction of the telephoneset and to optimize the receive part thereafter. This because, due to sidetone, signals demodulated by the transmit channel will be seen in the receiver channel.

In the text below it is supposed that the printed circuit board on which the TEA106X-UBA1702/A is built, is provided with a groundplane which is connected to VEE. It is preferred to place the components meant for EMC as close as possible to the pins, except when otherwise stated.

For more details on the EMC performance of the TEA106X, see [Ref. 3].

RF signals entering the printed circuit board at the a/b wires should be decoupled to the groundplane via capacitors, preferably placed as close as possible to the a/b connection. Since these capacitors will be in parallel with the set impedance, their value is limited. In practice, a total capacitance of 10nF between the a/b wires may be applied without degrading the balance return loss.

Between the a/b wires and the groundplane two capacitors of can be placed, as well as a series inductance of 22 μH in series with the diode bridge.

RF signals entering the inputs of the transmit channel (MIC+, MIC-) will be demodulated and amplified to the line. Because of the high gain (in the order of 50 dB) the transmit channel is very sensitive to RF signals. Therefore, decoupling at the inputs is essential.

The inputs MIC+ and MIC- can preferably be decoupled by 2 capacitors of 2.2 nF connected to the groundplane. Also series resistors can be applied which in combination with the capacitors form low pass filters towards the inputs. Resistors of 1 kΩ are advised (in case of an electret microphone) which reduce the gain setting with less than half a dB, depending on the input impedance of pins MIC+ and MIC-. This can be compensated by adapting the transmit gain adjustment resistor R7 of the TEA106X.

In case a handset microphone is connected to the TEA106X microphone inputs via a long cord, extra decoupling is needed. This can be done by adding two capacitors of 2.2 nF each, placed between the cord connection and the groundplane, preferably as close as possible to the cord connection.

RF signals entering the inputs of the receive channel (IR) will be demodulated and amplified to the earpiece and therefore decoupling at the input is essential.

At the input IR of the TEA106X a capacitor of 1 nF connected to the groundplane is advised.

In case an earpiece is connected to the TEA106X via a long cord, extra decoupling is needed. This can be done by adding two capacitors of 10nF each, placed between the cord connection and the groundplane, preferably as close as possible to the cord connection. To prevent the remaining RF signals from entering the earpiece output-stage of the TEA106X via the QR pin, a series resistor should be applied to create a high ohmic path. The value of this resistor is dependent on the type of earpiece capsule used. When a dynamic capsule of 150 Ω is used, a resistors of 22 Ω is advised. This will reduce the gain setting with 1.19 dB. This can be compensated by adapting the receive gain adjustment resistor R4 of the TEA106X.

Besides these essential measures some additional measures can be taken. A capacitor between GAS2 and the groundplane of 100 pF can improve EMC in the transmit direction. A series combination of a resistor of 365 Ω and a capacitor of 4.7 nF connected between STAB and the groundplane can improve EMC for both transmit and receive direction.

8. REFERENCES

[Ref. 1] Philips Semiconductors Data Handbook
Semiconductors for telecom systems - IC03
Philips Semiconductors, 1995

[Ref. 2] UBA1702/A Line Interrupter Driver and Ringer IC
Preliminary specification
October 1995

[Ref. 3] Measures to meet EMC requirements for TEA1060-family speech transmission circuits
by M. Coenen and K. Wortel
PCALE reportnumber: ETT89016

Extended application information on the speech transmission circuits can be found in:

[Ref. 4] TEA1060 family versatile speech/transmission ICs for electronic telephone sets
Designers' guide by P.J.M. Sijbers
July 1987, 12NC 939834110011

[Ref. 5] Application of the versatile speech/transmission circuit TEA1064 in full electronic telephone sets
by F. van Dongen and P.J.M. Sijbers
PCALE reportnumber: ETT89009

[Ref. 6] Application of the speech-transmission circuit TEA1062
by P.T.J. Biermans
PCALE reportnumber: ETT89008

More information on dialler / ringers can be found in:

[Ref. 7] PCD3330-1: A multi-standard repertory dialler/ringer with EEPROM (programmed PCD3353A)
Objective specification; July 1993

[Ref. 8] PCD3332-2: A multi-standard pulse/tone repertory dialler/ringer (programmed PCD3349A)
Objective specification; March 1994

APPENDIX A. TEA106X QUICK REFERENCE DATA

	DC-CHARACTERISTICS (with slope resistance R9 = 20 Ω)		
Member	V(LN-VEE) (in V)at Iline = 15 mA	V(LN-VEE) (in V) at R(REG-LN) = 68 kΩ	V(LN-VEE) (in V) at R(REG-SLPE) (in Ω)
TEA1060	4.45 ± 0.20	3.80 + 0.25 / -0.30	5.0 ± 0.30 at 39 k
TEA1062	4.00 + 0.25 / -0.45	3.50 ±	4.5 ± .... at 39 k
TEA1064B	3.50 ± 0.25	---	4.4 ± 0.35 at 20 k
TEA1067	3.90 ± 0.25	3.40 ± 0.30	4.5 ± 0.30 at 39 k
TEA1068	4.45 ± 0.25	3.80 ± 0.30	5.0 ± 0.35 at 39 k

	SENDING GAIN		RECEIVE GAIN (from IR to QR+)	
Member	Setting range (in dB)	Gain (in dB) with R7 = 68 kΩ	Setting range (in dB)	Gain (in dB) with R4=100 kΩ
TEA1060	44 - 60	52 ± 1	17 - 33	25 ± 1
TEA1062	44 - 52	52 ± 1.5	20 - 31	31 ± 1.5
TEA1064B	44 - 52	52 ± 1	20 - 39	31 ± 1
TEA1067	44 - 52	52 ± 1	20 - 39	31 ± 1
TEA1068	44 - 60	52 ± 1	17 - 33	25 ± 1

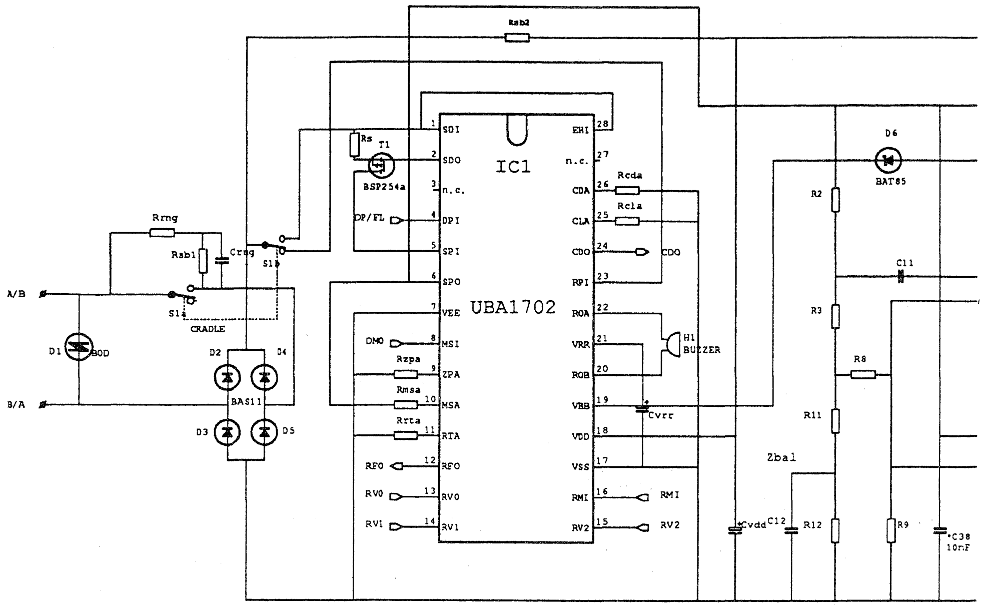
	SENDING NOISE	
Member	Noise (in dBmp) with R7 = 68 kΩ	Noise (in dBmp) with sending gain of 44 dB
TEA1060	-70	-78
TEA1062	-69	-77
TEA1064B	-72	-80
TEA1067	-72	-80
TEA1068	-72	-80

For more data see datahandbook [Ref. 1].

APPENDIX B. LIST OF ABBREVIATIONS AND DEFINITIONS

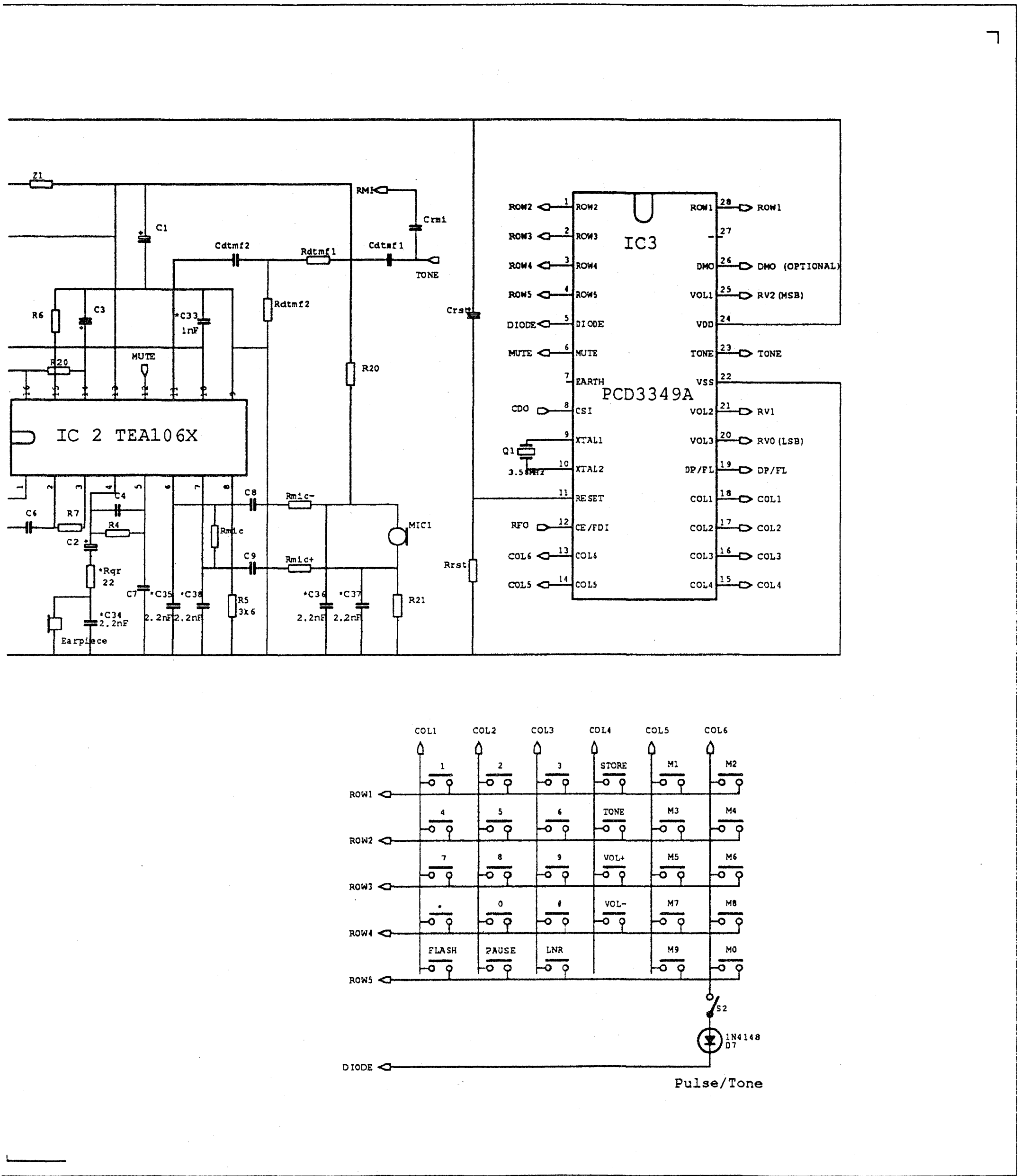
AGC	Automatic Gain Control: line loss compensation of the TEA106X
BRL	Balance Return Loss
CE/notT0	Chip enable / not test0 input of microcontroller, used for putting the device in active mode and for ringer frequency discriminator input
Crng	Ringer input capacitor
Crst	Capacitor for generating external reset for microcontroller
CSI	Cradle Switch Input, see T1
DMO	Dial Mode Operation, provision to guarantee a low set resistance during the make period of pulse dialling, also designated as Mute2 or NSA
DTMF	Dual Tone Multi Frequency (dialling system)
EHS	Electronic Hook Switch (required for on-hook dialling)
EMC	Electro Magnetic Compatibility: the collective noun for the susceptibility and the radiation of a circuit / apparatus
FDI	Frequency Discriminator Input, see CE/notT0
Flash	Timed break of (DC) line current for call transfer by the exchange
HOOK	Hook switch control input, see T1
IR	Receive input pin of the TEA106X
LN	Positive line terminal pin of the TEA106X
MIC	Microphone input pin
MIC+, MIC-	Microphone inputs pins on the TEA106X
MOSFET	Metal Oxide Field Effect Transistor
Mute	Mode which is operational during dialling
PCALE	Product Concept & Application Laboratory Eindhoven
PCB	Printed Circuit Board
Powerdown	Reduced current consumption mode during pulse dialling or flash operation
PTT	Telephone administration
QR-, QR+	Telephone earpiece output on TEA106X
Rcda	Resistor setting line current detection sensitivity
Rcla	Resistor setting line current limit value
RF	Radio Frequencies / Ringer Frequency input, see CE/notT0
Rmic	Resistor setting the microphone sensitivity
Rmsa	Resistor setting mute (DMO) voltage
Rrng	Ringer input resistor
Rrst	Resistor for generating external reset for microcontroller
Rrta	Resistor setting ringer threshold
Rsb1, Rsb2	Resistors for stand-by (trickle) current for memory retention
Rzpa	Resistor setting zener protection voltage
SLPE	DC slope pin on TEA106X
T1	Test1 input of microcontroller, used for on / off hook detection
TEA106X	IC of the TEA106X speech transmission family: TEA1060/61, TEA1062, TEA1063, TEA1064A, TEA104B, TEA1065, TEA1066, TEA1067, TEA1068
VCC	Supply pin of the TEA106X
VEE	Ground reference pin of TEA106X and UBA1702/A
VSS	Ground reference for microcontroller and logic reference for the UBA1702/A

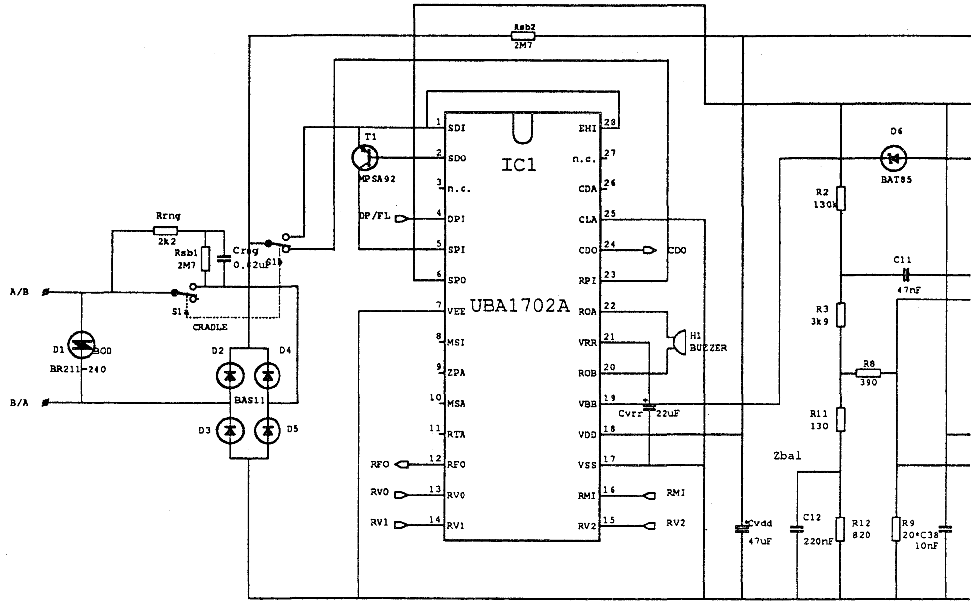
APPENDIX C APPLICATION DIAGRAMS



EMC components marked with *

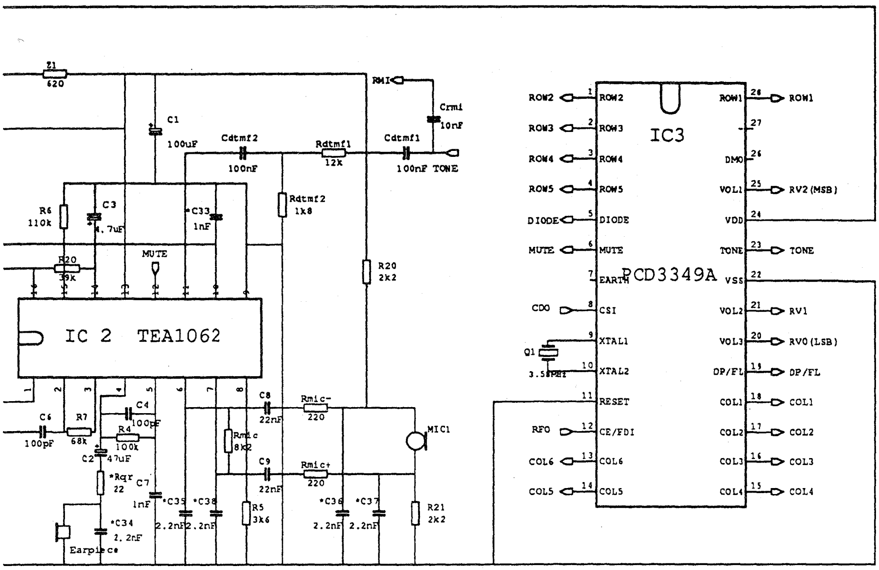
Figure C1 Basic application with TEA106X

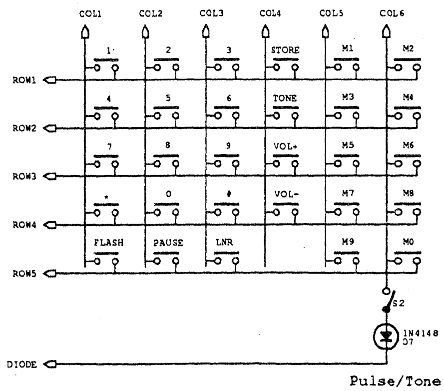

IC 2 TEA106X
IC3
PCD3349A
Pulse/Tone
MUTE
TONE
Earpiece
MIC1
1N4148
D7
S2
ROW1
ROW2
ROW3
ROW4
ROW5
COL1
COL2
COL3
COL4
COL5
COL6
DIODE
STORE
FLASH
PAUSE
LNR
VOL+
VOL-
DMO (OPTIONAL)
RV2 (MSB)
RV1
RV0 (LSB)
DP/FL

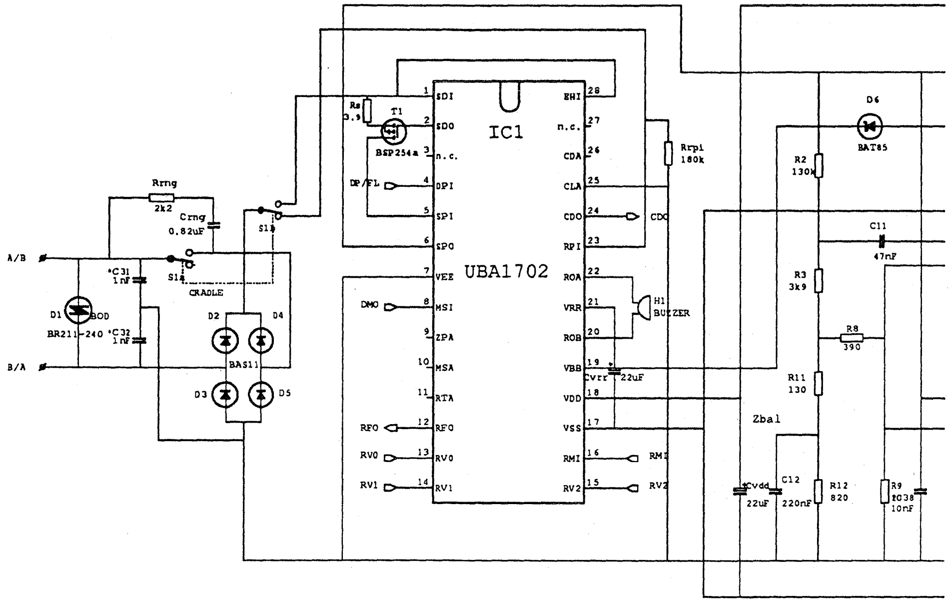


EMC components marked with *

Figure C2 Application UBA1702A with TEA1062

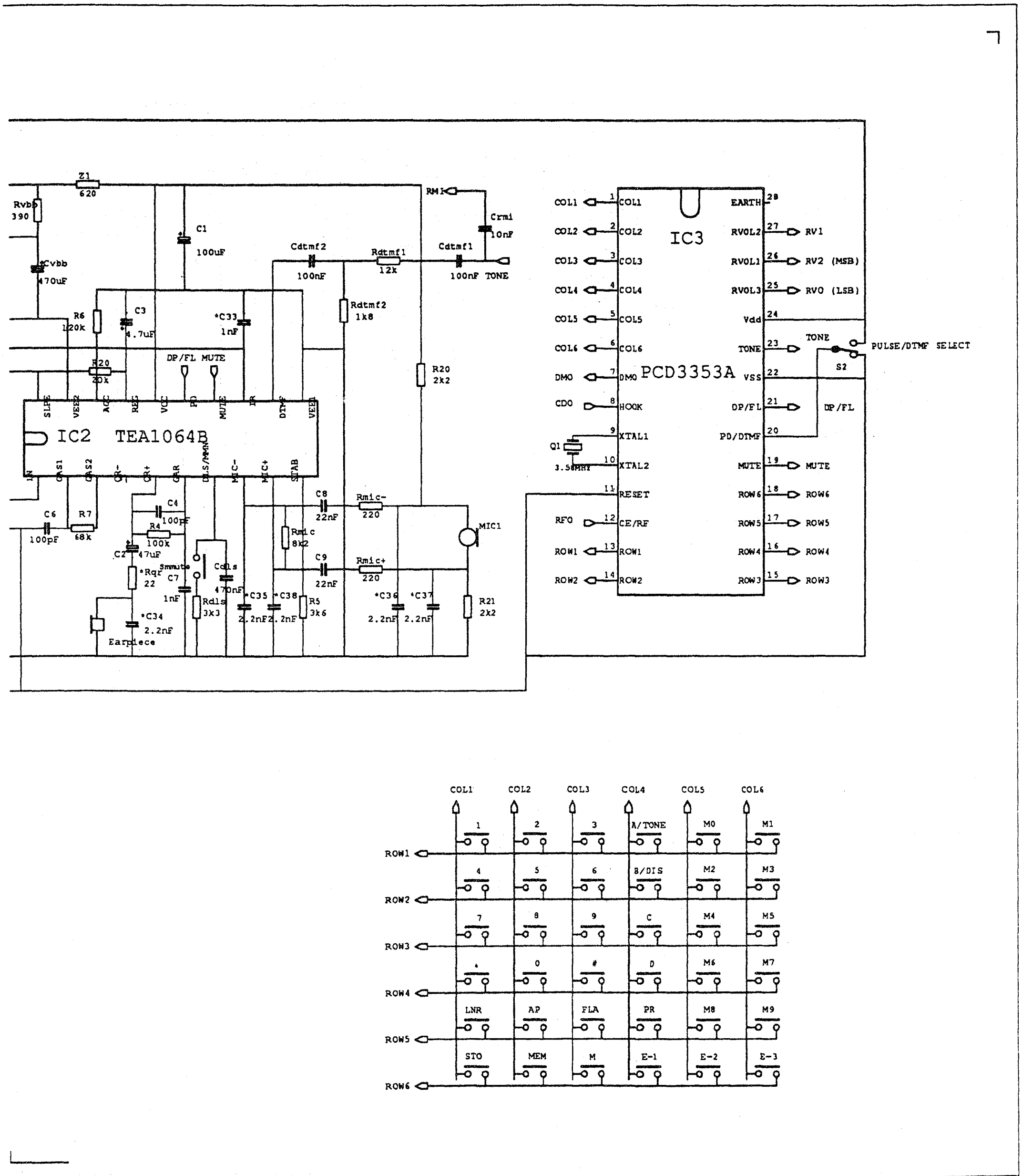

IC 2 TEA1062
IC3
PCD3349A
Pulse/Tone


COL1 COL2 COL3 COL4 COL5 COL6
ROW1 ROW2 ROW3 ROW4 ROW5
STORE TONE VOL+ VOL-
FLASH PAUSE LNR
DIODE
Pulse/Tone



EMC components marked with *

Figure C3 Application UBA1702 with TEA1064B



Z1
620
Rvbb
390
Cvbb
470uF
C1
100uF
Cdtmf2
100nF
Rdtmf1
12k
Cdtmf1
100nF
TONE
RM
Crmi
10nF
Rdtmf2
1k8
R6
120k
C3
4.7uF
*C33
1nF
R20
20k
DP/FL
MUTE
R20
2k2
SLPE
VEE2
AGC
REG
VCC
PD
MUTE
IR
DTMF
VEE1
IC2
TEA1064B
LN
GAS1
GAS2
QR-
QR+
GAR
DLS/MMUTE
MIC-
MIC+
STAB
C6
100pF
R7
68k
C4
100pF
R4
100k
C2
47uF
*Rqr
22
Smmute
C7
1nF
Cdls
470nF
Rdls
3k3
*C34
2.2nF
Earpiece
C8
22nF
Rmic-
220
Rmic
8k2
C9
22nF
Rmic+
220
*C35
*C38
2.2nF
R5
3k6
*C36
*C37
2.2nF
MIC1
R21
2k2
IC3
PCD3353A
COL1
COL2
COL3
COL4
COL5
COL6
DMO
CDO
HOOK
XTAL1
XTAL2
Q1
3.58MHz
RESET
RFO
CE/RF
ROW1
ROW2
ROW3
ROW4
ROW5
ROW6
EARTH
RVOL2
RVOL1
RVOL3
RV1
RV2 (MSB)
RV0 (LSB)
Vdd
TONE
VSS
DP/FL
PD/DTMF
MUTE
S2
PULSE/DTMF SELECT
1
2
3
A/TONE
M0
M1
4
5
6
B/DIS
M2
M3
7
8
9
C
M4
M5
*
0
#
D
M6
M7
LNR
AP
FLA
PR
M8
M9
STO
MEM
M
E-1
E-2
E-3

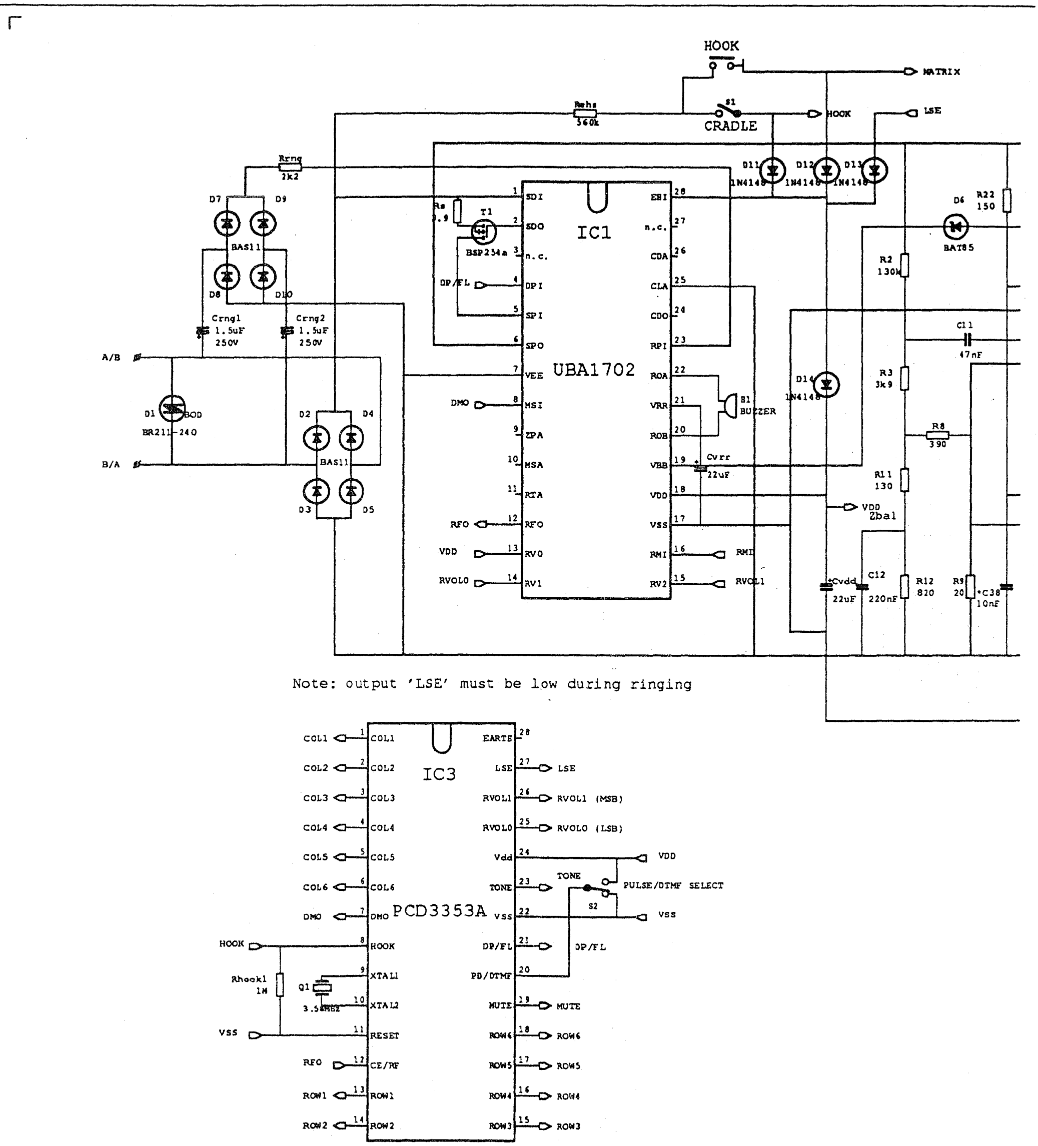
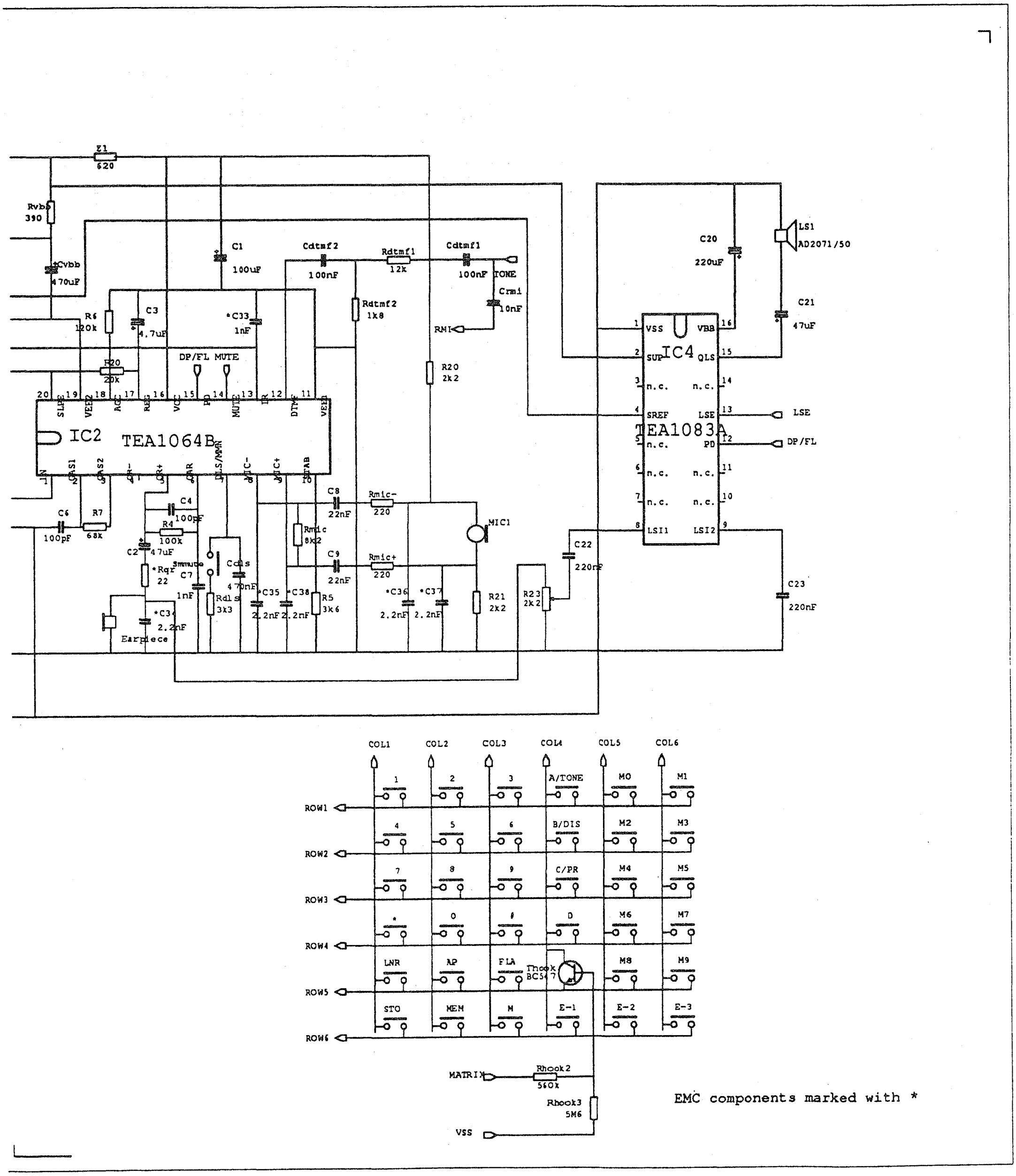


Figure C4 Application UBA1702 with TEA1064B/1083(A)


IC2
TEA1064B
IC4
TEA1083A
LS1
AD2071/50
MIC1
Earpiece
COL1
COL2
COL3
COL4
COL5
COL6
ROW1
ROW2
ROW3
ROW4
ROW5
ROW6
MATRIX
VSS
LSE
DP/FL
RMI
TONE
EMC components marked with *

3 SPEECH/TRANSMISSION, SUPPLY, LISTENING-IN AND HANDSFREE CIRCUITS

CONTENTS

APPLICATION NOTE Nr	9398 341 10011
TITLE	TEA1060 Family Versatile Speech/Transmission ICs for Electronic Telephone sets ‘Designers Guide’
AUTHOR	P. J. M. Sijbers
DATE	August 1992

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VERSATILE SPEECH/TRANSMISSION ICs FOR ELECTRONIC TELEPHONE SETS

The speech and transmission functions which form the heart of a subscriber telephone set are undoubtedly the most difficult part of the system to implement on a single integrated circuit. Such an IC must must provide an interface between the line and a pulse or a DTMF dialling circuit, yet be flexible enough to work in conjunction with many different types of microphone and earpiece. Furthermore, it must be adaptable to the different transmission requirements of the various telephone authorities. The versatile TEA1060 family is a series of five bipolar speech/transmission ICs which have been designed to meet the requirements of most major international PTTs in Europe, the USA, and the Far East.

Besides their use in fully electronic telephone sets (basic sets and feature phones) they can also form the logical link between the telephone line and the electronic dialling circuitry (DTMF or pulse dialling), and control circuits in automatic answering machines, facsimile equipment and cordless telephones. The ICs were designed primarily for the increasingly used common line-interface systems (with electronic switching between dialling and speech) but are also suited for separated speech systems (with a two-wire connection between the dialling base and the handset).

The TEA1060 family ICs are powered with current from the telephone line and you can use them with virtually any kind of microphone; the set impedance can be complex or real. An anti-sidetone function can be realized by incorporating either the TEA1060 family anti-sidetone bridge or the Wheatstone bridge configuration in the application. You'll find a detailed description of the anti-sidetone bridges in Appendix A.

The member of the TEA1060 family which you should use depends on the type of microphone capsule that's going into the telephone set and the local PTT requirements. For telephone sets with low-impedance dynamic and magnetic microphones which generally require a gain of between 44 dB and 60 dB, you can use the TEA1060 (2 x 4 kΩ input impedance).

For piezoelectric and electret-capacitor microphones with source followers or preamplifiers, you should use the TEA1061 because they require a high-impedance termination (2 x 20 kΩ) and a gain of between 30 dB and 46 dB. The TEA1066T has the input options of the TEA1060 and TEA1061 and so you can use it with high- or low-impedance microphones. This has been achieved by encapsulating the IC in a small outline 20-pin (SO-20) package, suitable for surface mounting.

The TEA1068 is a more flexible member of the TEA1060 family. With a gain of between 44 and 60 dB and an input impedance of 64 kΩ (32 kΩ unbalanced) the IC can be externally tailored to suit any combination of microphone impedance / sensitivity. It also has advantages in applications where very accurate microphone matching is necessary. When used with the German "Graue Mikrofon Schnittstelle" (Common Microphone Interface) most types of microphones can be used without the need to readjust circuit parameters. Appendix B gives a more detailed description of the German "Graue Mikrofon Schnittstelle".

If you need an IC that will allow parallel operation with conventional telephone sets, the low-voltage TEA1067 is the chip that you need. This parallel operation is mandatory for telephones used in the United States. Its microphone amplifier is suitable for medium sensitivity microphones which require between 44 and 52 dB of gain and its earpiece amplifier has the highest signal to noise ratio; 4 dB higher than the rest of the TEA1060 family. It can operate with a voltage drop as low as 1,6 V, but a slightly degraded performance will be inevitable at the lowest voltages. Both the TEA1067 and the TEA1068 are available in an 18-pin DIL and also in an SO-20 package for surface mounting.

All the ICs can provide a supply for peripheral circuitry, but the ability to supply current to the peripheral circuitry largely depends on the particular application. Generally, most dialling circuits and microcontrollers can be powered, but for applications like loudspeaking facilities you will need additional supply circuitry (see Appendix D - Application Circuit Examples). When used in conjunction with the TEA1080 supply circuit, the ability to supply current to peripheral circuitry can be extended considerably.

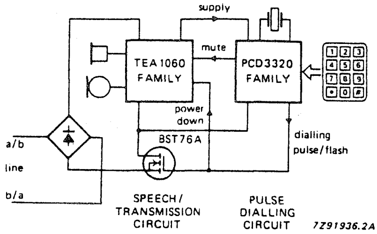
OVERVIEW OF THE TEA1060 FAMILY FEATURES

TEA1060 FAMILY FEATURE	TEA 1060	1061	1066T	1067	1068
Microphone inputs:					
low-sensitivity (52-60 dB gain) low-impedance dynamic or magnetic mikes	●		●		●
medium sensitivity (44-52 dB gain) low-impedance dynamic or magnetic mikes	●		●	●	●
electret with source follower or preamplifier (30-46 dB gain)		●	●	*	*
piezo-electric (30-46 dB gain)		●	●	*	*
large amplifier setting range	●	●	●	●	●
very accurate microphone matching				●	●
Receiver outputs:					
dynamic or magnetic	●	●	●	●	●
piezo-electric	●	●	●	●	●
large amplifier setting range (17-39 dB)	●	●	●		●
extra large amplifier setting range (20-45 dB)				●	
extra low-noise amplifier				●	
Electronic mute input	●	●	●	●	●
DTMF input	●	●	●	●	●
Voltage-regulator:					
adjustable d.c. voltage:					
at 15 mA range 3,85 to 5,05 V	●	●	●		●
at 15 mA range 3,3 to 4,5 V				●	
peripheral supply point	●	●	●	●	●
adjustable d.c.slope	●	●	●	●	●
Power-down input	●	●	●	●	●
Gain Control of mic. and rec. amp.					
control can be switched off	●	●	●	●	●
adapted to 600 Ω feed	●	●	●	●	●
adaptable to exchange supply voltage	●	●	●	●	●
SO-encapsulation			●	x	x
Parallel operation possible (>1,6 V)				●	

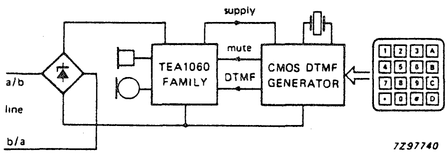
● available function

* attenuator at input neccessary for TEA1067 and TEA1068

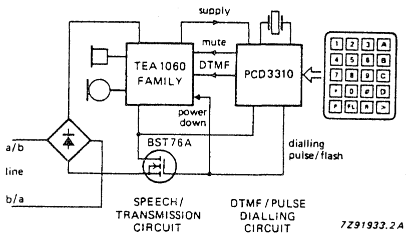
x available on special order



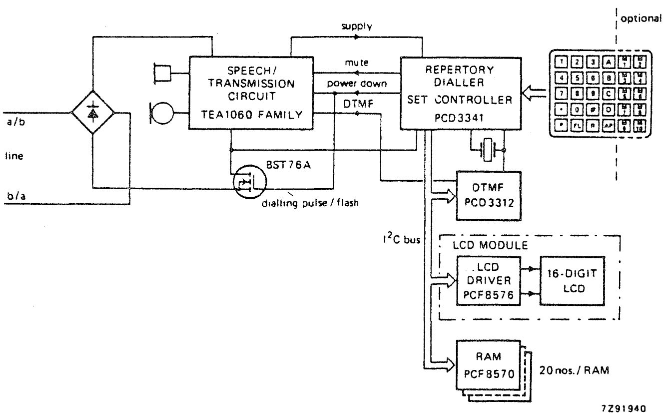
(a) Basic set with pulse dialling



(b) Basic set with DTMF dialling



(c) Set with pulse and DTMF dialling



(d) Top-of-the-range feature phone

Fig.1 Subscriber telephone sets with a common line-interface

SUBSCRIBER SET ARCHITECTURES

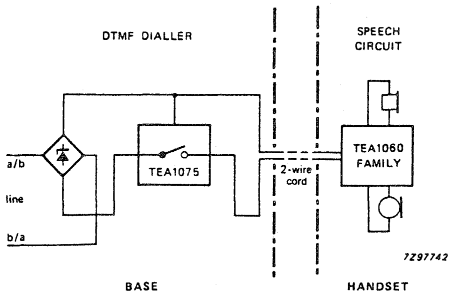
There are two basic types of architecture for electronic telephone subscriber sets. In the common line-interface architecture, the speech and dialling circuits are connected to the line by the same interface. In the architecture with separated speech and dialling circuits each is connected to the line by an individual interface.

A common line-interface has the advantage that only one voltage regulator and one transmitting stage are necessary to apply either speech, or dialling signals to the line. Switch-over from the dialling mode to the speech mode is performed electronically by the IC (Mute function) enabling virtually click-free operation. Sets with different dialling systems and/or features can make use of one common design for the transmission part. In this way a modular approach is possible.

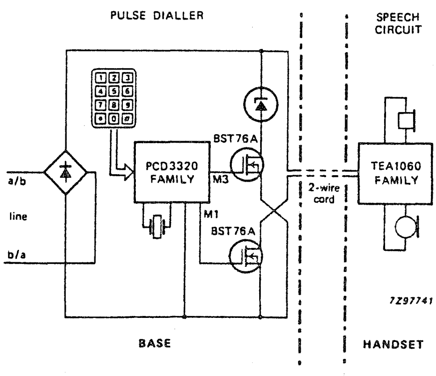
For separated speech and dialling circuits, both the speech circuits and the dialling circuits must have a voltage regulator and a transmission line interface. The switch-over from the dialling to the speech mode can be performed either by external electronic components, or by a common contact operated by every button on the keyboard. This has the advantage that the speech circuit can be placed closer to the microphone and earpiece to achieve better immunity to r.f. interference.

SETS WITH A COMMON LINE-INTERFACE

The versatility of the TEA1060 family allows them to be used with several types of subscriber set with common line-interface as shown in Fig.1. Figure 1(a) and (b) show two basic sets, one with pulse and one with DTMF dialling. Figure1(c) shows a set with both pulse and DTMF dialling. All the advantages of the ICs can be fully exploited if pulse and DTMF dialling are required in one set. A top-of-the-range feature phone with a microcontroller is shown in Fig.1(d).



(a) Basic set with DTMF dialling



(b) Basic set with pulse dialling

Fig.2 Subscriber telephone sets with separated speech and dialling functions

The advantages of a common line-interface architecture are:

- a well defined interface (supply, common, mute, power-down and interrupter) between speech functions, dialling functions and control functions.
- the switch-over from dialling to speech mode is virtually click-free because of the internal mute function in the TEA1060 family.
- the line-interface is not duplicated.
- it is possible to have a confidence tone in the earpiece during DTMF dialling.
- no additional peripheral components are needed for switching-over from dialling to speech mode.
- this modular approach results in considerable flexibility with regard to dialling features.

SETS WITH A SEPARATED SPEECH AND DIALLING CIRCUIT

The advantages of the architecture with separated speech and dialling circuits (Fig.2(a) and (b)) are:

- this type of architecture allows a member of the TEA1060 family to be used as a direct two-wire replacement for the conventional speech part of the telephone.
- the speech IC can be installed in the handset with only a two-wire connection to the base of the instrument.
- high immunity to r.f. interference because the wires between the microphone capsule and the speech/transmission IC will be very short.

The line-interface functions are duplicated, and because the speech and dialling circuits are alternately connected to the line it is more difficult to produce a confidence tone during DTMF dialling. Generally, the switch-over from the dialling mode to the speech mode is not click-free, but the clicks can be reduced to an acceptable volume by using techniques which require extra wires between the speech and dialling circuitry.

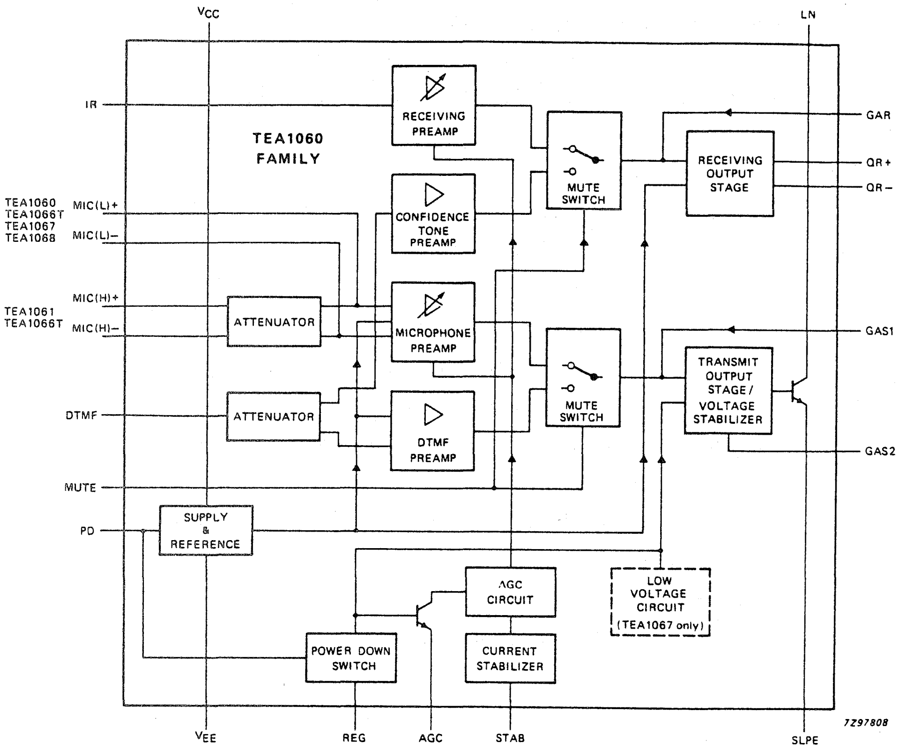


Fig.3 Internal circuitry of the TEA1060 family

FUNCTIONAL DESCRIPTION

The block diagram of the TEA1060 family is shown in Fig.3. The internal functions are:

- Voltage regulator with adjustable static resistance and the facility for adjusting the voltage drop externally (4,45 ± 0,6 V for the TEA1060/61/66T/68 and 3,9 ± 0,6 V for the TEA1067 at 15 mA).
- Supply connection for powering peripheral circuits. The capabilities of the supply depend on the d.c. voltage setting of the voltage-regulator, on external components and on the available line current.
- Microphone amplifier with a wide-range gain setting and a frequency roll-off with adjustable cut-off frequency (44 to 52 dB for the TEA1067, 30 to 46 dB for the TEA1061/66T(H), and 44 to 60 dB for the TEA1060/66T(L)/68).
- Low-impedance(8 kΩ) differential microphone inputs for dynamic and magnetic microphones on the TEA1060 & TEA1066T(L).
- High-impedance microphone inputs for piezo-electric microphones (balanced connection 40 kΩ) or for electret-condenser microphones with a source follower or pre-amplifier (unbalanced connection 20 kΩ) on the TEA1061 and TEA1066T(H).
- Very high-impedance microphone inputs (64 kΩ) for accurate microphone matching with external resistors; suitable for all types of microphones (TEA1067 and TEA1068).
- DTMF input.
- Confidence tone in the earpiece during DTMF dialling.
- Transmitting output stage.
- Receiver amplifier with two complementary outputs suitable for magnetic, dynamic or piezoelectric earpieces; the amplifier has wide gain setting range (TEA1067 20 to 45 dB, and the rest are 17 to 39 dB) and adjustable cut-off frequency.
- Line loss compensation facility (line current dependent) for microphone and earpiece amplifiers. The control curve has been optimized for a 600 Ω feeding bridge and is adaptable to various exchange supply voltages.

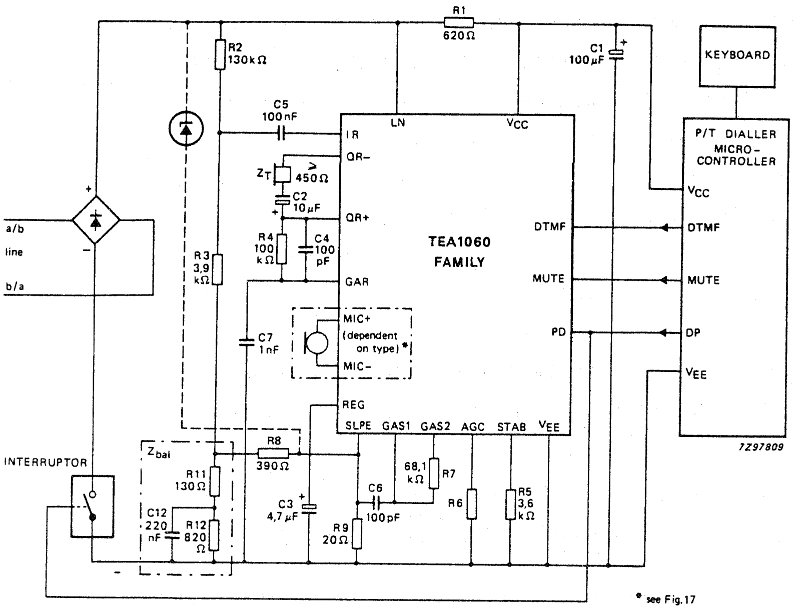


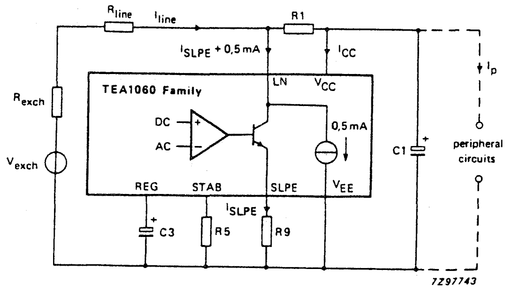
Fig.4 Basic application circuit for the TEA1060 family (not including protection)

- Mute input to inhibit both the earpiece amplifier and the microphone amplifier during dialling, and to permit DTMF input and confidence tone.
- Power-down input to minimize the internal supply current of the IC during line interruptions with pulse dialling or register recall (flash); the voltage-regulator capacitor is disconnected to prevent start-up delays after line interruptions to minimize the contribution of the IC to the shape of the current pulses during pulse dialling.
- Low-voltage circuit enabling parallel operation (>1,6 V TEA1067 only)

In the following text the members of TEA1060 family are described with reference to the basic circuit shown in Fig.4. The pinning/signal functions are given in the table below.

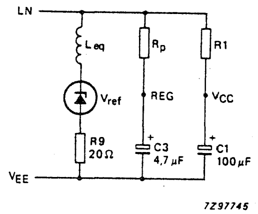
Signal Name	IC Pin Numbers TEA1060	TEA1061	TEA1066T	TEA1067/ TEA1068
LN Positive line terminal	1	1	1	1
GAS1 Gain adjustment Tx amplifier	2	2	2	2
GAS2 Gain adjustment Tx amplifier	3	3	3	3
QR- Inverting output Rx amplifier	4	4	4	4
QR+ Non-inverting output Rx amplifier	5	5	5	5
GAR Gain adjusment Rx amplifier	6	6	6	6
MIC(H)- Inverting mike i/p(20 kΩ)		7	8	
MIC(H)+ Non-inverting mike i/p(20 kΩ)		8	10	
MIC(L)- Inverting mike i/p(4 kΩ)	7		7	
MIC(L)+ Non-inverting mike i/p(4 kΩ)	8		9	
MIC - Inverting mike i/p(32 kΩ)				7
MIC + Non-inverting mike i/p(32 kΩ)				8
STAB Current stabilizer	9	9	11	9
V_{EE} Negative line terminal	10	10	12	10
IR Rx amplifier input	11	11	13	11
PD Power-down input	12	12	14	12
DTMF Dual-tone multi-frequency input	13	13	15	13
MUTE Mute input	14	14	16	14
V_{CC} Positive supply decoupling	15	15	17	15
REG Voltage regulator decoupling	16	16	18	16
AGC Automatic gain control	17	17	19	17
SLPE Slope (dc resistance) adjustment	18	18	20	18

Note: (H)and (L) refer to the high- and low-impedance microphone amplifier inputs, respectively, of the TEA1066T.



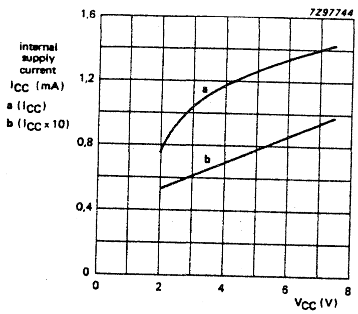
$V_{LN} = V_{ref} + (I_{SLPE} \times R_9)$ (with $I_{SLPE} > I_{TH}$ for TEA1067)
where: V_{ref} = the internal reference voltage of 4,2 V (3,6 V for the TEA1067)
$I_{SLPE} = I_{line} - I_{CC} - 0{,}5\ \text{mA} - I_p$

Fig.5 Power supply arrangements



$L_{eq} = R_p \times R_9 \times C_3$ = 1,65 H (1,52 H for TEA1067)
R_p = 17,5 kΩ (16,2 kΩ for the TEA1067)

Fig.6 Equivalent impedance.



(a) Normal operating condition; PD = low.
(b) Power down condition; PD = high.

Fig.7 Internal supply current I_{CC} as a function of V_{CC}

SUPPLY CONSIDERATIONS

Supply and set impedance

The IC is supplied with current from the line as shown in Fig.5. The equivalent impedance of the circuit is shown in Fig.6. For effective operation of the telephone circuitry the IC must have a low resistance to d.c. and a high impedance to the speech signal. This is done by incorporating an artificial inductor in the IC, L_{eq}. The value L_{eq} is R_p x R_9 x C_3. The value of C_3 also determines the start-up time of the d.c. voltage regulator and has been chosen such that the voltage regulator starts-up as soon as the V_{CC} smoothing capacitor has been charged. The value of L_{eq} should be adjusted by changing C_3 (taking into account a different start-up time) rather than changing the value of R_9 because the latter influences several other parameters as described later.

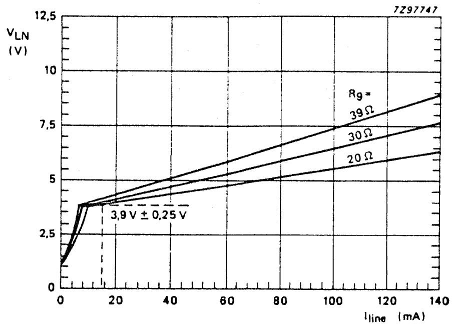
The impedance of the whole circuit, to audio frequencies, is determined by the value of $R_1//R_p$.The network R_1C_1 provides a smoothed voltage V_{CC} for the IC (typically I_{CC} = 1 mA at V_{CC} = 2,8 V) and for the peripheral circuits (I_p). Typical I_{CC} as a function of V_{CC} is shown in Fig.7.

The direct current which flows into the set is determined by:

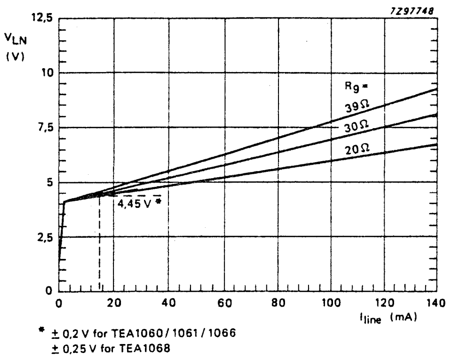
- the exchange supply voltage (V_{exch})
- the resistance of the feeding bridge (R_{exch})
- the resistance of the subscriber line (R_{line})
- the d.c. voltage across the subscriber set, including the polarity guard

If the line current exceeds (I_{CC} + 0,5 mA + I_p) then the voltage regulator diverts the excess current through pin LN. The equation for the voltage drop across the IC, V_{LN}, is given in Fig.5. For the TEA1067 this equation can only be used for line currents exceeding I_{TH} (the threshold current in the low-voltage range, typ. 9 mA).

The internal reference voltage is temperature-compensated, giving a low temperature-coefficient for the line voltage V_{LN}; about -2mV/K at I_{line} = 15 mA (-1 mV/K for TEA1067). I_{SLPE} is normally much greater than the sum of I_{CC} + 0,5 mA + I_p. The equivalent circuit for d.c. conditions is therefore that of a 4,2 V voltage regulator diode in series with resistor R_9 (3,6 V for the TEA1067 where $I_{line} > I_{TH}$ to satisfy the U.S. requirement for a line voltage of 6 V at 20 mA).

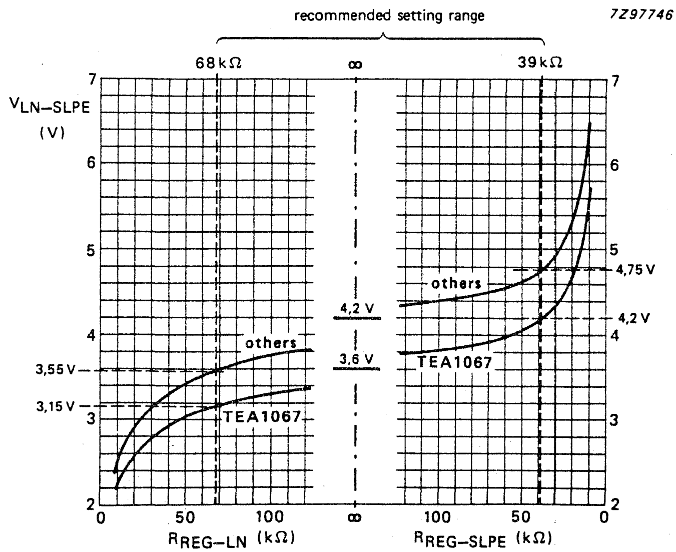


(a) TEA1067 d.c. characteristics



(b) TEA1060 family d.c. characteristics (not TEA1067)

Fig.8 DC characteristics



With line currents of between 11 and 140 mA the d.c. voltage;
$V_{LN} = V_{LN-SLPE} + (I_{LINE} - 1,5\ mA) \times R_9$.

Fig.9 Internal reference voltage $V_{LN-SLPE}$ as a function of resistors $R_{REG-SLPE}$ and R_{REG-LN}.

The typical d.c. voltage V_{LN} is shown in Fig.8 (a) (TEA1067) and (b) (rest of family) as a function of line current. The gradient of the line is determined by R_9. The optimum value for R_9 is 20 Ω, as used in the basic application circuit example (Fig.4). Another value for R_9 is sometimes necessary to rebalance the anti-sidetone circuit e.g. when a set impedance other than 600 Ω is chosen. Any change in the choice of value of R_9 will cause:

- a change of microphone gain and DTMF gain
- a shift of the gain-control characteristic
- the sidetone will also be affected
- a decrease of the maximum output swing on LN if the value of R_9 exceeds 30 Ω (especially at high line currents and high ambient temperature)
- shift of the low-voltage threshold current I_{TH} (TEA1067 only).

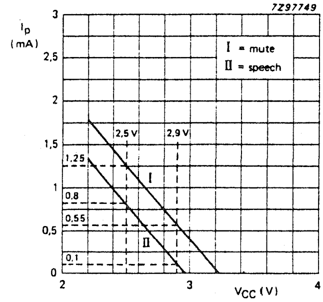
If a change in the value of R_9 is necessary, the design procedure in Appendix C should be followed.

Increasing the d.c. Slope

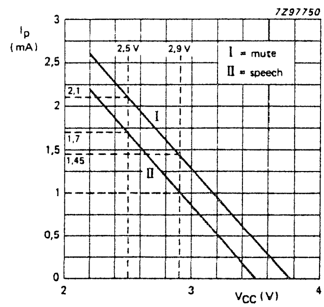
The gradient of the d.c. characteristic can be increased by connecting a resistor between pin LN and node [R_1, R_2]. This resistor doesn't influence the set impedance, but it slightly decreases the maximum output swing on the line. Another alternative in tone dialling applications, is to increase the value of protection resistor R_{10}. See "Transient Suppression and the polarity guard" and Fig.31.

Adjusting the d.c. Voltage Drop

Adjusting the d.c. voltage drop across the circuit will change the maximum output swing of the sending and receiving amplifiers, and the supply current available for peripherals. The voltage drop across the circuit ($V_{LN-SLPE}$) can be increased by connecting a resistor between pins REG and SLPE or decreased by connecting a resistor between pins REG and LN ($R_{REG-SLPE}$ and R_{REG-LN} respectively). These resistors set the internal reference voltage of the voltage stabilizer ($V_{ref} = V_{LN-SLPE}$). However, there will be a small change in the temperature-coefficient and set impedance and there will be a slight increase in the spread of the voltage drop. Figure 9 shows how $V_{LN-SLPE}$ changes with the value of $R_{REG-SLPE}$ and R_{REG-LN}. Only one of the resistors should be used.



(a) V_{LN} = 3,9 V (typical for TEA1067, or other members with R_{REG-LN} = 75 kΩ



(b) V_{LN} = 4,45 V (TEA1067 with $R_{REG-SLPE}$ = 39 kΩ and typical value for all other members)

- Speech mode: V_{LN} = 1,4 V_{rms} (THD < 2 %)
  V_{QR+}= 150 mV_{rms} across 150 Ω single ended load (THD < 2%)
- Mute mode: V_{LN} = 1 V_{rms}

Fig.10 TEA1060 Family Typical current I_p available from V_{CC} at I_{line} = 15 mA, (T_{amb} = 25°C)

Parallel operation with TEA1067

At line currents below the low-voltage threshold current I_{TH} (typically 9 mA), the internal reference voltage is automatically adjusted to a lower value. At 1 mA a typical voltage drop of 1,6 V is obtained. This facilitates operation of the circuit with more telephone sets connected in parallel, with line voltages down to a minimum of 1,6 V inside the polarity guard. The sending and receiving amplifiers will of course have reduced gain and reduced output swing in the low-voltage range and the peripheral supply will be reduced.

Supply for peripheral circuits

The voltage available at V_{CC} can be used to supply peripheral circuits such as a pulse dialler or a DTMF dialler. However, the current I_p and the voltage V_{CC} available from the circuit are dependent on the values of external components used with the IC and on the line current available.

Figure 10 (a) and (b) show the typical available current I_p as a function of V_{CC} with a line current of 15 mA. The typical available current I_p and the corresponding voltage V_{CC} as a function of line current are shown in Fig.11(a) and (b) for the speech mode, and in Fig.12(a) and (b) for the mute mode. Because the TEA1067 has a reduced line voltage compared with the rest of the TEA1060 family, the supply capabilities are also reduced, but with $R_{REG-SLPE}$ = 39 kΩ (resulting in V_{ref} = 4,2 V), the supply possibilities of the TEA1067 are the same as for the rest of the family.

The lowest power is available at minimum line current. At higher values of line current, the limit on I_p is imposed by the requirement to maintain at least the minimum permitted voltage between pins V_{CC} and SLPE (minimum instantaneous voltage; $V_{CC} - V_{SLPE}$ is at least 1,5 V). If this condition is not met, the maximum sending level on LN will be limited.

If 15 mA is the minimum line current under normal operating conditions, the available current I_p will be determined by the minimum supply voltage required for the peripheral circuits. For Philips CMOS telephony circuits, the minimum supply voltage is 2,5 V. With V_{ref}= 4,2 V, I_p will be typically 2,1 mA when V_{CC} is 2,5 V (worst case 1,75 mA). With V_{ref} = 3,6 V, I_p will be typically 1,25 mA when V_{CC} is 2,5 V (worst case 0,9 mA).

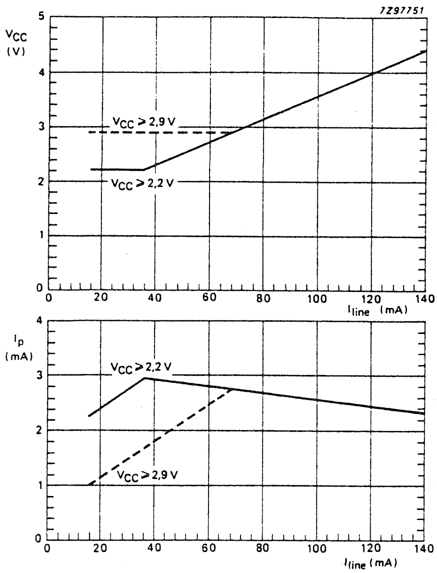


Fig.11 (a) Typical V_{CC} and I_p as a function of line current in the speech mode with V_{ref}=4,2 V. Signal conditions otherwise as in Fig.10.

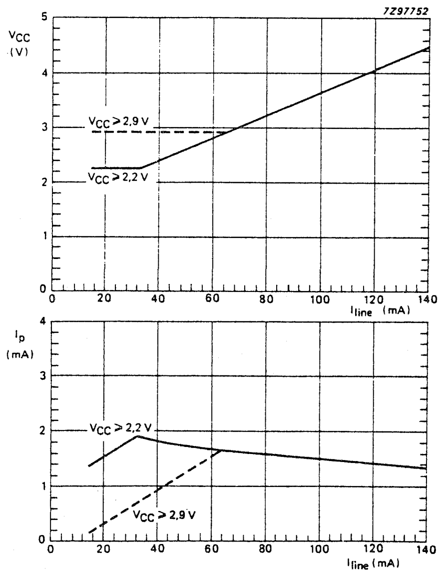


Fig.11 (b) Typical V_{CC} and I_p as a function of line current in the speech mode with V_{ref} = 3,6 V. Signal conditions otherwise as in Fig.10.

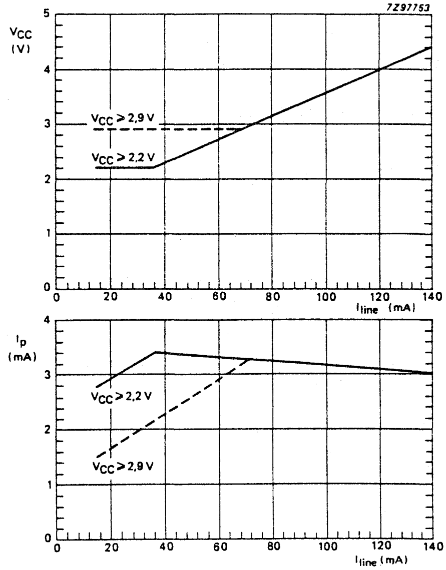


Fig.12 (a) Typical V_{CC} and I_p as a function of line current in the mute mode with V_{ref} = 4,2 V. Signal conditions otherwise as in Fig.10.

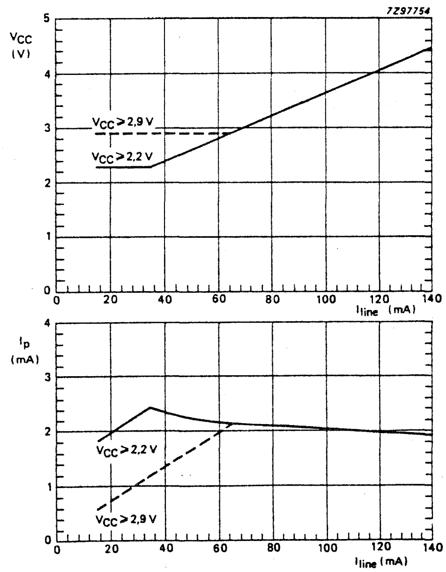


Fig.12 (b) Typical V_{CC} and I_p as a function of line current in the mute mode with V_{ref}=3,6 V. Signal conditions otherwise as in Fig.10.

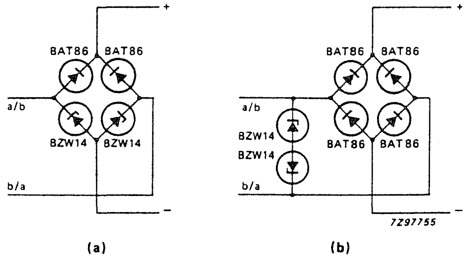


Fig.13 Schottky diode polarity guard with protection giving less voltage drop than the normal 1,4 V polarity guard.
(a) typically 0,5 V less drop.
(b) typically 1 V less drop.

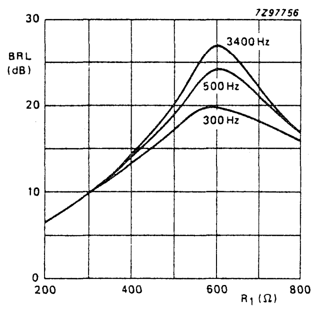
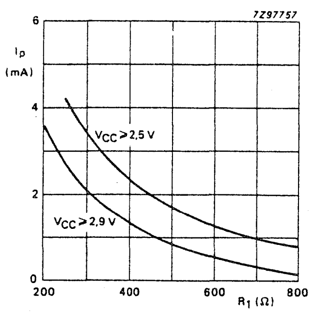


Fig.14 (a) Balance return loss as a function of R_1 (TEA1067)



I_{line} = 15 mA, V_{LN} = 3,9 V

Fig.14 (b) Typical supply current I_p as a function of R_1 in the mute mode with V_{ref} = 3,6 V (For the TEA1067 and the rest of the TEA1060 family with R_{LN-REG} = 75 kΩ)

In the speech mode, the available current is very much dependent on the received signal strength because of the class B receiving amplifier output stage. With an extremely high and continuous drive to the receiving amplifier, the available current will be typically 0,8 mA when V_{ref} is 3,6 V and 1,7 mA when V_{ref} is 4,2 V. In practice the receiving amplifier will not be driven continuously so the available supply current will be higher under normal speech conditions.

To prevent discharge of a battery used for memory retention, a diode should be connected between V_{CC} and the power pin of the peripheral circuit. Taking into account a forward voltage drop for a Schottky diode (BAT85; V_F is less than 0,32 V at 25 ^oC at 1 mA) the minimum value of V_{CC} is about 2,9 V. This results in a typical available current of 0,55 mA (worst case 0,2 mA) with V_{ref} = 3,6 V and 1,45 mA (worst case 1,1 mA) with V_{ref} = 4,2 V in the mute condition.

The power available from the supply point in the standard application under typical conditions is sufficient for low power circuits such as pulse diallers and the preamplifiers in electret microphones. Most of our CMOS DTMF diallers and microcontrollers can be powered, but under worst case conditions (e.g. long line using V_{ref} = 3,6 V (typical value for the TEA1067) and a tone dialler) the available current is not sufficient, so measures should be taken to increase the supply capabilities. Several ways to improve the supply capabilities of the TEA1067 (which also apply in general to the rest of the TEA1060 family) are given in the following paragraphs. The best choice depends on the requirements of the local PTT and the telephone manufacturer. In Ref.5 more details are given about how to meet the USA requirements (RS470) with the TEA1067 while solving the supply problem. The methods summarised below are more fully described in Ref.7.

Extending the supply capabilities

Increasing the line voltage

The supply voltage may be increased simply by setting the voltage drop across the circuit to a higher value. However, this will raise the line voltage and this can conflict with local PTT requirements. The line voltage can be reduced while maintaining a higher voltage across the IC, in DTMF dialling sets (without flash), by using a polarity guard with Schottky diodes (see Fig.13). For further information see Ref.5.

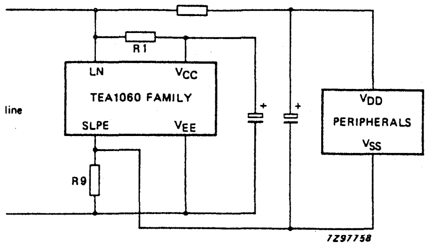


Fig.15 Increasing the supply capability by inserting an RC smoothing filter between LN and SLPE

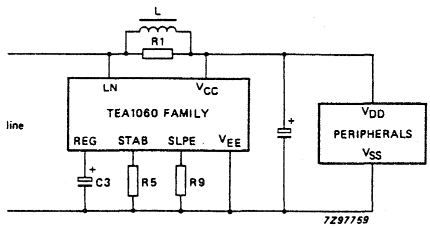


Fig.16 Increasing the supply capability by using an inductor in parallel with R_1

Compromise between set impedance and supply

The TEA1067 gives a very good balance return loss (BRL) with respect to a 600 Ω reference impedance. Where the value of BRL is better than the PTT requirement, a smaller value of R1 can increase the supply capabilities, while retaining a satisfactory BRL. Figure 14(a) shows the BRL as a function of R_1 and Figure 14 (b) shows the typical available supply current as a function of R_1 in the mute condition.

RC smoothing filter between LN and SLPE

An RC filter connected between LN and SLPE can be used to increase the peripheral supply current up to 3 mA (see Fig.15). This method provides a supply voltage which is independent of line current variation because it makes use of the internal reference voltage V_{ref}. However, several changes in the application are necessary which are more fully described in Ref.3 and Ref.8.

Inductor in parallel with R_1

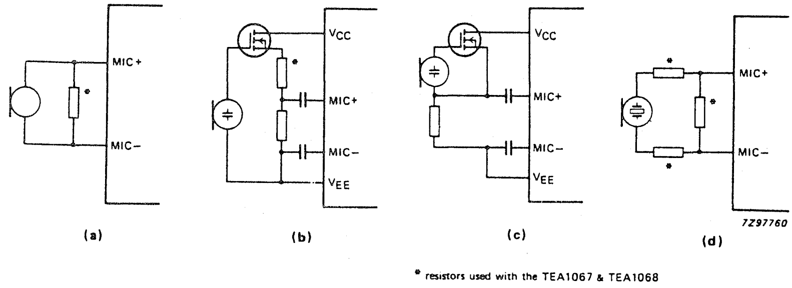
An inductor in parallel with R_1 will increase the supply capabilities, but it must have a value of more than 2,5 H to avoid influencing the BRL significantly (see Fig.16). As the size of the coil must increase with an increase in the current requirement, an electronic solution may be more desirable (next paragraph).

Electronic inductor using the TEA1080 supply circuit

The TEA1080 supply circuit can be connected either between LN and V_{EE}, or between LN and SLPE, to extend the peripheral supply capabilities of the TEA1060 family. This combination is recommended for loudspeaking listening-in and hands-free applications, where a relatively large power is needed. Reference 2 gives a description of the TEA1080, Ref.3 gives the application with the TEA1060 family and references 1 and 4 give application examples of listening-in and hands-free.

Parallel operation with a conventional set

If a conventional telephone set is connected in parallel with a member of the TEA1060 family, on a long loop with low line current, the line voltage will drop below the zener voltage of the voltage stabilizer for the transmission circuit. Only the TEA1067 will function at voltages below its zener voltage because it automatically decreases its zener voltage if the current drawn from the line drops below the threshold current I_{TH}.



* Resistors used with the TEA1067 and TEA1068
- (a) Dynamic or magnetic microphone
- (b) Electret-capacitor microphone with source follower
- (c) Electret-capacitor microphone with a preamplifier
- (d) Piezo-electret microphone

Fig.17 Methods of connecting microphones to the TEA1060 family of ICs

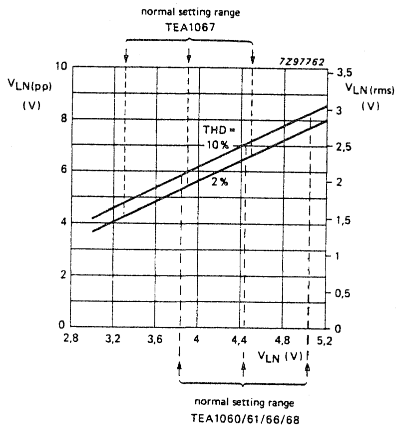
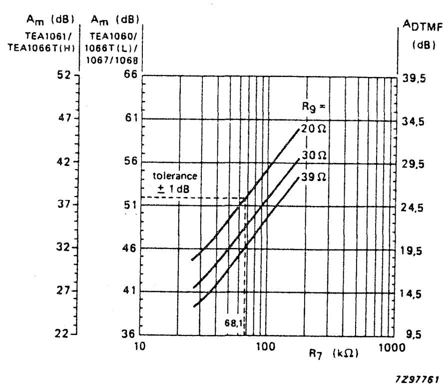


Fig. 18 Maximum swing of the transmit output stage as a function of d.c. voltage V_{LN} adjusted with R_{VA} (I_{line} = 15 mA)



R_1 = 620 Ω
R_L = 600 Ω

Fig.19 Microphone gain and DTMF gain as a function of the value of R_7.
Note: For the TEA1067 R_7 must not be more than 68,1 kΩ.

If a 200 Ω conventional set is connected in parallel with the TEA1067, with 20 mA available line current, the line voltage will drop to 3,2 V. This will leave 4 mA of line current at a voltage of 2 V for the TEA1067, inside the polarity guard, assuming that the current used by the peripherals can be neglected at such a low-voltage. This will normally be the case for pulse diallers, DTMF diallers in the standby mode and for microcontrollers (running on the battery). If the peripheral supply current is not negligible (e.g. a DTMF dialler in the operating mode) less current remains available for the TEA1067 resulting in a degraded performance. This can be avoided by switching the peripherals to a low power mode when there is insufficient line voltage.

MICROPHONE AMPLIFIER

Low-impedance microphone amplifier in the TEA1060 and TEA1066T

The TEA1060 and TEA1066T have a low-impedance (2 x 4 kΩ) symmetrical microphone amplifier input. If a lower input impedance is necessary, an external resistor can be added between pins MIC+ and MIC-. Dynamic or magnetic microphones can be connected as shown in Fig.17(a); the value of the resistor shown depends on the input impedance required. The microphone input stage accepts signals up to 17 mV for 2 % THD with internal soft limiting. The gain of the microphone amplifier, measured between the microphone inputs and the transmitting amplifier output LN, is:

$$A_m = 1{,}356 \times \frac{R_7 + r_d}{R_5\ R_9} \times \frac{R_i R_L}{R_i + R_L}$$

where...

$R_i = R_1//17{,}5$ kΩ (the dynamic resistance of the circuit)

R_L = load resistance at LN during measurement

$r_d = 3{,}47$ kΩ (the dynamic resistance of the internal circuitry)

For a practical circuit such as that shown in Fig.4, if we use the values $R_7 = 68{,}1$ kΩ, $R_5 = 3{,}65$ kΩ, $R_9 = 20$ Ω, $R_1 = 620$ Ω and $R_L = 600$ Ω, then:

$$20 \log A_{m(TEA1060)} = 52 \pm 1\text{dB}.$$

High-impedance microphone amplifier in the TEA1061 and TEA1066T

The TEA1061 and TEA1066T have a reduced sensitivity, high-impedance (2 x 20 kΩ) symmetrical, microphone amplifier input. The gain for this input is 14 dB lower than that of the TEA1060. The connection of an electret-capacitor microphone is shown with a source follower in Fig.17(b), and with a preamplifier in Fig.17(c). The connection of a piezo-electric microphone is shown in Fig.17(d). The microphone input stage accepts signals up to 85 mV for 2 % THD with internal soft limiting.

High-impedance microphone amplifier in the TEA1067 and TEA1068

The TEA1067 and TEA1068 have a high-impedance (2 x 32 kΩ), symmetrical amplifier input. The gain equation and the maximum input signal are the same as for the TEA1060 (for the TEA1067 $R_i = R_1//16,2$ kΩ). Dynamic, magnetic, piezoelectric or electret-condenser microphones can be used. The microphone arrangements are shown in Fig.17. To minimize noise, the microphone inputs must always be loaded. The equivalent noise voltage (psophometrically weighted; P53 curve) at the microphone inputs is typically 0,65 μV_{rms-p} with 8,2 kΩ across the inputs (0,45 μV_{rms-p} with 200 Ω across the input).

Asymmetrical drive of microphone amplifiers

If the microphone inputs are asymmetrically driven, care should be taken to ensure that the impedances between the common rail, and pins MIC+, and MIC- are equal. Otherwise any residual line signal present on the supply point (V_{CC}) will cause inaccuracy in gain. To prevent low frequency regeneration (motorboating) care should be taken that the value of the capacitor connected to MIC- is always smaller than that connected to MIC+ (including the tolerances of capacitors).

Gain adjustment

The gain of the microphone amplifier can be adjusted over a range of ±8dB (+0 and -8dB for the TEA1067) by means of R_7. Figure 19 shows the microphone gain as a function of R_7 and that the static resistance (R_9) directly influences the gain of the transmitting channel. Resistor R_5 (3,65 kΩ) determines the current in an internal current stabilizer and no alternative value is permissible. Any change in the value of R_9 will influence many other circuit parameters (except for the microphone gain) as described in "Supply and Set impedance" and in Appendix C.

Maximum output swing for the transmit output stage

Where the line current is sufficient, clipping of the bottom of the output signal at LN can occur when the internal output transistor saturates (V_{LN} - V_{SLPE} = 0,9 V). The top of the output signal can be clipped because of a lack of collector current in the output transistor. Symmetrical clipping can be obtained by connecting a zener diode between pins LN and SLPE the value of which is dependent on the d.c. reference voltage setting (Fig.4). Figure 18 shows the maximum output swing of the transmit output stage as a function of the d.c. line voltage V_{LN} with a line current of 15 mA.

Stability and frequency roll-off

A 100 pF capacitor C_6, connected between pins GAS1 and SLPE, is necessary to ensure the stability of the microphone amplifier. Higher values of C_6 can be used to filter off high frequency signals (the cut-off frequency is determined by the time constant R_7C_6). For example, if R_7 = 68.1 kΩ and C_6 = 220 pF, the cut-off frequency, $f_{3\ dB}$, is about 10 kHz.

Parallel operation with the TEA1067

In the case of parallel operation of sets, the operating voltage of the TEA1067 can drop below the internal reference voltage, but the circuit will adapt the internal reference voltage. This will influence the performance of the microphone amplifier.

Figure 20 shows the maximum output voltage at pin LN as a function of the line current flowing into the TEA1067, with a 600 Ω impedance telephone set connected in parallel with the circuit shown in Fig.4. Transmitting gain is 52 dB in the case of a normal 600 Ω load; however, with a 600 Ω telephone set connected in parallel, the gain decreases to about 48,5 dB. The maximum output swing at low line current is not determined by the d.c. voltage at pin LN but by the available current in the output stage of the TEA1067.

Figure 21 shows the transmit gain as a function of the d.c. voltage at pin LN (V_{LN}). The reduction in gain starts at V_{LN} = 2,2 V. At V_{LN} = 2 V the reduction is about 2 to 3 dB, and about 12 dB at V_{LN} = 1,6 V. The results are valid for a typical sample in the basic application circuit of Fig.4. Changing component values will influence the results.

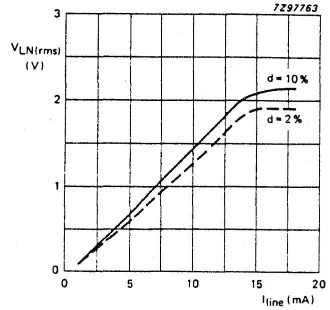


Fig.20 Maximum output voltage (TEA1067) of the transmitting output stage as a function of I_{line} in low line current range

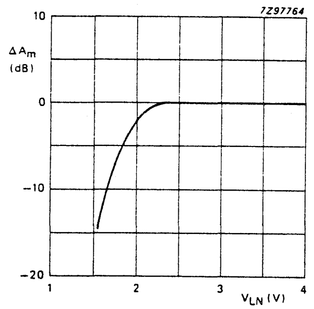


Fig.21 Typical transmit gain (TEA1067) as a function of d.c. voltage V_{LN} in low voltage range.

DTMF AMPLIFIER

A dual-tone multi-frequency dialling signal can be applied to the IC's DTMF input which has an impedance of 20 kΩ. The DTMF gain depends on the values of R1, R5, R7, R9 and R_L in the same way as the microphone gain (see Fig.19). The voltage gain between the DTMF input and the transmitter output at pin LN is 26,5 dB less than the microphone gain of the TEA1060. For a microphone gain of 52 dB for a TEA1060, the DTMF gain is 25,5 dB but the choice of gain to suit another particular microphone capsule will alter the DTMF gain. The dialling tones must therefore be adjusted to the appropriate level before they are fed to the DTMF amplifier. The DTMF input accepts signals up to 170 mV r.m.s. for 2 % THD.

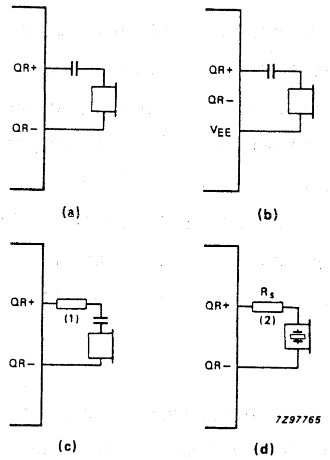
Temperature dependence

The DTMF amplifier is internally temperature-compensated. However, because it is asymmetrically driven, some influence can be expected from the residual a.c. line voltage on supply pin V_{CC}, the level of which depends on the effectiveness of the supply voltage filter R_1C_1 (especially on the value of C_1). The temperature-dependence of C_1 has some influence on the DTMF gain via an internal feedback mechanism, so it should have a low-temperature coefficient. The DTMF gain deviation was measured in the basic application circuit of Fig.4 with a 100 μF, 25 V capacitor (Catalogue no. 2222 030 36101) with respect to a nominal 25 °C at the following temperatures:

- -0,5 dB at -25 °C
- -0,2 dB at -10 °C
- +0,1 dB at +55 °C
- +0,2 dB at +70 °C

RECEIVING (EARPIECE) AMPLIFIER

The input of the receiving amplifier is pin IR. The receiving amplifier has two complementary Class B outputs; non-inverting output QR+, and inverting output QR-. The output can either be a single-ended or differential configuration (see Fig.22), depending on the application.



(a) Dynamic earpiece with Z_t > 450 Ω
(b) Dynamic earpiece with Z_t < 450 Ω
(c) Magnetic earpiece with Z_t > 450 Ω
(d) Piezoelectric earpiece

Fig.22 Connection of earpieces:

Z_T = 450 Ω

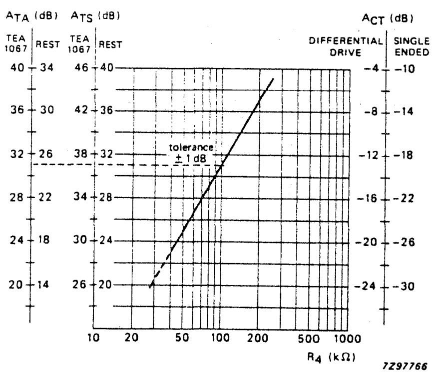


Fig.23 Gain of the receiving amplifier (A_{TA} and A_{TS}) and confidence tone as a function of R_4. The broken region of the graph is applicable only to the TEA1067.

The receiving amplifier can drive dynamic, magnetic, or piezoelectric earpieces. A low-impedance dynamic or magnetic earpiece, with an impedance up to 450 Ω, must be driven in the single-ended configuration as shown in Fig.22(b). It is possible to use differential drive to achieve maximum efficiency with dynamic, magnetic or piezoelectric earpieces which have impedances of more than 450 Ω (see Fig.22(a), (c) and (d)). To prevent distortion of the output signal in the event of the output stage running out of current with an inductive load (magnetic earpiece), an additional resistor ((1) in Fig.22(c)) may be required.

A piezoelectric earpiece presents a capacitive load. The maximum permitted capacitive load between pins QR+ and QR- is 100 nF, but the decrease in phase margin must be compensated by a series resistor ((2) in Fig.22(d)) (for example if C_L = 100 nF, R_S must be 50 Ω). This is described in more detail in Ref.6.

With an asymmetrical load, the gain (A_{TA}) of the receiving amplifier, between the input IR and the output QR+ is:

$$A_{TA} = 0{,}657 \times \frac{R_4}{R_5} \times \frac{Z_T}{Z_T + r_o} \qquad \text{(plus 6 dB for TEA1067)}$$

where:

Z_T = impedance of the earpiece,

r_o = output impedance of receiving amplifier ($\simeq$ 4 Ω).

For the values R_4 = 100 kΩ, R_5 = 3,65 kΩ, and Z_T = 450 Ω:

$20 \log A_{TA} = 25 \pm 1$ dB (For TEA1067 31 ± 1 dB)

If both outputs QR+ and QR- are used for drive, the gain A_{TS} is increased by about 6 dB:

$$A_{TS} = 1,314 \times \frac{R_4}{R_5} \times \frac{Z_T}{Z_T + 2r_o} \qquad \text{(plus 6 dB for the TEA1067)}$$

If we insert the values for R_4, R_5 and Z_T which were used above, this gives:

$$20 \log A_{TS} = 31 \pm 1 \text{ dB} \qquad \text{(For TEA1067 37} \pm 1 \text{ dB)}$$

The gain of the receiving amplifier can be adjusted over a range of ± 8 dB (-11 and +8 dB for the TEA1067) by means of R_4. Figure 23 shows the gain (A_{TA} and A_{TS}) and confidence tone (A_{CT}) as a function of the value of R_4, for both symmetrical and asymmetrical drives. The total receive gain between line and earpiece can be found by subtracting the attenuation of the anti-sidetone circuit circuit (32 dB) from A_{TA} or A_{TS}.

The signal received on the line is attenuated by the anti-sidetone network, before it enters the amplifier. In the basic application, this attenuation is about 32 dB and almost flat over the whole audio frequency range when using the TEA1060 family anti-sidetone bridge. The amplifier input signal IR, is symmetrically soft-limited internally, to 34 mV r.m.s. (17 mV for the TEA1067) for 2 % THD, and to 106 mV r.m.s. (53 mV for the TEA1067) for 10 % THD.

The equivalent noise at input IR of the receiving amplifier (psophometrically weighted; P53 curve) is typically 2 μV_{rms-p} (1,25 μV_{rms-p} for the TEA1067). With the anti-sidetone circuit connected to the input the noise generated at the line pin LN will add, via the anti-sidetone circuit, to the equivalent input noise of the receiving amplifier. The total noise generated at the earpiece output depends on microphone gain that has been set, and on the actual sidetone suppression. Any additional circuitry connected to pin LN (for example an artificial inductor to extend the peripheral supply capability) can give a noise contribution.

Stability and frequency roll-off

Stability is ensured by two capacitors C_4 and C_7 (see Fig.4) connected between pins QR+ and GAR, and between pins GAR and V_{EE} respectively. The value of C_7 must be ten times that of C_4. Generally C_4 is 100 pF and C_7 is 1 nF. Larger values of C_4 can be applied to filter off high frequency signals. The cut-off frequency is determined by the time constant R_4C_4. For example, if C_4 = 150 pF and R_4 = 100 kΩ, the cut-off frequency, $f_{3\ dB}$, is about 10 kHz, and C_7 must be 1500 pF.

Maximum output swing and parallel operation with the TEA1067

As with the microphone amplifier, the performance of the receiving amplifier is dependent of the d.c. line voltage V_{LN}. Figure 24 shows the maximum output swing of the receiving output stage as a function of d.c. line voltage V_{LN} for I_{line} = 15 mA. The maximum output swing of the receiving output stages depends on the load and on the d.c. voltage drop across the circuit.

As discussed earlier, the d.c. line voltage is considerably influenced by a parallel connected telephone set and will result in a degraded performance of the receiving amplifier. Figure 25 shows the maximum output swing for three load values, versus the d.c. voltage drop in the low voltage range of V_{LN} (d_{tot} = 10%). With a line voltage V_{LN}, of 2 V, an output swing of 15 mV_{rms} with a 150 Ω load, can be attained. At about 1,6 V, the receiving amplifier is totally cut-off.

Figure 26 shows the receiving amplifier gain as a function of the line voltage V_{LN}. Gain decrease starts when V_{LN} is about 3 V, and by 2 V, the gain has been decreased by about 13 dB. The results are valid for a typical sample in the basic application circuit of Fig.4. Changing components will have influence on the results.

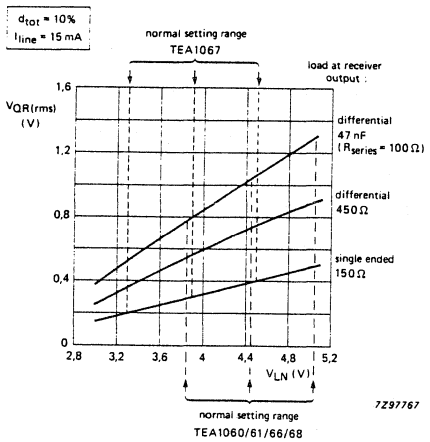


Fig. 24 Maximum output swing receiving amplifier as a function of voltage V_{LN} adjusted with R_{VA}

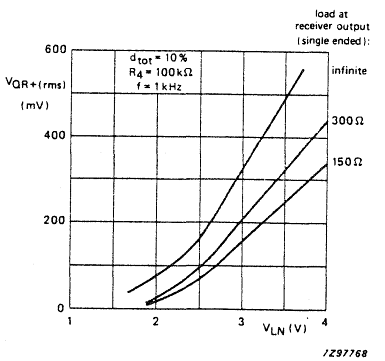


Fig.25 Maximum output swing receiving amplifier as a function of V_{LN} in low voltage range.

Confidence tone

During DTMF dialling, the dialling tones can be heard at a low-level in the earpiece. The volume of the dialling tone depends both on the gain of the receiving amplifier and on the level of tone applied to the DTMF input. The gain A_{CT} between the DTMF input and the telephone output is:

$$20 \log A_{CT} = 20 \log A_T - 44 \text{ dB} \quad (20 \log A_T - 50 \text{ dB for TEA1067})$$

where A_T is a general term for telephone gain and can be replaced by either A_{TA} for the gain with the asymmetric load, or by A_{TS} for the gain with the symmetric load. This is shown in Figure 23.

LINE-CURRENT-DEPENDENT GAIN CONTROL

The gains for the microphone amplifier and the receiving amplifier which were derived in the preceeding sections are applicable only when the a.g.c. is inoperative; that is, with pin AGC open-circuit. When the resistor R_6 is connected between pins AGC and V_{EE}, the line current dependent gain control of both the microphone amplifier and the receiving amplifier become operative without affecting the DTMF amplifier.

Below a specific value of line current ($I_{line-start}$) the gain is equal to the values calculated with the equations given earlier. If $I_{line-start}$ is exceeded, the gain of both controlled amplifiers decreases with increasing d.c. line current. Gain control stops when another value of line current ($I_{line-stop}$) is exceeded. The gain control range of both amplifiers is typically 6 dB. This corresponds to a line length of 5 km of 0,5 mm diameter copper twisted-pair cable which has a resistance of 176 Ω/km with an average ac attenuation of 1,2 dB/km. The slope of the gain control characteristic has been chosen for optimum tracking between the line attenuation and the required amplifier gain (typical tracking error 0,8 dB max.) for a system with a 2 x 300 Ω feeding bridge. In the case of lines with other parameters, a small additional tracking error will be introduced.

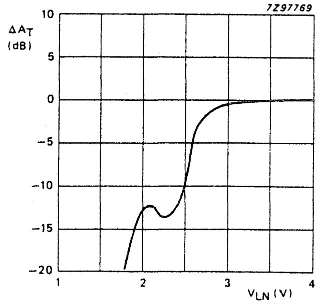


Fig.26 Typical receive gain as a function of V_{LN} in low voltage range.

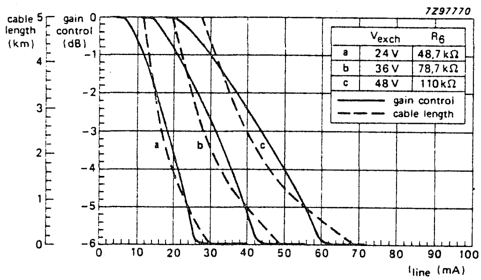


Fig.27 Gain Control characteristics with a 600 Ω feeding bridge, R_9 =20 Ω and cable consisting of 0,5 mm diameter copper twisted pair, having resistance of 176 Ω/km, and an average a.c. attenuation of 1,2 dB/km The 24 V curve is not applicable to the TEA1067.

Correction for variation in exchange supply voltage

The value of resistor R_6 must be chosen to suit the exchange supply voltage. In Fig.27, the control curves are shown for a V_{exch} of 24 V, 36 V and 48 V with a feeding bridge resistance of 2 x 300 Ω. The relationship between line length and line current is also shown in Fig.27. These 'ideal' curves have been calculated on the assumption that the voltage drop across the circuit has been set for a V_{LN} of 4,45 V at 15mA ($R_{REG-SLPE}$ = 39 kΩ for the TEA1067; typical value for the rest of the family) and assuming a polarity guard with 1,4 V voltage drop. Other parameters will give slightly different optimum values for R_6.

Correction for resistance of feeding bridge

The value of resistor R_6 must be changed if the feeding bridge has a resistance other than 600 Ω. This will slightly increase the automatic gain control tracking error ($<$1,2 dB) because the a.g.c. characteristic has been optimized for a 600 Ω feeding bridge. Figure 28 shows the control characteristics of a 400 Ω feeding bridge with exchange supply voltages of 24 V, 36 V, and 48 V. Figure 29 and Fig.30 respectively show the characteristics of an 800 Ω bridge and a 1 kΩ bridge at 36 V, 48 V and 60 V.

The optimum values for R_6 with various values of exchange supply voltage and exchange feeding bridge resistance are shown in the following table.

	R_{exch} (Ω)			
	400	600	800	1000
V_{exch}(V)	R_6 (kΩ) with R_9 = 20 Ω			
24*	61,9*	48,7*	-	-
36	100	78,7	68,0*	60,4*
48	140	110	93,1	82,0
60	-	-	120	102

* not for TEA1067

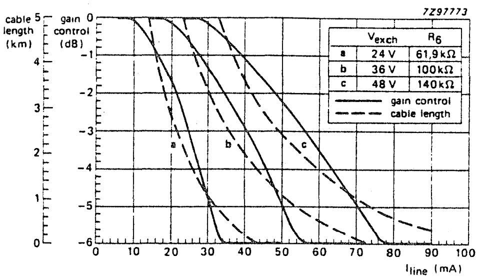


Fig.28 Gain control characteristics, with 400 Ω feeding bridge.
Other parameters as defined in Fig.27.
The 24 V curve is not applicable to the TEA1067.

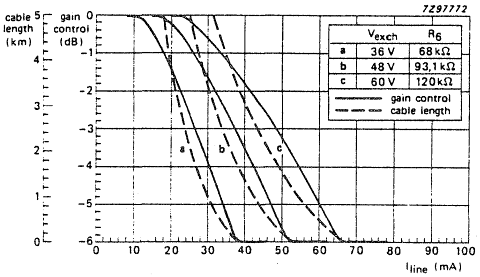


Fig.29 Gain control characteristics with 800 Ω feeding bridge.
Other parameters as defined for Fig.27.
The 36 V curve is not applicable to the TEA1067.

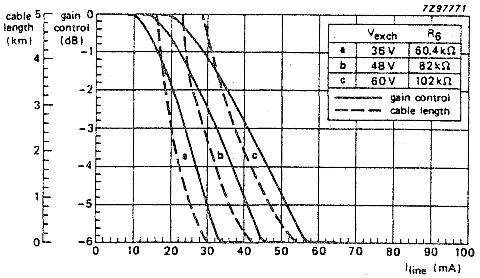


Fig.30 Gain control characteristics with 1000 Ω feeding bridge.
Other parameters as defined in Fig.27.
The 36 V curve is not applicable to the TEA1067.

MUTE INPUT

The MUTE input is used to switch between the dialling and speech modes. If MUTE is HIGH ($\geqslant$ 1.5 V $\leqslant$ 15 µA), both the microphone and receiving amplifier inputs are inhibited, and the DTMF input enabled. The converse situation is obtained if MUTE is LOW ($\leqslant$ 0,3 V) or open circuit. Switching causes negligible clicking at the earpiece and on the line.

For the TEA1067, if the supply voltage at V_{CC} drops below 2 V (as with no external load at V_{CC}, V_{LN} < 2,5 V and I_{line} < 6 mA) the mute function becomes inoperative and signals applied to either the microphone inputs or the DTMF input will be put on to the line. Normally, dialling will not take place under these low-voltage conditions which only occur during parallel operation of sets under worst-case (end of long line) conditions.

POWER-DOWN INPUT

The power-down input PD, is for use in pulse dialling and in register recall applications (where the telephone line current is interrupted, leaving the set without continuous power). During these interruptions, the transmission IC and the peripheral circuits must be supplied by the charge stored in the V_{CC} smoothing capacitor C_1. The discharge time of this capacitor may be increased if the power-down function is used and the ripple on V_{CC} will be reduced.

When input PD is HIGH ($\geqslant$ 1,5, $\leqslant$ 10 µA), the internal supply current I_{CC} is reduced from 1 mA to typically 55 µA at V_{CC} = 2,8 V. Furthermore, the voltage regulator capacitor C_3 at pin REG is internally disconnected to prevent it from being discharged during line interruptions. This allows the voltage regulator to start without delay after each line interrupt, at the same d.c. line voltage as before the interruption and so minimizes the distortion of the current pulses during pulse dialling. Of course, with a highly inductive exchange feeding bridge, the inductors mainly determine current waveform. Under these conditions the voltage regulator (active circuit) may have some switch-on delay and cause a voltage overshoot at the line connection pin LN of the IC.

If the voltage drop across the circuit is increased by means of $R_{REG-SLPE}$, the power down function will be affected. This will change the shape of the current pulses.

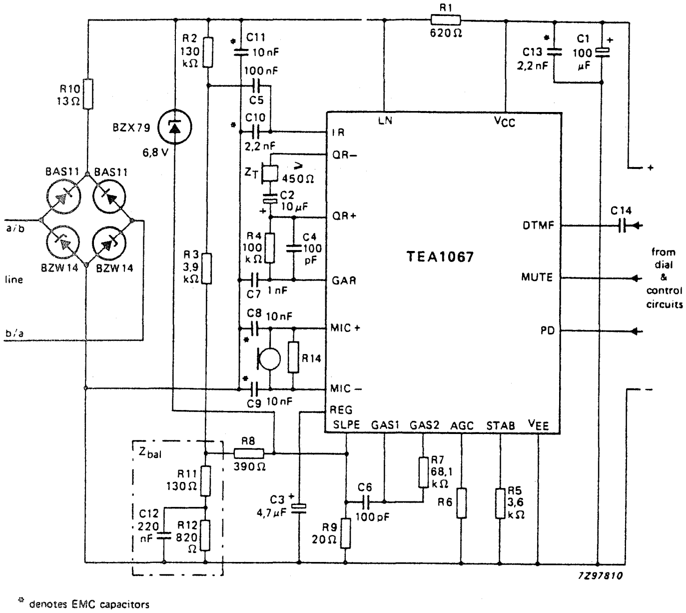


Fig.31 Basic application diagram of TEA1067 in sets with DTMF dialling. Protection is also shown; RFI capacitors are marked with a *.

IMMUNITY TO R.F. SIGNALS

In the presence of high-intensity electromagnetic fields, it is possible for common-mode, amplitude modulated, r.f. signals to be induced in the telephone lines. These common-mode signals can sometimes become differential-mode signals as a result of asymmetrical parasitic capacitance to ground (for example, through the hand of the subscriber holding the handset). Preventive measures must be taken to avoid the possibility of these signals being detected and appearing as an unwanted signal at the earpiece or on the line.

Small discrete capacitors can help suppress spurious r.f. signals before they enter the circuit. In Fig.31, C_8 and C_9 at the microphone inputs, C_{10} at the receiver amplifier input IR, C_{13} at the supply point V_{CC}, and C_{11} at the positive line terminal LN are included for this purpose. All the capacitors are connected to the common V_{EE}.

The layout of the pcb has a strong influence on r.f. immunity. The copper ground area should be as large as possible. Earth loops must be avoided and print tracks kept as short as possible. RFI suppression capacitors should be mounted as close as possible to the IC pins. In practice, two inductors of between 68 µH and 1 mH, connected in series with the telephone lines, will improve the r.f. immunity considerably and a closed copper ´guard´ ring around the circuit will give adequate protection from magnetic fields.

Because the TEA1067 and TEA1068 have very high impedance microphone inputs, a low-pass filter can be connected in series with both microphone inputs without affecting gain accuracy. It should be mounted as close as possible to pins MIC+ and MIC-.

A low resistance termination, across the microphone inputs, will reduce pick-up of unwanted r.f. via the handset cord.

TRANSIENT SUPPRESSION AND THE POLARITY GUARD

Unprotected speech/transmission ICs might be destroyed by excessive current surges on the telephone lines if preventive measures are not taken. Those sets with only DTMF dialling require different protection to those with only pulse dialling or DTMF dialling with 'flash' (register recall by means of a timed line interruption).

With DTMF dialling only, the bridge rectifier (Fig.31) which normally acts as a polarity guard, also incorporates two transient suppressor diodes (such as BZW14). Under normal operating conditions, one of the two transient suppressor diodes conducts and the other is reversed-biased. If the voltage across the set temporarily exceeds the reference voltage of the previously reversed-biased diode, it will conduct and limit the voltage across the set. The maximum permissible continuous voltage across a member of the TEA1060 family is 12 V. During switch-on and line interruptions the maximum permissible voltage is 13,2 V allowing the use of a 12 V voltage reference diode in the polarity guard.

Further protection is provided by R_{10} in series with the bridge rectifier which limits the current that can be drawn by the IC. The maximum transient voltage allowed across the circuit including the protection resistor (R_{10} = 13 Ω and R_9 = 20 Ω) is 28 V for 1 ms with a repetition time of 5 s. This corresponds to a 50 A surge on the BZW14 zener diodes, used in the polarity guard.

For DTMF dialling with flash, or for pulse dialling, a different protection arrangement is necessary. The line current must be zero during line interruptions, so the bridge rectifier must be able to withstand a voltage of around 200 V. A polarity guard using four BAS11 diodes is suitable for this purpose. Protection against line current surges can then be obtained using a VDR connected directly between the telephone lines (See Fig.45).

The speech circuit can be protected by a 12 V zener diode connected between pins LN and V_{EE}. Alternatively, if a current limiter is used (e.g. combined with the interruptor) the 8,2 V (6,8 V for the TEA1067) voltage-regulator diode connected between pins LN and SLPE will protect the speech circuit. The latter method also provides symmetrical clipping of the sending signal.

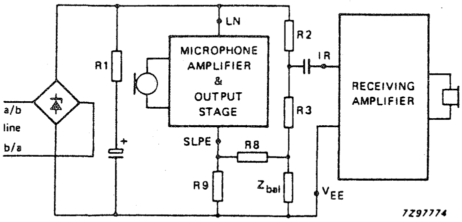
ANTI-SIDETONE CIRCUIT

To avoid the reproduction of microphone signals in the telephone transducer, an anti-sidetone circuit uses the microphone signal from pin SLPE to cancel the microphone signal at the input of the receiving amplifier (IR). The TEA1060 family anti-sidetone bridge or the conventional Wheatstone bridge may be used as the basis for the design of the anti-sidetone circuit, see Fig.32.

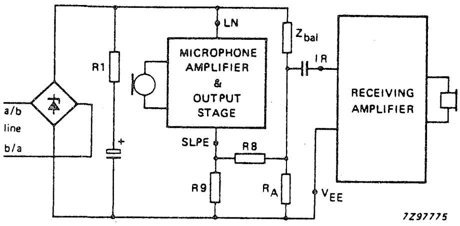
The TEA1060 family anti-sidetone bridge has the advantage of a relatively flat transfer function in the audio-frequency range between pins LN and IR, both with real and complex set impedances. Furthermore, the attenuation of the bridge for the received signal is independent of the value chosen for Z_{bal} after the set impedance has been fixed and the condition $R_9R_2 = R_1(R_3 + (R_8//Z_{bal}))$ is met. Therefore, readjustment of receive gain is not necessary in many cases.

The Wheatstone bridge has the advantages of needing one resistor fewer than the TEA1060 family anti-sidetone bridge and requires only a small capacitor (about 10 nF) in Z_{bal}. Calculation of the values is easier. The disadvantages include the dependence of the attenuation of the bridge on the value chosen for Z_{bal} and the frequency dependence of that attenuation. This necessitates a readjustment of the receive gain.

Appendix A gives a more detailed comparison of the TEA1060 family anti-sidetone bridge and the Wheatstone bridge. Either bridge can be used with a real or a complex set impedance.

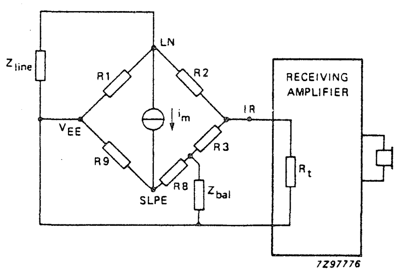


(a) TEA1060 family bridge

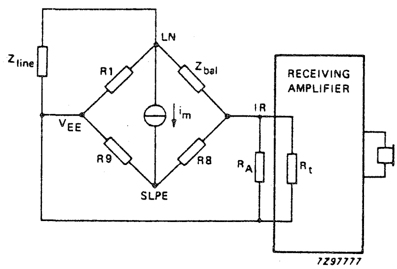


(b) Wheatstone bridge

Fig.32 Anti-sidetone circuits



(a) TEA1060 family bridge anti-sidetone circuit



(b) Wheatstone bridge anti-sidetone circuit

Fig.33 Equivalent circuits

TEA1060 family anti-sidetone bridge

The equivalent circuit of the TEA1060 family anti-sidetone bridge is shown in Fig.33(a). Optimum suppression of the sidetone signal occurs when:

(a) $R_9R_2 = R_1(R_3 + (R_8//Z_{bal}))$

(b) $\dfrac{Z_{bal}}{Z_{bal} + R_8} = \dfrac{Z_{line}}{Z_{line} + R_1}$

If the correct fixed values are chosen for R_1, R_2, R_3, and R_9, then condition (a) will always be fulfilled provided that $|R_8//Z_{bal}| << R_3$.

To obtain the optimum sidetone suppression (b) has to be fulfilled such that:

$$Z_{bal} = \frac{R_8}{R_1} Z_{line} = k\ Z_{line}$$

where k is the scale factor : $k = \dfrac{R_8}{R_1}$

The scale factor must be chosen to meet the following criterea:

- compatability with a standard capacitor from the E_6 or E_{12} range for Z_{bal}
- $|Z_{bal}//R_8| << R_3$ necessary to fulfill condition (a) to ensure correct operation of the anti-sidetone circuit
- $|Z_{bal} + R_8| >> R_9$ to avoid influencing the microphone gain

In practice Z_{line} varies considerably with the length and type of line. Consequently a value for Z_{bal} has to be chosen that corresponds with an average line length giving satisfactory sidetone suppression with short and long lines. The suppression also depends on the accuracy with which Z_{bal} matches this average line impedance.

In the basic application circuit, Fig.31, Z_{bal} has been optimized for a line length of 5 km, 0,5 mm diameter copper twisted-pair with an average attenuation of 1,2 dB/km, a d.c. resistance of 176 Ω/km and a capacitance of 38 nF/km. The approximate equivalent impedance is shown in Fig.34. The scale factor k has been chosen in accordance with the above criteria, resulting in the following practical values k = 0,636, R_{11} = 130 Ω, R_{12} = 820 Ω, C_{12} = 220 μF and R_8 = 390 Ω. With the line current dependent gain control activated a roughly equal sidetone level results (acoustically measured) for telephone sets connected directly to the exchange and connected with a line of 10 km. Where no a.g.c. is used, the anti-sidetone has to be optimized for a shorter line length, in order to obtain acoustically equal sidetone levels for a set connected with a very short line and a set connected with a 10 km line to the exchange. (Practical values for a line length of 2 km are: R_{11} = 130 Ω, R_{12} = 1 kΩ, C_{12} = 100 nF and R_8 = 620 Ω). The overall sidetone suppression is worse in this case compared with the situation where a.g.c. is activated. In practice a compromise is chosen between the transmit and receive gain and the sidetone level; sending and receiving gain will therefore be reduced.

The attenuation of the received line signal between pins LN and IR is:

$$\frac{V_{IR}}{V_{LN}} = \frac{R_t//R_3}{R_2 + (R_t//R_3)}$$

where R_t is the input impedance of the receiving amplifier (typically 20 kΩ). This attenuation is about 32 dB for the basic application circuit shown in Fig.31. Frequency dependence of the input attenuation is negligible over the audio frequency range. However, to prevent high frequency components from entering the receiving amplifier, a frequency roll-off can be obtained by connecting a capacitor between pins IR and V_{EE}.

Complex set impedance

A complex network can be used instead of R_1. Normally the bridge can be rebalanced by readjusting the values of R_8 and Z_{bal}, and either R_2 or R_9. Changing R_9 affects many other parameters and the range of possible values is limited, so the design procedure given in Appendix C should be considered. Changing R_2 influences the attenuation of the received signal between pins LN and IR. This necessitates readjustment of the receiving gain. Note that changing R_1 also influences the peripheral supply capabilities.

In some cases, calculating the optimum condition is not very useful because a compromise must be made to meet the sidetone requirement in several conditions. In these cases, an empirical method is more practical and probably faster. This involves performing acoustic measurements and changing the values of Z_{bal} and R_8 until the requirements are met. A more detailed analysis of the TEA1060 family anti-sidetone bridge is given in Appendix A.

Wheatstone bridge

In the Wheatstone bridge (equivalent circuit shown in Fig.33(b)) optimum sidetone suppression is given by:

$$Z_{bal} = \frac{R_8}{R_9} \times \frac{R_1\ Z_{line}}{R_1 + Z_{line}} \quad \text{provided that } R_8 / R_9 >> 1$$

Also for this bridge type a value for Z_{bal} has to be chosen to correspond with the average line length. The attenuation of the received line signal between pins LN and IR is given by:

$$\frac{V_{IR}}{V_{LN}} = \frac{R_8//R_t//R_A}{Z_{bal} + (R_8//R_t//R_A)}$$

where R_t = input impedance of the receiving amplifier at pin IR (typically 20 kΩ)

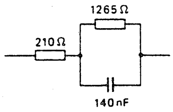


Fig.34 Equivalent line impedance for optimum sidetone suppression.

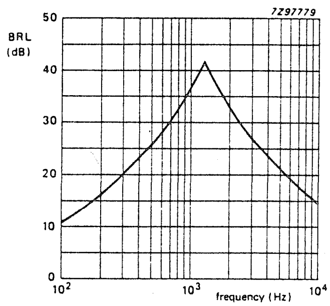


Fig.35 Balance Return Loss as a function of frequency.

$$BRL = 20 \times 10 \log \left| \frac{Z + Z_o}{Z - Z_o} \right| \quad \text{with } Z_o = 600\ \Omega$$

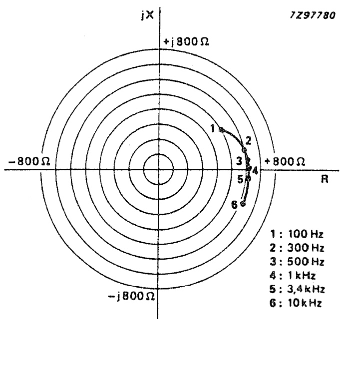


Fig.36 Polar plot of impedance between telephone lines.

A practical circuit could have the following values; R_8 = 820 Ω, R_1 = 620 Ω and Z_{bal} optimized for the line impedance as shown in Fig.34 with R_A = infinite and a 600 Ω load at the line, the attenuation varies typically from about 24 dB to 27,5 dB over the audio-frequency range; the lower attenuation occurs at the upper frequencies. R_A is used to adjust the bridge attenuation; its value doesn´t influence the balance of the bridge.

If complex set impedances are used with the Wheatstone bridge, it can be rebalanced by altering the values of Z_{bal}. However the frequency dependence of the transfer function between pins LN and IR will increase. A more detailed description is given in Appendix A.

HINTS FOR PRINTED CIRCUIT BOARD LAYOUT

To avoid heavy current flow through p.c.b. tracks near sensitive input tracks, resistors R_9 and R_6 must be close to pin V_{EE}, and the ground connection of the earpiece should be at a point where no large line current is flowing. The copper tracks connecting R_7 and R_4 to the IC should be as short as possible. The ground connection of all RFI capacitors should be large and as close as possible to the pins that have to be decoupled. The ground plane must also be as large as possible.

PERFORMANCE

The following measurements have been made with the basic application circuit (DTMF dialling only) including RFI capacitors as shown in Fig.31. This gives an indication of the performance of the TEA1060 family. A typical TEA1067 sample was used.

BALANCE RETURN LOSS (BRL)

The result of the BRL measurement is shown in Fig.35. The impedance of the circuit is shown in Fig.36. Different values chosen for C_3 and for R_9 will have influence on the impedance and the BRL of the circuit as well as some other parameters.

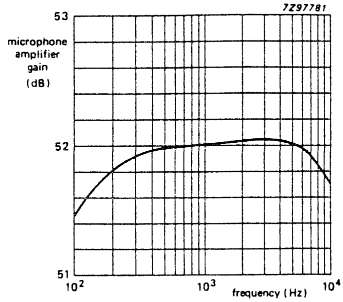


Fig.37 Frequency characteristic of microphone amplifier.

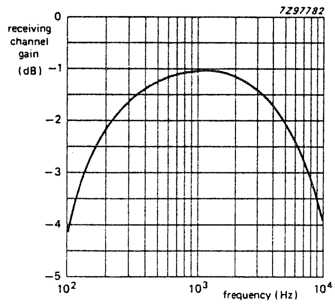


Fig.38 Frequency characteristic of the receiving channel.

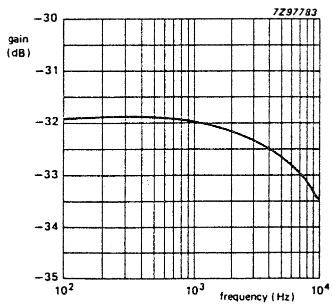


Fig.39 Frequency characteristic of the anti-sidetone circuit between pins LN and IR.

FREQUENCY CHARACTERISTICS

Figure 37 shows the frequency-response of the sending channel measured between the microphone inputs and the transmitter output at pin LN with a 600 Ω load. The microphone gain is set by R_7 to 52 dB (R_7 = 68,1 kΩ). The upper cut-off frequency is about 24 kHz (mainly determined by the time-constant R_7C_6). Note that if a complex set impedance is chosen, this will influence the frequency characteristic.

Figure 38 shows the frequency response of the receiving channel measured between pin LN and the QR+ output loaded with 150 Ω (single ended drive; 10 μF d.c. blocking capacitor). With R_4 = 100 kΩ the transfer ratio is -1 dB at 1 kHz. The lower cut-off frequency (120 Hz) is determined by load resistor R_L and the d.c. blocking capacitor C_2. The upper cut-off frequency (about 9,5 kHz) is determined partly by R_4C_4 (15 kHz) and partly by the cut-off frequency of the anti-sidetone circuit (18 kHz).

The frequency response of the anti-sidetone circuit (LN to IR) is given in Fig.39. The cut-off frequency is about 18 kHz. This is determined by the 2,2 nF capacitor between pins IR and V_{EE} (necessary for r.f. suppression).

The transfer ratio as a function of frequency measured from the microphone inputs to a 150 Ω asymmetric load at the receive output QR+ (10 μF d.c. blocking capacitor) is shown in Fig.40. This represents the electrical sidetone. The sending signal at pin LN is also shown. Furthermore the receive signal at the earpiece is shown with the same signal at pin LN in the receiving condition. The difference between the wanted receive signal and unwanted sidetone signal at the earpiece, with equal levels on pin LN in both the receive and the sending conditions, is the electrical sidetone suppression. So for this application, the electrical sidetone suppression for telephone sets connected directly to the exchange is about 7,3 dB at 1 kHz. The result depends mainly on the balance of the anti-sidetone circuit. In this case, the balance impedance Z_{bal} has been optimized for a 5 km line with 0,5 mm diameter, 176 Ω/km and 38 nF/km.

Electrical sidetone suppresion is not dependent on whether or not gain control is used because both amplifiers (microphone and receive) are affected by the gain control function.

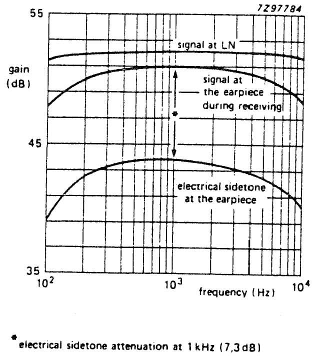


Fig.40 Frequency characteristic of the electrical sidetone for telephone sets connected to the exchange by a very short line. The distance between the curve of the signal at the earpiece and the electrical sidetone at the earpiece gives the electrical sidetone attenuation.

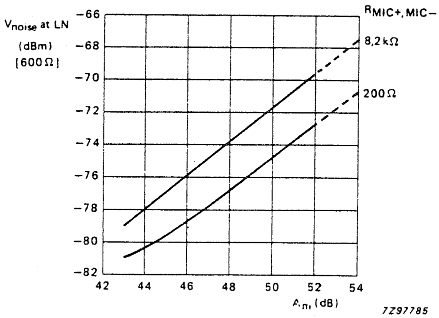
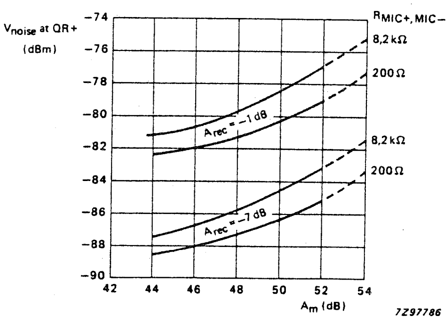


Fig.41 Psophometrically weighted noise on pin LN as a function of microphone amplifier gain.



A_m = microphone amplifier gain
A_{rec} = Overall Receive Gain = (A_{TA} - 32 dB)

Fig.42 Psophometrically weighted noise at the receiver output as a function of microphone gain.

NOISE

The typical psophometrically weighted (P53-curve) noise level measured on pin LN with a 600 Ω load is given as a function of microphone gain in Fig.41. The microphone input was loaded with a 200 Ω and an 8,2 kΩ resistor.

Figure 42 shows the psophometrical weighted noise level at the receive output (single-ended 300 Ω load) as a function of microphone gain. Parameters are the overall receive gain A_{REC}, which includes the anti-sidetone attenuation, and the resistor across the microphone inputs.

EFFECT OF POWER-DOWN FUNCTION

Figures 43 and 44 show the line voltage V_{LN} and the line current I_{line} with and without the power down function in the application of Fig.45. A resistor bridge was used in the exchange.

ACOUSTIC MEASUREMENTS; BSI (British) AND FTZ121R8 (German) SPECIFICATIONS

Circuits like those in Fig.31, have been shown to meet the electrical requirement of BS6317 for sets with a real impedance and with a complex impedance. Furthermore, similar circuits using the TEA1067/8 can be designed, for telephones with complex impedance, which meet or exceed the German Post Office requirement, FTZ121R8.

The results of OREM A measurements are given in Fig.46. They are only valid for the microphone, telephone transducers and cable used in the test. Other types of cable require a different balance network Z_{bal} and a different R_6 to correct the gain control curve.

- Bruel & Kjaer 3354 test system.
- Artificial cable (0-10 km) Ø 0,5 mm, 176 Ω/km, 38 nF/km
- Exchange supply voltage 48 V
- Feeding bridge 2 x 300 Ω
- Application of TEA1067 shown in Fig.31
  with R_4 = 44,2 kΩ: resulting in A_{TA} = 23,8 dB
  with R_7 = 54,9 kΩ: resulting in $A_{M\ 1061}$ = 36,2 dB
- Microphone capsule : Electret with FET preamplifier
  6,3 mV/Pa at 1 kHz
- Telephone capsule : dynamic Z_T = 150 Ω
  30 dBPa/V at 1 kHz

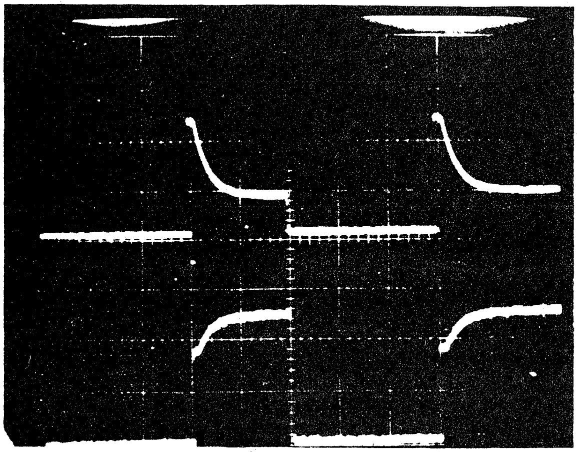


Fig.43 Line voltage and line current without power-down function. The upper trace shows the voltage at pin V_{LN} with 5 V per division along the vertical axis and zero volts along the middle horizontal line. The lower trace shows the line current I_{line} with 20 mA per division along the vertical axis and zero amperes along the bottom of the display. The horizontal axis is in divisions of 20 ms per division.

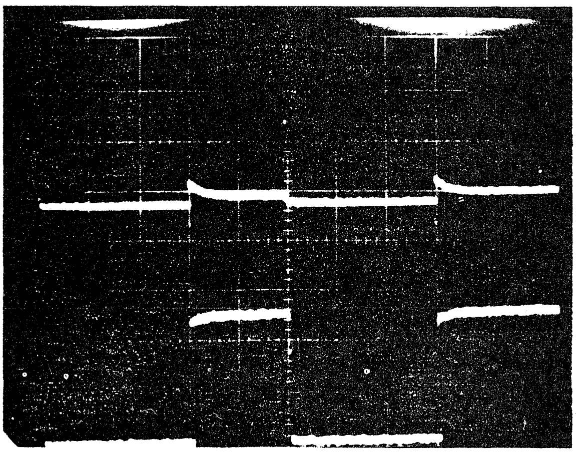


Fig.44 Line voltage and line current with power-down function. The upper trace shows the voltage at pin V_{LN} with 5 V per division along the vertical axis and zero volts along the middle horizontal line. The lower trace shows the line current I_{line} with 20 mA per division along the vertical axis and zero amperes along the bottom of the display. The horizontal axis is in divisions of 20 ms per division.

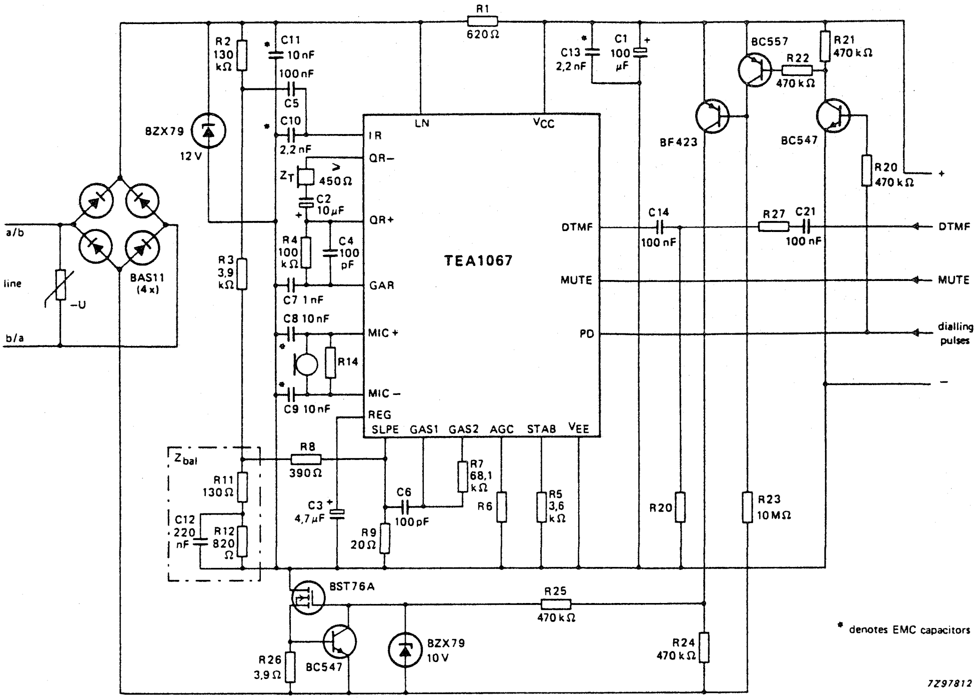
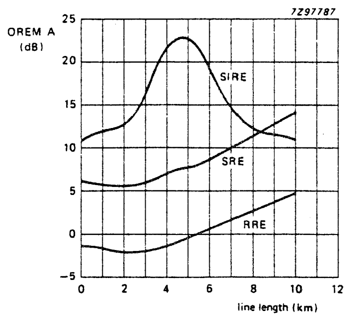


Fig.45 Basic application diagram of TEA1067 in sets with combined pulse and tone dialling including interruptor with interface. RFI capacitors are marked with a *. Protection of the IC is by means of the VDR in combination with a 12 V zener between LN and V_{EE}. This also provides symmetrical clipping of the sending signal.



RRE - Receive Reference Equivalent
SRE - Send Reference Equivalent
SIRE - Sidetone Reference Equivalent

Fig.46 OREM A Characteristics

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APPENDIX A - ANTI-SIDETONE BRIDGE CALCULATIONS

The following calculations are necessary to achieve optimum sidetone suppresion with the TEA1060 family. The calculations are given for both the TEA1060 family anti-sidetone bridge and the Wheatstone bridge for both real and complex impedance.

TEA1060 FAMILY ANTI-SIDETONE BRIDGE

The conditions during transmission with the TEA1060 family anti-sidetone bridge are defined for the circuit shown in Fig.47 and the equivalent circuit in Fig.48. From normal bridge considerations, we can write the following. For the sending signal on LN if $R_2 \gg |R_1//Z_{line}|$, then:

$$V_{LN} = -i_m \frac{R_1 Z_{line}}{R_1 + Z_{line}}$$

and for the signal on SLPE, if $|R_8 + Z_{bal}| \gg R_9$, then, $V_{SLPE} = i_m R_9$

SENDING CONDITIONS

Sending conditions are first considered in relation to the equivalent circuit shown in Fig.49 and then redrawn, using Thévenin´s theorem, to give the circuit shown in Fig.50, where:

$$V_A = -i_m \frac{R_1 Z_{line}}{R_1 + Z_{line}}$$

$$V_B = i_m R_9 \frac{Z_{bal}}{Z_{bal} + R_8}$$

If we can say that $R_2 \gg |R_1//Z_{line}|$, then the equivalent circuit can be redrawn as in Fig.51. This in turn, can be rearranged as shown in Fig.52, where:

$$V_C = V_A \frac{R_t}{R_t + R_2} = -i_m \frac{R_1 Z_{line}}{R_1 + Z_{line}} \times \frac{R_t}{R_2 + R_t}$$

If the superposition theorem is applied, the signal at the input of the receiving amplifier is given by:

$$V_{IR} = V_B \frac{R_2//R_t}{(R_2//R_t) + R_3 + (R_8//Z_{bal})} + V_C \frac{R_3 + (R_8//Z_{bal})}{(R_2//R_t) + R_3 + (R_8//Z_{bal})}$$

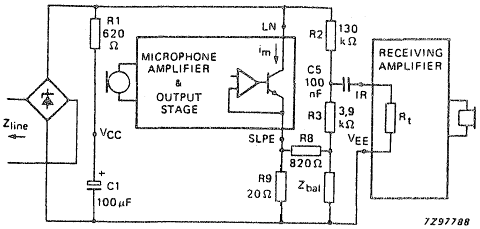


Fig.47 TEA1060 family bridge anti-sidetone bridge

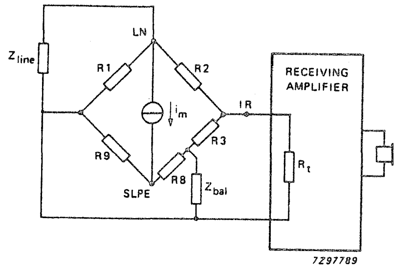


Fig.48 Equivalent circuit to TEA1060 family bridge derived from Fig.47

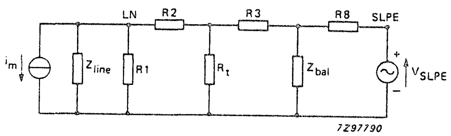


Fig.49 Equivalent circuit to the TEA1060 family bridge under sending conditions.

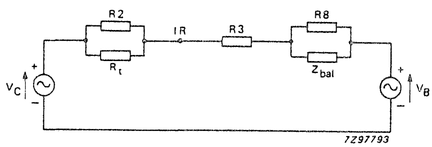


Fig.50 Equivalent circuit to Fig.49 after application of Thévenin's theorem.

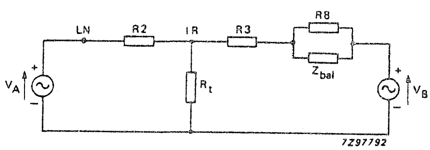


Fig.51 Modified equivalent circuit derived from Fig.50 with assumption that $R_2 >> |R_1 \;//\; Z_{line}|$

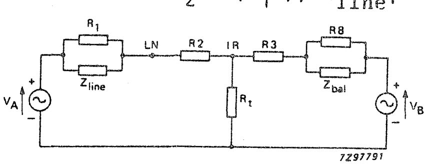


Fig.52 Rearrangement of circuit shown in Fig.51

By substitution of V_B and V_C we obtain:

$$V_{IR} = i_m R_9 \frac{Z_{bal}}{Z_{bal}+R_8} \times \frac{R_2//R_t}{(R_2//R_t)+R_3+(R_8//Z_{bal})}$$

$$- i_m \frac{R_1 Z_{line}}{R_1+Z_{bal}} \times \frac{R_t}{R_2+R_t} \times \frac{R_3+(R_8//Z_{bal})}{(R_2//R_t)+R_3+(R_8//Z_{bal})}$$

$$= \frac{i_m}{(R_2//R_t)+R_3+(R_8//Z_{bal})} \times [R_9 \frac{Z_{bal}}{Z_{bal}+R_8} (R_2//R_t)$$

$$- R_1 \frac{Z_{line}}{Z_{line}+R_1} \times \frac{R_t}{R_2+R_t} \{R_3+(R_8//Z_{bal})\}]$$

$$= \frac{i_m}{(R_2//R_t)+R_3+(R_8//Z_{bal})} \times \frac{R_t}{R_2+R_t} \times [R_9 \frac{Z_{bal}}{Z_{bal}+R_8} R_2 - R_1 \frac{Z_{line}}{Z_{line}+R_1} \{R_3+(R_8//Z_{bal})\}]$$

The signal at the input of the receiving amplifier is completely cancelled when:

(a) $R_9 R_2 = R_1 [R_3 + (R_8//Z_{bal})]$

and

(b) $$\frac{Z_{bal}}{Z_{bal}+R_8} = \frac{Z_{line}}{Z_{line}+R_1}$$

If correct fixed resistor values are chosen for R_1, R_2 and R_9, then condition (a) will always be fulfilled provided that $|R_8//Z_{bal}| << R_3$.

To obtain optimum sidetone suppression, condition (b) has to be fulfilled, resulting in:

$$Z_{bal} = \frac{R_8}{R_1} Z_{line} = k\ Z_{line}$$

where k is the scale factor and $R_8 = k\ R_1$

There are several points that should be noted in relation to the use of these equations in practical circuitry:

1. In practice, Z_{line} varies considerably with the line-length and cable type. Consequently, an average value of Z_{bal} has to be chosen.
2. The impedance Z_{bal} will normally be complex, and it is therefore necessary that $|R_8//Z_{bal}| \ll R_3$ to fulfill condition (a) above. The scale factor k is used for this purpose.
3. The input impedance of the receiving amplifier (R_t) has no influence on the bridge balance.

RECEIVING CONDITIONS

The receiving conditions for the TEA1060 family anti-sidetone bridge can be defined in relation to the equivalent circuits shown in Fig.53. The signal at the input of the receiving amplifier is given by the following equation:

$$V_{IR} = \frac{R_t//[R_3+(R_8//Z_{bal})]}{R_2+[R_t//\{R_3+(R_8//Z_{bal})\}]} V_{LN}$$

If we assume that $|R_8//Z_{bal}| \ll R_3$, then we can write the following:

$$V_{IR} = \frac{R_t//R_3}{R_2+(R_t//R_3)} V_{LN}$$

so the attenuation of the circuit is independent of frequency, and of the value of Z_{bal} under the conditions stated.

Example: If R_2 = 130 kΩ, R_3 = 3,92 kΩ, and Rt = 20 kΩ, then
20 log (V_{IR}/V_{LN}) (attenuation of the bridge) is -32,2 dB.

It is therefore possible to select a value of Z_{bal} to give optimum sidetone suppression, with any type of cable, without substantially affecting either the gain or the frequency response. In practice, the value of $R_8//Z_{bal}$ cannot be entirely ignored, resulting in a small anti-sidetone attenuation error of around of 0,2 dB over the frequency range 300 Hz to 3,4 kHz. The practical value for the attenuation will be slightly smaller than the calculated value.

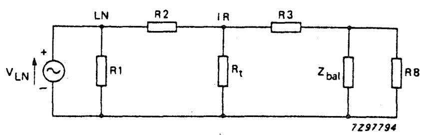


Fig.53 Equivalent circuit to TEA1060 family bridge under receiving condition.

FREQUENCY DEPENDENCE ON RECEIVING CONDITIONS

The theoretical variation in attenuation over the full frequency range from d.c. to infinity can be derived from calculations based on the equivalent circuit shown in Fig.54. The frequency dependence of the circuit is due to the impedance of the capacitor C because:

$$Z_{bal} = R_a + \frac{R_b}{1 + j\omega R_b X_C}$$

At low frequencies the value of Z_{bal} will be $R_a + R_b$ and at high frequencies will be R_a.

$$V_D = V_{LN} \frac{R_t}{R_2 + R_t}$$

and

$$V_{IR} = V_D \frac{R_3 + (R_8//Z_{bal})}{(R_2//R_t) + R_3 + (R_8//Z_{bal})}$$

So the attenuation of the circuit can be calculated from the following:

$$\frac{V_{IR}}{V_{LN}} = \frac{R_3 + (R_8//Z_{bal})}{(R_2//R_t) + R_3 + (R_8//Z_{bal})} \times \frac{R_t}{R_2 + R_t}$$

Example: At R_2 = 130 kΩ, R_3 = 3,92 kΩ, R_8 = 390 Ω, R_t = 20 kΩ, R_a = 130 Ω, R_b = 820 Ω, and C = 220 nF, we get an attenuation of -31,7 dB at d.c. and an attenuation of -32 dB for $f \rightarrow \infty$, indicating that in practice there is very little variation even over the maximum possible frequency range.

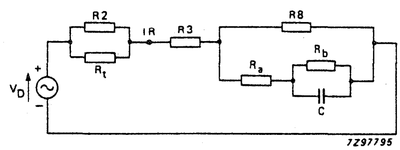


Fig.54 Equivalent circuit to the TEA1060 family bridge for calculating frequency dependence under receiving condtions.

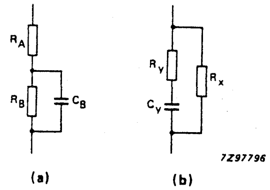


Fig.55 Equivalent circuit:
(a) complex impedance to replace a real impedance in bridge circuit.
(b) transformed version for calculation.

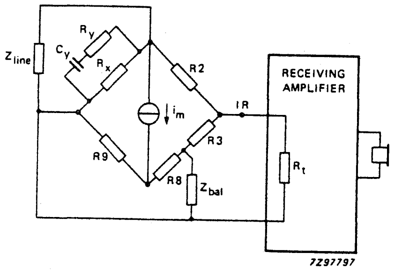


Fig.56 TEA1060 family bridge with complex impedance.

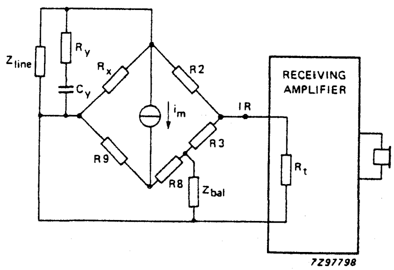


Fig.57 TEA1060 family bridge as in Fig.56 with R_y and C_y re-arranged

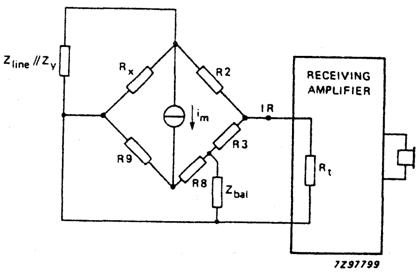


Fig.58 Anti-sidetone circuit of Fig.57 simplified.

COMPLEX IMPEDANCE

For those applications where PTT authorities require the usual real telephone set impedance has to be replaced by a complex impedance, the calculations require that R_1 is replaced by a suitable network Z_C, such as that in Fig.55(a), and we can write the following three equations:

- o $R_A = R_x//R_y$
- o $R_B = R_x^2/(R_x+R_y)$
- o $C_B = C_y[(R_x+R_y)/R_x]^2$

For the transformed version of this network, shown in Fig.55(b) we can write:

- o $R_x = R_A + R_B$
- o $R_y = \frac{R_A}{R_B}(R_A + R_B)$
- o $C_y = C_B[R_B/(R_A + R_B)]^2$

Recalculation for balancing the bridge to optimize the sidetone suppression is necessary in order to take into account the complex impedance (see Fig.56). Fig.56 can be redrawn as Fig.57 using the transformation. The calculation is best handled, regarding R_y and C_y as making a constant contribution to the line impedance. Substitution of Z_y by $R_y+(1/j\omega C_y)$ results in Fig.58.

The complex impedance bridge can be balanced in the same way as the real impedance bridge, but with R_x substituted for R_1 and $Z_{line}//Z_y$ substituted for Z_{line}. We then have the equation below:

$$V_{IR} = \frac{i_m}{(R_2//R_t)+R_3+(R_8//Z_{bal})} \times \frac{R_t}{R_2+R_t}$$

$$\times \left[R_9 \frac{Z_{bal}}{Z_{bal}+R_8} R_2 - R_x \frac{Z_{line}//Z_y}{(Z_{line}//Z_y)+R_x} \{R_3+(R_8//Z_{bal})\}\right]$$

The signal at the input of the receiving amplifier is completely cancelled when:

(a) $R_9R_2 = R_x(R_3 + [R_8//Z_{bal}])$

and

(b)
$$\frac{Z_{bal}}{Z_{bal} + R_8} = \frac{Z_{line}//Z_y}{(Z_{line}//Z_y) + R_x} \Rightarrow Z_{bal} = \frac{R_8}{R_x}(Z_{line}//Z_y)$$

which, with the substitution of $k = (R_8/R_x)$ as a scale factor, gives; $Z_{bal} = k(Z_{line}//Z_y)$. Generally, the resistance R_x is larger than the original 600 Ω taken for R_1. This means that condition (a) above can be fulfilled only by changing the value of either R_3, R_2, or R_9. If we assume that $R_x > R_1$, then the three possible changes can be considered as follows.

Decreasing R_3 will also increase the attenuation of the bridge in the receiving direction, and will also make it more difficult to meet the condition $|R_8//Z_{bal}| \ll R_3$. Decreasing R_3 is therefore not recommended.

Increasing R_2 also increases the bridge attenuation in the receiving direction, thus reducing the maximum attainable gain. This reduction is normally acceptable when a sensitive earpiece transducer is used, but may give excessive noise with an insensitive transducer.

Increasing the value of R_9 is the most preferable option. This will slightly decrease the maximum obtainable gain in the transmitting direction (microphone and DTMF gain) but will cause difficulty only if a very insensitive microphone transducer is used; an associated shift in the start and stop points of the gain control characteristic can easily be corrected with the resistor connected to the AGC pin. There is also a small increase of slope in the d.c. characteristic but this will be acceptable in most cases.

Thus, with a complex impedance in the TEA1060 family anti-sidetone bridge, the circuit designer can select either the second or third option outlined above. If neither of these approaches is preferred, then the Wheatstone bridge can be considered as an alternative approach but this will present a frequency-dependent transfer characteristic between the line and the input of the receiving amplifier.

WHEATSTONE BRIDGE

The TEA1060 family can be used with the conventional Wheatstone bridge configuration shown in Fig.59. The equivalent circuit is given in Fig.60. As a more familiar circuit it may seem easier to design, but it has frequency-dependent attenuation. If $R_8 \gg R_9$, then for the sending signal on SLPE, we can write: $V_{SLPE} = i_m R_9$

It should be noted that, in this circuit, the function of R_A is to adjust the attenuation of the anti-sidetone circuit under receiving condition. In the following calculations, R_A is neglected so the attenuation calculated is the minimum value.

SENDING CONDITIONS

The sending conditions can be considered in relation to the equivalent circuit of Fig.61. By using Thévenin's theorem, this can be replaced by the circuit in Fig.62, where:

$$V_A = -i_m \frac{R_1 Z_{line}}{R_1 + Z_{line}}$$

$$V_B = i_m R_9 \frac{R_t}{R_8 + R_t}$$

Using the superposition theorem we can write:

$$V_{IR} = i_m R_9 \frac{R_t}{R_8 + R_t} \times \frac{Z_{bal} + (R_1 // Z_{line})}{Z_{bal} + (R_1 // Z_{line}) + (R_8 // R_t)}$$

$$-i_m \frac{R_1 Z_{line}}{R_1 + Z_{line}} \times \frac{R_8 // R_t}{Z_{bal} + (R_1 // Z_{line}) + (R_8 // R_t)}$$

$$V_{IR} = \frac{i_m}{Z_{bal} + (R_1 // Z_{line}) + (R_8 // R_t)} \times \frac{R_t}{R_8 + R_t} \times \left[R_9 (Z_{bal} + (R_1 // Z_{line})) - \frac{R_1 Z_{line}}{R_1 + Z_{line}} R_8\right]$$

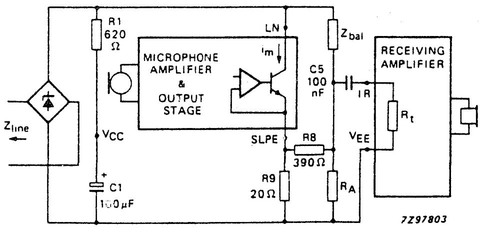


Fig.59 Wheatstone bridge anti-sidetone circuit

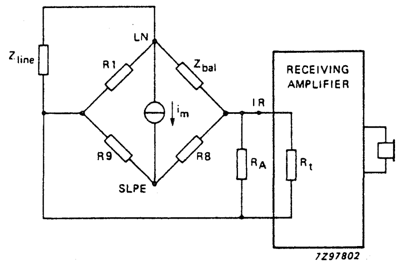


Fig.60 Equivalent circuit to Wheatstone bridge anti-sidetone

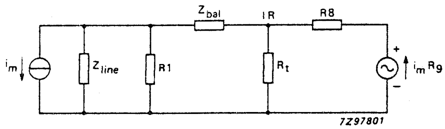


Fig.61 Equivalent circuit to Wheatstone bridge under sending conditions

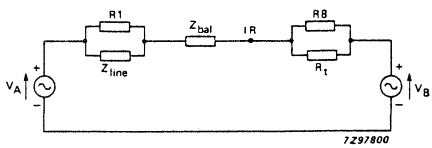


Fig.62 Equivalent circuit to Fig.61 after the application of Thévenin's theorem

The signal at the input of the receiving amplifier is completely cancelled when the following conditions are fulfilled:

$$R_9[Z_{bal}+(R_1//Z_{line})] = \frac{R_1 Z_{line}}{R_1+Z_{line}} R_8$$

$$Z_{bal}+(R_1//Z_{line}) = \frac{R_8}{R_9} x \frac{R_1 Z_{line}}{R_1+Z_{line}}$$

$$Z_{bal} = \left[\frac{R_8}{R_9} -1\right] x \frac{R_1 Z_{line}}{R_1+Z_{line}}$$

Then, if $(R_8/R_9)>>1$:

$$Z_{bal} = \frac{R_8}{R_9} x \frac{R_1 Z_{line}}{R_1+Z_{line}}$$

Hence, defining the optimum conditions for balance of the Wheatstone bridge.

RECEIVING CONDITIONS

The receiving conditions can be defined in relation to the equivalent circuit shown in Fig.63. The signal at the input of the receiving amplifier is given by:

$$V_{IR} = \frac{R_8//R_t}{Z_{bal}+(R_8//R_t)} V_{LN}$$

So the attenuation of the anti-sidetone circuit is given by:

$$\frac{V_{IR}}{V_{LN}} = \frac{R_8//R_t}{Z_{bal}+(R_8//R_t)}$$

For any given value of Z_{bal}, the attenuation is frequency dependent because Z_{bal} is complex. If the impedance Z_{bal} is changed to match a different type or length of cable, the characteristic of the frequency dependence must also vary. This affects the frequency response of the receiving channel, and the attenuation of the anti-sidetone circuit so the gain will need readjustment.

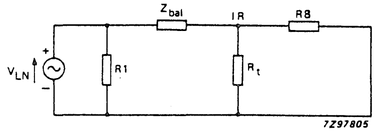


Fig.63 Equivalent circuit to Wheatstone bridge under receiving conditions.

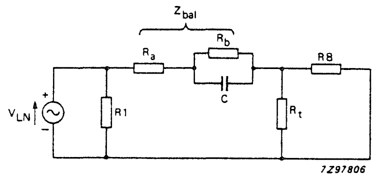


Fig.64 Equivalent circuit to Wheatstone bridge for calculating frequency-dependence under receiving conditions.

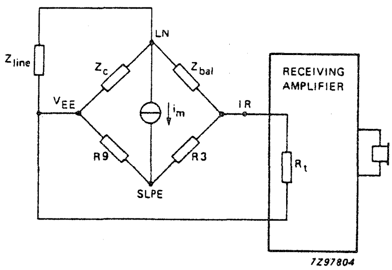


Fig.65 Wheatstone bridge anti-sidetone circuit with complex impedance

FREQUENCY DEPENDENCE ON RECEIVING CONDITIONS

The theoretical variation in attenuation, over the full frequency range, can be derived from calculations based on the equivalent circuit in Fig.64. The attenuation of the circuit is given by:

$$20 \log \frac{V_{IR}}{V_{LN}} = 20 \log \frac{R_8//R_t}{Z_{bal} + (R_8//R_t)}$$

$$Z_{bal} = R_a + \frac{R_b}{1 + j\omega R_b C}$$

at $f = 0$, $Z_{bal} = R_a + R_b$ and at $f \rightarrow \infty$, $Z_{bal} = R_a$

So for the practical values of; R_a = 9,1 kΩ, R_b = 9,1 kΩ, R_8 = 820 Ω, R_t = 20 kΩ, and C = 10 nF the attenuation is -27,6 dB at d.c., and -22 dB as $f \rightarrow \infty$. This indicates that there is a significant variation in attenuation (5,6 dB) between the extreme limits of frequency. At f = 300 Hz the attenuation is -27,5 dB, and the value, for f = 3,4 kHz, is -23,9 dB (a difference of 3,6 dB). These figures are the minimum attenuation values disregarding R_A in parallel with the R_t (see Fig.59) and if we now include R_A with the value 330 Ω, the two limiting values are; -38 dB for d.c., and -32,2 dB as $f \rightarrow \infty$, (a variation of 5,8 dB).

COMPLEX IMPEDANCE

For those applications where PTT authorities require the usual telephone set impedance to be replaced by a complex impedance, the calculations require that R_1 be replaced by a suitable network representing Z_C such as in Fig.55(a) and Fig.55(b). Changing R_1 to a complex impedance necessitates recalculation to optimize the sidetone suppression. However, for the bridge in Fig.65 we can adapt the balance condition derived earlier by replacing R_1 with Z_C, and again by assuming that $(R_8/R_9) >> 1$ we get the following balance condition:

$$Z_{bal} = \frac{R_8}{R_9} \times \frac{Z_C Z_{line}}{Z_C + Z_{line}}$$

The occurrence of Z_C in the equation for Z_{bal} indicates the increased capacitance of Z_{bal} and it indicates the frequency-dependent nature of the transfer function between the line, LN, and the input to the receiving amplifier, IR.

APPENDIX B - GRAUE MIKROFON SCHNITTSTELLE

The TEA1067 and TEA1068 have been designed to meet the German "Graue Mikrofon Schnittstelle" requirements. This requirement is a definition of a two-wire mirophone input circuit which is suitable for several different types of microphone without the need to readjust circuit parameters. This means that microphones must have a certain sensitivity for a given load (8,2 kΩ//22 nF). Electret-condenser microphones may also be used, so it is be necessary to apply a d.c. voltage at the microphone inputs (Tonaderspeisung).

Figure 66 gives a microphone input circuit example according to the "Graue Mikrofon Schnittstelle" requirements.

The supply voltage is applied to microphone input terminals V1 and V2,via the resistors R_3, R_4 and R_5, making it possible to power electret-condenser microphones. Capacitors C_1 and C_2 prevent this supply voltage from reaching the TEA1067/8 inputs MIC + and MIC-. Capacitor C_3 determines the input capacitance of the circuit (22nF) while resistors R_1, R_2, R_3, R_4 and R_6 determine the input resistance:

$$\frac{1}{R_{in}} = \frac{1}{R_1 + R_2} + \frac{1}{R_3 + R_4} + \frac{1}{R_6}$$

.....resulting in R_{in} = 8,2 kΩ for the example given.

The tolerance of R_{in} need not be greater than 5% because the input impedance is mainly determined by accurate external resistors. The tolerance of R_{in} therefore results in only a small variation in gain between the input and the line.

An application proposal according to the "Graue Mikrofon Schnittstelle" with the TEA1068 is shown in Appendix D.

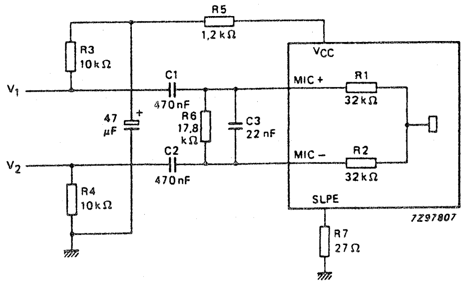
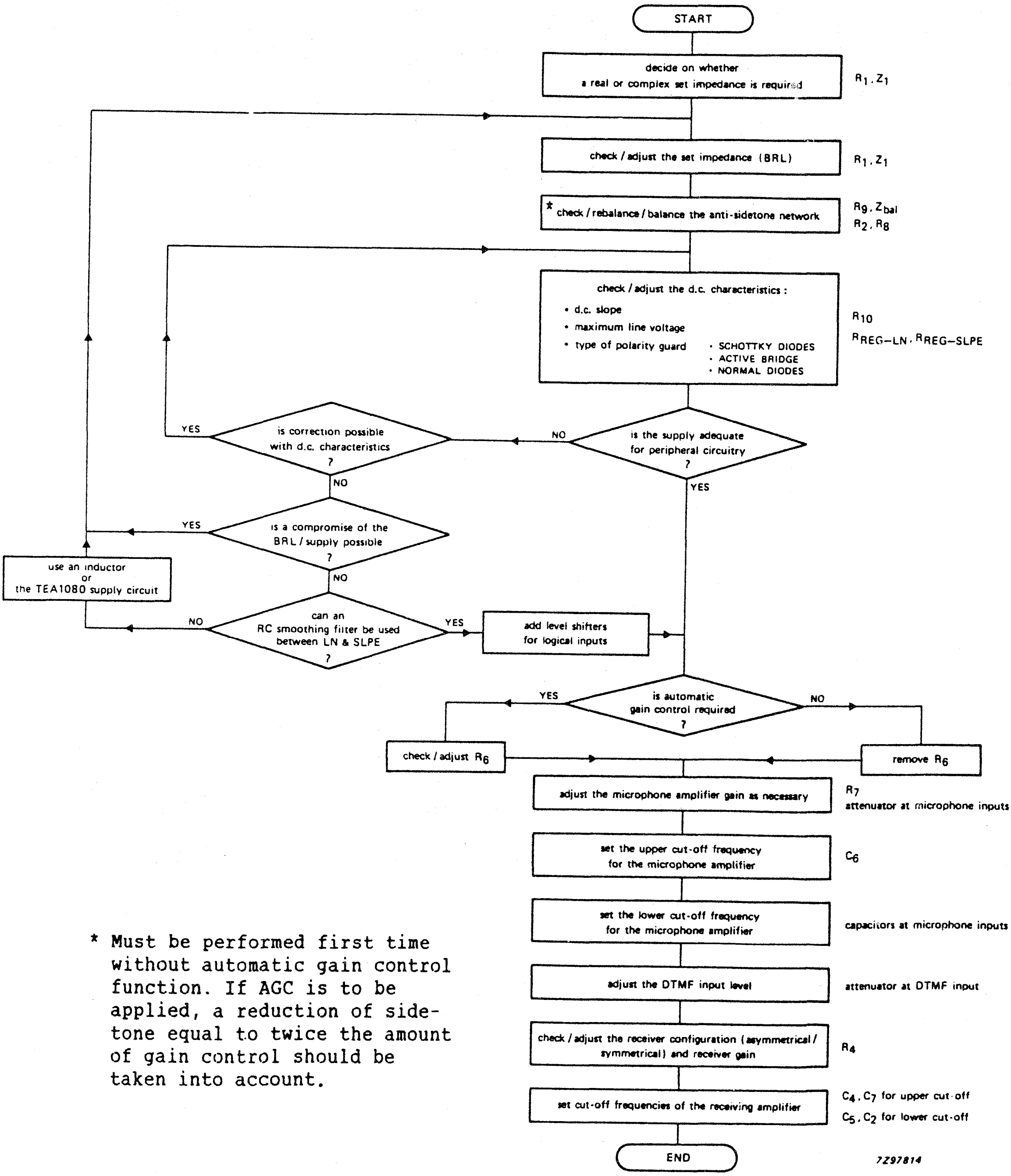


Fig.66 Graue Mikrofon Schnittstelle input circuit

APPENDIX C - ADJUSTING PARAMETERS FOR THE TEA1060 FAMILY



* Must be performed first time without automatic gain control function. If AGC is to be applied, a reduction of side-tone equal to twice the amount of gain control should be taken into account.

APPENDIX D - APPLICATION CIRCUIT EXAMPLES

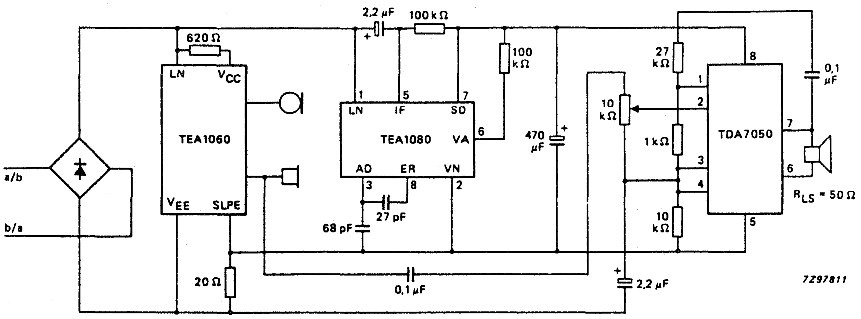


Fig.67 Circuit diagram for listening-in application with TEA1060, TEA1080 and TDA7050

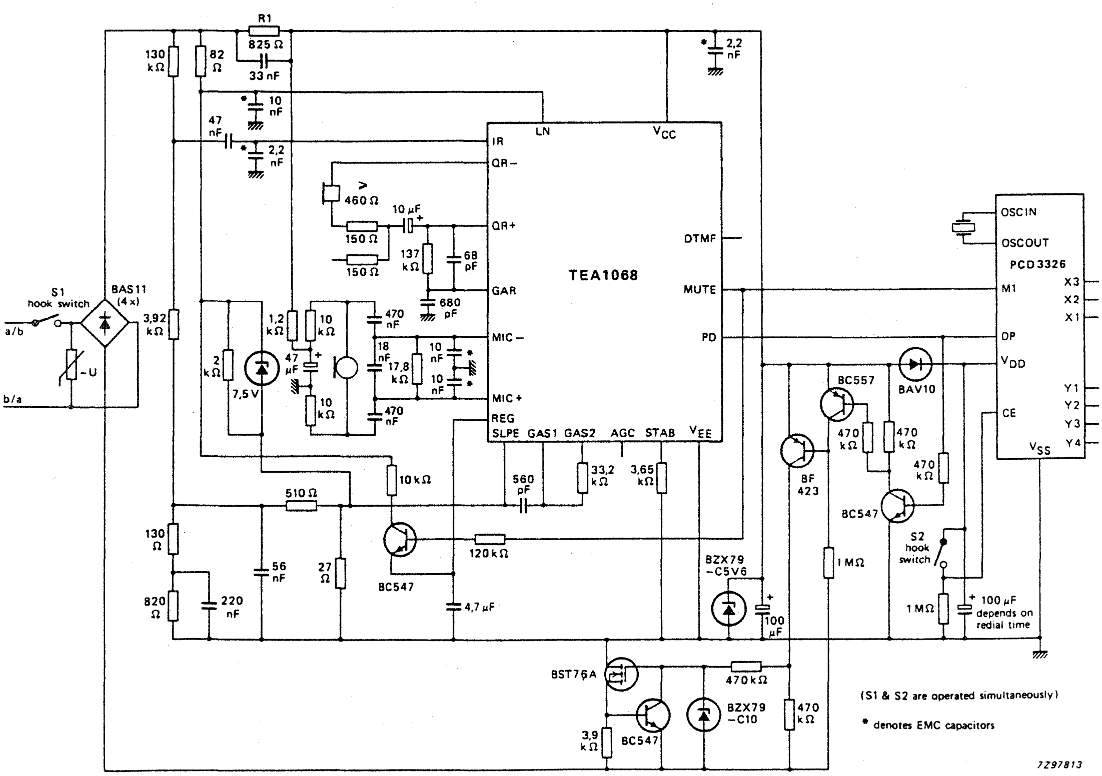


Fig.68 Circuit Diagram for Pulse-dialling with the Graue Hörkapsel Schnittstelle and also incorporating the Graue Mikrofon Schnittstelle

APPLICATION NOTE Nr	ETT/AN89008
TITLE	Application of the speech-transmission circuit TEA1062
AUTHOR	P. T. J. Biermans
DATE	October 1989

Summary:

This report describes differences between the low-voltage speech-transmission IC's TEA1062 and TEA1067. The report should be used in combination with the TEA1060 family designers' guide (lit.1).
The TEA1062 has a modified performance and less features compared to the TEA1067.

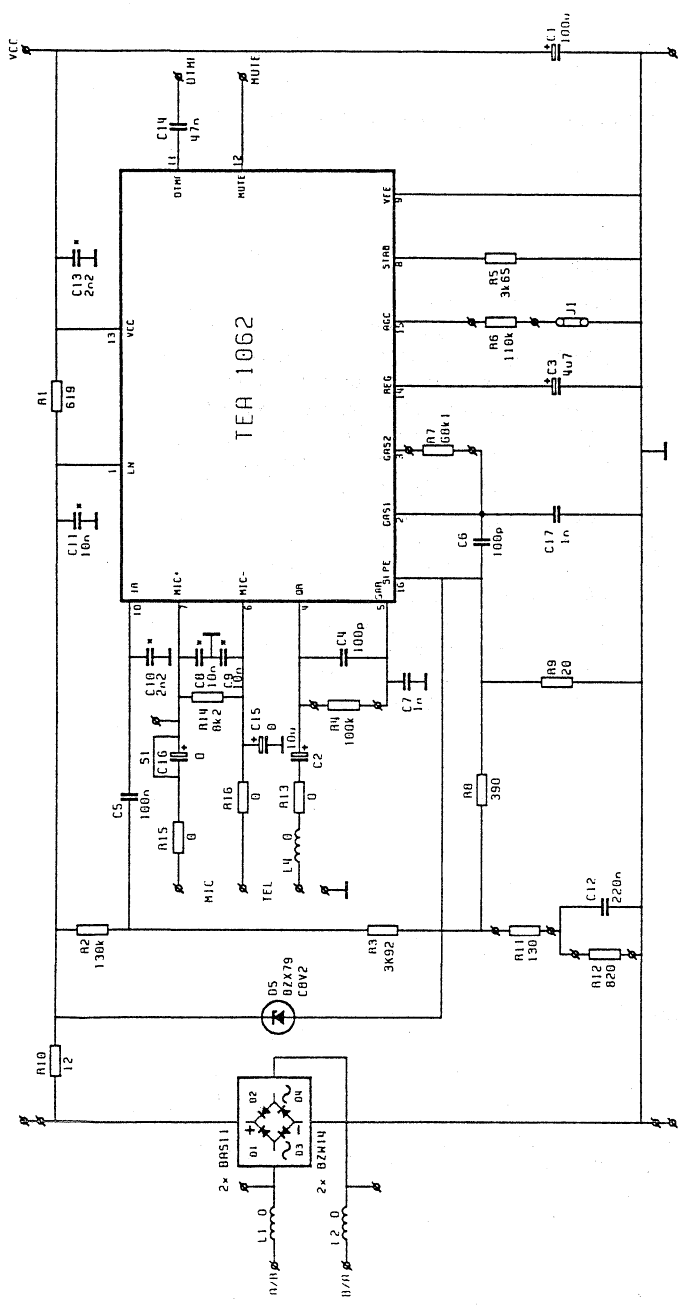
Table of Contents:

1. INTRODUCTION

The TEA106X family consists of a range of bipolar integrated circuits performing all speech and line interface functions required in fully electronic telephone sets.
The TEA1060 family designers guide (lit.1) provides information on most of the members of the TEA106X. The TEA1062 is not described in this guide. Detailed information about this circuit can be found in the datasheet (lit.2).
The TEA1062 is a low-voltage speech-transmission IC able to operate down to a dc line voltage of 1.6 V to facilitate the use of more telephone sets in parallel. The TEA1062 has a modified performance and less features compared to the TEA1067.

In this report differences between the TEA1062 and the TEA1067 as described in lit.1 are elucidated. When applying the TEA1062 this report should be used in combination with the designers guide.

The figures given in this report all refer to the TEA1062.
The TEA1062 is described in this report with respect to the basic application circuit shown in fig. 1 (lit.3).



CIRCUIT DIAGRAM OF PRINTED CIRCUIT BOARD CAB3422

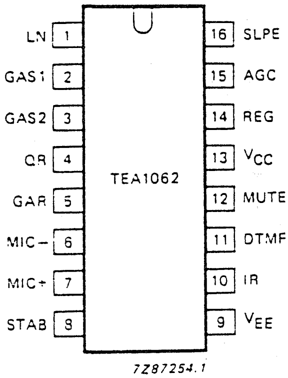
(Note : Capacitors with capacitance 0 are not mounted.
Resistors with resistance 0 and coils with
inductance 0 are replaced by a short.
Capacitors marked '*' are used to provide basic EMC performance)

Figure 1: Basic application circuit of the TEA1062

2. DESCRIPTION OF THE CIRCUIT

2.1 Pinning

The pinning/signal functions of the TEA1062 are in given in fig. 2.



PINNING

1	LN	positive line terminal
2	GAS1	gain adjustment; transmitting amplifier
3	GAS2	gain adjustment; transmitting amplifier
4	QR	non-inverting output, receiving amplifier
5	GAR	gain adjustment; receiving amplifier
6	MIC−	inverting microphone input
7	MIC+	non-inverting microphone input
8	STAB	current stabilizer
9	V_{EE}	negative line terminal
10	IR	receiving amplifier input
11	DTMF	dual-tone multi-frequency input
12	MUTE	mute input
13	V_{CC}	positive supply decoupling
14	REG	voltage regulator decoupling
15	AGC	automatic gain control input
16	SLPE	slope (DC resistance) adjustment

Figure 2: Pinning diagram

The TEA1062 has two pins less compared to the TEA1067:
- PD : power-down input;
- QR-: inverting output receiver amplifier.

2.2 Line voltage and peripheral supply

In fig. 3 the internal supply current I_{CC} is shown as a function of V_{CC}.
For V_{CC}= 2.8 V I_{CC}= 0.8 mA for the TEA1062 and I_{CC}= 1.0 mA for the TEA1067.

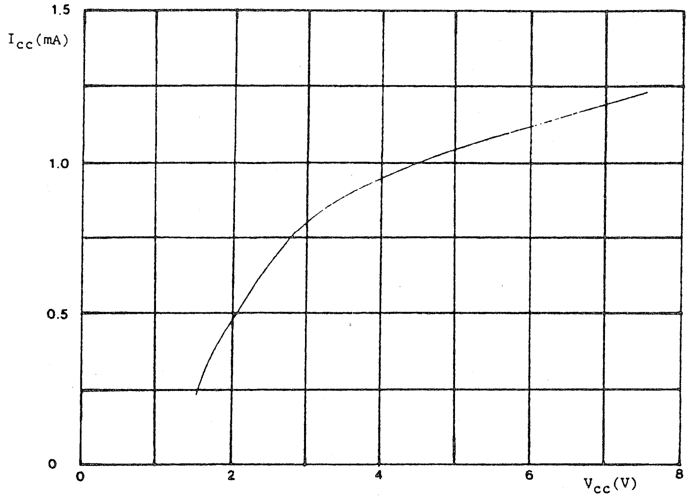


Figure 3: Internal supply current I_{cc} as a function of V_{cc}

In fig. 4 the dc line voltage V_{LN} is given as a function of the line current I_{line}. For a line current of 15 mA V_{LN}= 4.0 V, whereas for the TEA1067 V_{LN}= 3.9 V. Another difference between both IC's is the minimum and maximum value of the line voltage for I_{line}= 15 mA: for the TEA1062 the minimum value equals 3.55 V and the maximum 4.25 V, whereas for the TEA1067 these values are 3.65 V and 4.15 V. For a I_{line}= 1 mA V_{LN}= 1.6 V for both devices.

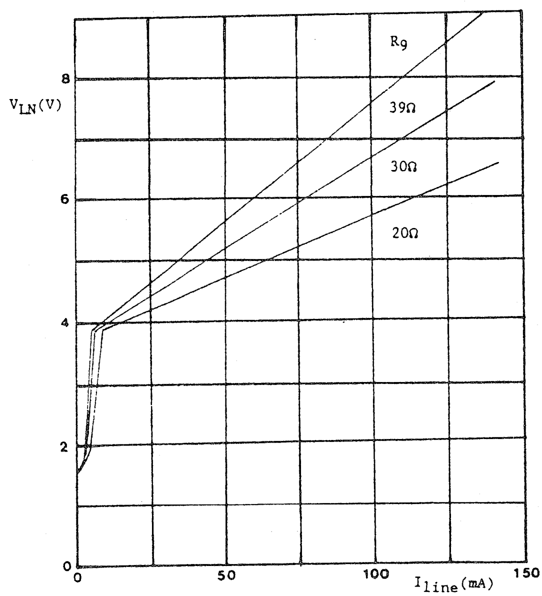


Figure 4: V_{LN} as a function of the line current I_{line}.

The dc line voltage can be increased by connecting a resistor between the pins REG and SLPE and decreased by connecting a resistor between the pins REG and LN. The influence of these resistors is slightly different compared to the figures given in lit.1 because of the slightly higher reference voltage of the TEA1062.
Fig. 5 shows how $V_{LN\text{-}SLPE}$ changes with the value of $R_{REG\text{-}SLPE}$ and $R_{REG\text{-}LN}$. The line current I_{line}= 15 mA.

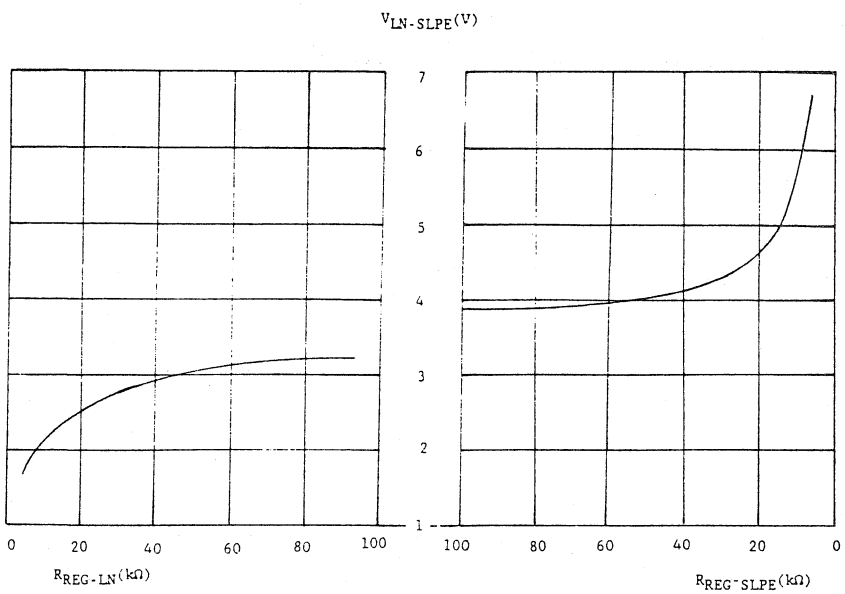


Figure 5: Internal reference voltage $V_{LN\text{-}SLPE}$ as a function of $R_{REG\text{-}SLPE}$ and $R_{REG\text{-}LN}$

If a resistor of 68 kΩ is connected between LN and REG V_{LN}= 3.5 V ($V_{LN\text{-}SLPE}$ + V_{SLPE}) for the TEA1062 and V_{LN}= 3.4 V for the TEA1067.

Fig. 6 shows the available peripheral current I_p as a function of V_{cc} for a line current of 15 mA for both the speech mode and the mute mode.
In the mute mode for the TEA1062 I_p= 1.55 mA for V_{cc}= 2.5 V, whereas for the TEA1067 I_p= 1.25 mA. An increase of 0.3 mA is obtained with the TEA1062 to supply peripheral circuits. This is caused by a lower I_{cc} and a higher reference voltage V_{ref}.
In the speech mode for the TEA1062 the available current I_p= 1.15 mA for V_{cc}= 2.5 V, and for the TEA1067 I_p= 0.8 mA.

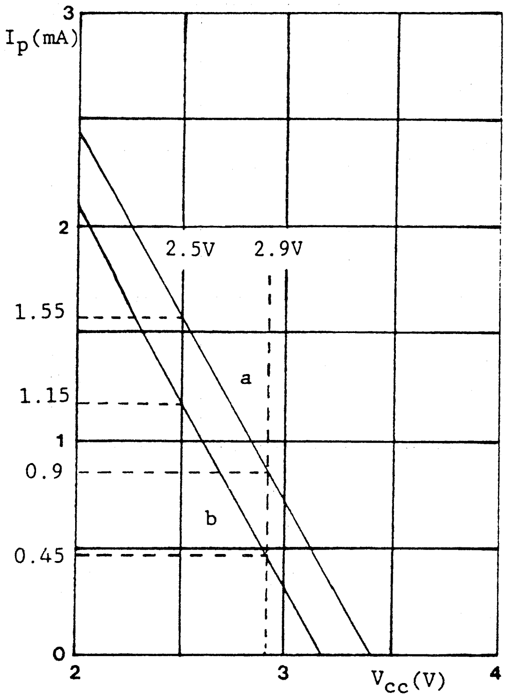
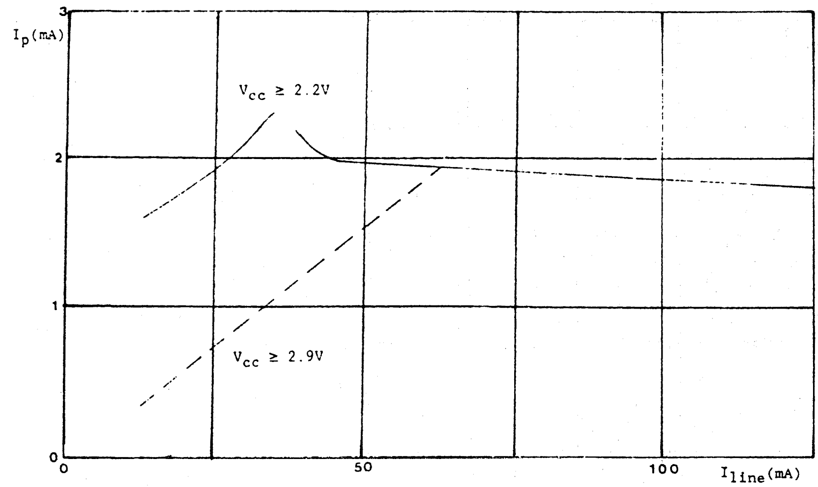
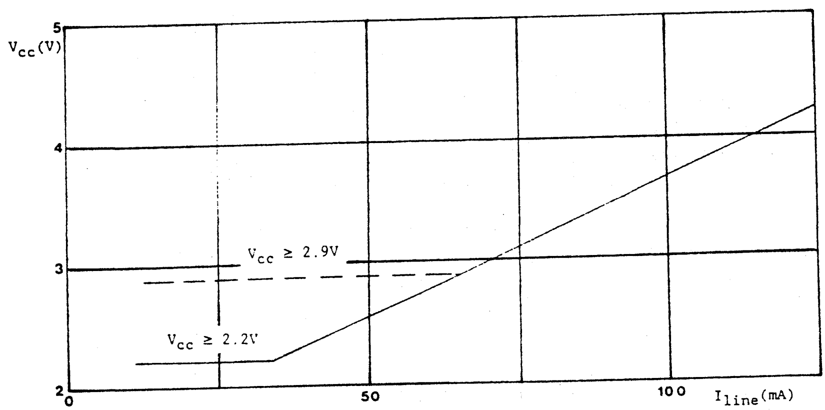


Figure 6: I_p as a function of V_{cc} (I_{line}= 15 mA)
a: mute mode: V_{LN}=1.0 V_{rms}
b: speech mode: V_{LN}=1.4 V_{rms} (THD < 2%)
V_{QR}=150 mV_{rms} across 150 Ω (THD < 2%)

The available current I_p and the corresponding voltage V_{cc} as a function of the line current are shown in fig. 7a and b for the speech mode, and in fig. 8a and b for the mute mode.

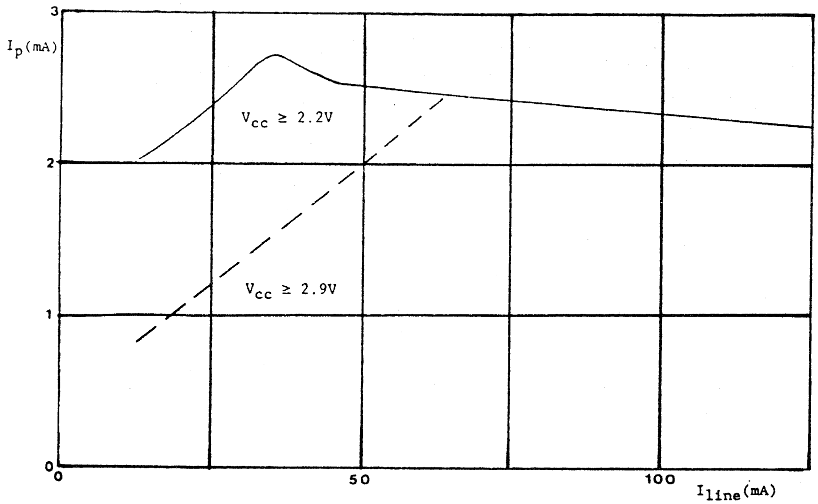


(a)

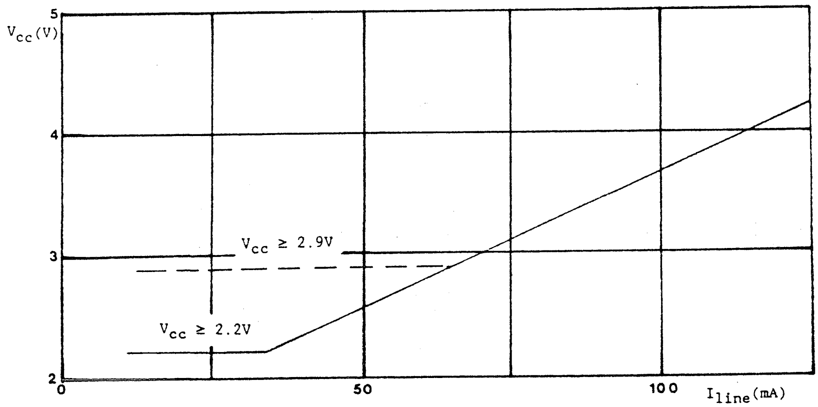


(b)

Figure 7: I_p (a) and V_{cc} (b) as a function of the line current in the speech mode



(a)



(b)

Figure 8: I_p (a) and V_{cc} (b) as a function of the line current in the mute mode

Fig. 9 shows the available supply current I_p as a function of R1 in the mute condition. Decreasing R1 leads to an increase in supply capability. A smaller value of R1 can be used in countries where the balance return loss (BRL) is better than the PTT requirements.

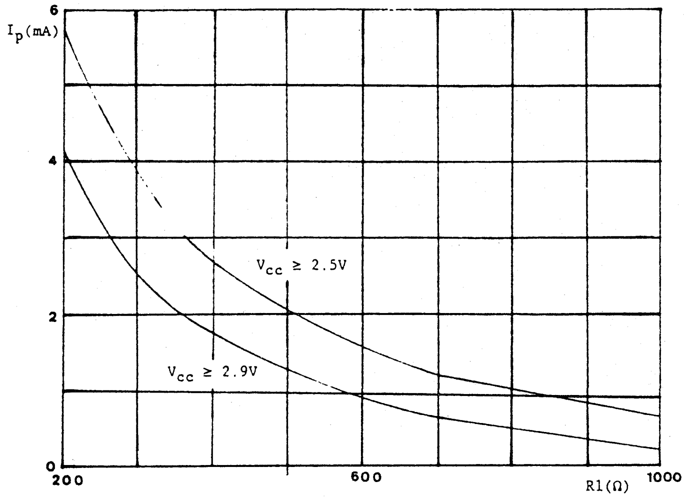


Figure 9: Supply current I_p as a function of R1 in the mute mode

2.3. Microphone amplifier

The gain of the microphone amplifier of the TEA1062 is comparable to that of the TEA1067. In fig. 10 the microphone gain is shown as a function of R7 for different values of R9. The spread on the nominal value of 52 dB (R7= 68.1kΩ) is 1.5 dB for the TEA1062 and 1.0 dB for the TEA1067.

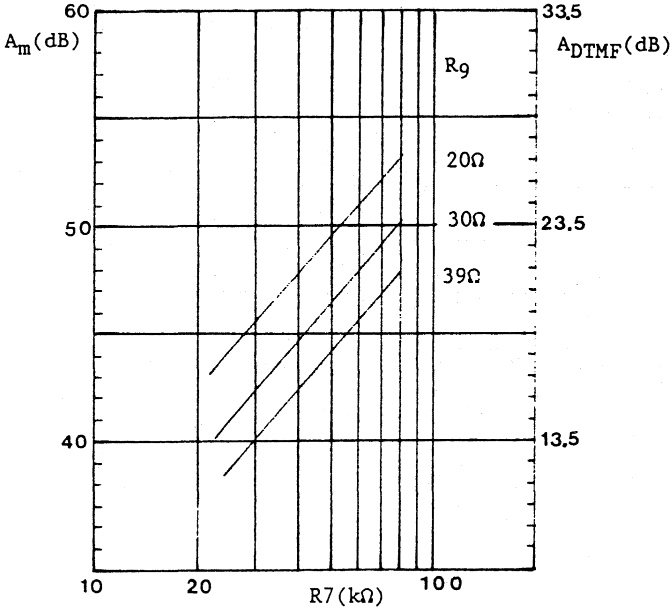


Figure 10: Microphone gain as a function of R7

To ensure stability of the microphone amplifier of the TEA1062 an additional capacitor C17 is connected between GAS1 and VEE. C17 should be 10 times C6. However, this also influences the set impedance. The set impedance has to be multiplied in this case by

$$\frac{1 + jwR7C6}{1 + jwR7(C6 + C17)}$$

Consequently, an additional zero and pole are introduced. In this application with C6 = 100 pF, C17 = 1 nF, and R7= 68.1kΩ a pole is present at f_{3dB}= 2.1 kHz and a zero at f_{3dB}= 23 kHz. The BRL is affected by this modification of the set impedance as is shown in chapter 3.

Fig. 11 shows the maximum output swing of the transmitter stage as a function of the line voltage V_{LN} for a line current of 15 mA. The results for the TEA1062 and TEA1067 are comparable.

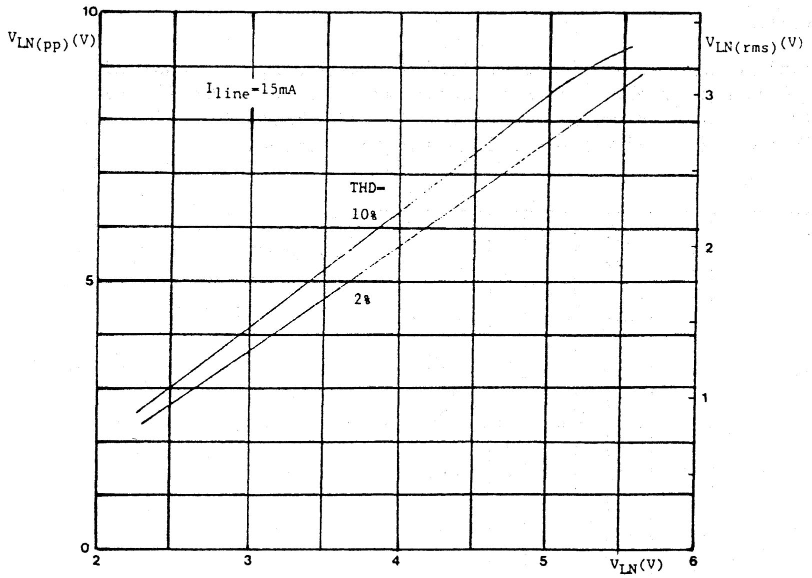


Figure 11: Maximum output swing of the transmitter stage as a function of the line voltage V_{LN} (I_{line}= 15 mA)

Fig. 12 shows the maximum output voltage at pin LN as a function of the line current I_{line} in the low current range.

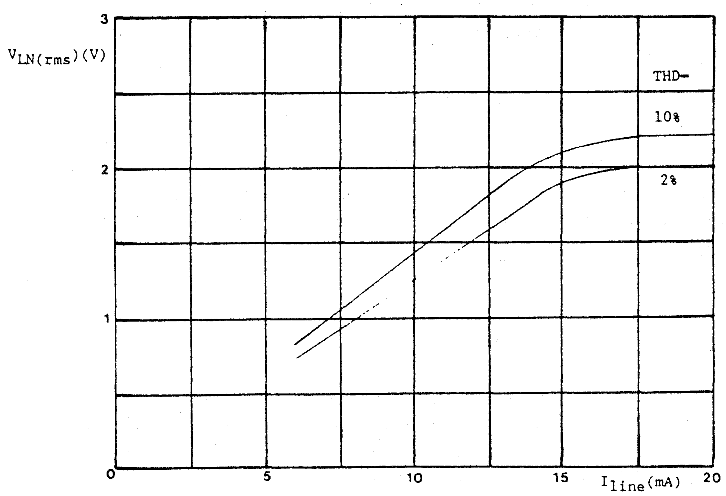


Figure 12: Maximum output voltage $V_{LN(rms)}$ as a function of I_{line}

For a line current of 15 mA and a total harmonic distortion of 10% $V_{LN(rms)}$= 2.1 V for the TEA1062 and $V_{LN(rms)}$= 2.2 V for the TEA1067. However, the typical value for the TEA1062 is $V_{LN(rms)}$= 2.3 V.

The typical maximum output swing of the TEA1062 is slightly higher compared to that of the TEA1067 because of the reduced internal current consumption I_{cc} and the slightly higher reference voltage V_{ref}.

Fig. 13 shows the transmit gain as a function of V_{LN}. A significant difference is obtained with respect to the TEA1067: at V_{LN}= 2.0 V the reduction is approximately 0.5 dB for the TEA1062 and 2 dB for the TEA1067 and for V_{LN}= 1.8 V the reduction is 5 dB for the TEA1062 and 8 dB for the TEA1067.

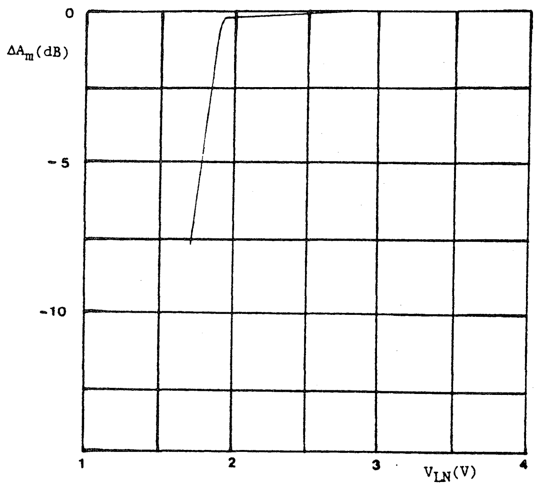


Figure 13: Transmit gain ΔA_m as a function of the line voltage V_{LN}

2.4. DTMF amplifier

The gain of the DTMF amplifier of the TEA1062 is identical to that of the TEA1067. However, the spread on the nominal value of 25.5 dB (R7= 68.1 kΩ) is 1.5 dB for the TEA1062 and 1.0 dB for the TEA1067.

2.5. Receiver amplifier

With the TEA1062 only a single-ended configuration is possible: the inverting output QR- of the TEA1067 is not present in this IC.

The gain of the TEA1062 is comparable to that of the TEA1067. In fig. 14 the receiver gain is shown as a function of R4. To ensure stability of the receiver amplifier at low line currents when it is loaded with a low-ohmic impedance (150 Ω) the maximum gain from pin IR to QR should be limited to 31 dB for the TEA1062, instead of the 31+8 dB for the TEA1067. The spread on the receiver gain is 1.5 dB for the TEA1062 and 1.0 dB for the TEA1067.

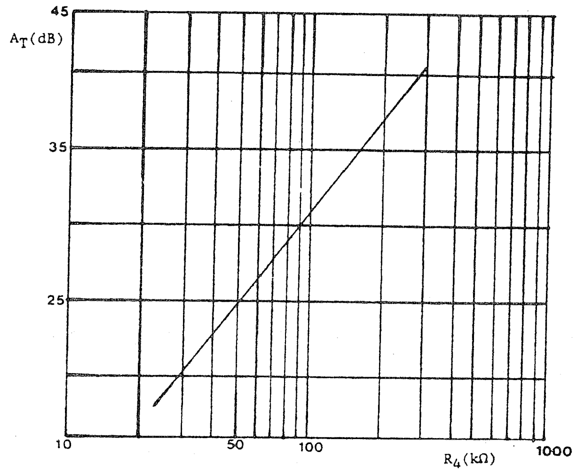


Figure 14: Receiver gain as a function of R4

In fig. 15 the maximum output swing of the receiver output stage $V_{QR(rms)}$ is shown as a function of the dc voltage V_{LN} for I_{line}= 15 mA and a total harmonic distortion of 10%. For V_{LN}= 4.0 V $V_{QR(rms)}$= 380 mV for the TEA1062 and 320 mV for the TEA1067 for a 150 Ω load.

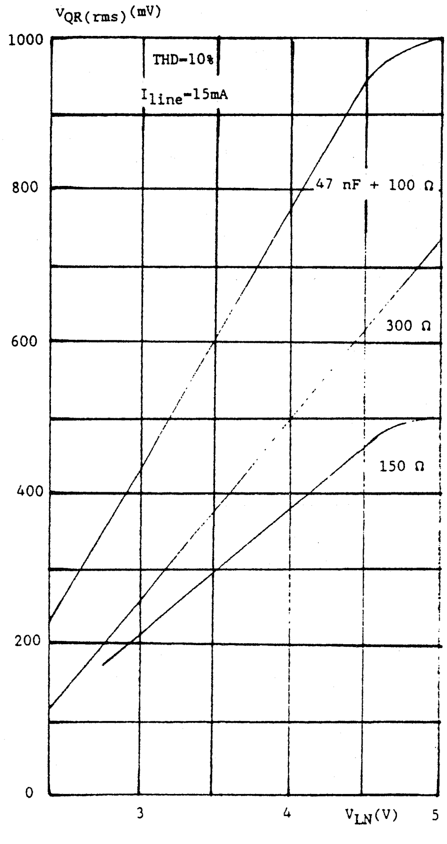


Figure 15: Maximum output swing $V_{QR(rms)}$ as a function of V_{LN}
(V_{LN} is set by means of R_{VA})

Fig. 16 shows the maximum output swing of the receiver output stage $V_{QR(rms)}$ as a function of the dc voltage V_{LN} in the low-voltage range for three load values and for a total harmonic distortion of 10%.

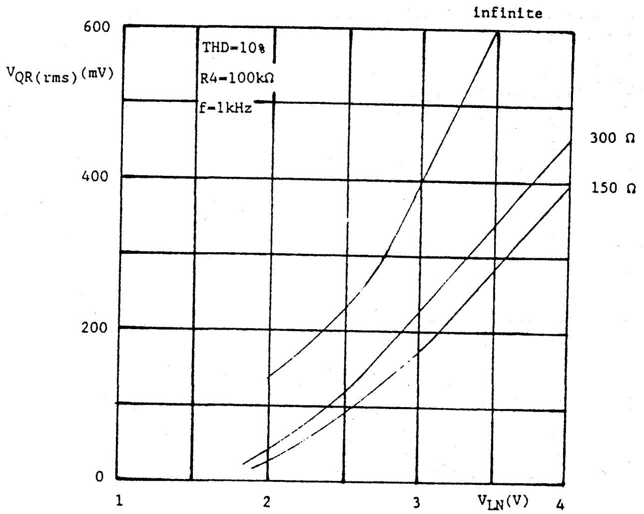


Figure 16: Maximum output swing $V_{QR(rms)}$ as a function of V_{LN} in the low-voltage range

With V_{LN}= 2.0 V for a 150 Ω load an output swing of 20 mV is obtained with the TEA1062 and 15 mV with the TEA1067. However, the typical values of both IC's are identical.

Fig. 17 shows the receiver amplifier gain as a function of V_{LN}.

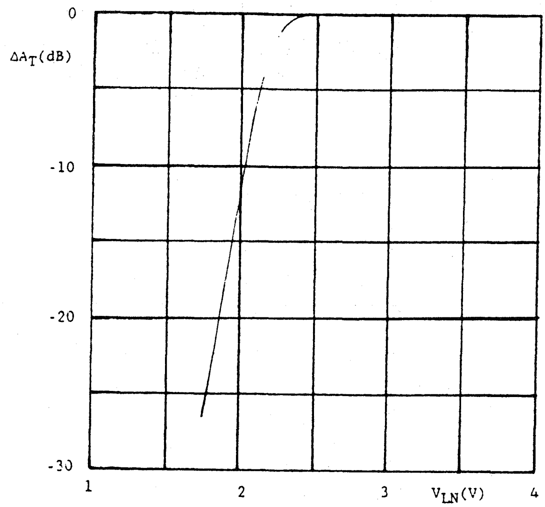


Figure 17: Receiver amplifier gain ΔA_T as a function of V_{LN}

The gain starts to decrease when V_{LN}= 2.3 V for the TEA1062, whereas for the TEA1067 the decrease starts for V_{LN}= 3.0 V. The local minimum in the receive gain of the TEA1067 for V_{LN}= 2.3 V has disappeared. For V_{LN}= 2.0 V the reduction in gain is approximately 13 dB for both devices.

2.6. Automatic gain control

The line-current-dependent gain-control characteristics of the TEA1062 and the TEA1067 are comparable. The gain-control range is 5.8 dB for the TEA1062 instead of 5.9 dB for the TEA1067. Refer to lit.1 for details.

2.7. MUTE input

The MUTE-input characteristics of the TEA1062 and the TEA1067 are similar. Refer to lit.1 for details.

2.8. PD input

The TEA1062 does not have a power-down (PD) input. This input is used in the TEA1067 to reduce the internal supply current during interruptions.

2.9. Anti-sidetone circuit

The anti-sidetone characteristics of the TEA1062 and the TEA1067 are similar. Refer to lit.1 for details.

3. PERFORMANCE

3.1. Impedance

The impedance of the circuit is shown in fig. 18. The dotted lines represent a balance return loss of 20 dB, 14 dB, and 8 dB (with respect to a 600 Ω impedance).

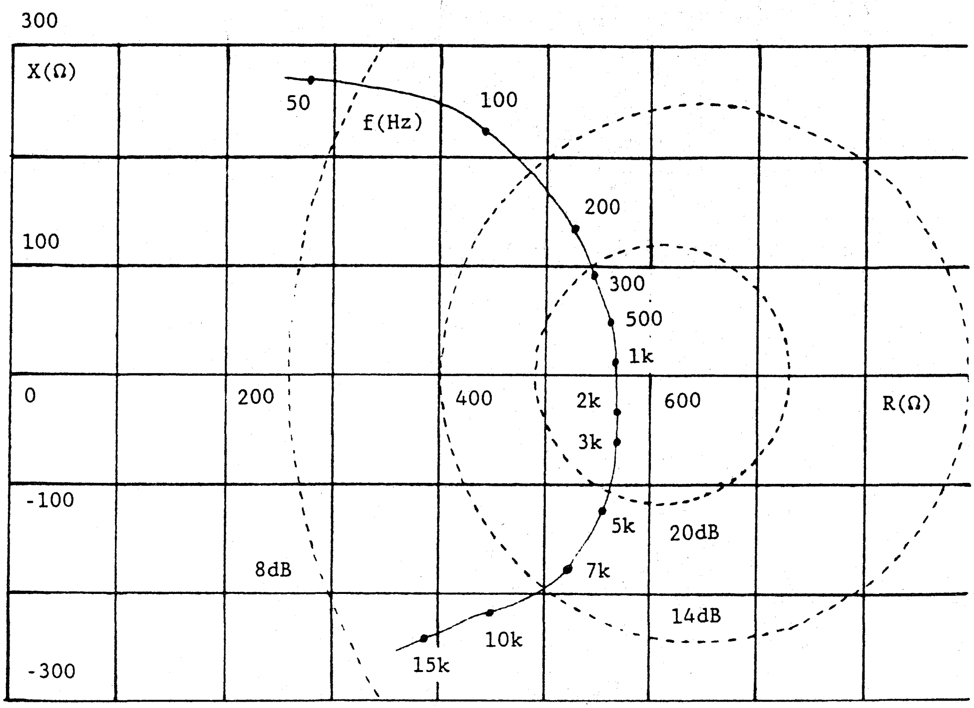


Figure 18: Polar plot of the impedance between the telephone lines

The BRL of the circuit is larger than 20 dB for the telephony audio frequency range. For a frequency of 1 kHz the impedance does not equal the 600 Ω real impedance. This is caused by the capacitor C17 of 1 nF between GAS1 and VEE.

3.2 Frequency characteristics

Fig. 19 shows the frequency response of the transmitting channel measured between the microphone inputs and the transmitter output at pin LN with a 600 Ω load.

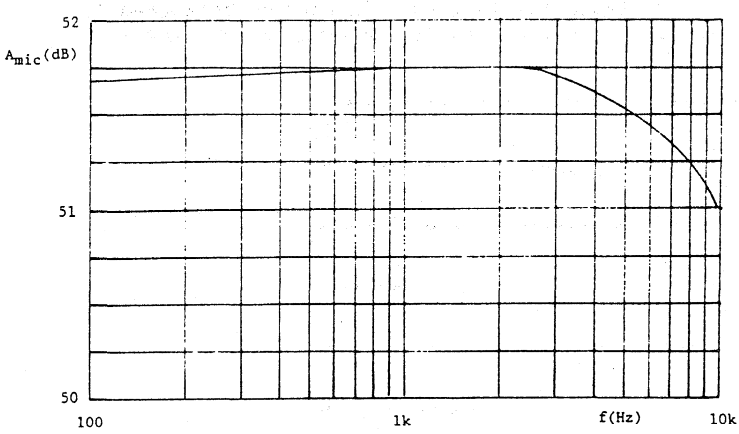


Figure 19: Frequency characteristic of the microphone amplifier

The microphone gain is set to 51.8 dB by R7 (68.1 kΩ). The upper cut-off frequency is approximately 21 kHz (predominantly determined by the timeconstant R7C6).

Fig. 20 shows the frequency response of the receiving channel measured between the input IR and the output QR loaded with 150 Ω.

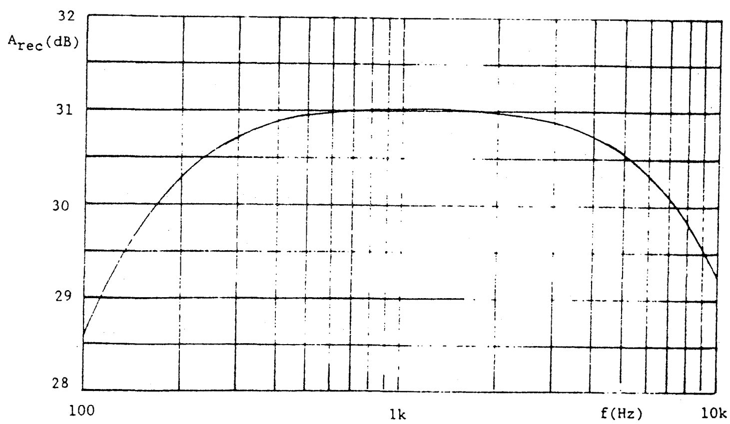


Figure 20: Frequency characteristic of the receiving channel measured between IR and QR

For 1 kHz the gain is 31 dB. The upper cut-off frequency is approximately 14 kHz (determined by the timeconstant R4C4).

Fig. 21 shows the frequency response of the receiving channel measured between pin LN and the QR output loaded with 150 Ω. The transfer ratio is -1 dB at 1 kHz.
In this case the upper cut-off frequency (9.5 kHz) is partly determined by the time constant R4C4 and partly by the cut-off frequency of the anti-sidetone circuit.

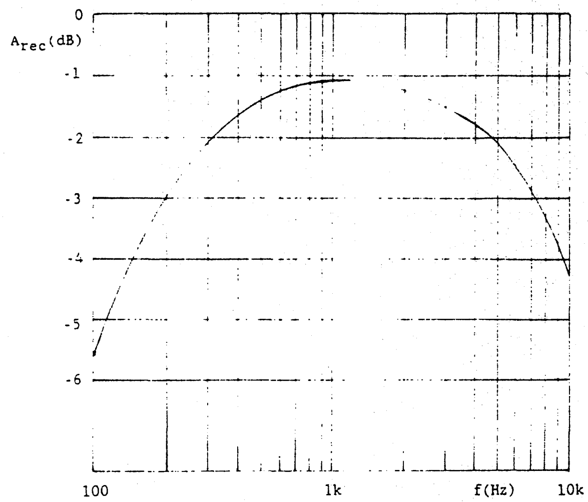


Figure 21: Frequency characteristic of the receiving channel measured between LN and QR

3.3. Noise

The psophometrically-weighted (P53-curve) noise level measured on pin LN with a 600 Ω load is given as a function of the microphone gain in fig. 22. The microphone input is loaded with a 200 Ω and a 8.2 kΩ resistor.

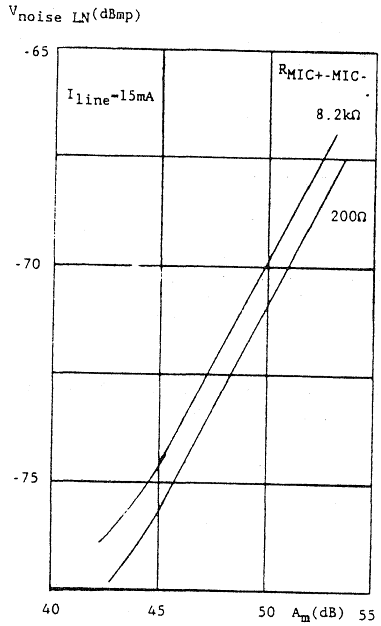


Figure 22: Psophometrically weighted (P53-curve) noise on pin LN as a function of the microphone gain

For a 200 Ω microphone input load and a microphone gain of 52 dB the TEA1062 has a noise level of -69 dBmp and the TEA1067 of -72 dBmp. The difference between a 200 Ω and 8.2 kΩ microphone load is 1dBmp for the TEA1062 and 3dBmp for the TEA1067.

Fig. 23 shows the noise output level on pin QR loaded with a 300 Ω resistor as a function of the receiver gain. For a gain of 31 dB the noise output voltage is 50 μV_{rms}, which is comparable to the TEA1067 noise level.

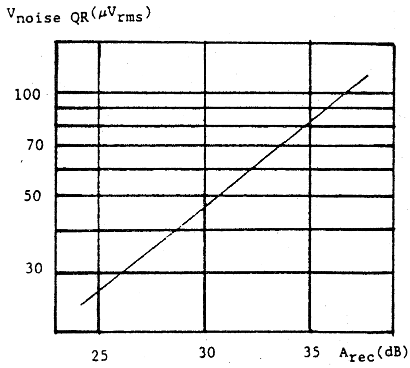


Figure 23: Psophometrically weighted (P53-curve) noise level on pin QR as a function of the receiver gain

4. REFERENCES

1) Components publication, "TEA1060 family: Versatile speech/transmission ICs for electronic telephone sets: Designers' guide", P.J.M. Sijbers, 12nc. 9398 341 10011

2) Components datasheet TEA1062 1989

3) PCALE Application Note ETT/AN89006, "Documentation for the TEA1062 printed circuit board CAB3422", P.T.J. Biermans and P.J.M. Sijbers

APPLICATION NOTE Nr	ETT/89009
TITLE	Application of the versatile speech/transmission circuit TEA1064 in full electronic telephone sets
AUTHOR	F. van Dongen, P.J.M. Sijbers
DATE	August 1989

Table of Contents:

List of tables and figures

1. INTRODUCTION

The TEA1064 speech transmission IC is based on the well-known TEA1067. Besides all features of the TEA1067 the TEA1064 also incorporates a dynamic limiter in the sending amplifier and an improved and more flexible supply point for peripheral circuits.
Performance of the mute, DTMF and AGC circuitry have been optimised further.

The application of this circuit is even more flexible than the rest of the TEA1060-family. A guide to show all the possibilities of the IC is very important. This report is intended as such a guide to provide all the necessary information to the set-designer. An extensive description of the IC is given and a flow chart giving the correct order to adjust parameters.

2. DESCRIPTION OF THE CIRCUIT

2.1. BLOCKDIAGRAM

The blockdiagram of the TEA1064 is shown in Fig. 1. The internal functions are:

- Voltage regulator with low voltage drop (internal reference voltage is 3.3 V). The voltage drop can be increased externally (to Vref=4.3 V max). The static resistance is adjustable.
- Supply for powering peripheral circuits with two options:
  * unregulated supply; regulated line voltage
  * stabilized supply; line voltage varies with peripheral supply current.
- Dynamic limiting of the sending signal (controlled by speech) prevents distortion and limits the maximum level of both the transmitted line signal and the sidetone.
- Microphone amplifier with a wide-range gain setting (44 to 52 dB) and a frequency roll-off with adjustable cut-off frequency.
- Very high-impedance microphone inputs (64kΩ) for accurate microphone matching with external resistors; suitable for all types of microphone transducers.
- DTMF input.
- Confidence tone in the earpiece during DTMF dialling.
- Transmitting output stage.
- Receiver amplifier with two complementary outputs suitable for magnetic, dynamic or piezoelectric earpieces; the amplifier has a wide gain setting range (20 to 45 dB) and adjustable cut-off frequency.
- Line loss compensation facility (line current dependent) for the microphone and earpiece amplifiers. The control curve has been optimised for a 600Ω feeding bridge and is adaptable to various exchange supply voltages.
- Mute input to inhibit both the earpiece amplifier and the microphone amplifier during dialling, and to enable the DTMF input and the confidence tone.
- Power-down input to minimize the internal current consumption of the IC during line interruptions with pulse dialling or register recall (flash); the voltage regulator capacitor is disconnected to prevent start-up delays after line interruptions to minimize the contribution of the IC to the shape of the current pulses during pulse dialling.
- Low-voltage circuit enabling parallel operation (> 1.7V).

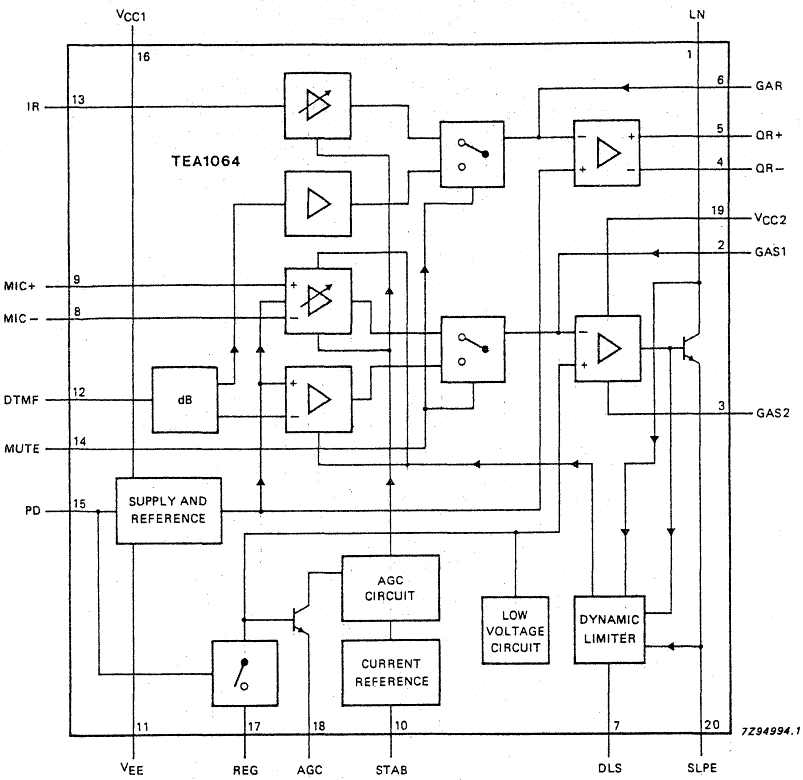
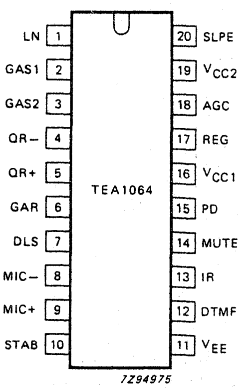


Fig. 1. Blockdiagram



1	LN	positive line terminal
2	GAS1	gain adjustment; transmitting amplifier
3	GAS2	gain adjustment; transmitting amplifier
4	QR−	inverting output, receiving amplifier
5	QR+	non-inverting output, receiving amplifier
6	GAR	gain adjustment; receiving amplifier
7	DLS	decoupling for transmit amplifier dynamic limiter
8	MIC−	inverting microphone input
9	MIC+	non-inverting microphone input
10	STAB	current stabilizer
11	V_{EE}	negative line terminal
12	DTMF	dual-tone multi-frequency input
13	IR	receiving amplifier input
14	MUTE	mute input
15	PD	power-down input
16	V_{CC1}	internal supply decoupling
17	REG	voltage regulator decoupling
18	AGC	automatic gain control input
19	V_{CC2}	reference voltage with respect to SLPE
20	SLPE	slope adjustment for DC curve/reference for peripheral circuits

Fig. 2. Pinning diagram

- Automatic disabling of the DTMF amplifier in extremely low voltage conditions.

The anti-sidetone circuit is implemented outside the IC by means of discrete components, thus allowing maximum flexibility of circuit design.
The pinning is shown in Fig. 2 together with a list of the pin functions. These abbreviations are used throughout the chapters that follow. Fig. 3 shows the basic application diagram.

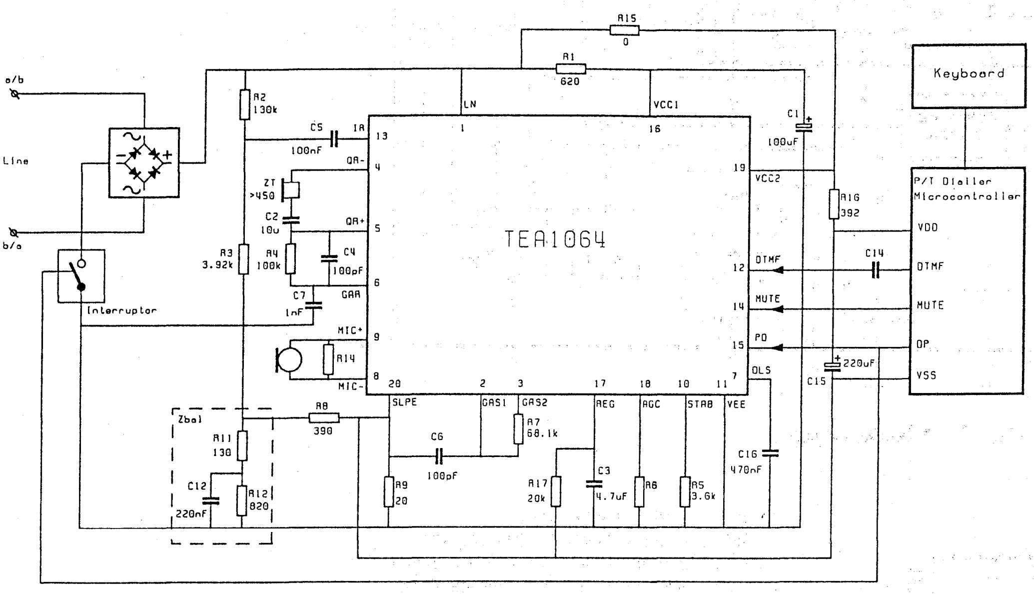


Fig. 3. Basic application of the TEA1064 with regulated line voltage (not including protection).
For the basic application giving stabilized supply voltage for peripherals the above circuit is changed as follows:
R15 must be 392Ω; the value of R16 is changed to 56Ω; the value of C3 is changed to 470 nF; R17 is removed.

2.2. SUPPLY AND SET IMPEDANCE

The IC is supplied with current from the telephone line. For effective operation of the telephone circuitry it must have a low resistance to d.c. and a high impedance to speech signals (300 to 3400 Hz). This is done by incorporating a voltage regulator in the IC in series with an artificial inductor. The total equivalent impedance of the circuit depends on the supply mode that is chosen (regulated line voltage or stabilized supply voltage).
The internal voltage regulator generates a temperature compensated reference voltage that is available between pins VCC2 and SLPE [$V_{ref} = V_{CC2} - V_{SLPE} = 3.3$ V (typ.)]. This internal voltage regulator must be decoupled by a capacitor between REG and VEE.

The reference voltage can be used:
* To regulate directly the line voltage (stabilized $V_{LN} - V_{SLPE} = V_{CC2} - V_{SLPE}$).
* To stabilize the supply voltage for peripherals. (Stabilized $V_{CC2} - V_{SLPE}$; varying $V_{LN} - V_{SLPE}$).

This means that two options are available for the supply. Both options have their own merits. The best choice between them depends on local PTT requirements and on the requirements imposed by the set designer. Some design considerations (features) are given below.

Regulated line voltage
- Same behaviour as the well known TEA1060/61/67/68.
- The line voltage does not depend on peripheral supply current.
- The supply capabilities for peripherals do not depend on the line current (when Iline > 11 mA).
- The supply voltage for peripherals decreases with peripheral supply current (internal resistance is in the order of 400 Ω).
- First order control loop for the voltage regulator; stability is always guaranteed.
- Requirements for set impedance (Balance Return Loss) can be met with high margin (the basic application with a 600Ω set gives a BRL=21 dB at 300 Hz and BRL=6dB at 50 Hz); increasing this margin is possible.
- High degree of freedom in the choice of set impedances; adjusting the set impedance is rather easy.

Stabilized supply voltage for peripherals
- Line voltage increases with supply current for peripherals.
- The supply capabilities for peripherals do not depend on the line current (when Iline > 11 mA).
- The supply voltage for peripherals is virtually independent of peripheral supply current (internal resistance is in the order of 50 Ω).
- Second order control loop for the voltage regulator; a compromise between stability and set impedance becomes necessary.
- As a consequence the requirements for set impedance are met with less margin at low frequencies (the basic application with a 600Ω set gives a BRL=17 dB at 300 Hz and BRL=3dB at 50 Hz)
- Less flexibility w.r.t. the choice of set impedance; adjusting the set impedance imposes a new compromise between BRL and stability (readjustment of some component values).

Both supply options are described in the following paragraphs.

2.2.1. REGULATED LINE VOLTAGE.

In this mode the VCC2 pin is connected to the LN pin. The general supply arrangement is shown in Fig. 4. This configuration gives a stabilized voltage across pins LN and SLPE (same as for TEA1060/61 TEA1067 and TEA1068 transmission IC's). Now this voltage between LN and SLPE is passed through a lowpass filter (consisting of R16 and C15) to provide a supply to the peripheral circuits ($VDD-V_{SLPE} = Vp$). The peripheral supply voltage Vp is independent of the line current (provided that Iline > 11 mA) and depends only on the peripheral supply current (internal resistance is equal to R16).

REMARK: The reference ground level for the peripheral circuitry is SLPE and <u>not</u> VEE!

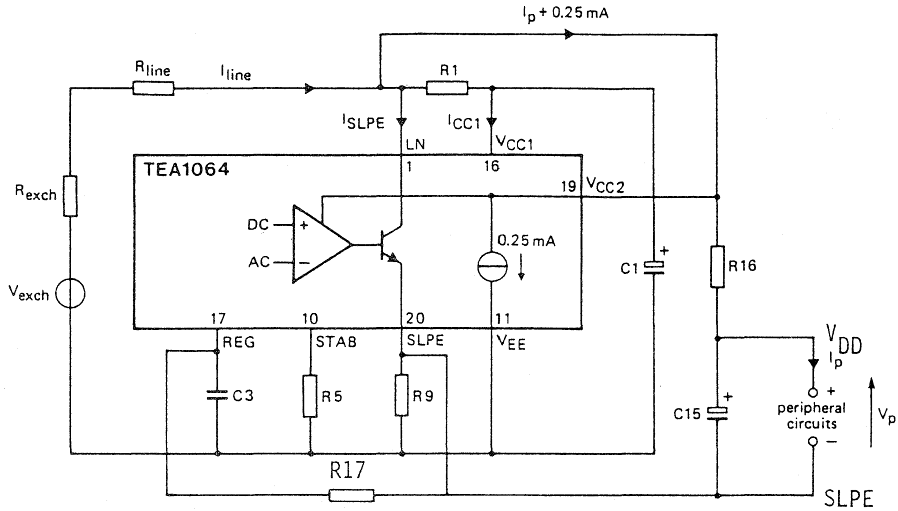


Fig. 4. Application with regulated line voltage (stabilized $V_{LN}-V_{SLPE}$). The voltage $V_{LN}-V_{SLPE}$ is fixed to Vref=4.3±0.35V with R17=20kΩ; Vref=3.3±0.25V without R17.
The line voltage:

$V_{LN} = V_{LN}-V_{SLPE} + (I_{SLPE}*R9)$

$V_{LN} = Vref + ([Iline-I_{CC1}-0.25mA]*R9)$

$V_{LN} \simeq Vref + ([Iline-1.5mA]*R9)$

The equivalent impedance of the circuit is shown in Fig. 5. The value of the artificial inductor Leq = Rp.R9.C3. The value of C3 also determines start-up time of the DC-voltage regulator and has been chosen such that the voltage regulator starts-up as soon as the VCC1 smoothing capacitor and the peripheral supply smoothing capacitor C16 have been charged. If necessary, the value of Leq should be adjusted by changing C3 (taking into account a different start-up time) rather than changing the value of R9, because the latter influences several other parameters (as described in 2.2.3.).

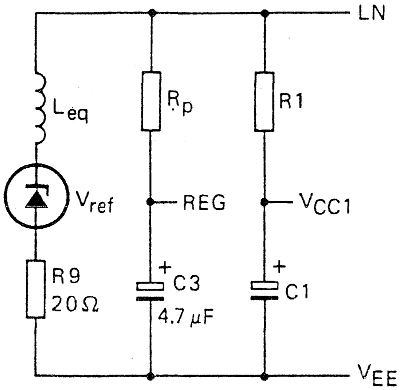


Fig. 5. Equivalent impedance between LN and VEE in the application with regulated line voltage (stabilized V_{LN}-V_{SLPE}):
R15=0Ω
Leq=C3*R9*Rp
Rp=15.5kΩ

The impedance of the whole circuit, to audio frequencies, is mainly determined by the value of R1//Rp. The value of the artificial inductor Leq determines the margin with which the BRL requirements at low frequencies are fulfilled.
The network R1C1 provides a smoothed voltage VCC1 for the IC itself (typically I_{CC1} = 1.2 mA at VCC1 = 2.8 V). The typical internal current consumption I_{CC1} as a function of VCC1 is shown in Fig. 6.

The impedance of the network between LN and SLPE is not seen in this equivalent impedance, because the output stage of the TEA1064 compensates it. See Ref. 1 for a complete analysis of the influence of a load between LN and SLPE on transmission parameters.

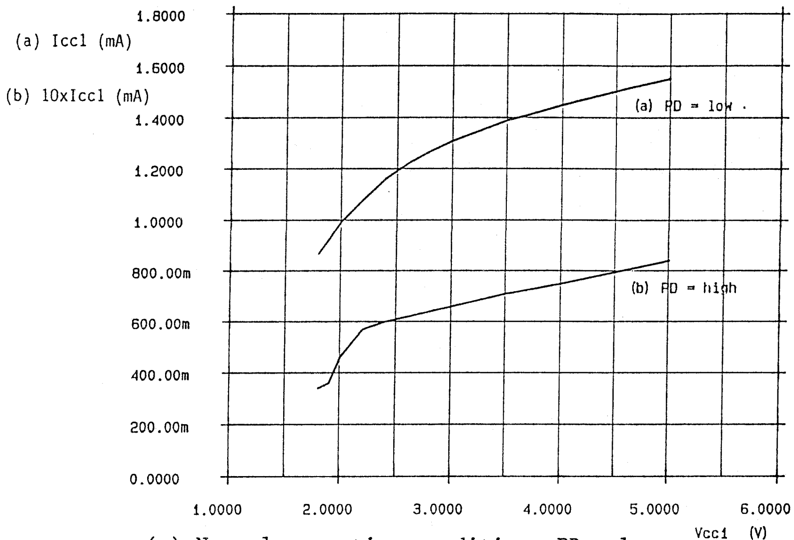
DC characteristics

The direct current which flows into the set is determined by (see Fig. 4):
* The exchange supply voltage (Vexch)
* The resistance or the feeding bridge (Rexch).
* The resistance of the subscriber line (Rline)
* The DC-voltage across the subscriber set, including the polarity guard.

If the line current exceeds (I_{CC1} + 0.25 mA) then the voltage regulator diverts the excess current through pin LN to SLPE. The equation for the voltage drop V_{LN} across the circuit, is given in Fig. 4. This equation can only be used for line currents exceeding I_{TH} (the threshold current of the low-voltage range, typ. 9 to 11 mA). Note that the voltage drop V_{LN} does not depend on the current flowing to the peripheral circuitry.

The current I_{SLPE} is normally much greater than the sum of I_{CC1} + 0.25 mA. Therefore the equivalent circuit for DC-conditions is that of a voltage regulator diode in series with resistor R9. The voltage drop across the voltage regulator diode can be set by means of a resistor R17 between REG and SLPE, between 3.3V (without R17) and 4.3V with R17=20kΩ (see 2.2.3.). The typical DC-voltage V_{LN} as a function of line current is shown in Fig. 7, with R17 = 20 kΩ and without R17. The slope of the DC-characteristic is determined by R9. The optimum value for R9 is 20 Ω, as used in the basic application circuit (Fig. 3). The effect of any change in R9 is described in 2.2.3.

It is recommended to use R17=20kΩ (to obtain the maximum value for Vref) in the basic application to ensure sufficient supply possibilities for peripheral circuitry and to guarantee the best possible performance of the speech amplifiers.



(a) Normal operating condition; PD = low.
(b) Power down condition; PD = high.

Fig. 6. Internal current consumption I_{CC1} as a function of VCC1.

Supply for peripheral circuits

The voltage available between VCC2 and SLPE (= V_{LN}-V_{SLPE}) can be used to power peripheral circuits via a lowpass filter R16, C15. The value of R16 and the level of DC voltage V_{LN}-V_{SLPE} determine the supply capabilities. In the basic application R16 = 392 Ω and C15 = 220 uF. The supply voltage as a function of supply current is shown in Fig. 8. The supply capability depends on the DC voltage V_{LN}-V_{SLPE}= V_{CC2}-V_{SLPE} that can be adjusted by means of a resistor R17 between pins REG and SLPE (See 2.2.3 and Fig. 21). To increase the supply capabilities further, the value of R16 can be decreased.
For Philips CMOS telephony dialers, the minimum supply voltage is 2.5 V. When a battery is used for memory retention, a diode should be connected between VDD and the supply pin of the peripheral circuit. Taking into account a forward voltage drop for a Schottky diode, the minimum value of Vp is about 2.9 V. Fig. 9 shows the typical supply current Ip as a function of R16 with Vp=2.9V and Vp=2.5V in two situations: with R17=20kΩ and without R17.

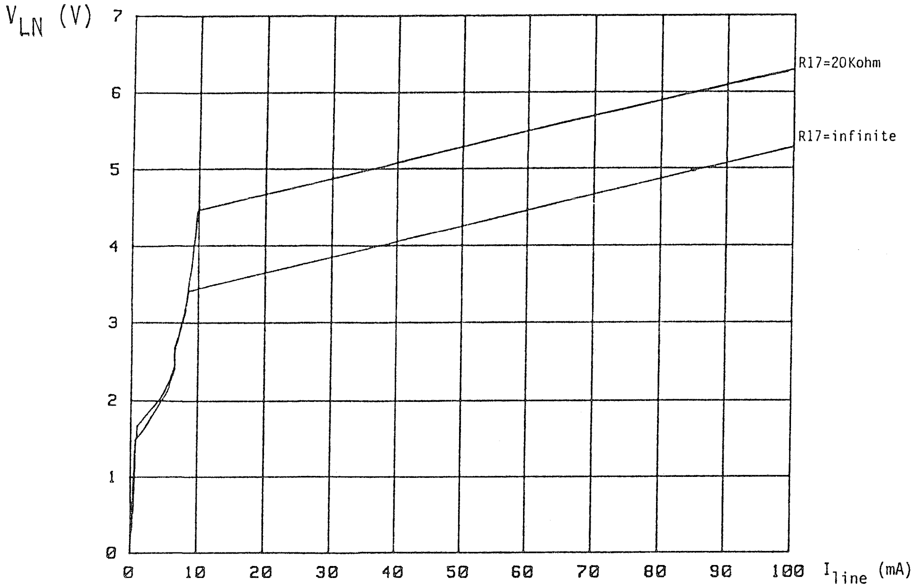


Fig. 7. DC characteristics (application with regulated line voltage).

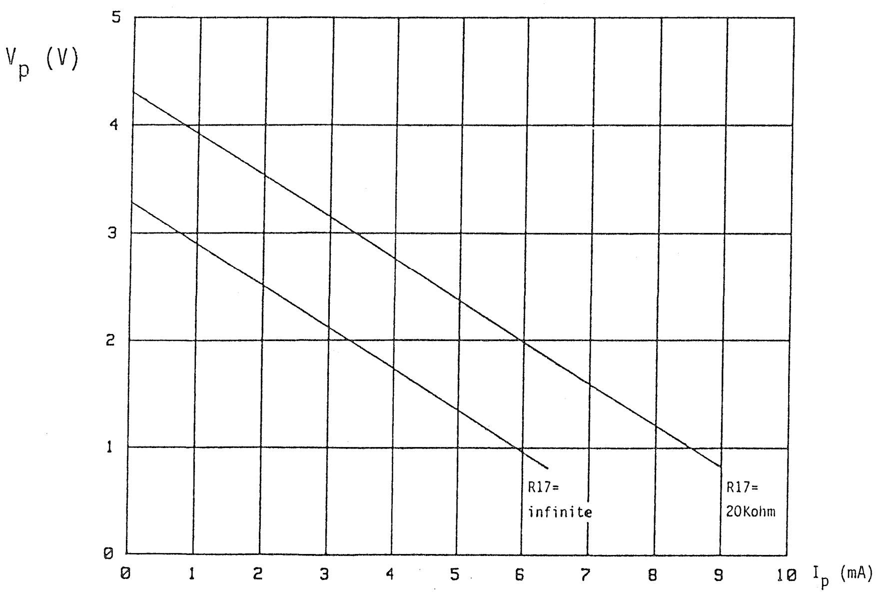


Fig. 8. Peripheral supply; Vp as a function of Ip (application with regulated line voltage).

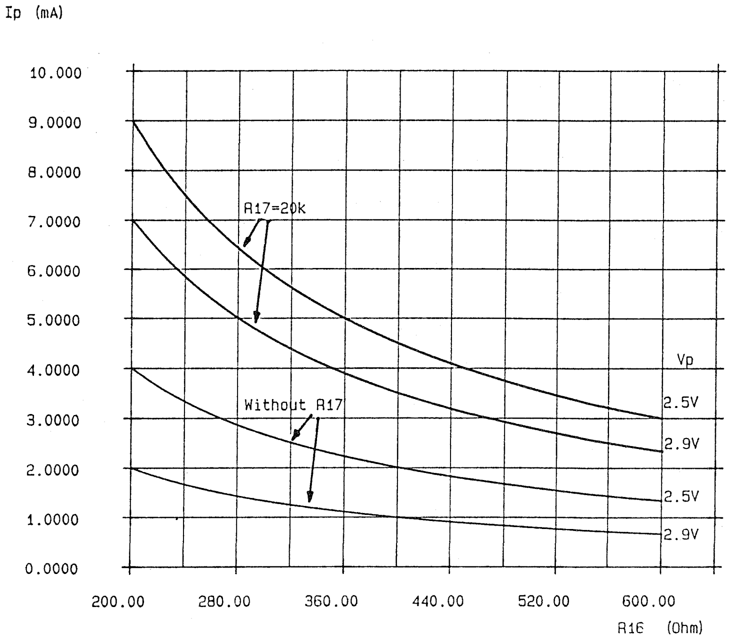


Fig. 9. Peripheral supply current as a function of R16.
(Application with regulated line voltage).

Keep in mind that the maximum possible sending level on the line at low line current will decrease by lowering R16. Also the peripheral supply current Ip has influence on the maximum possible sending level.

Maximum sending level

Two mechanisms limit the maximum possible sending level on the line: limitation by current or limitation by voltage.
At low line current where current limitation occurs, the maximum output swing on the line is influenced by R16. At high line currents the maximum sending level is limited by the DC voltage V_{LN}-V_{SLPE}. In both these situations, the internal dynamic limiter in the sending channel prevents distortion when the microphone input is overdriven (See chapter 2.4 and Appendix A.3.).
The maximum peak to peak output swing on the line in sending direction for the basic application (R16 = 392Ω) is shown in Fig. 10 without R17 and with R17=20kΩ. The areas where the signal level is limited by current or voltage can be distinguished clearly.
Practical values for R16 are from 200 to 600 Ω. The maximum sending level with R16 = 200 Ω is shown in Fig. 11.

Calculation of the maximum possible sending level can be done as explained in 2.4.2. and Appendix A.3.

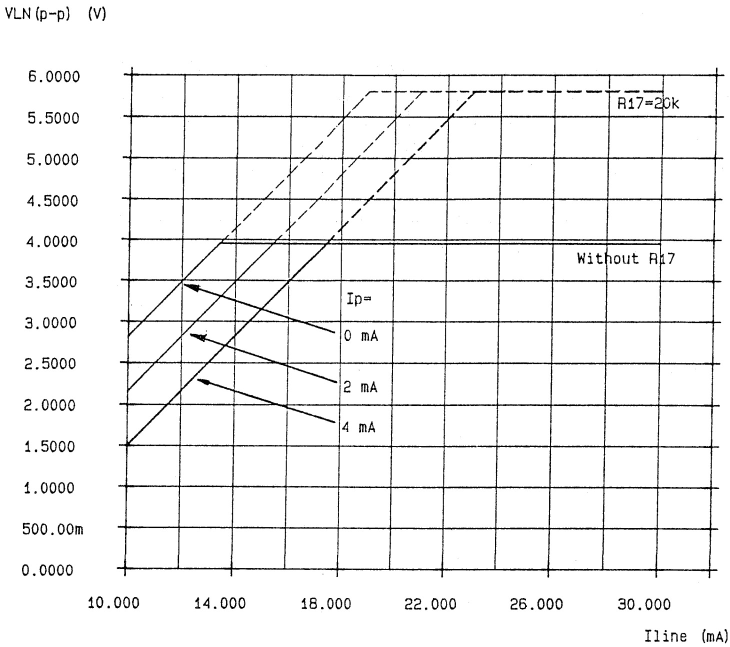


Fig. 10. Maximum AC sending level on the line as a function of line current with peripheral supply current as a parameter; (application with regulated line voltage; R15=0Ω; R16=392Ω).

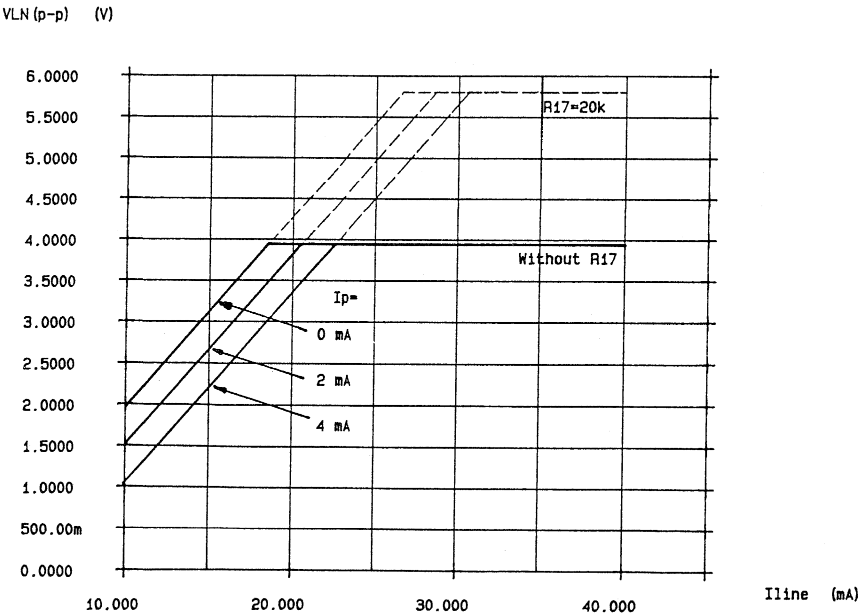


Fig. 11. Maximum AC sending level on the line as a function of line current with peripheral supply current as a parameter; (application with regulated line voltage; R15=0Ω; R16=200Ω).

Set impedance higher than 600 Ω.

If a set impedance higher than 600 Ω is required (which is generally the case with complex set impedance), then R1 (See Fig. 3) has to be replaced by the wanted impedance (keep in mind that the anti-sidetone bridge has to be rebalanced). In the application with regulated line voltage, this has no influence on the peripheral supply possibilities. Also the rest of the application is virtually not affected. However the supply possibilities of VCC1 (e.g. for active microphones; see 2.2.4.) are affected and therefore it is recommended to increase the voltage drop by means of R17 to the maximum allowed value.

2.2.2. STABILIZED PERIPHERAL SUPPLY VOLTAGE.

This configuration is shown in Fig. 12. It provides a stabilized voltage (of 3.3 V without R17) across pins VCC2 and SLPE for peripheral circuits. The DC voltage V_{LN} now varies with the peripheral supply current.
The V_{CC2}-V_{SLPE} supply must be decoupled by means of capacitor C15. The voltage regulator control loop must be completed by resistor R15 between LN and VCC2. The control loop of the voltage regulator now is a second order loop and therefore resistor R16 is introduced (connected between VCC2 and SLPE in series with C15) for stable loop operation.

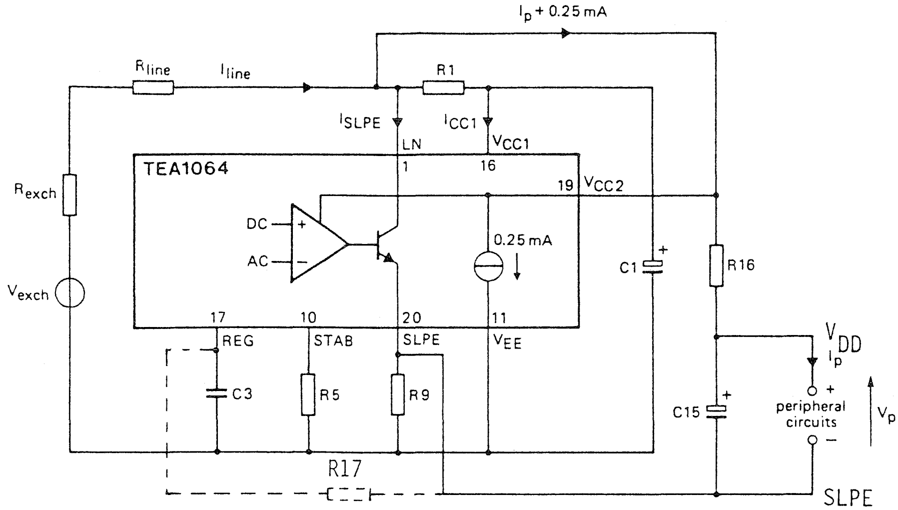


Fig. 12. Application with stabilized supply voltage for peripheral circuits; R15=392Ω; R16=56Ω; C3=470nF.
The DC line voltage on LN is:

$V_{LN} = V_{LN}\text{-}V_{SLPE} + (I_{SLPE}*R9)$

$V_{LN} = V_{CC2}\text{-}V_{SLPE} + V_{LN}\text{-}V_{CC2} + (I_{SLPE}*R9)$

$V_{LN} = Vref + ([Ip+0.25*10^{-3}]*R15)+([Iline-I_{CC1}-0.25mA]*R9)$

$V_{LN} \approx Vref + ([Ip+0.25*10^{-3}]*R15)+([Iline-1.5mA]*R9)$

where:
Vref is the internal reference voltage between VCC2 and SLPE; (can be adjusted with R17; Vref = 3.3V typ. without R17).

The ratio R15/R16 is now important for stable loop operation with sufficient phasemargin and for satisfactory set impedance. The optimum value of R15/R16 depends on the actual value of the set impedance (determined by R1) and the value of the voltage regulator capacitor at pin REG (C3).

<u>Application with 600Ω set impedance</u>

It has been calculated that with R1 = 619 Ω, C3 = 470nF, R15 = 392 Ω and C15 = 220 uF, R15/R16 must be smaller than 9.5 for stable loop operation with sufficient phase margin, and R15/R16 must be higher than 4 for satisfactory set impedance at the lowest audio frequency (BRL>=14dB at 300Hz).

In the basic application R15/R16=7 resulting in R16 = 56 Ω.

The equivalent impedance of the circuit between LN and VEE is calculated in Appendix A.1. For audio frequencies the equivalent impedance can be simplified as shown in Fig. 13.
So for the basic application Req = 124kΩ and Leq = 1.17 H. The impedance of the total circuit is therefore mainly determined by R1 and Leq (with f>300Hz).

$$Req = Rp\left(\frac{R15}{R16} + 1\right)$$

$$Leq = C3 * R9 * Req$$

with Rp = 15.5kΩ

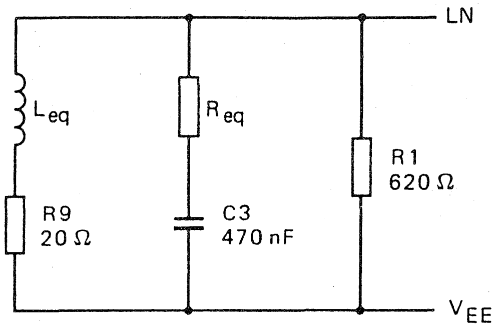
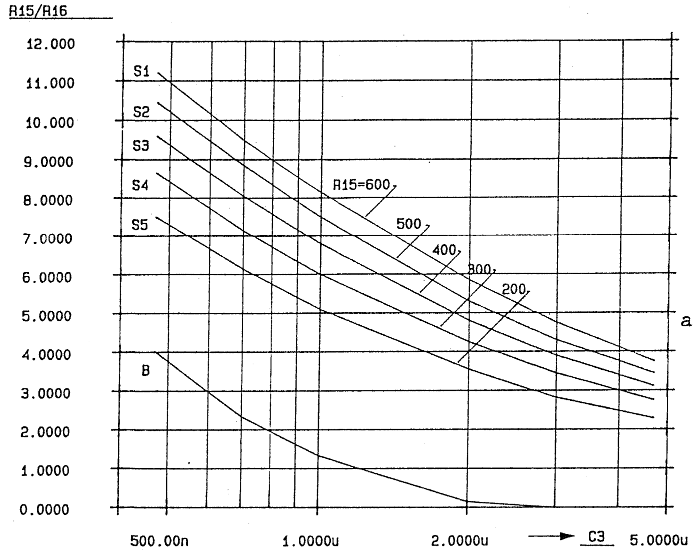


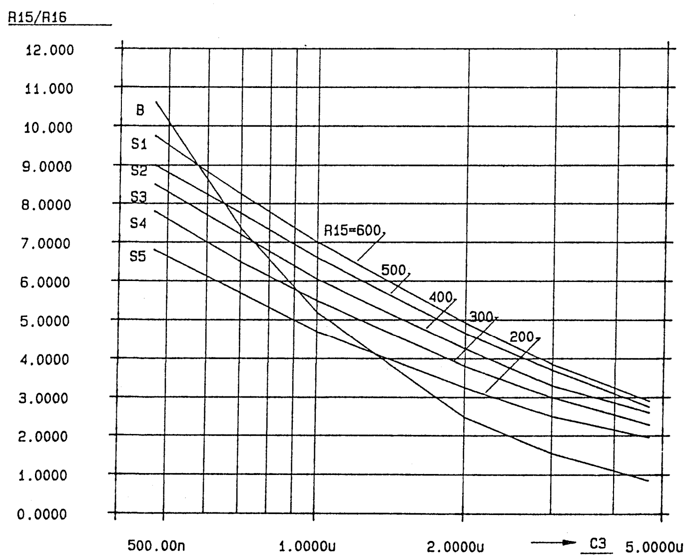
Fig. 13. Equivalent impedance between LN and VEE at f>300Hz in the application with stabilized supply voltage for peripheral circuits.

DC start-up time is now mainly determined by the timeconstant (R15+R16+R9)*C15 and partly by Rp*C3 (Rp being an internal resistor of 15.5kΩ connected between VCC2 and REG).

The network R1C1 provides a smoothed voltage VCC1 for the IC itself (typically I_{CC1} = 1.2 mA at VCC1 = 2.8 V). Typical I_{CC1} as a function of VCC1 is shown in Fig. 6.
The impedance of the network between LN and SLPE is not seen in the equivalent impedance between LN and VEE, because the output stage of the TEA1064 compensates it (provided that the choice of R15/R16 and C3 has been made correctly). Please see Ref. 1 for a complete analysis of the influence of a load between LN and SLPE on transmission parameters. Appendix A.1. gives the calculation of set impedance for the TEA1064 with stabilized supply.



a) R1=600Ω



b) R1=400Ω

Curves

S1 to S5: Maximum value of R15/R16 with R15 as a parameter to guarantee stable operation (phase margin = 50°)

B : minimum value of R15/R16 for BRL=14dB at 300Hz with Zref=600Ω

Fig. 14. Allowed range for the ratio R15/R16 in the application with stabilized supply for peripherals; setimpedance Zref=600Ω; a) R1=600Ω; b) R1=400Ω.

Adjusting the artificial inductor Leq

If necessary the value of Leq can be adjusted by changing C3. Alternative is changing the value of R9, but this influences several other parameters (see 2.2.3.). Changing the value of C3 has influence on the DC start-up time and also on the stability (phase margin; see Appendix A.2.) of the voltage regulator control loop. Therefore the ratio R15/R16 must be adapted to the new value of C3.

For other values than C3=470nF and R15=392Ω, the ratio R15/R16 has to be chosen from Fig. 14a. This figure shows the required ratio R15/R16 as a function of C3. The curves marked with 'S' indicate the maximum value of R15/R16 (with R15 as a parameter) for a phase margin of 50 ^{o}. The curve marked with 'B' gives the minimum value of R15/R16 for a BRL=14dB at 300 Hz (with 600 Ω as a reference). For a selected value of C3, a ratio R15/R16 must be chosen that lies between one of the curves 'S' (depending on R15) and curve 'B'.

Decreasing R1

For sets (with 600 Ω impedance) requiring a low DC voltage drop it can be useful to decrease the value of R1 (keep in mind that the anti-sidetone bridge must be rebalanced). In this way the supply possibilities of the VCC1 supply are improved (see 2.2.4) and the output capabilities of the earpiece amplifier are increased. However maintaining a correct BRL necessitates readjustment of both C3 and the ratio R15/R16. Fig. 14b gives the necessary information to do this when R1=400 Ω. With R15=400 Ω practical values are C3= 2.2 uF and 2.3< R15/R16 <4; this results in a value of R16 between 184 Ω and 100 Ω.

DC-characteristics

The direct current which flows into the set is determined by (see Fig. 12):
* The exchange supply voltage (Vexch)
* The resistance or the feeding bridge (Rexch).
* The resistance of the subscriber line (Rline)
* The DC-voltage across the subscriber set, including the polarity guard.

If the line current exceeds (I_{CC1} + 0.25 mA) then the voltage regulator diverts the excess current through pin LN to SLPE. As the voltage V_{CC2}-V_{SLPE} is kept constant, the DC voltage drop between LN and SLPE now depends on the current for the peripherals Ip.

The equation for the voltage drop V_{LN} across the circuit, is given in Fig. 12. This equation can only be used for line currents exceeding I_{TH} (the threshold current of the low-voltage range, about 9 to 11 mA). Note that the voltage drop V_{LN} now depends on the current flowing to the peripheral circuitry.

The current I_{SLPE} is normally much greater than the sum of I_{CC1} + 0.25 mA. Therefore the equivalent circuit for DC-conditions is that of a voltage regulator diode with value Vref + ([Ip + 0.25*10-3]*R15) in series with resistor R9. The typical DC-voltage V_{LN} as a function of line current with Ip as a parameter is shown in Fig. 15a. The slope of the DC-characteristic is determined by R9. The optimum value for R9 is 20 Ω, as used in the basic application circuit (Fig. 3). If R9 is to be changed, see 2.2.3.

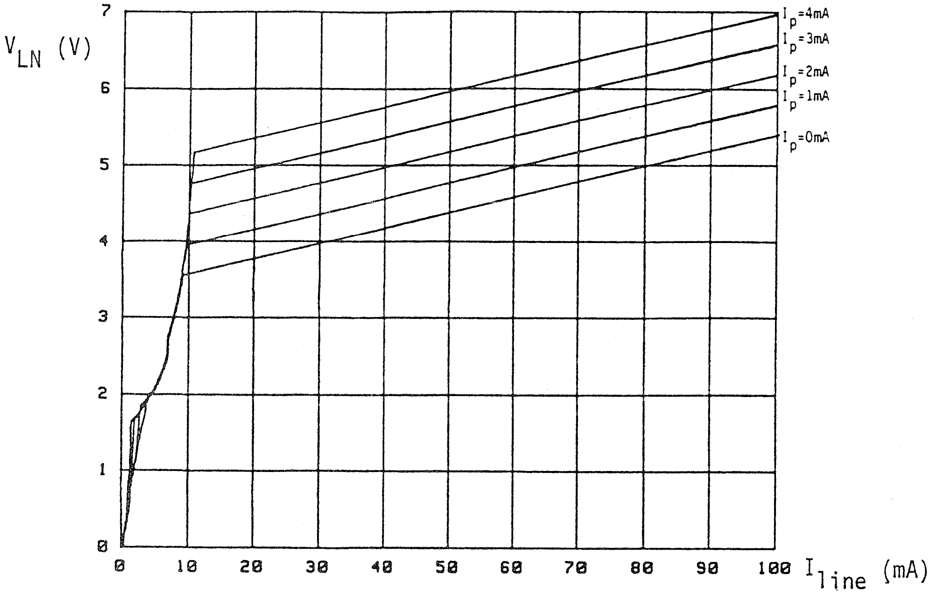


Fig. 15a. DC voltage drop V_{LN} as a function of line current in the application with stabilized supply; parameter is Ip. R15=392Ω, R16=56Ω.

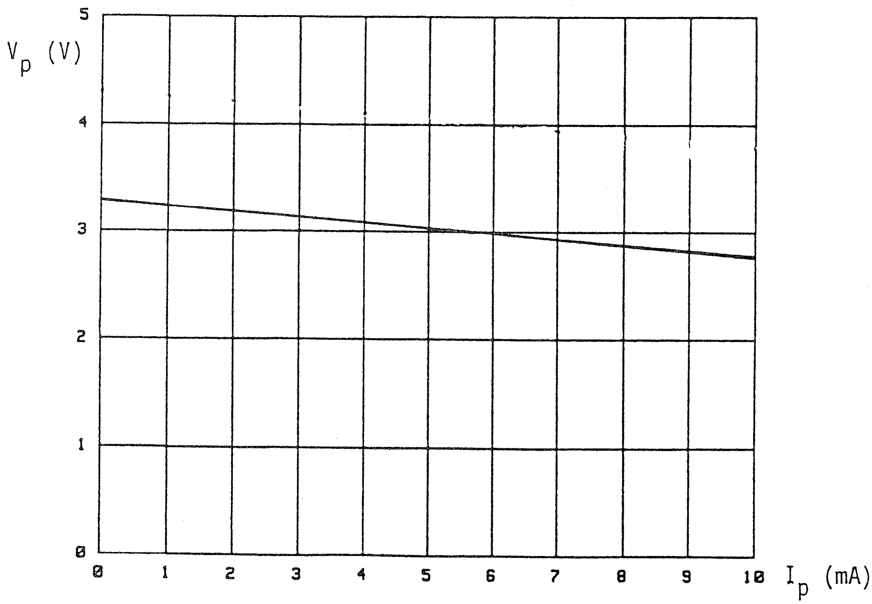


Fig. 15b. Peripheral supply voltage Vp as a function of peripheral supply current Ip with stabilized supply. R15=392Ω, R16=56Ω.

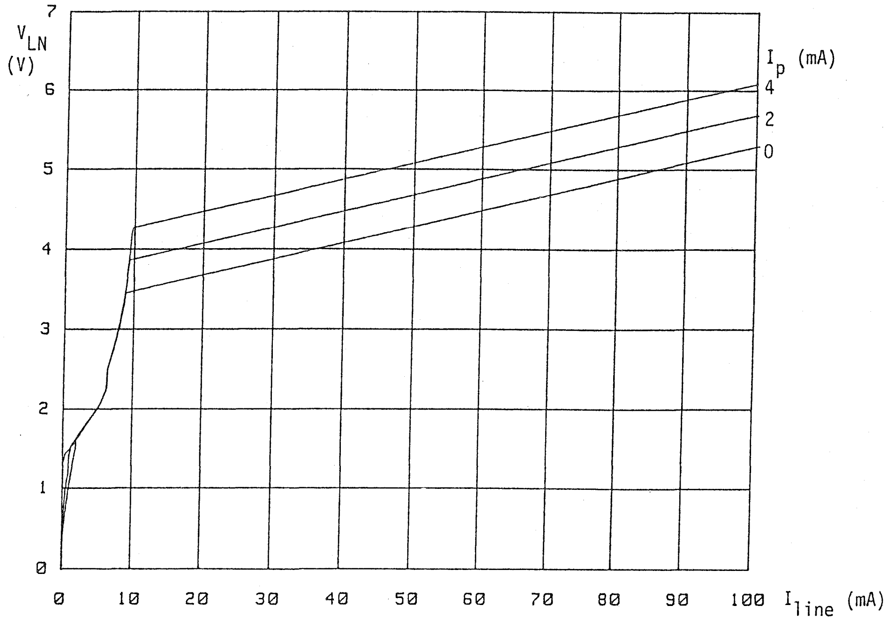


Fig. 16a. DC voltage drop V_{LN} as a function of line current in the application with stabilized supply; parameter is Ip; R15=200Ω, R16=33Ω.

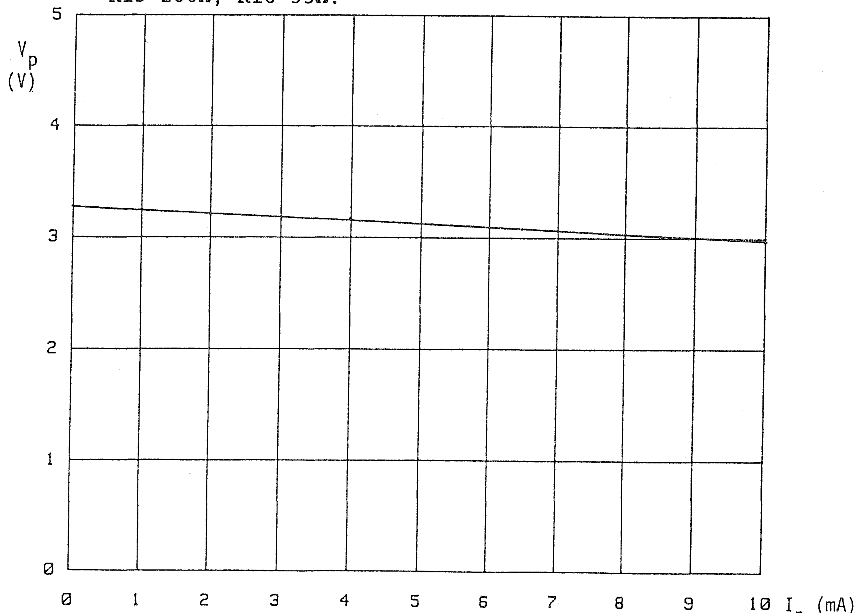


Fig. 16b. Peripheral supply voltage Vp as a function of peripheral supply current Ip with stabilized supply; R15=200Ω, R16=33Ω

Supply possibilities

Now the voltage between VCC2 and SLPE provides a stabilized supply (3.3 V without R17) for peripheral circuits via resistor R16 and smoothing capacitor C15. The supply possibilities with R16=56 Ω are shown in Fig. 15b. Keep in mind that the line voltage increases with the supply current Ip.
In case the maximum allowed line voltage is exceeded, or if the supply possibilities must be increased further, the value of R15 can be decreased and the ratio R15/R16 has to be reduced also (See Fig. 14a and 14b). Practical values for R15 are from 200 to 600 Ω.
The typical DC-voltage V_{LN} (with R15 = 200Ω and R16 = 33Ω) as a function of line current with Ip as a parameter is shown in Fig. 16a; the peripheral supply voltage Vp as a function of peripheral supply current is shown in Fig. 16b.

Keep in mind that the maximum possible sending level on the line at low line current is affected by lowering [R15 + R16]. Also the peripheral supply current Ip has influence on the maximum possible sending level.

Maximum sending level

Two mechanisms limit the maximum possible sending level on the line: limitation by current or limitation by voltage.
At low line currents where current limitation occurs, the maximum output swing on the line is influenced by R15+R16. At high line currents the maximum sending level is limited by the DC voltage V_{LN}-V_{SLPE}. As the voltage between LN and SLPE increases with peripheral supply current, the maximum output swing also does. In both these situations, the internal dynamic limiter in the sending channel prevents distortion when the microphone input is overdriven (See chapter 2.4.2 and Appendix A.3.).
The maximum peak to peak output swing on the line in sending direction for the basic application (R15 = 392Ω, R16 = 56Ω) is shown in Fig. 17 without R17. The areas where the signal level is limited by current or voltage can be distinguished clearly.
Practical values for R15 are from 200 to 600 Ω. The maximum sending level with R15 = 200 Ω and R16 = 33Ω is shown in Fig. 18.

Calculation of the maximum possible sending level can be done as explained in 2.4.2. For the complete calculation see Appendix A.3.

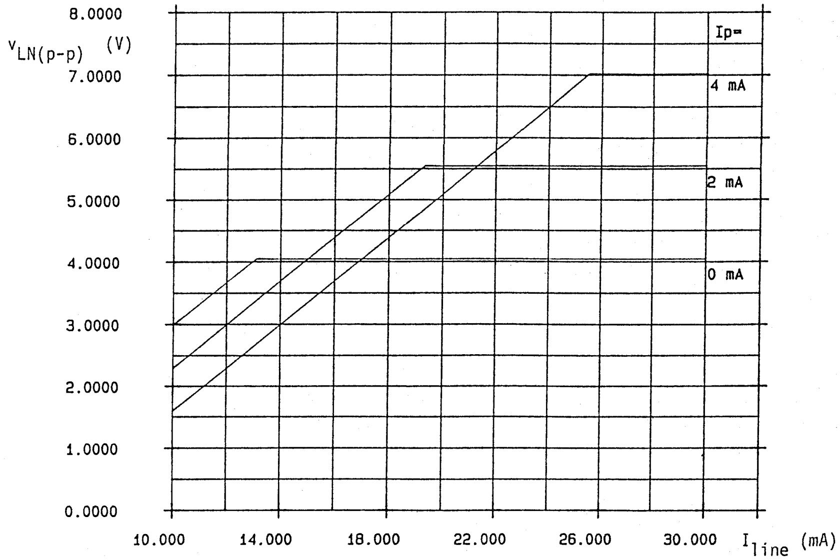


Fig. 17. Maximum AC sending level on the line as a function of line current with peripheral supply current as a parameter; (application with stabilized supply; R15=392Ω, R16=56Ω, without R17).

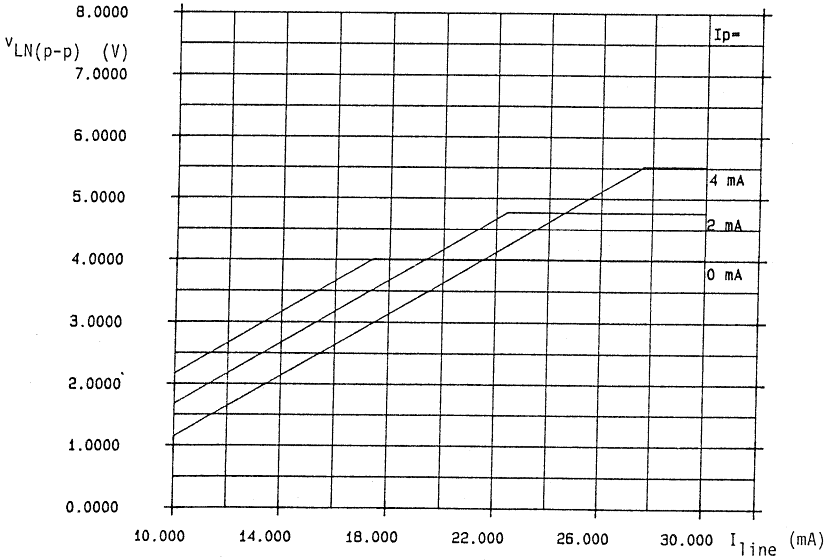


Fig. 18. Maximum AC sending level on the line as a function of line current with peripheral supply current as a parameter; (application with stabilized supply; R15=200Ω; R16=33Ω; without R17).

Set impedance higher than 600Ω.

If a set impedance higher than 600 is required (which is generally the case with complex set impedance), then R1 has to be replaced by the wanted impedance (and the anti-sidetone bridge must be rebalanced). In the application with stabilized supply voltage for peripherals, this results in a different choice of the optimum values for the ratio R15/R16 and the voltage regulator decoupling capacitor C3.
A compromise must always be made between BRL and stability of the voltage regulator control loop.

In Fig. 19 the supply system with a complex set impedance is represented. For the compromise between BRL and stability (represented by the value of phase margin) Ra is the most important component. Both Rb and Cb normally have a small influence on BRL at 300Hz and their influence on phase margin is negligible. Keep in mind that Rb and Cb of course have a considerable influence on the sending frequency characteristic (roll-off at 3400Hz) and normally Rb and Cb must have values such that both the sending frequency curve is within the limits and the BRL is acceptable at higher frequencies 3400Hz - 4000 Hz. In practice this means that Rb can have a very low value (sometimes zero) and Cb must be as small as possible.

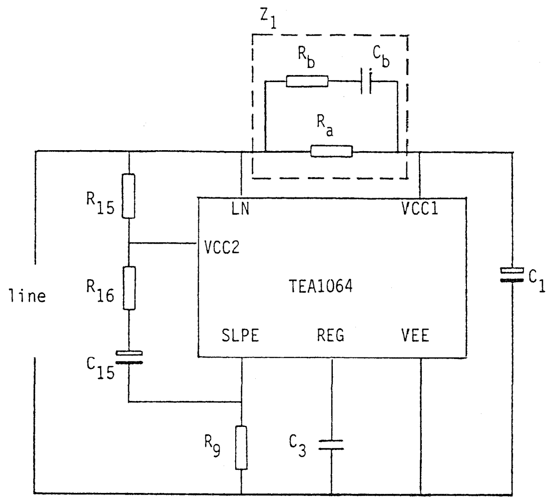


Fig. 19. TEA1064 with complex set impedance.

For the considerations w.r.t. the phase margin of the supply system Rb and Cb are neglected and only Ra is used. So in fact Ra can be considered as being equal to R1 of Fig. 12. For the considerations w.r.t. the BRL several options for the impedance Z1 are used with the German and British requirements as a reference.

If Ra is fixed at a value between 800 Ω and 1200 Ω, the value of R15/R16 as a function of C3 can be determined from Fig. 20a or 20b. In both figures the maximum value of R15/R16 as a function of C3 (for a phase margin of >50°) is

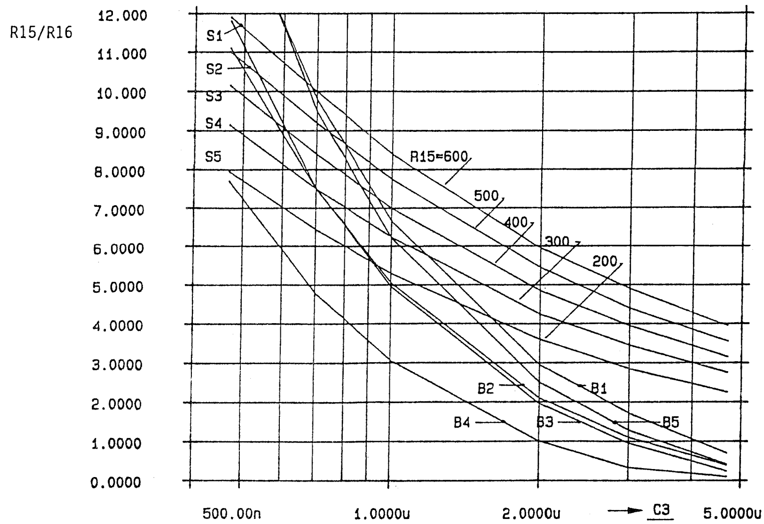
shown with R15 as a parameter (curves marked with 'S' in Fig. 20a and 20b). Remark: The curves for phase margin are only valid for the conditions as described in Appendix A.2.
The minimum value of R15/R16 as a function of C3 is determined by the BRL requirements (curves marked with 'B' in Figs. 20a and 20b). In Fig. 20a R15/R16 for a BRL=14 dB at 300 Hz with reference to the German complex set impedance (220Ω + 820Ω//115nF) is given as a function of C3 and with the set-impedance Z1 (see Fig. 19) as a parameter. In Fig. 20b R15/R16 for a BRL=16 dB at 200 Hz with reference to the British complex set impedance (370Ω + 620Ω//310nF) is given as a function of C3 and with the set-impedance Z1 (see Fig. 19) as a parameter. Table 1 gives an overview of the BRL-curves marked with 'B' as used in Figs. 20a and b.

<table>
<tr><th>Fig.</th><th>Curve</th><th>Setimpedance Z1</th><th>Conditions</th></tr>
<tr><td>20a</td><td>B1
B2
B3
B4
B5</td><td>800Ω
800Ω//33nF
1kΩ
220Ω+820Ω//115nF
1.2kΩ</td><td>German complex impedance:
Zref = 220Ω+820Ω//115nF
BRL=14dB at 300 Hz</td></tr>
<tr><td>20b</td><td>B1
B2
B3
B4</td><td>800Ω
800Ω//(470Ω+68nF)
1kΩ
370Ω+620Ω//310nF</td><td>British complex impedance:
Zref = 370Ω+620Ω//310nF
BRL=16dB at 200 Hz</td></tr>
</table>

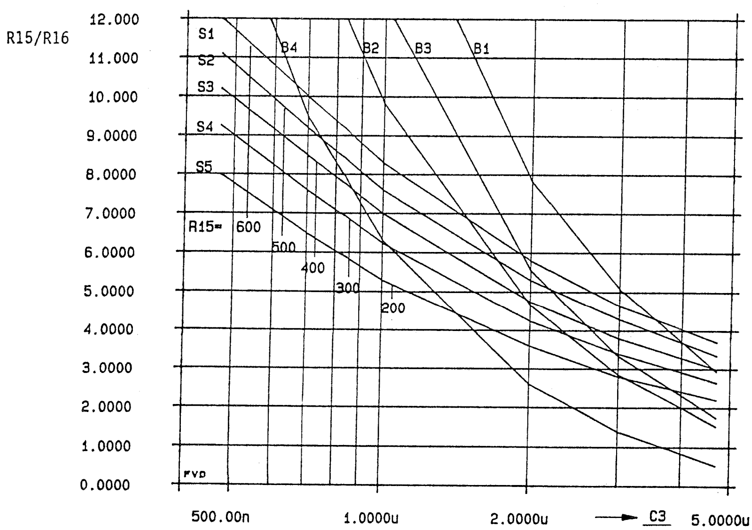
Table 1. Overview of BRL-graphs shown in Figs. 20a and 20b

The procedure is now as follows:
A set impedance Z1 (see Fig. 19) is selected (preferred is a value for Ra as low as possible to give correct BRL at the lowest frequency). A value for R15 can be chosen taking into account maximum DC line voltage (see 2.2.1. under DC-characteristics) and maximum transmit output swing (see 2.4.2. and Appendix A.3.). C3 must be chosen as small as possible to give a correct requirement in Fig. 20a or 20b for the minimum (curve 'B') and the maximum R15/R16 (one of the curves 'S'). So $(R15/R16)_{'B'} < (R15/R16)_{'S'}$. Then a value R16 must be chosen to give correct R15/R16.

Example:
A set for the German market has to be designed. Starting values are R15=400Ω and Z1=800Ω. As shown in Fig. 20a, the minimum value of C3 is then about 0.95uF (cross-point curve S3 and B1). In practice C3=1.5uF can be a good choice and this results in a requirement 4<R15/R16<5.5. However with a completely resistive setimpedance Z1=800Ω the BRL requirement at 3400 Hz cannot be met. Therefore Z1=800Ω//33nF is a better choice giving satisfactory BRL in the range 300 Hz to 3400Hz and allowing a wider range for R15/R16, which has to be between 3 and 5.5 in this case (curves B2 and S3 in Fig. 20a). So R16 must be chosen between 133Ω and 72Ω.



a) German complex setimpedance.
B1 to B5: minimum value of R15/R16 for a BRL=14dB at 300Hz with Zref=220Ω+820Ω//115nF and Z1 as a parameter (see Table 1)



b) British complex set impedance.
B1 to B4: minimum value of R15/R16 for a BRL=16dB at 200Hz with Zref=370Ω+620Ω//310nF and Z1 as a parameter (see Table 1).

Fig. 20. Allowed range for the ratio R15/R16 in the application with stabilized supply for peripherals;
a) German setimpedance Zref=220Ω+820Ω//115nF
b) British setimpedance Zref=370Ω+620Ω//310nF.
Curves S1 to S5: Maximum value of R15/R16 with R15 as a parameter to guarantee stable operation (phase margin = 50^o)

2.2.3. ADJUSTMENTS.

Several adjustments are possible and they are discussed below. They are applicable with both supply modes.

Increasing the slope of the DC-characteristic

This can be done by connecting a resistor between LN and node ([R1, R2] in Fig. 3). This resistor doesn't influence the set impedance, but it slightly affects the maximum output swing on the line. Another alternative in tone dialling applications, is to increase the value of the protection resistor R10. See chapter 4 and Fig. 41.

Changing of R9

In the basic application, the value of R9 is 20Ω. This is the optimum value for R9.
Another value for R9 can sometimes be necessary to rebalance the anti-sidetone circuit e.g. when a set impedance other than 600Ω is chosen. Any change in the value of R9 will cause:

* a change in gain of the microphone and DTMF amplifiers.
* a shift of the gain-control characteristic.
* a change of the sidetone (bridge balance).
* a shift of the low-voltage threshold current I_{TH}.
* a change in the DC line voltage.
* a change of the gradient (slope) of the DC-characteristic.

If a change in the value of R9 is necessary, the design procedure in appendix D should be followed.

Adjusting the DC-voltage drop

This will affect the maximum output swing of the sending and receiving amplifiers, and the supply voltage and current available for the peripherals. The voltage drop across the circuit can be increased by connecting a resistor R17 between pins REG and SLPE. This resistor sets the internal reference voltage of the voltage stabilizer (Vref = V_{CC2}-V_{SLPE}). However there will be a small change in the temperature coefficient and the set impedance and there will be a slight increase in the spread of the voltage drop. Fig. 21 shows how V_{CC2}-V_{SLPE} changes with the value of R17.

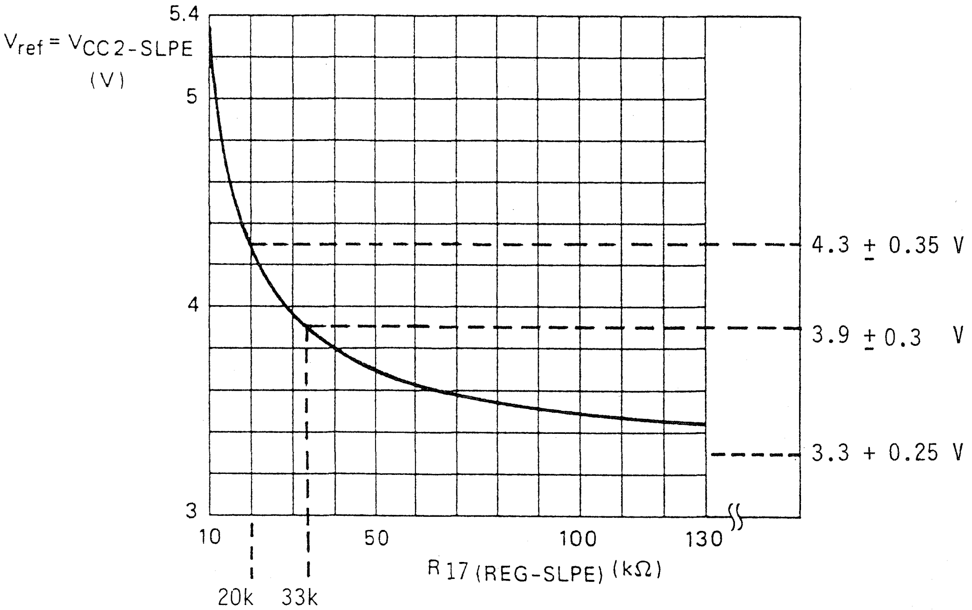


Fig. 21. Internal reference voltage V_{CC2}-V_{SLPE} = Vref as a function of R17 for line currents between 11 and 140 mA

2.2.4. SUPPLY PIN VCC1

The voltage available at VCC1 is normally used to supply only the internal circuitry of the TEA1064. However a small current can be drawn to supply external circuitry having as a ground reference VEE (e.g. an active microphone; see Fig. 22). **However circuits that provide interface signals to the TEA1064 inputs DTMF, MUTE or PD must be supplied between VCC2 and SLPE (from Vp).**
The current Im and the voltage VCC1 available from the internal supply pin VCC1 depend on the DC settings of the IC (Vref, R1, R9), on the line current available and on the required AC signal level at the line and receiver outputs.

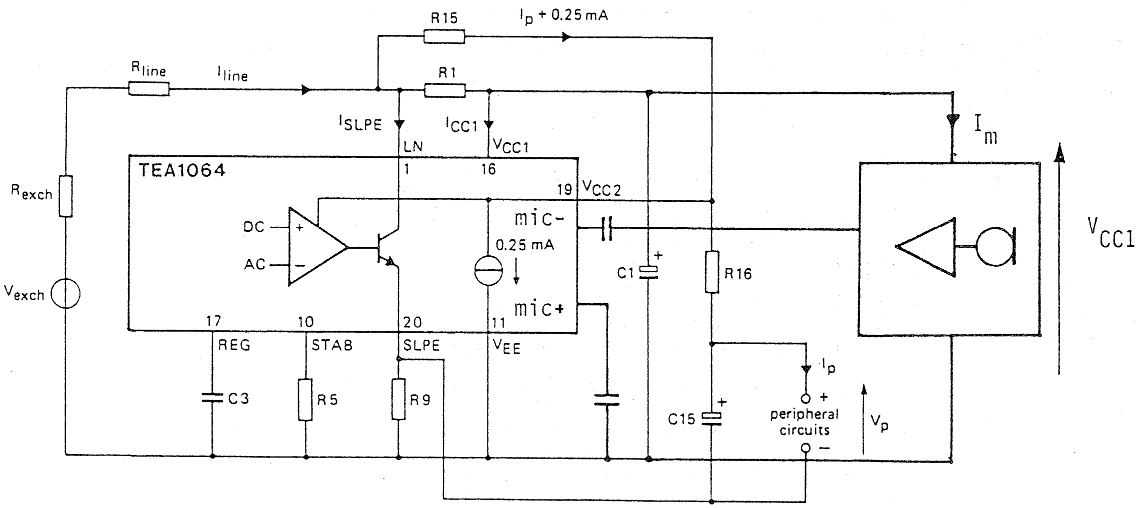


Fig. 22. Supply configuration using VCC1 to supply an active microphone

The minimum permitted voltage on VCC1 depends on the DC voltage at SLPE and on a voltage Vmin that consists of the sum of a DC-voltage of 2.1 V and the AC-peak voltage on SLPE.

In formula:

$$VCC1min \simeq Vmin + (Iline - 1.5*10^{-3}) * R9 \qquad [1]$$

with:

$$Vmin = 2.1 + \frac{\hat{v}LN * R9}{(R1//RL)} V \qquad [2]$$

being the minimum instantaneous voltage between VCC1 and SLPE with an AC-peak voltage $\hat{v}$LN on pin LN

and

R1 = resistor between LN and VCC1 determining the set-impedance.
RL = AC line impedance .
R9 = resistor between SLPE and VEE (slope of DC-curve).

Generally the following formula for the maximum permitted Im can be given:

$$Im_{max} = \frac{Vref - Vmin + (Ip+0.25*10^{-3})*R15}{R1 - R9} - (I_{CC1} + Irec) * \left(\frac{R1}{R1 - R9}\right) \qquad [3]$$

in which:

- Vref = the internal DC-reference voltage V_{CC2}-V_{SLPE} (adjustable with R17; see Fig. 21)
- Ip = the peripheral supply current taken from Vp
- I_{CC1} = internal DC current consumption of TEA1064 (See Fig. 6)
- Irec = internal current necessary to power the earpiece amplifier to realize an AC-peak voltage $\hat{v}T$ across the earpiece impedance RT

$$\text{asymmetrical drive:} \quad Irec_A = \frac{\hat{v}T}{RT*\pi} \qquad [4]$$

$$\text{symmetrical drive:} \quad Irec_S = \frac{\hat{v}T*2}{RT*\pi} \qquad [5]$$

For example:

Application with regulated line voltage (Fig. 3)

With R15=0, R17=20kΩ ---> Vref= 4.3V ±0.35V, R1=619Ω, R9=20Ω, RL=600Ω, RT=150Ω single ended drive, Iline = 15 mA;

a) Mute mode with vLN = 1Vrms ---> $\hat{v}$LN=1.414V; Irec can be supposed to be zero.

So it can be calculated from [2] that Vmin = 2.17 V and from [1] we find

$\boxed{VCC1_{min} \approx 2.44\ V}$

From Fig. 6 we can read that I_{CC1} = 1.18 mA with VCC1 = 2.44 V

From [3] we now find that with a typical Vref =4.3V $\boxed{Im_{max} = 2.34\ mA}$

b) Speech mode with vLN = 1.4 Vrms and vT = 150 mVrms

This results in $\hat{v}$LN = 2V and $\hat{v}$T = 212 mV; from [2] we find Vmin = 2.23 V; from [4]: IrecA = 0.45 mA

Now from [1] we find $\boxed{VCC1min \approx 2.5\ V}$ and from Fig. 6 this yields I_{CC1}=1.2 mA

From [3] we now find that with a typical Vref= 4.3V $\boxed{Im_{max} = 1.77\ mA}$

For comparison with the rest of the TEA1060 family see Fig. 10. of Ref. 2.

REMARKS:

- The limit on Im is imposed by the requirement to maintain at least the minimum permitted voltage between VCC1 and SLPE (minimum instantaneous voltage ; VCC1-V_{SLPE} is at least 2.1 V). If this condition is not met the maximum DTMF level on LN and the maximum receive level on the QR outputs will be limited; furthermore the receive gain will decrease.
- In an application with stabilized supply voltage for peripherals, the current Immax depends on the current Ip because Vref varies with Ip. Therefore Immax must be calculated in the condition with minimum Ip.
- In an application with low DC voltage drop, it can be useful to decrease the value of R1 (compromising between set-impedance [BRL] and VCC1 supply capabilities). See paragraph 2.2.2. if the application with stabilized supply is used and Ref. 2 (paragraph 'extending the supply capabilities' if the application with regulated line voltage is used.

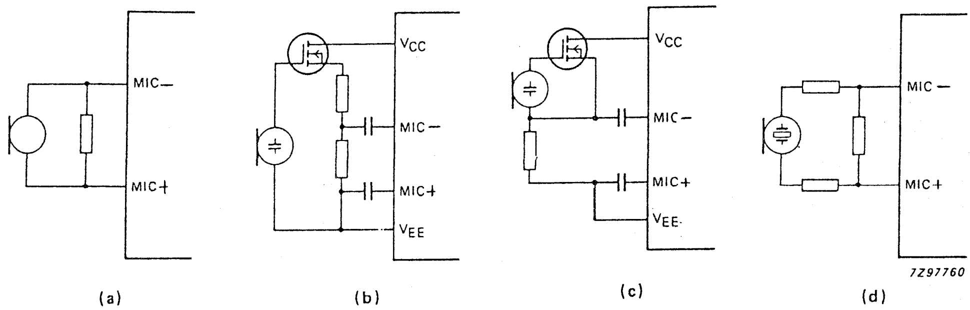
2.2.5. PARALLEL OPERATION WITH THE TEA1064

At line currents below the low-voltage threshold current I_{TH} (typ. 9 to 11mA; see Figs. 7 and 15a), the internal reference voltage is automatically adjusted to a lower value. At 2 mA a typical voltage drop of 1.7V is obtained. This facilitates operation of the circuit with more telephone sets connected in parallel, with line voltages inside the polarity guard down to a minimum of 1.7 V. The sending and receiving amplifiers will of course have reduced gain and reduced output swing capabilities in the low voltage range. Furthermore the peripheral supply capabilities will be reduced.

If a 200Ω conventional set is connected in parallel with the TEA1064, with 20mA available line current, the line voltage will drop to 3.2V. This will leave 4mA of line current at a voltage of 2V for the TEA1064, inside the polarity guard. The current used by the peripherals has no influence because this current will flow through R9.

2.3. MICROPHONE AMPLIFIER.

The TEA1064 has symmetrical very high impedance microphone inputs; this is identical to the TEA1067 and TEA1068. The input impedance is typically 64kΩ (2x32kΩ) with maximum tolerances of ± 20%. With this high input impedance it is possible to determine the matching of several microphone types very accurately by means of external components. The circuit is suitable for dynamic, magnetic or piezoelectric microphones with symmetrical drive; electret microphones with built-in source follower or preamplifier can be used in asymmetrical mode. The arrangements with several microphone types are shown in Fig. 23.



(a) Dynamic or magnetic microphone
(b) Electret-capacitor microphone with source follower
(c) Electret-capacitor microphone with a preamplifier
(d) Piezoelectric microphone

Fig. 23. Methods of connecting microphones.

The gain of the microphone amplifier (between the MIC inputs and the line output LN) is given by the following equation (referring to Fig. 3):

$$Am = 1.356 * \frac{R7 + rd}{R5*R9} * \frac{Ri*RL}{Ri + RL} \qquad [6]$$

where,

Ri = R1//Req , the dynamic impedance of the circuit; $Req=Rp(1 + \frac{R15}{R16})$

Rp = 15.5 kΩ

R_L = load resistance at LN during the measurement; normally 600 Ω

r_d = dynamic resistance of the internal circuitry (3.47kΩ)

R5 = 3.65kΩ; fixed external resistor determining the current of an internal current stabilizer.

If, for a practical circuit such as shown in Fig. 3, we insert in the above equation the following realistic values R7=68.1kΩ, R5=3.65kΩ, R9=20 Ω, R1=620 Ω and RL=600 Ω and R15= 0, R16=392Ω (regulated line voltage) then:

$20\log A_m = 52$ dB.

In the case of stabilized supply voltage (R15=392Ω, R16=56Ω), the gain increases slightly (0.15dB).

2.3.1. GAIN ADJUSTMENT

For various microphone sensitivities, the gain can be set between 44dB and 52dB by means of R7; this takes values between 25kΩ and 68.1kΩ. The microphone gain is shown as a function of R7 in Fig. 24.
It will be clear that any different choice of R9 (static resistance of the DC-characteristic) will directly influence the gain of the transmitting channel. The value of R9 also has influence on other parameters (See 2.2.3.). The value used in the basic application diagram is 20 Ω. If this value is to be changed, the consequences should be considered carefully and the design procedure as given in Appendix D must be followed.

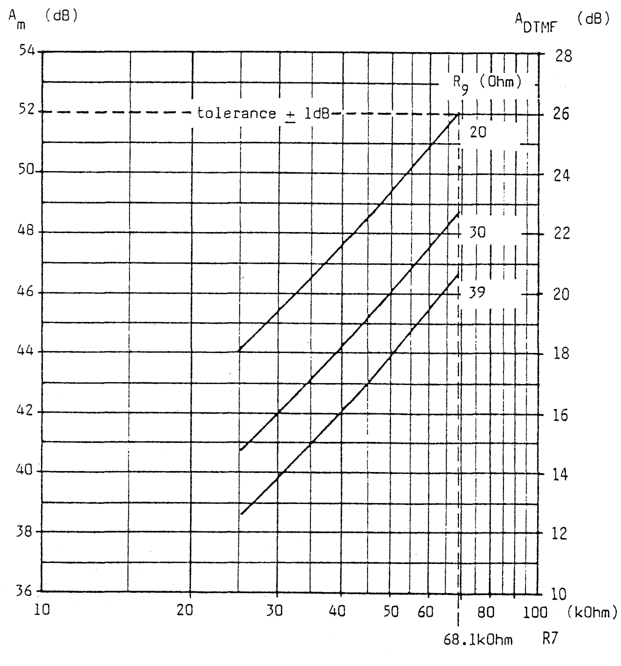


Fig. 24. Microphone gain and DTMF gain as a function of the value of R7.

2.3.2. NOISE, MAX. INPUT SIGNAL

Noise

To obtain optimum noise performance the microphone inputs must be loaded. The equivalent noise-voltage (psophometrically weighted; P53-curve) at the microphone input is typically 0,65 $uV_{(rms)p}$ with 8,2 kΩ across the microphone inputs. With 200 Ω across the inputs the equivalent noise at the input measures typically 0,45 $uV_{(rms)p}$.

Maximum input signal

The internal microphone preamplifier accepts signals up to 17 mV_{rms} for a 2% level of total harmonic distortion (THD=2%) because of the internal soft limiting. This means that the minimum possible gain of the microphone amplifier measured between the inputs and the line is 44dB with clipping of the line signal being determined fully by the transmit output stage. In case a lower gain is necessary the input signal must be attenuated before entering the preamplifier otherwise the input stage will be overloaded and causes extra distortion (soft clipping) of the line signal.

2.3.3. ASYMMETRICAL DRIVE

In case asymmetrical drive of the microphone inputs is used, the **MIC-** input should be used as a signal input. Care should be taken that both inputs MIC+ and MIC- see equal impedances to the common, otherwise residual line signals being present on the supply point (VCC1) will cause inaccuracy in gain and sometimes (with a large DC-blocking capacitor connected to MIC-) even low-frequency hicking (motorboating) may occur. In practice the signal source (microphone connected to the capacitor at MIC-) has a certain impedance, therefore realising equal impedances to ground for both the MIC- and MIC+ input is only possible if a resistor equal to the signal source impedance is used in series with the capacitor at MIC+.
To prevent this extra component a more practical solution is to choose the value of the capacitor connected to MIC- smaller than that connected to MIC+ (taking into account the tolerances of these capacitors.)

2.3.4. STABILITY AND FREQUENCY ROLL-OFF

The 100pF external capacitor C6 connected between GAS1 and SLPE is necessary for ensuring the stability of the transmitting amplifier. Larger values can be applied, and these will then operate as a first-order low-pass filter, the cut-off frequency of which is determined by the time-constant R7C6. This gives f_{3dB}=23kHz with R7=68.1kΩ and C6=100pF.

2.3.5. PARALLEL OPERATION

In case of parallel operation of sets, the operating voltage of the TEA1064 can drop below the internal reference voltage and the circuit automatically adjusts this voltage to a lower value. Of course this will have influence on the performance of the microphone amplifier.

In Fig. 25 the maximum output voltage at pin 1 (LN) is shown as a function of line current flowing into the TEA1064, with a 600 Ω impedance telephone set connected in parallel with the basic application circuit shown in Fig.3. Transmit gain is 52dB in case of a normal 600 Ω load; however with a 600 Ω set in parallel, gain decreases with about 3.5dB. The maximum output swing is not determined by the DC-voltage at pin LN but by the available current in the output stage of the TEA1064 and the dynamic limiter prevents distortion.

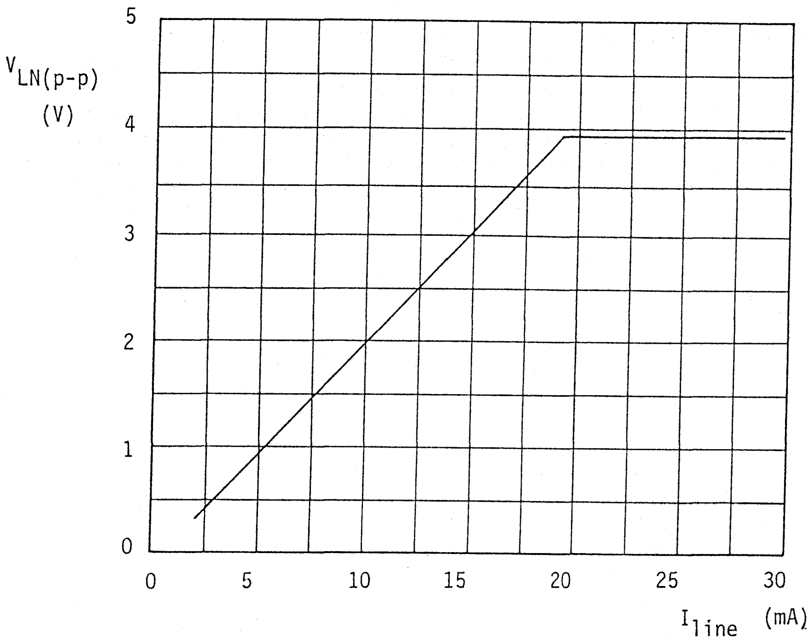


Fig. 25. Maximum output voltage of the transmitting output stage versus I_{line} in low line current range (with a 600 Ω telephone set connected in parallel).

In Fig. 26 the transmit gain versus the DC-voltage at pin LN is shown. Gain decrease starts at V_{LN}=2.2V. At V_{LN}=2V the decrease is about 2-3B and about 8 dB at V_{LN}=1.6V.

The results given are valid for a typical sample in the basic application circuit of Fig. 3. Changing component values will influence the results.

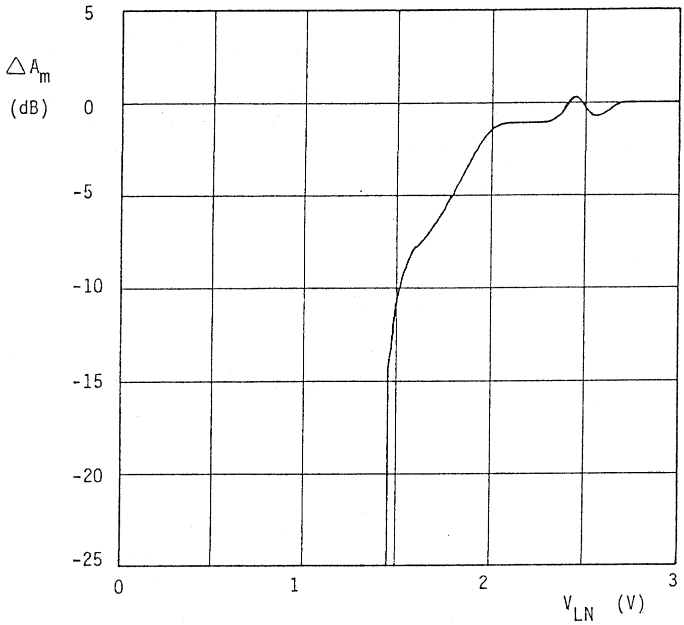


Fig. 26. Typical change of transmit gain as a function of DC voltage V_{LN} in the low voltage range (parallel operation with Iline = 2-11 mA).

2.4. DYNAMIC LIMITER IN SENDING DIRECTION.

The maximum level of the sending signal at LN is limited by the DC-voltage drop across the IC or by the DC-current flowing in the transmit output stage. To prevent distortion (clipping) of the transmitted signal, a dynamic limiter is incorporated. The use of such a limiter also improves the sidetone performance considerably (less distortion and a limited sidetone level in overdrive conditions).
When peaks of the transmitted signal on the line exceed an internally-determined threshold, the gain of the sending amplifier is reduced rapidly. The time in which gain reduction is effected (the attack time) is very short. The circuit stays in the gain-reduced condition until the peaks of the transmitted signal remain below the threshold level. The sending gain then returns to normal after a time determined by the capacitor connected to DLS (the release time).

2.4.1. ATTACK AND RELEASE TIMES

In Fig. 27 the definitions of both attack and release times are shown. Both the attack and the release times are proportional to the value of the capacitor connected to DLS (C16). The attack time is determined by fast discharge of C16 via an internal 1kΩ resistor to VEE during the times when the threshold is exceeded. The release time is determined by a slow recharge

of C16 with a constant current of 1.3 uA.
The recommended value for C16 is 470 nF and this is used in the basic application.
In order to damp the control loop of the dynamic limiter a small resistor (practical value 100Ω) can be used in series with C16. This will result in a different dynamic behaviour (less undershoot during attack time).

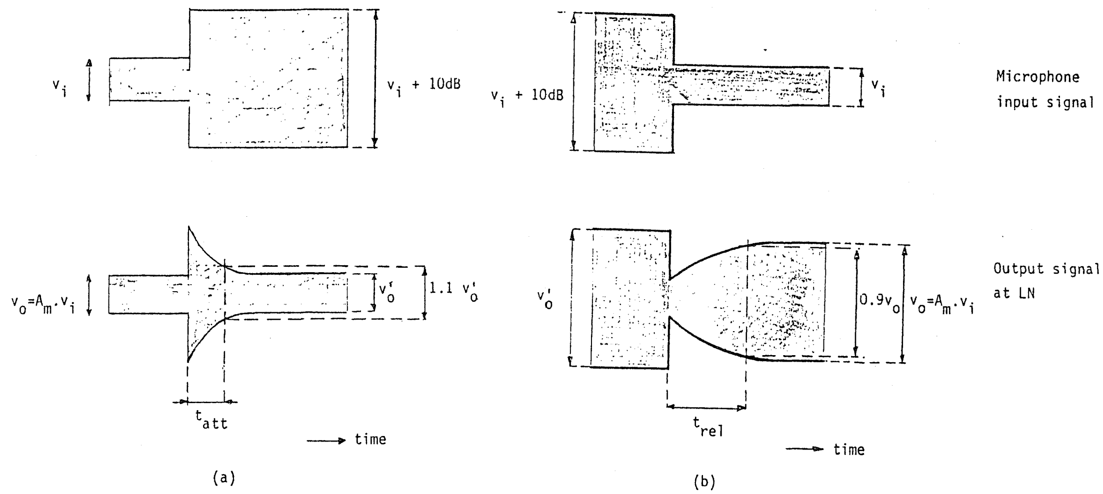


Fig. 27. Dynamic behaviour of the dynamic limiter; definition of attack (a) and release times (b)

2.4.2. MAXIMUM SENDING OUTPUT LEVEL

The internal detector threshold of the dynamic limiter adapts automatically to the DC voltage setting of the circuit (the DC voltage V_{LN}-V_{SLPE}). This means that the maximum transmit output swing on the line will be higher if the DC voltage dropped across the circuit is increased. This can be the case when the voltage adjust resistor R17 is changed or when the peripheral supply current increases in the application with stabilized supply for peripherals.
The internal threshold is lowered automatically if the DC current in the transmit output stage is insufficient to modulate the line. This prevents distortion of the sending signal in case the TEA1064 is operated on a long line (low line current) or when the TEA1064 is operated in parallel with another telephone set.
The maximum possible peak to peak output swing can be read from the graph shown in Fig. 28. or can be calculated by means of the formulas given below. A detailed calculation is given in Appendix A.3.

Current limitation:

$$v_{LN(p-p)} = 2 * (I_{line} - I_{CC1} - I_p - 0.25mA) * \frac{(R1//RL)*(R15 + R16)}{(R1//RL) + R9 + R15 + R16} \qquad [7]$$

where:

I_{CC1} is the internal current consumption of TEA1064 (typ. 1.2mA at VCC1=2.8V; see also Fig. 6).
Ip is the peripheral supply current.
R1 is the resistor between LN and VCC1 determining the set impedance.
RL represents the load resistance at LN.
R9 is the resistor between SLPE and VEE.

Voltage limitation:

$$v_{LN(p-p)} = 2 * \frac{V_{LN-SLPE} - 1.2}{1 + \frac{R9}{R1//RL}} \qquad [8]$$

where:

$V_{LN-SLPE}$ is the DC voltage drop between LN and SLPE, that is constant in the application with regulated line voltage and varies with the peripheral supply current Ip in the case of stabilized supply voltage for peripherals.

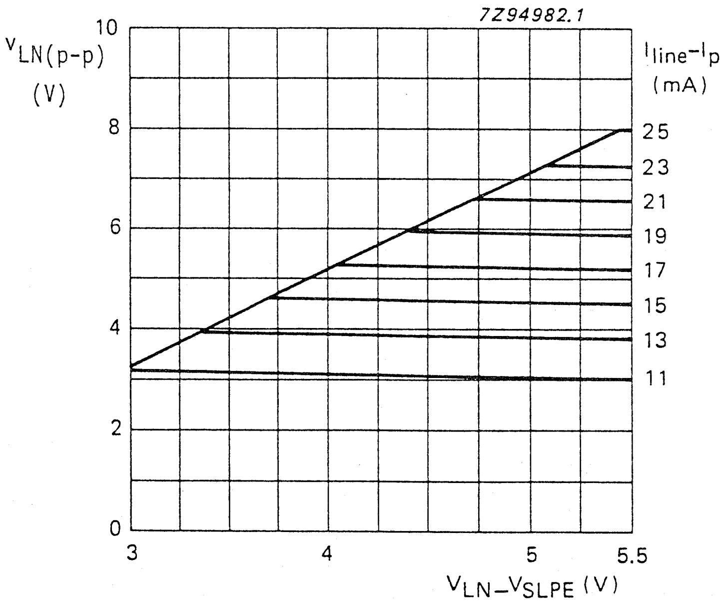


Fig. 28. Maximum sending output swing on LN as a function of the DC voltage drop V_{LN}-V_{SLPE} with Iline-Ip as a parameter; R15=392Ω, R16=56Ω; or R15=0Ω and R16=392+56=448Ω.

2.4.3. AC START-UP TIME

During start-up of the TEA1064, the capacitor connected to DLS (C16) must be charged to a voltage of 1.5V before the nominal gain of the sending amplifiers is reached. As the internal charge current is only 1.3 uA this means that the AC start-up time of the sending amplifier will be about 540msec with C16=470nF. To speed-up AC starttime, an external charge current can be applied to the DLS capacitor during the start-up phase. A simple possibility for this is shown in Fig. C2 of Appendix C (T6, R26, R27).

2.5. DTMF AMPLIFIER

REMARK: The reference used for the DTMF input is SLPE.

A dual tone multi-frequency dialling signal can be applied to the IC between the DTMF input and SLPE. Input impedance is typically about 20kΩ. The voltage gain measured between the DTMF-SLPE input and the transmitter output at LN-VEE is 26 dB less than that of the microphone amplifier. Thus:

$$20\log A_{DTMF} = 20\log A_m - 26 \text{ dB} \qquad [9]$$

The DTMF gain depends on the values of R1, R5, R7, R9 and R_L in the same way as the microphone gain (see Fig. 24). With the component values of the basic application (microphone gain of 52 dB), the DTMF gain is 26 dB.
The choice of gain to suit one particular microphone capsule will also predetermine the DTMF gain. The dialling tones must therefore be adjusted to the appropriate level before they are applied; the DTMF input accepts signals up to $340mV_{rms}$ for THD=2% with internal soft limiting of the input stage.

Details, how to design the coupling network between a Philips DTMF generator and the transmission circuit can be found in Ref. 3 (as an example PCD3311/12 has been taken). Also an analysis is given in this reference about the total spread of the dial tones that can be expected on the line.

<u>Temperature dependence</u>
The DTMF amplifier is internally temperature compensated. The external decoupling capacitor at VCC1 has no influence on the DTMF gain (unlike the rest of the TEA106X-family).
The following typical values with respect to 25°C were found:
-0.2 dB at -25°C
+0.1 dB at +75°C.

<u>Parallel operation</u>
At low line current when the voltage at VCC1 drops below 2.1 V (V_{LN} = 2.8 V in the basic application without DC-load at VCC1) the DTMF amplifier is automatically disabled. This means that during extremely low voltage conditions (e.g. parallel operation) no dialling tones can be sent to the line. This is done to obtain optimum performance of the speech amplifiers during parallel operation on long lines when the DC-voltage across the set drops to extremely low values and dialling is not possible anyway.

2.6. RECEIVING AMPLIFIER

The input of the receiving amplifier is pin IR. Input impedance is approximately 20kΩ. The amplifier has two complementary Class B outputs, the non-inverting output QR+, and the inverting output QR-. The outputs can be used either for single ended drive or for symmetrical drive, depending on the impedance, sensitivity and type of earpiece used.
It can drive either dynamic, magnetic or piezoelectric earpieces as shown in Fig. 29. Earpieces with an impedance up to 450 Ω must be driven in single ended mode (low-impedance dynamic or magnetic capsules). This is shown in Fig. 29a. For impedances above 450 Ω, with a high-impedance dynamic, magnetic or piezoelectric capsule, differential drive is possible, as shown in Fig. 29b, c, d. The additional series resistor (1) shown in Fig. 29c can be used to prevent distortion of the output signal when the output stage runs out of current (causing a dI/dt in an inductive load).

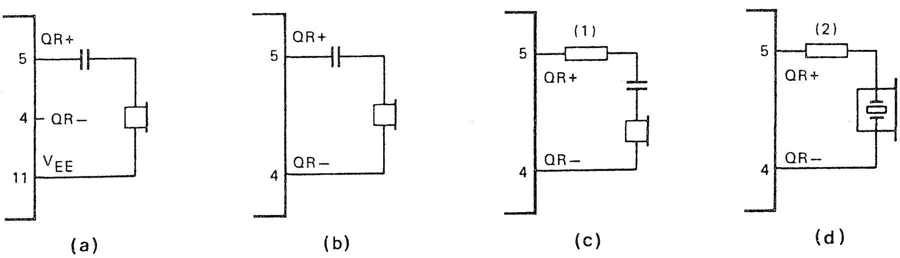


Fig. 29. Alternative receiver arrangements: a) dynamic earpiece with an impedance less than 450Ω; b) dynamic earpiece with an impedance more than 450Ω; c) magnetic earpiece with an impedance more than 450Ω, resistor (1) may be connected to prevent distortion (inductive load); d) piezo-electric earpiece, resistor (2) is required to ensure sufficient phase margin (stability with capacitive load).

A piezoelectric earpiece represents a capacitive load. Capacitive loading of the receiving output stage is permitted up to a maximum of 100nF between QR+ and QR-. However the decrease of phase margin (could give lead to instabilities) must be restored by means of the series resistor ((2)in Fig. 29d.) (for example with C_L=100nF, $R_{(2)}$=50 Ω). More information can be found in Ref. 4.

With an asymmetric load the gain A_{TA} of the receiving amplifier, measured between the input IR and the output QR+ is given by (referring to Fig. 3):

$$A_{TA} = 1.314 * \frac{R4}{R5} * \frac{ZT}{ZT + ro} \qquad [10]$$

Where, ZT = earpiece impedance
r_o = output impedance of the receiving amplifier (typically 4 Ω).

If we insert the values R4=100kΩ, R5=3.65kΩ and Z_T=450 Ω, this results in:

$20\log A_{TA}$ = 31dB ±1dB.

With both outputs QR+ and QR- being used in symmetrical mode, the gain A_{TS} is increased by 6dB and is given by:

$$A_{TS} = 2.628 * \frac{R4}{R5} * \frac{ZT}{ZT + ro} \qquad [11]$$

This results with the values for R4, R5 and ZT which were used above in:

$20\log A_{TS}$=37dB ±1dB.

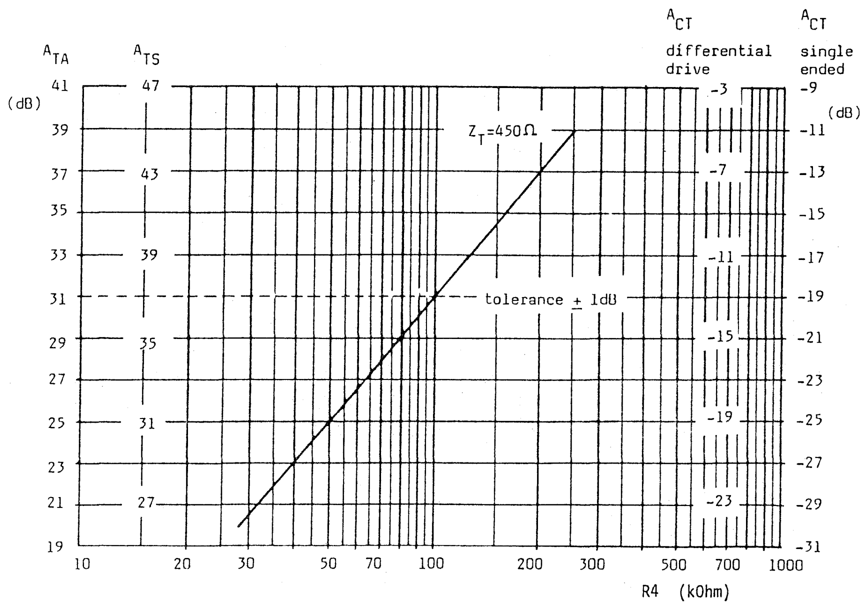


Fig. 30. Gain of the receiving amplifier (ATA and ATS) and the confidence tone gain (ACT) as a function of R4.

2.6.1. GAIN ADJUSTMENT

The gain of the receiving amplifier can be adjusted by means of R4 between 20dB and 39dB with single ended drive and between 26dB and 45dB in case of differential drive. This takes values of R4 between 28kΩ and 250kΩ. The gains A_{TA} and A_{TS} together with the confidence tone (explained in chapter 2.7) as a function of R4 are shown in Fig. 30.

REMARK:
The signal received on the line is attenuated by the anti-sidetone network, before it enters the receiving amplifier. In the basic application this attenuation is about 32 dB and is almost flat over the whole audio frequency range when using the TEA1060 family anti-sidetone bridge.

2.6.2. NOISE, MAX. INPUT SIGNAL

Noise
The equivalent noise at the input IR of the receiving amplifier (psophometrically weighted; P53-curve) is typically 1,25 $uV_{(rms)p}$ (input open). With the anti-sidetone circuit connected to the input, the noise generated at the line pin LN will add via the anti-sidetone circuit to the equivalent input noise of the receiving amplifier. The total noise generated at the earpiece output depends on the microphone gain that has been set and on the actual sidetone suppression; furthermore extra circuitry connected to pin LN (for example an artificial inductor to extend supply possibilities) can give a noise contribution.

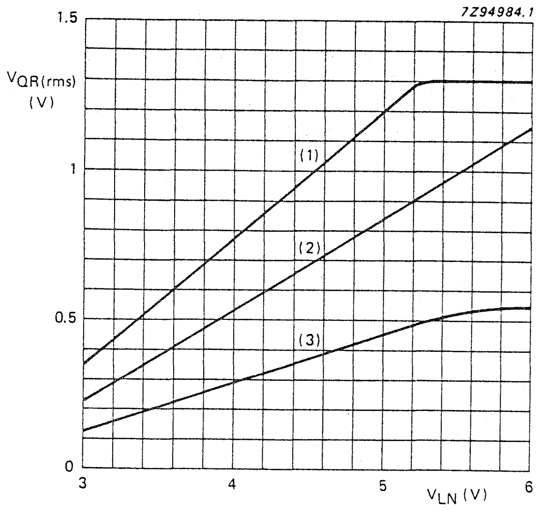
Maximum input signal
The signal received on the line is attenuated by the anti sidetone network before it enters the input IR of the receiving amplifier. In the basic application circuit (Fig. 3) this attenuation is about 32dB. Frequency response between the line and the input IR is almost flat in the audio frequency range when using the special TEA1060-family bridge configuration. The signal at the input IR of the amplifier is internally limited by symmetrical soft limiting to $17mV_{rms}$ for THD=2% and to $53mV_{rms}$ for THD=10%.

2.6.3. MAXIMUM OUTPUT SIGNAL

The maximum output swing of the receiving output stages depends on the impedance of the earpiece and the DC-voltage drop across the circuit.
The maximum output swing versus the DC-line voltage V_{LN} is shown in Fig. 31 with I_{line}=15mA. The maximum output swing will be higher under speech conditions where the ratio of peak to the RMS value is higher.

2.6.4. STABILITY AND FREQUENCY ROLL-OFF

Stability is ensured by the use of the two discrete capacitors C4 and C7 in Fig. 3. Capacitor C4 is connected between QR+ and GAR and capacitor C7 is connected between GAR and VEE. The value of C7 is recommended to be ten times greater than that of C4 and the values are generally C4=100pF and C7=1nF. A larger value of C4 may be chosen so as to obtain a first-order low-pass frequency characteristic, the cut-off frequency being determined by the time-constant R4C4. In this case the ratio of 10:1 for C7:C4 must be preserved. With C4=100pF and with R4=100kΩ, the cut-off frequency f_{3dB}=16kHz.



THD=2%; Iline = 15 mA
Curves:
(1) differential load of 47nF (series resistor=100Ω); f=3400Hz.
(2) differential load of 450Ω; f=1kHz.
(3) single ended load of 150Ω; f=1kHz.

Fig. 31. Maximum output swing of the receiving amplifier as a function of the DC voltage drop V_{LN} with the load at the receiver output as a parameter; Valid for both supply options. (R1=619Ω; no load at VCC1)

2.6.5. PARALLEL OPERATION

Similar to the microphone amplifier, the possibilities of the receiving amplifier will be decreased under low voltage conditions occurring during parallel operation of sets.
Fig. 32 shows the maximum output swing of the receiving amplifier (THD=10%) versus the line voltage V_{LN} with different loads in the low-voltage part. The maximum output swing naturally decreases with the DC-voltage at LN. At V_{LN}=2V typically an output swing of 15mV_{rms} with a 150 Ω load can be obtained. At about 1.6V the receiving amplifier is totally cut-off.

Fig. 33 shows the receive gain as a function of the DC-line voltage V_{LN}. Gain decrease starts at about V_{LN}=2.9V; at V_{LN}=2.5V the gain has been decreased by about 15dB.
The results are valid for a typical sample in the basic application circuit of Fig. 3 with regulated line voltage and R17=20kΩ. Changing components will have influence on the results.

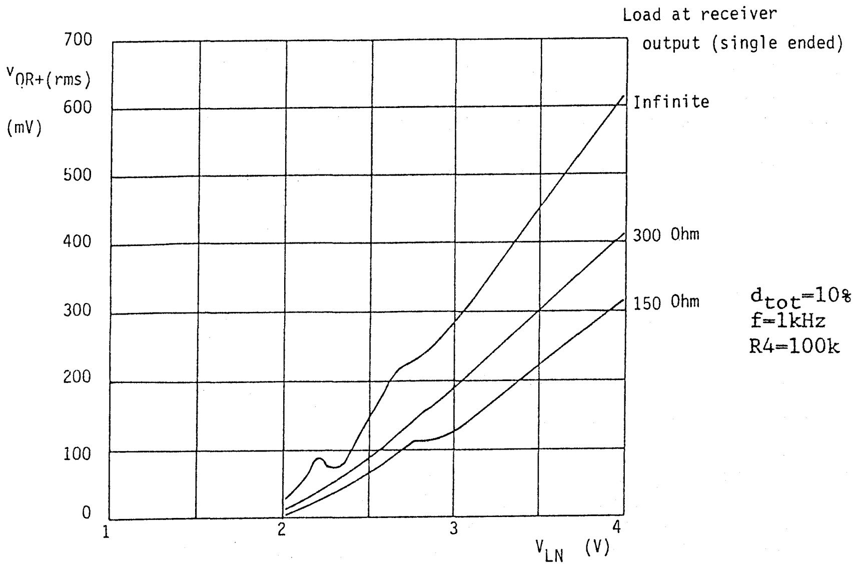


Fig. 32. Maximum output swing receiving amplifier versus V_{LN} in low voltage range. (Parallel operation with Iline =2 - 11 mA). (R1=619Ω; no load at VCC1)

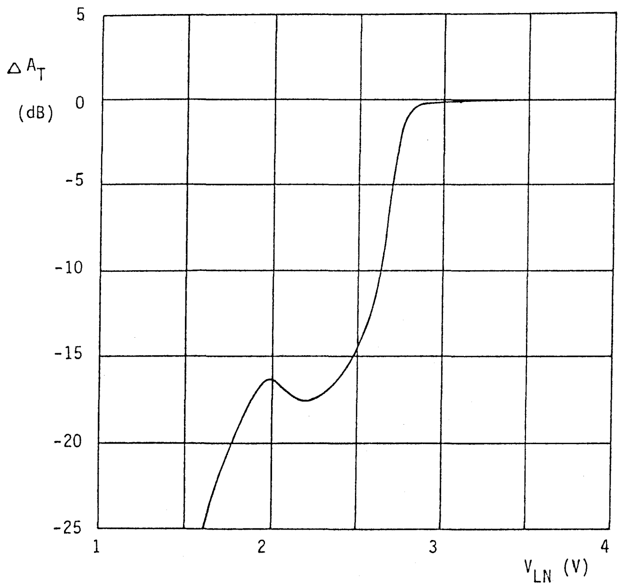


Fig. 33. Typical receive gain versus V_{LN} in low voltage range (parallel operation with Iline = 2-11 mA). (R1=619Ω; no load at VCC1)

2.7. CONFIDENCE TONE

During DTMF dialling, the dialling tones can be heard at a low level in the earpiece. The level of the tones at the receiving output depends on the gain that has been set for the receiving amplifier and on the tone level applied to the DTMF input.
The gain A_{CT} between the DTMF input (with reference SLPE) and the receiving output is given by:

$$20\log A_{CT} = 20\log A_T - 49\ dB \qquad [12]$$

in which A_T is a general term for telephone gain and can be replaced by either A_{TA} (single-ended drive) or A_{TS} (symmetrical drive). This is shown in Fig. 30.

2.8. LINE CURRENT DEPENDENT GAIN CONTROL

The gain figures of the microphone amplifier and the receiving amplifier which were derived in the preceding chapters are applicable only when the AGC is inoperative; that is, with pin AGC is open circuit.
When the resistor R6 is connected between AGC and VEE, the line current dependent gain control of both microphone amplifier and the receiving amplifier become operative; the DTMF amplifier is not affected.

Below a specific value of line current, $I_{line\text{-}start}$, the gain is equal to the values calculated with the formulas given before (2.3 and 2.6). If the current $I_{line\text{-}start}$ is exceeded, the gain of both the controlled amplifiers decreases with increasing DC line current. Gain control stops when another value of line current ($I_{line\text{-}stop}$) is exceeded. The gain control range of both amplifiers is typically 6dB. This corresponds with a line length of 5km of 0.5mm diameter copper twisted-pair cable with a DC-resistance of 176 Ω/km and an average AC-attenuation of 1.2dB/km. The slope of the gain control characteristic has been chosen for an optimum tracking between the line attenuation and the required amplifier gain (typical error <=0.8dB) for a system with a 2x300 Ω feeding bridge. In case lines with other parameters are used, a small additional tracking error will be introduced.

The internal AGC-circuitry in the TEA1064 has been optimised to minimize spread. This has resulted in AGC curves that differ slightly from the TEA1067 AGC-curves. Also the optimum values for R6 differ from the rest of the TEA1060-family.

2.8.1. CORRECTION FOR EXCHANGE SUPPLY VOLTAGE

The value of resistor R6 must be chosen in accordance with the supply voltage in the exchange. In Fig. 34 the control curves are shown for V_{EXCH}=36V and 48V with a feeding bridge resistance of 2x300 Ω.
Also the calculated relationship between line length and line current is shown in Fig. 34. These ideal curves have been calculated with the assumption that an increased voltage drop across the circuit has been set (V_{LN}=4.45V at 15mA; R17=22kΩ) and assuming a polarity guard with 1.4V voltage drop. Other parameters will give slightly different results; giving slightly different optimum values for R6.

2.8.2. CORRECTION FOR FEEDING BRIDGE RESISTANCE

In case the feeding bridge in the exchange has a resistance other than 600 Ω, then R6 must be adjusted. This will introduce a minor increase in tracking error because the slope of the gain control curve has been optimised for a 600 Ω feeding bridge. With a 1000 Ω feeding bridge the typical tracking error that can be expected is <=1.2dB.
Fig. 35 shows the control curves for a 400 Ω feeding bridge with exchange supply voltages of 36V and 48V. Fig. 36 and 37 respectively show the characteristics for an 800 Ω bridge and a 1 kΩ bridge with 48V and 60V.

The optimum values of R6 for the various values of exchange supply voltage and exchange feeding bridge resistance, with a 1.4V diode bridge, R9=20 Ω and increased line voltage V_{LN}=4.45V at 15mA (R17=22kΩ in the application with regulated line voltage or without R17 and with Ip = 2mA in the application with stabilized supply) are as follows:

	R_{EXCH} (Ω)			
	400	600	800	1000
V_{EXCH}(V)	R6 (kΩ) with R9=20 Ω			
36	90.9	71.5	-	-
48	127	100	82.5	71.5
60	-	-	105	90.9

V_{LN}=4.45V
at I_{line}=15mA

In case a value for R9 is used different from 20 Ω the value for R6 must be adapted.

Table 2. Optimum values for AGC resistor R6.

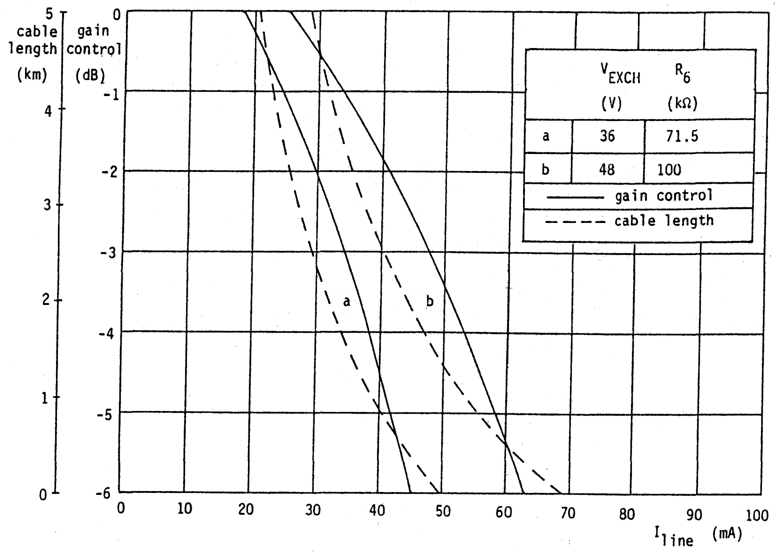


Fig. 34. Gain control characteristics; 600 Ω feeding bridge.

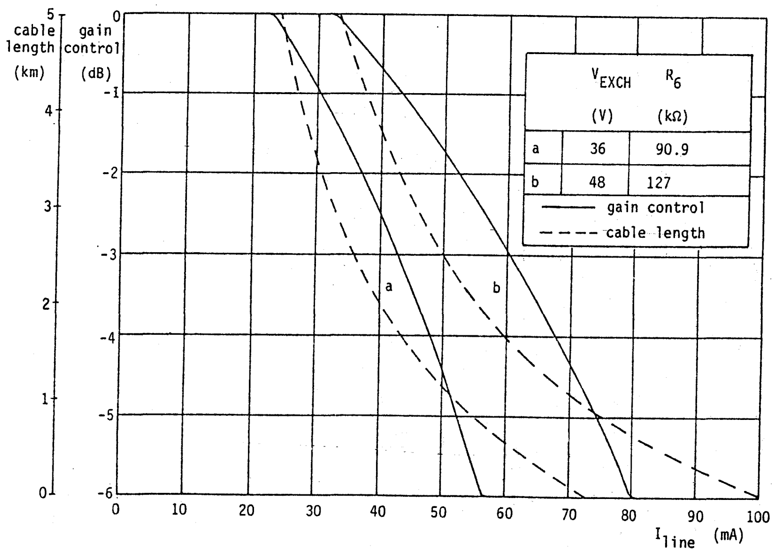


Fig. 35. Gain control characteristics; 400 Ω feeding bridge

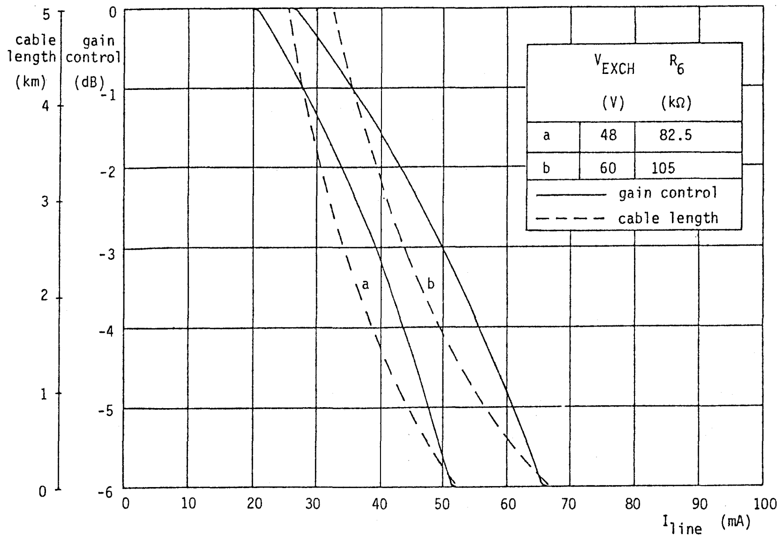


Fig. 36. Gain control characteristics; 800 Ω feeding bridge

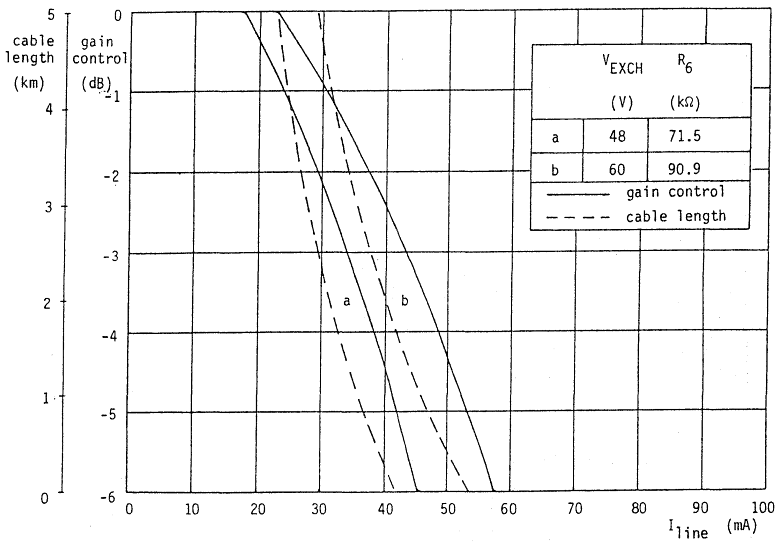


Fig. 37. Gain control characteristics; 1000 Ω feeding bridge

2.9. MUTE INPUT

REMARK: The reference used for the MUTE input is SLPE.

Electronic switching between dialling and speech can be obtained by controlling the MUTE input. If a high level (>=V_{SLPE}+1.5V, <=20uA) is applied to the MUTE input, then both the microphone and receiving amplifier inputs are inhibited and the DTMF input is simultaneously enabled. The converse situation, with DTMF inhibited and the microphone and earpiece amplifier both enabled is obtained by either applying a low-level input (<=SLPE+0.3V) to MUTE, or by leaving MUTE input open. The internal switching takes place with negligible clicking at the earpiece outputs and on the line.

At low line currents, the MUTE function is operational down to VCC1=2.1V (V_{LN}=2.8V without DC-load at VCC1). Below this voltage it is not possible to enable the DTMF amplifier; in this way optimum performance of the speech amplifiers is guaranteed under extremely low voltage conditions (parallel operation). Disabling of the microphone and receiving amplifiers is possible down to V_{LN}=1.7V (R1=619Ω; no load at VCC1).

Voltage and current limitations

The maximum voltage allowed at the MUTE input is VCC1+0.4 V. In practice the supply terminal VDD for peripherals can be at a higher potential (w.r.t. VEE) than VCC1. Therefore during its high state the mute output of the dial IC can be at a higher potential than VCC1. In such cases a resistor in series with the MUTE pin is necessary to limit the maximum input current to 30 uA)

2.10. POWER DOWN INPUT

REMARK: The reference used for the PD input is SLPE.

The power-down input PD is available for use in pulse dialling and in register recall applications, where the telephone line is interrupted. During these interruptions the telephone set is without continuous power and the transmission IC must be supplied by the charge available in the smoothing capacitor C1 connected to VCC1; the peripheral circuits must be supplied by the charge available in the smoothing capacitor C15 in Fig. 3. The discharge time of these capacitors will be longer in case the power-down function is used; this results in less ripple on VCC1 and Vp.

When a high level input (>=V_{SLPE}+1.5V, <10uA) is applied to the PD pin, the internal supply current I_{CC1} is reduced from about 1.2mA to typically 60uA at V_{CC1}=2.8V. Furthermore, the voltage regulator capacitor C3 at REG is internally disconnected to prevent it from being discharged during line interrupts. This means that after each line interrupt, the voltage regulator is able to start without delay at the same DC-line voltage as before the interrupt. This minimizes the contribution of the IC to the shape of the current pulses during pulse dialling. Of course with a highly inductive exchange feeding bridge, the inductors mainly determine current waveform. Under these conditions the voltage regulator may have some switch-on delay and cause a voltage overshoot at the line connection (LN) of the IC.

If the voltage drop across the circuit is increased by means of R17, the power down function will be affected because C3 is connected now via R17 to SLPE. This will slightly affect the shape of the current pulses.

When the power-down facility is not required, the PD pin can be left open-circuit or connected to SLPE (PD = LOW is defined as applying a voltage <=SLPE+0.3V).

Voltage and current limitations

The maximum voltage allowed at the PD input is VCC1+0.4 V. In practice the supply terminal VDD for peripherals can be at a higher potential (w.r.t. VEE) than VCC1. Therefore during its high state the output of the dial IC used to drive the PD pin, can be at a higher potential than VCC1. In such cases a resistor in series with the PD pin is necessary to limit the maximum input current to 30 uA)

2.11. ANTI-SIDETONE CIRCUIT

To avoid the reproduction of microphone signals in the telephone transducer, the anti-sidetone circuit uses the microphone signal from pin SLPE to cancel the microphone signal at the input IR of the receiving amplifier. The TEA1060-family anti-sidetone bridge (Fig. 38a.) or the conventional Wheatstone bridge (Fig. 38b.) may be used as the basis for the design of the anti-sidetone circuit.

The TEA1060 family anti-sidetone bridge has the advantage of a relatively flat transfer function in the audio-frequency range between pins LN and IR, both with real and complex set impedances. Furthermore, the attenuation of the bridge for the received signal (between LN and IR) is independent of the value chosen for Zbal after the set impedance has been fixed and the condition R9R2=Ri(R3+R8) is met. Therefore, readjustment of the overall receive gain is not necessary in many cases.

The Wheatstone bridge has the advantages of needing one resistor fewer than the TEA1060 family anti-sidetone bridge and requires only a small capacitor (about 10nF) in Zbal. Calculation of the values is generally considered to be easier. The disadvantages include the dependence of the attenuation of the bridge on the value chosen for Zbal and the frequency dependence of that attenuation. This necessitates a readjustment of the overall receive gain.

In Ref. 2 a detailed comparison of both bridge types together with the bridge calculations of the TEA1060-family anti-sidetone bridge and the Wheatstone bridge is given.
The TEA1060 family anti-sidetone bridge have been reconsidered mathematically in Appendix B; this results in more accurate bridge formulas.

2.11.1. TEA1060-FAMILY BRIDGE

The equivalent circuit of the TEA1060-family bridge is shown in Fig. 38a. Optimum suppression of the sidetone signal is obtained when the following conditions are fulfilled:

a) $$R9*R2=Ri*(R3+R8) \qquad [13]$$

where $Ri = R1//Req$ is the dynamic impedance of the circuit

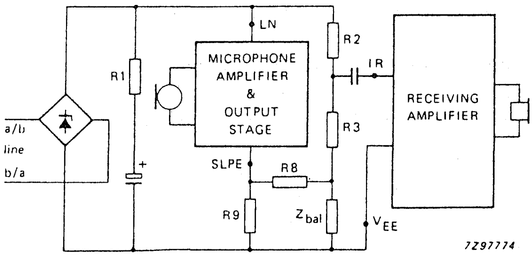
$$Req = Rp*(1+\frac{R15}{R16}) \qquad \text{and } Rp = 15.5k\Omega$$

b) $$Zbal = \frac{R3*(R8+R9)}{R2*R9} * Z_{line} = k * Z_{line} \qquad [14]$$

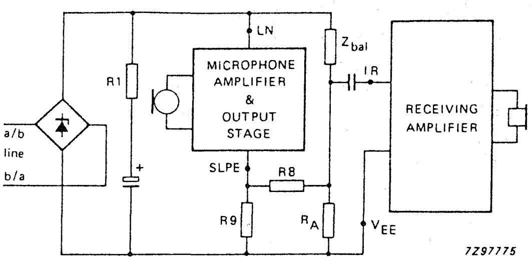
Where k is a scale factor

$$k = \frac{R3*(R8+R9)}{R2*R9} = \frac{R3*(R8+R9)}{Ri*(R3+R8)} \qquad [15]$$

Normally R1 and R9 are fixed and R8, R2 and R3 can be chosen by the designer.

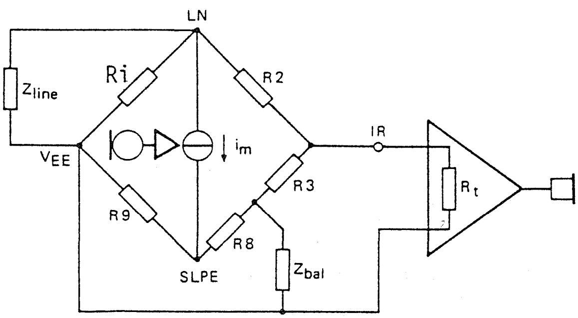


(a) TEA1060 family bridge

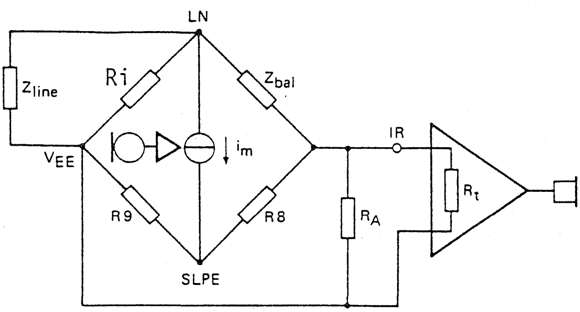


(b) Wheatstone bridge

Fig. 38. Anti-sidetone circuits



(a) TEA1060 family bridge



(b) Wheatstone bridge

Fig. 39. Equivalent circuits

Scale factor k (in fact the values of R8 and R3) and the value of R2 must be chosen to meet the following criteria:

- compatibility with a standard capacitor from the E6 or E12 range for the capacitor used in Zbal.
- respect condition a) to ensure optimum suppression of the anti-sidetone bridge.
- |Zbal+R8|>>R9 to avoid influence on microphone gain
- (|R3 + [R8//Zbal]|)<< 20k to avoid influence of the input impedance at IR (Rt =20kΩ+20%) on the bridge attenuation. (In practice R3<<20kΩ).
- |R8//Zbal|<<R3 to avoid influence of Zbal on the overall receive gain.

In practice Z_{line} varies strongly with the line length and line type. Consequently a value for Zbal has to be chosen that corresponds with an average line length giving satisfactory sidetone suppression with short and long lines. The suppression further depends on the accuracy with which Zbal equals this average line impedance.
In the basic application of Fig. 3, Zbal has been optimised for a line length of 5km 0.5mm diameter copper twisted pair with an average attenuation of 1.2dB/km, a DC-resistance of 176 Ω/km and a capacitance of 38nF/km. The approximate equivalent line impedance is shown in Fig. 40.

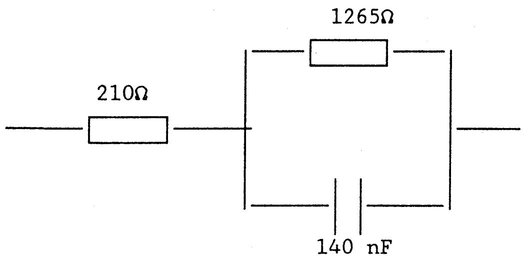


Fig. 40. Equivalent line impedance for optimum sidetone suppression of a 5 km of 0.5mm copper twisted pair; 176 Ω/km, 38 nF/km.

For compatibility of the capacitor value in Zbal with a standard capacitor from the E6 series (220 nF) :

$$k = \frac{140 \text{ nF}}{220 \text{ nF}} = 0.636$$

For R3 a value of 3.92kΩ has been chosen. So Zbal, R8 and R2 can be calculated resulting in the following practical values R11=130 Ω, R12=820 Ω, C12=220nF, R8=390 Ω and R2=130kΩ.

Remark: With these practical values the exact value of the scale factor in the basic application differs a little from the initially taken one and the values of R15 and R16 slightly influences k. This has a minor effect on the sidetone characteristics.

With the line current dependent gain control activated a roughly equal sidetone level results (acoustically measured) for telephone sets connected directly to the exchange (0 km line length) and connected with a line of 10 km.
In case no AGC is used the sidetone has to be optimised for a shorter line length in order to obtain equal (acoustical) sidetone levels for a set connected with a very short line and a set connected with a 10km line to the exchange. (Practical values for a line length of 2 km are: R11=130 Ω, R12=1 kΩ, C12=100nF, R8=620 Ω and R2=130kΩ). Of course overall sidetone suppression is worse in that case compared with the situation where AGC is activated. In practice normally a compromise is chosen between loudness of the set and sidetone level; this means that sending and receiving gain will be reduced somewhat.

The attenuation of the received line signal between LN and IR can be derived from:

$$\frac{v_{IR}}{v_{LN}} = \frac{Rt//R3}{R2+(Rt//R3)} \qquad [16]$$

if $|R8//Zbal| \ll R3$

Where Rt is the input impedance of the receiving amplifier (typically 20 kΩ). This attenuation is about 32dB with the basic application as shown in Fig. 3. Frequency dependence of the input attenuation is negligible in the audio frequency range. However a frequency roll-off can be obtained by means of a capacitor connected between IR and VEE to prevent high frequency components from entering the receiving amplifier.

<u>Complex set impedance</u>

Complex set impedances can be realized by using a complex network instead of R1, and normally the bridge can be rebalanced by re-adjusting the values of R8 and Zbal, and either R2 or R9. This is described in Ref. 2. Changing R9 has also consequences on other parameters (See 2.2.3). Therefore the design procedure as given in Appendix D should be followed. Changing R2 has influence on the attenuation of the received signal between LN and IR; this necessitates a re-adjustment of the receiving gain. Note that changing R1 also has influence on the capabilities of the VCC1-supply for active microphones (see 2.2.4.).

A detailed analysis of the TEA1060 anti-sidetone bridge in the case of complex set impedance is given in Ref. 2.

<u>A software tool to optimize the TEA1060 anti sidetone bridge</u>

In some cases calculating the bridge components in the condition to give optimum sidetone suppression is not very useful. Normally a compromise must be chosen to meet sidetone requirements in several conditions. In those cases help can be provided by means of a software tool named 'BRIDGE' developed in PCALE (Ref. 5). It is designed for use with the IBM PC or compatibles. The program described generates values for Zbal to meet a certain sidetone suppression. The values of R1, R9, R2, R3 and R8 have to be entered into the PC first. For the correct order of adjusting components see the Appendix D.

2.11.2. WHEATSTONE BRIDGE

The conditions in the Wheatstone bridge (equivalent circuit in Fig. 39b for optimum sidetone suppression are given by:

$$Zbal = \frac{R8}{R9} * \frac{Ri*Z_{line}}{Ri+Z_{line}} \quad \text{provided that } R8/R9>>1 \qquad [17]$$

where:

$$Ri = R1//[15.5*10^3(1 + \frac{R15}{R16})]$$ is the dynamic impedance of the circuit

Also for this bridge type a value for Zbal has to be chosen that corresponds with an average line length.

The attenuation of the received line signal between LN and IR is given by:

$$\frac{V_{IR}}{V_{LN}} = \frac{R8//Rt//R_A}{Zbal+(R8//Rt//R_A)} \qquad [18]$$

Where Rt= input impedance of the receiving amplifier at IR
Typically 20 kΩ

A practical circuit could have the following values:
R8=820 Ω, R1=620 Ω and Zbal optimised for the line impedance as shown in Fig. 40. With R_A=infinite and a 600 Ω load at the line, the attenuation varies typically from about 24dB to 27.5dB over the normal audio frequency range; the lower attenuation occurs at the upper frequencies. R_A is used to adjust the bridge attenuation; its value does not have influence the balance of the bridge.

Complex set impedance

If complex set impedance are used with the Wheatstone bridge, it can be re-balanced by adapting the values of Zbal. However the frequency dependence of the transfer function between LN and IR will increase.

A detailed analysis of the Wheatstone bridge is given in Ref. 2.

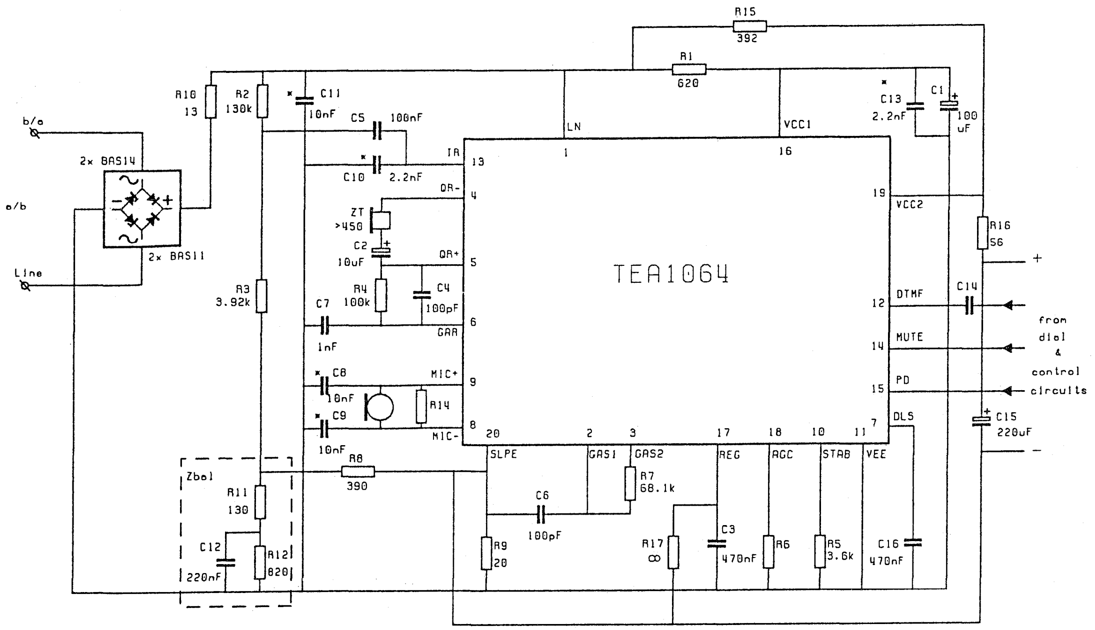


Fig. 41. Application of TEA1064 (stabilized supply), with basic RFI protection in sets with DTMF dialling (without flash); overvoltage protection is also shown; RFI capacitors are marked with a *.

3. IMMUNITY TO R.F. SIGNALS

In the presence of high-intensity electromagnetic fields, it is possible for common-mode amplitude modulated R.F. signals to be induced in the a/b lines. These common-mode signals can sometimes become differential-mode signals as a result of asymmetrical parasitic impedances to ground (for example through the hand of the subscriber holding the handset). Preventive measures have to be taken to avoid the possibility of these signals being detected and the low-frequency modulation appearing as unwanted signal at the earpiece or on the line.

3.1. BASIC PROTECTION

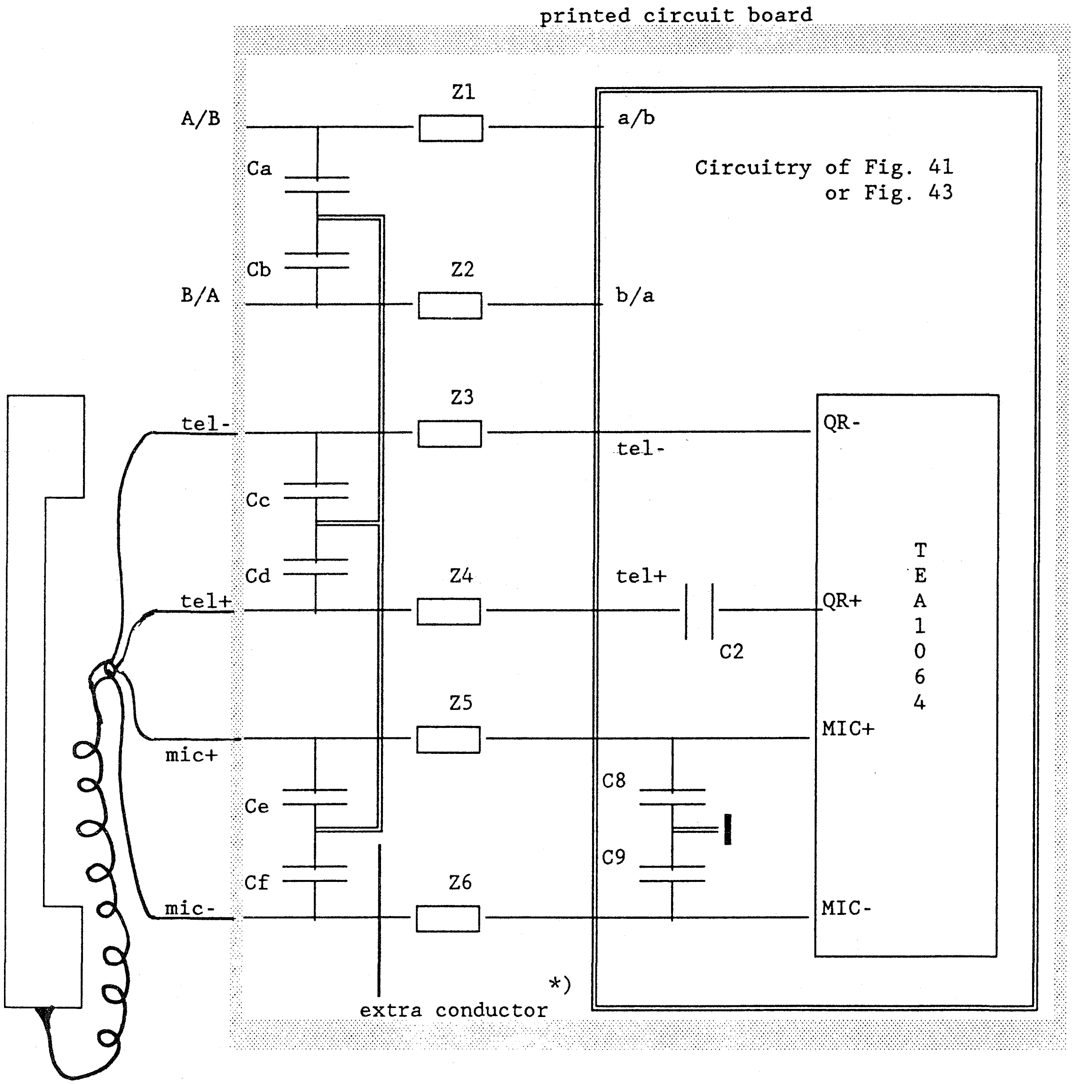
Small discrete capacitors are necessary to suppress the unwanted R.F. signals before they can enter the circuit. Capacitor types suitable for high frequencies must be used, for example ceramic types. In Fig. 41 they have been added to the basic application circuit: C8 and C9 at the microphone inputs, C10 at the receiving input IR, C13 at the supply point VCC1, and C11 at the transmitter output LN. All of the capacitors are connected to the common VEE.

Furthermore the lay-out of the printed circuit board has influence on R.F.-immunity. The copper ground area should be kept as large as possible. Earth planes or grids must be created and print tracks must be kept as short as possible. RFI-capacitors must be mounted as close as possible to the IC-pins and to the ground grid.

3.2. ENHANCED PROTECTION

In practice it has been shown that a so-called RF-guard, which by means of ceramic capacitors couples the RF signal from the a/b-lines to the leads of the handset-cord, gives further improvement of immunity. The realization of such an 'RF-guard' is shown in Fig. 42. How the 'RF-guard' works can be explained as follows:
The RF-signals are mainly picked up by the leads connected to the telephone base set (handset cord and base cord). The induced RF current flows now through the extra conductor, marked with *) in Fig. 42, instead of through the reference of the telephony circuitry.
Care must be taken that the re-radiation of the RF-currents through this extra conductor will not be picked up by the telephony circuitry again. Therefore a open space must be realised on the printed circuit board between the circuitry and the extra conductor to reduce mutual coupling between the two.

The 'RF-guard' should be used in combination with the impedances Z1 up to Z6. Z1 and Z2 are two inductors (chokes with a value of about 20 μH).
Because the TEA1064 has a very high microphone input impedance, it is possible to use low-pass filtering in series with both microphone inputs, without affecting gain accuracy (impedances Z5 and Z6 (both 1kΩ resistors) in combination with C8 and C9). If a low ohmic termination across the microphone inputs is used (R14), a small correction of the microphone gain by means of R7 will be necessary. The RC-filter should be positioned as close as possible to pins 7 (MIC-) and 8 (MIC+).



Ca through Cf = 4.7 nF; Ca and Cb should be 500 Volt types.
Z1, Z2 = 22 μH with high RF-losses (low Q)
Z3, Z4 = 22 Ω or 2.2 μH
Z5, Z6 = 1 kΩ

*) If other peripherals are used, the same network must be continued.

Fig. 42. 'RF-guard' positioned at the handset cord and base cord incoming lines.

The TEA1064 earpiece output stages have very low output impedances. Therefore the influence of the impedances Z3 and Z4 on overall receive gain depends mostly on the impedance of the earpiece transducer. In most cases only a minor correction of the receive gain (R4) will be necessary.

Some remarks:

* The decoupling capacitor (C11=10 nF in Fig. 41) between LN and VEE can be removed. This can be necessary to improve the BRL at high frequency (3400Hz). Capacitors Ca and Cb will affect the BRL.

* The capacitors connected to the IC-pins MIC, REG should be connected as near as possible to pin VEE. Local (RF-)currents highly influence the immunity behaviour of the circuit. The looparea enclosed by the components connected to these points to ground shall be kept as small as possible.

* The extra conductor <u>does not need to be</u> connected to VEE. If a connection has to be made then at one spot only !! Avoid ground loops !

* All connections which will be made to peripherals (a,b-lines, handset, second earshell) by means of long leads (> 200 mm) shall be decoupled close to one-another in the same way as shown in Fig. 42.

4. POLARITY GUARD AND TRANSIENT SUPPRESSION

The transmission IC can be destroyed by excessive current surges on the telephone lines if no proper measures are taken. The type of protection differs for sets with only DTMF dialling or sets with either pulse-dialling or DTMF dialling with "flash" (register recall by means of a timed line interrupt).
With DTMF dialling only, the bridge rectifier, which normally acts as a polarity guard also can incorporate two voltage reference diodes (such as BZW14). Under normal operating conditions, one of the two voltage reference diodes conducts while the other is non-conducting. If the voltage across the set temporarily exceeds the reference voltage of the previously mentioned non-conducting diode, it will conduct and limit the voltage across the set. The maximum permissible voltage across the transmission circuit is 12V continuously and is determined by the collector-emitter breakdown voltage of the IC-process used. During switch-on and line interrupts the maximum permissible voltage is 13.2V allowing the use of a 12V voltage reference diode in the polarity guard.
Further protection is offered by the resistor R10 (see Fig. 41) in series with the bridge rectifier, which limits the current that can be drawn by the IC. The maximum allowed transient voltage on the circuit including the protection resistor R10 being 13 Ω and with R9=20 Ω is 28V during 1msec with a repetition time of 5 sec. This corresponds with a 50A surge current into the BZW14 zener diodes used in the polarity guard.

For DTMF dialling with flash, or for pulse dialling, a different protection arrangement is necessary because during line interruption, the line current must be zero. This means that the bridge rectifier must be able to withstand a relatively high voltage of the order of 200V. A polarity guard using four diodes with typenumber BAS11 is appropriate for this purpose. Protection against line current surges can then be obtained by means of a suitable VDR or a BOD (Philips type range BR211) connected between the ab lines in front of the polarity guard. The speech circuit is protected against overvoltages that may occur, for example during switching-in, by means of a 12V regulator diode connected between LN and VEE, or in case a current limiter is used (e.g. combined with the interrupter), by a 6.8V voltage regulator diode connected between LN and SLPE.

Fig. 43. shows an application of the TEA1064 with an interrupter circuit. For a special interrupter circuit suitable for very low DC-line voltages please see Ref. 6.

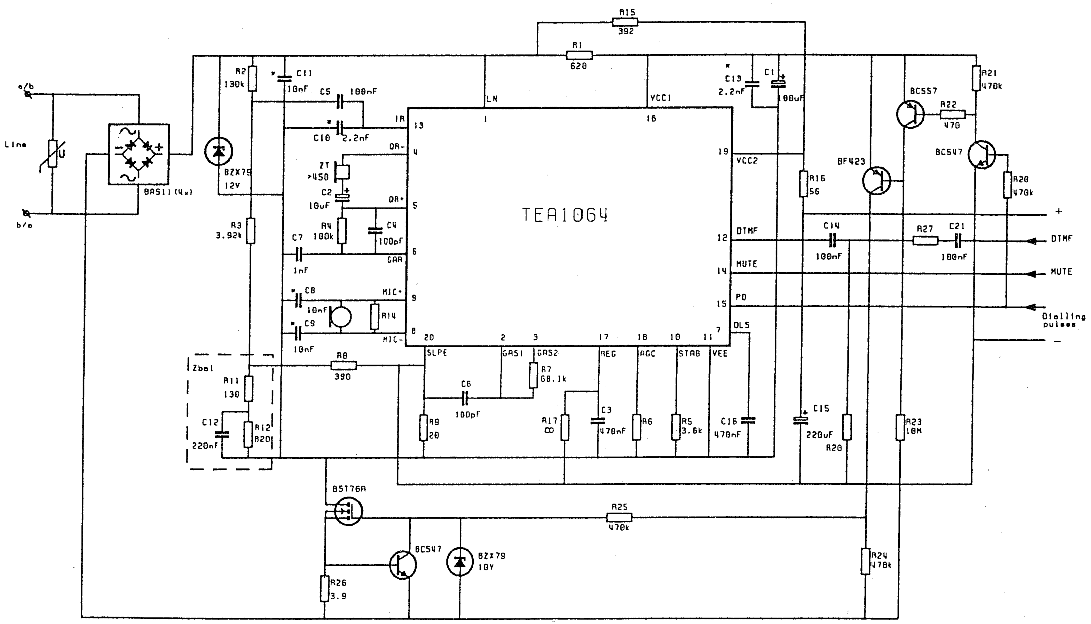


Fig. 43. Basic application of TEA1064 (with stabilized supply) in sets with combined pulse and tone dialling including interrupter with interface. RFI capacitors are marked with a *. Protection of the IC is by means of the VDR in combination with a 12V zener between LN and VEE.

5. HINTS FOR PRINTED CIRCUIT BOARD LAY OUT

Care must be taken to avoid that the large line current flows into common ground tracks to which sensitive points are connected.
For this reason resistors R5 connected between STAB and VEE and R6 connected between AGC and VEE must be situated on the p.c.b. as close as possible to pin VEE.
Also the ground connection of the earpiece should preferably be realized at a point where no large line current is flowing.

The copper tracks connecting R7 and R4 to the corresponding IC-pins (most important are GAS2 and GAR) should be kept as short as possible.

The ground connection of all RFI-capacitors should be realized by means of as large as possible copper planes or grids. RFI-capacitors must be connected as close as possible to the pins that have to be decoupled.

The ground plane on the circuit board must be kept as large as possible where every copper area must be connected to the ground-plane (or grid) on at least two points.

An 'RF-guard' as described in paragraph 3.1 can be realised as shown in Fig. 42 and furthermore the remarks related to this (end of 3.2) should be followed.

An example of a printed circuit board layout for the TEA1064 can be found in Ref. 9.

6. PERFORMANCE

Some measurements have been done with the basic application circuit with stabilized supply voltage for peripherals (including RFI-capacitors), as shown in Fig. 41. (R15=392Ω, R16=56Ω and C3=470nF). This gives an indication of the performance of the TEA1064. A typical sample has been used for the measurements.
No acoustical measurements were done because results will be comparable to the TEA1067 test results as described in Ref. 2.

6.1. BALANCE RETURN LOSS (BRL)

The result of the BRL measurement with a 600Ω reference impedance is shown in Fig. 44. A polar plot of the set-impedance is shown in Fig. 45. Different values for C3 and the ratio R15/R16 will strongly affect the results.

6.2. FREQUENCY CHARACTERISTICS

Fig. 46 shows the frequency response of the sending channel measured between the microphone inputs and the output LN with a 600 Ω load. The microphone gain is set by R7 to 52 dB (R7=68.1kΩ). The upper cut-off frequency is about 24 kHz and is determined mainly by the timeconstant R7C6. The lower cut-off frequency is about 30 Hz and is fixed by the choice made for R15/R16 and C3. Note that if a complex set impedance is chosen, this will influence the sending frequency characteristic.

Fig. 47 shows the frequency response of the receiving channel measured between pin LN and a 150 Ω load resistor connected in single ended mode to the earpiece output pin QR+ via a 10 uF DC-blocking capacitor. With R4 = 100 kΩ a transfer ratio of -1.3 dB at 1 kHz has been measured. The lower cut-off frequency (≈ 150 Hz) is determined by the 150 Ω load and the DC-blocking capacitor C2. The upper cut-off frequency (≈ 9.5 kHz) is determined partly by R4C4 (15kHz) and partly by the cut-off frequency [1/(2π*R3*C10)] of the anti-sidetone circuit (LN to IR) being about 18 kHz.

The transfer ratio as a function of frequency measured from the microphone inputs to a 150 Ω asymmetrical load at the receive output QR+ (10uF DC-blocking capacitor) is shown in Fig. 48. This represents the electrical sidetone at 0 km of telephone line (600 Ω load at LN). The measured sending signal at LN is shown also. Furthermore the receive signal at the earpiece is shown with the same signal at pin LN in receiving condition.
The difference between wanted receive signal and unwanted sidetone signal at the earpiece, with equal signal levels on pin LN in both the sending and the receiving conditions, is the electrical sidetone suppression. This means that for this application the electrical sidetone suppression for telephone sets connected directly to the exchange (0km of line length) is about 7.5dB at 1kHz. The result depends strongly on the balance of the anti-sidetone circuit. In this case the balance impedance Zbal has been optimised for 5km line length with 0.5mm diameter, 176 Ω/km and 38nF/km.

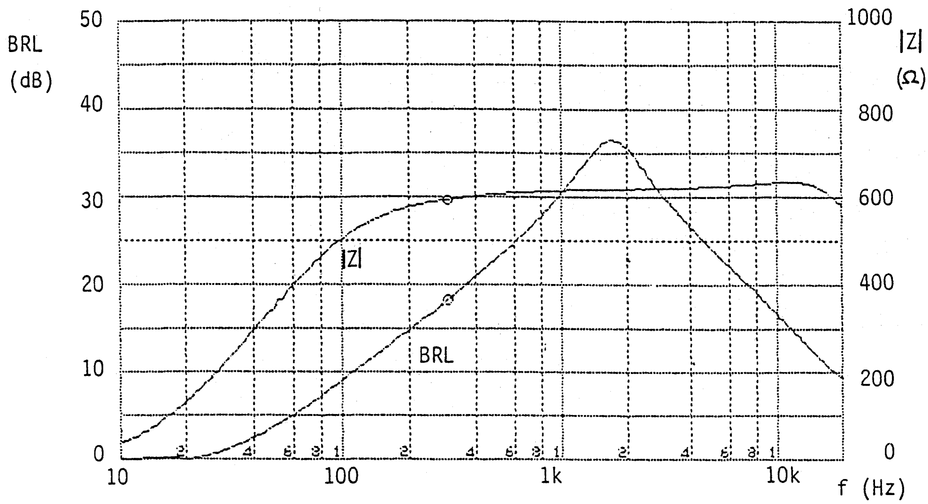


Fig. 44. Balance Return Loss and magnitude of set impedance between 10 Hz and 20 kHz.

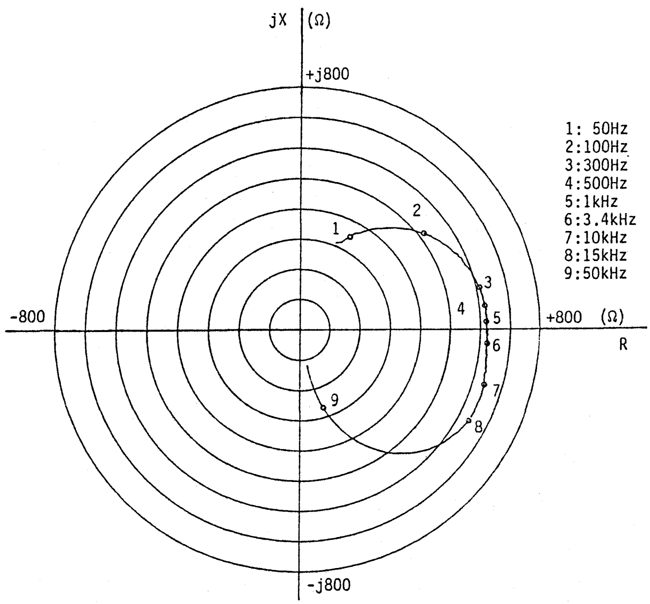


Fig. 45. Polar plot of the set impedance between a/b terminals.

Electrical sidetone suppression is not dependent on whether or not gain control is used, because both amplifiers (microphone and receive) are affected by the gain control function.

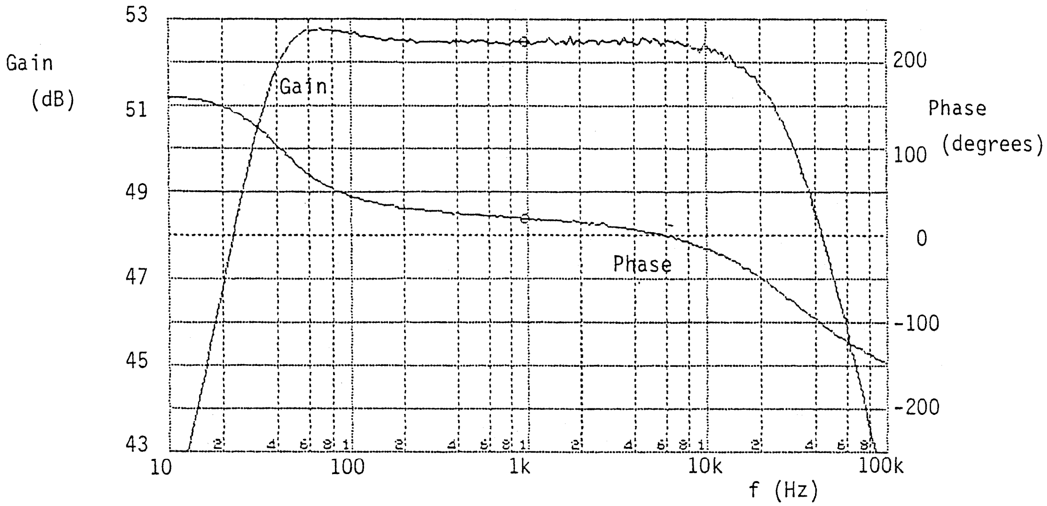


Fig. 46. Frequency characteristic of the microphone amplifier. (MIC to LN).

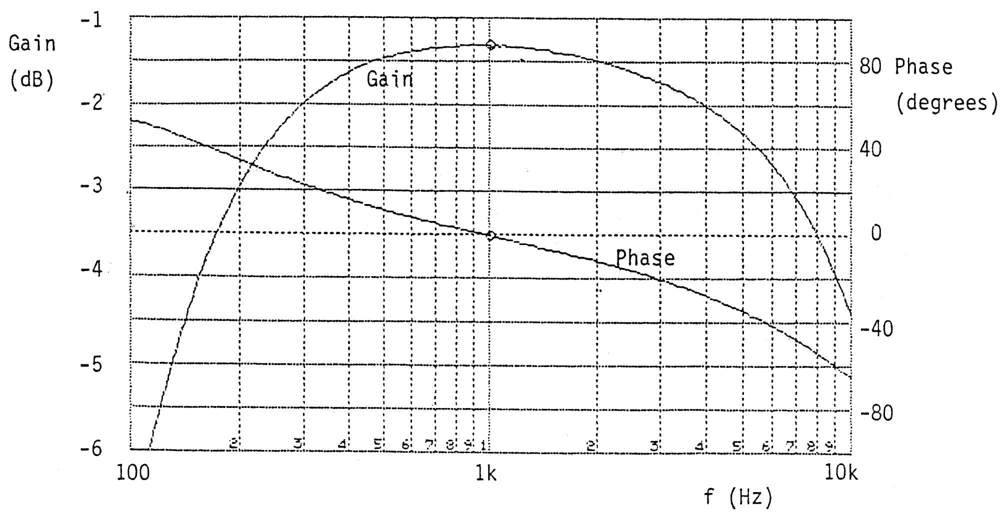


Fig. 47. Frequency characteristic of the receiving channel. (LN to QR+)

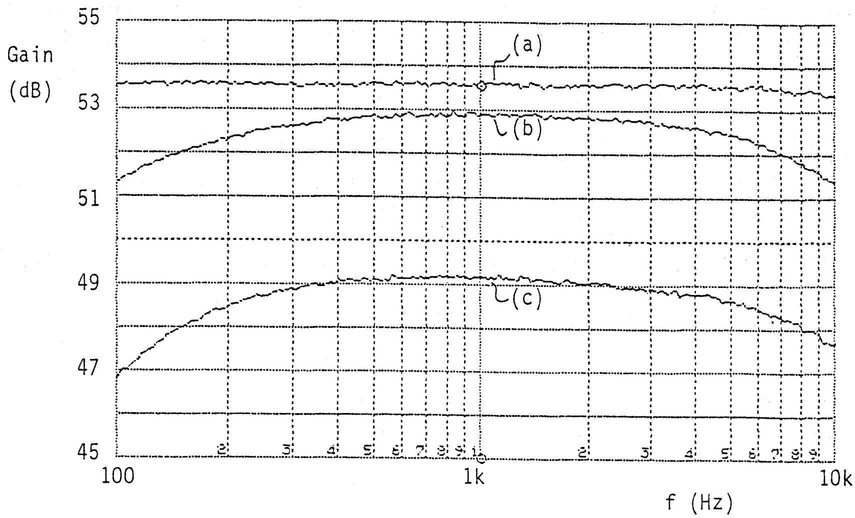


Fig. 48. Frequency characteristic of the electrical sending (a), receiving (b) and sidetone (c) in the case of a telephone set connected via a very short line to the exchange. The difference between curves (b) and (c) is the electrical sidetone suppression.

6.3. DYNAMIC LIMITER

The dynamic behaviour of the dynamic limiter that prevents clipping of the sending signal is shown in Fig. 49. Fig. 49a shows the line signal on pin LN with a signal applied to the microphone inputs that jumps between an rms level of 2mV and 40 mV. The short attack time and relatively long release time can be distinguished clearly. Fig. 49b shows the 'attack' with a timebase of 1msec/div.

Fig. 50. shows the maximum sending signal on the line output LN and its total harmonic distortion, as a function of the microphone input signal with Iline as a parameter. It can be seen that the gain control range is rather high; at about 20 dB overdrive of the microphone inputs is the distortion of the line signal THD < 10 %. The sidetone level on the earpiece is proportional to the line signal. So also the sidetone level is controlled by the dynamic limiter and therefore the level is limited and distortion is reduced (For measurement results see paragraph 3.10.5 of Ref. 8).

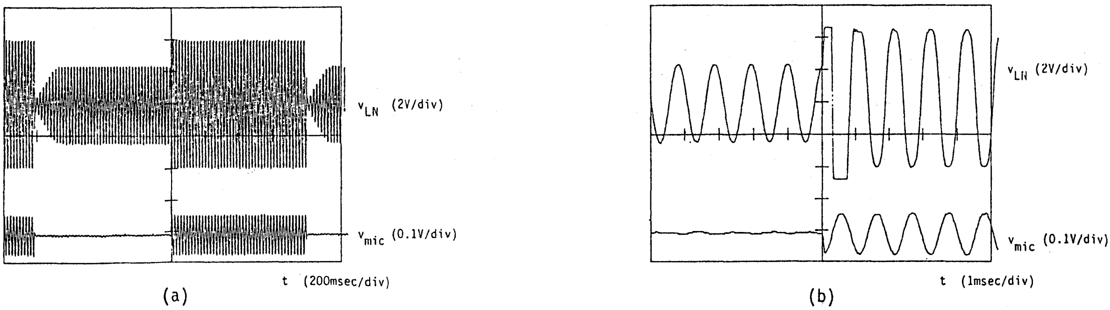


Fig. 49. Dynamic behaviour of the dynamic limiter; horizontal time base: (a) 200 msec/div, (b) lmsec/div.

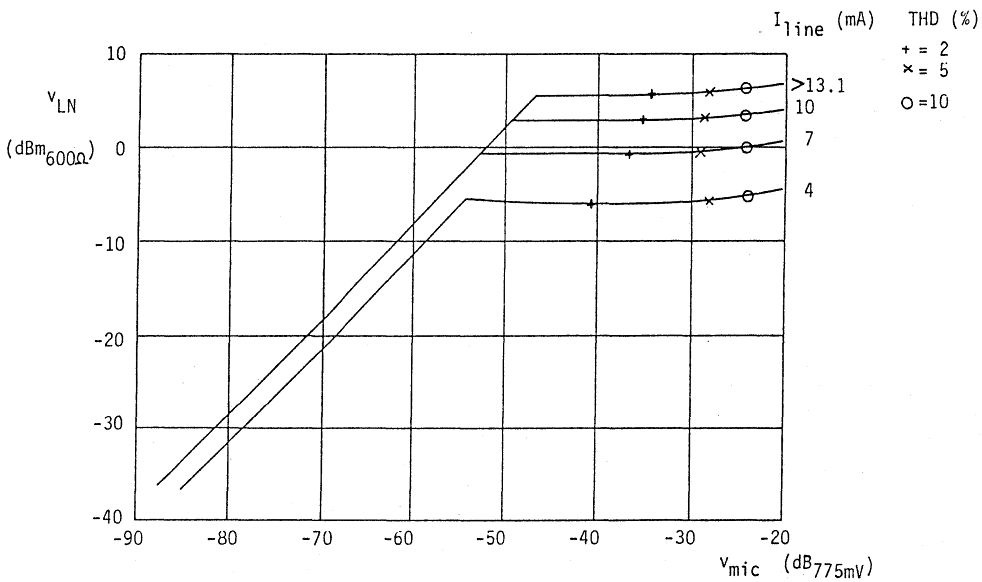


Fig. 50. Static behaviour of the dynamic limiter; sending signal on LN as function of the microphone input signal.

6.4. NOISE

The typical psophometrically weighted (P53-curve) noise level measured on the pin LN with a 600 Ω load is given as a function of microphone gain in Fig. 51. The microphone input is loaded with a 200 Ω and an 8.2kΩ resistor.

Psophometrical weighted noise level at the receive output (single ended 300 Ω load) as a function of microphone gain is shown in Fig. 52. Parameters are the overall receive gain A_{REC} (LN to the earpiece), and the resistor across the microphone inputs.

6.5. EFFECT OF POWER DOWN FUNCTION

Fig. 53 shows the line voltage V_{LN} and the line current I_{line} with and without the power down function in the application of Fig. 43. A resistor bridge was used in the exchange.

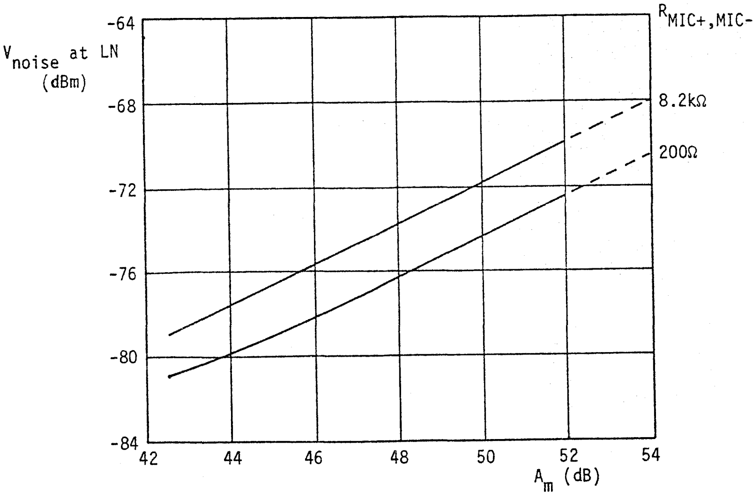
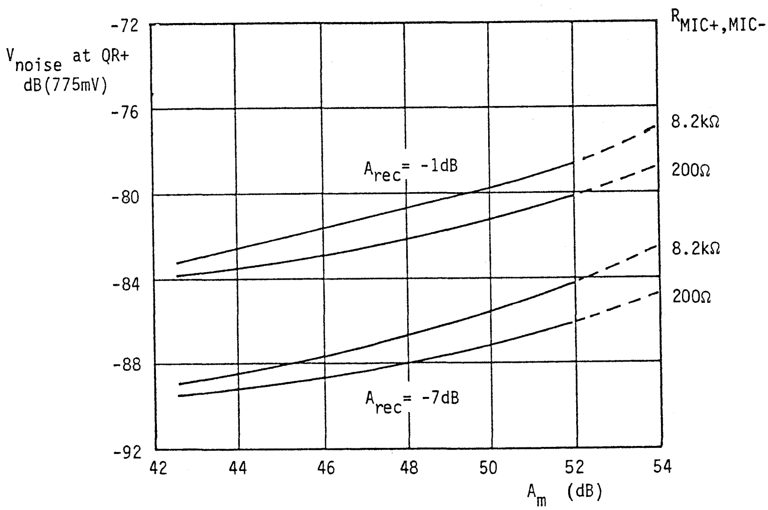


Fig. 51. Psophometrically weighted noise on the transmitter output LN as a function of microphone amplifier gain.



Arec = overall receive gain = A_{TA} - 32 dB

Fig. 52. Psophometrically weighted noise at the QR+ earpiece output as a function of microphone amplifier gain.

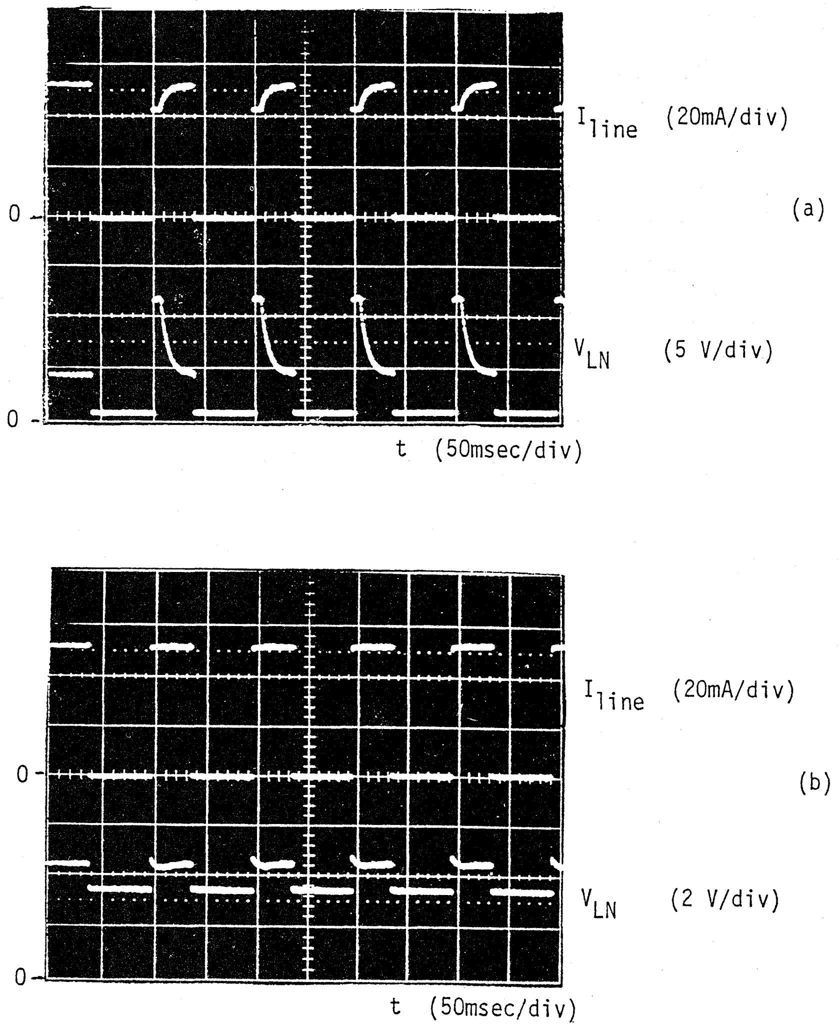


Fig. 53. Line voltage and line current during pulse dialling without the use of power down (a) and with power down function (b).

7. REFERENCES

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'TEA1081: a Supply IC for peripheral circuits in electronic telephone sets' by F. v. Dongen.
12nc 9398 058 30011

Ref. 2 Philips Components publication,
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Ref. 3 Philips Components Application Report ETT8503,
'The coupling network between the DTMF generator PCD3311/12 and the transmission circuit TEA1060/61' by J. Mulder.

Ref. 4 Philips Components Application report ETT8603,
'Stability analysis of the receiving amplifier of the TEA1060/61' by P.J.M. Sijbers.

Ref. 5 Philips components Laboratory report ETT8711,
'Bridge: A software tool for optimising the TEA1060 anti-sidetone bridge' by J.A.C.M. v. Tiggelen.

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Ref. 7 Philips Components Laboratory report ETT89008,
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Ref. 8 Philips Components Laboratory report ETT8802,
'Application Proposal for the German Market of the TEA1064 Electronic Speech Transmission Circuit' by K. Wortel.
12nc 9398 063 00011

Ref. 9 Philips Components Application Note ETT/AN89011,
'Documentation for TEA1064 printed circuit board CAB3458'
by P. Sijbers and K. Wortel

8. ACKNOWLEDGEMENT

The authors would like to thank Paul Queudot and Jean Pierre Coulmance (PCALP; Application lab Paris) for their analysis of the TEA1060 family anti sidetone bridge. Furthermore many thanks to the design team of Philips Composants/RTC Caen (Alain Charlot, Alain Gauthier and Philippe Jouen) for their work during the development phase of the TEA1064.

APPENDIX A: CIRCUIT CALCULATIONS.

This chapter gives the results of calculations done on the transmit/voltage regulator stage of the TEA1064 with stabilised supply voltage for: circuit impedance (see A.1.), stability of the voltage regulator stage (1B) and maximum transmission level which can be generated on the line (1C). Fig. A1 shows the equivalent circuit diagram in which are defined the electrical symbols used in the expressions hereafter. Z1 and Z_{line} can be replaced by R1 in case of 600 Ω set impedance respectively by RL in case of 600 Ω test load.

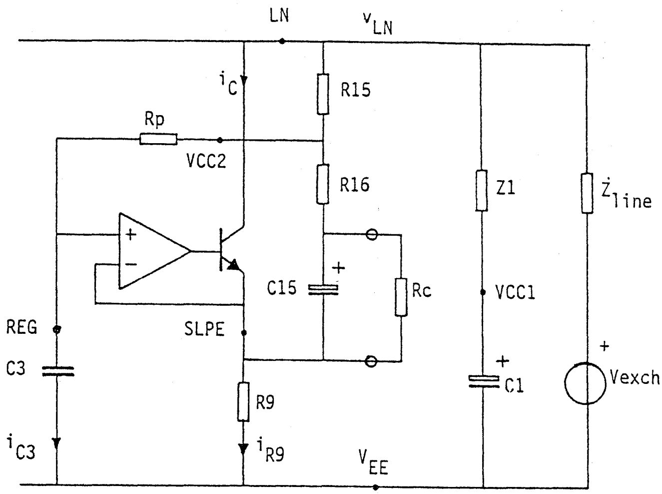


Fig. A1. Equivalent circuit diagram of transmit/voltage regulator stage of the TEA1064 with stabilised supply voltage.

A.1. CIRCUIT IMPEDANCE Z_{LN}.

Calculation of the impedance between LN and VEE, without taking into account Z1 and C1, is derived as follows:

$$Z_{LN} = \frac{v_{LN}}{i_{C3} + i_{R9}} = \frac{v_{LN}}{v_{SLPE}*(p*C3 + 1/R9)} \qquad [A1]$$

$$p = j*\omega \qquad \omega = 2*\pi*f \; ; \qquad v_{SLPE} = v_{REG} = \frac{v_{CC2}}{1 + p*C3*Rp} \qquad [A2]$$

$$i_{R15} = i_{C3} + i_{R16} \text{ thus } \frac{v_{LN} - v_{CC2}}{R15} = \frac{p*C3}{1 + p*C3*Rp} * v_{CC2} + \frac{v_{CC2} - v_{SLPE}}{Zc} \qquad [A3]$$

where Zc is the external network between VCC2 and SLPE consisting of R16, smoothing capacitor C15 and the peripheral circuits represented by Rc:

$$Zc = R16 + \frac{Rc}{1 + p*C15*Rc} \qquad [A4]$$

Deriving v_{CC2} from [A3] and substituting in [A2] gives v_{SLPE} which has to be substituted in [A1]. The resulted general formula (expression A5) for the circuit impedance is:

$$Z_{LN} = \frac{R9}{1 + p*C3*R9} *(1 + p*C3*(R15 + Rp*(1 + \frac{R15}{R16 + Rc} * \frac{1 + p*C15*Rc}{1 + p*C15*Rc//R16})))$$

With the assumption that Rp >> R15 and Rc >> R16 this expression can be converted in an equivalent network, see diagram Fig. A2, consisting of the following elements:

$$Ls = C3*R9*Rp*(1 + \frac{R15/Rc + \omega^2*C15^2*R15*R16}{1 + \omega^2*C15^2*R16^2}) \qquad [A6]$$

$$Rs = - \frac{\omega^2*C3*C15*R9*R15*Rp}{1 + \omega^2*C15^2*R16^2} \qquad [A7]$$

$$Rx = Rp*(1 + \frac{R15/Rc + \omega^2*C15^2*R15*R16}{1 + \omega^2*C15^2*R16^2}) \quad [A8] \qquad Lx = \frac{C15*R15*Rp}{1 + \omega^2*C15^2*R16^2} \quad [A9]$$

The total set impedance between LN and VEE can be calculated by:

$$Z_{SET} = \frac{(p*Ls + Rs + R9) * (p*Lx + Rx + 1/(p*C3))}{p*Ls + Rs + R9 + p*Lx + Rx + 1/(p*C3)} // (Z1 + 1/(p*C1)) \qquad [A10]$$

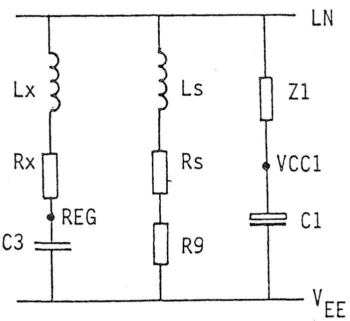


Fig. A2. Equivalent circuit impedance between LN and VEE for TEA1064 application with stabilised supply voltage.

The circuit impedance according expression [A5] can be simplified for audio frequencies. For f > 300 Hz at which ω*C15*Rc >> 1 and ω*C15*R16 >> 1 and assuming that Rp >> R15 and Rc >> R16 the circuit impedance is as follows:

$$Z_{LN} = \left(\frac{R9}{1 + p*C3*R9} * \left(1 + p*C3*Rp*\left(1 + \frac{R15}{R16}\right)\right)\right) // R1 \qquad [A11]$$

The equivalent circuit diagram for this expression is shown in Fig. 13 at which R1 >> 1/(p*C1). The elements of this equivalent circuit diagram are:

$$Req = Rp * \left(1 + \frac{R15}{R16}\right) \quad \text{and} \quad Leq = C3*R9*Req$$

The total set impedance can be calculated by:

$$Z_{SET} = \frac{(p*Leq + R9) * (Req + 1/(p*C3))}{p*Leq + R9 + Req + 1/(p*C3)} // R1 \qquad [A12]$$

A.2. STABILITY.

Application of the TEA1064 with stabilised supply point, from which the principal circuit diagram is shown in Fig. A1, introduces a second order control loop of the voltage regulator system which is unstable (LF oscillations) for some value combinations of external components.
Open loop simulations have been done to calculate which combinations of component values for this application can be advised in order to offer enough phase margin for stable operation taking into consideration the in this chapter given conditions.

Calculated is the phase margin as a function of C3 with R15/R16 as parameter. This is carried out with fixed values for R9 and C15 and with different values for R15 and supply load Rc. Set impedance Z1 is varied between 400 and 1200 Ω which covers most of the applications. For complex set impedances only the real part of Z1 is used. Stability problems (can) occur at f < 50 Hz at which the imaginary part of Z1 has no influence. Supply bridge networks as used by PTT organizations, according table A1, are applied as line load.

Results.
A part of the simulation results are given in the Figs. 14 and 20 where for a fixed phase margin of 50° the ratio R15/R16 is shown as a function of C3 (curves "S") with R15 as a parameter value. The curves for 50° phase margin are selected from the calculated phase margin figures at worst case conditions with respect to supply load Rc and line loads. Worst case figures appear at Rc is infinite and zero line length.
The phase margin varies a lot for the different supply bridges with the rest of the conditions remaining the same, while it varies also a lot for the individual bridges under different conditions. For the shown results is used the phase margin (as a function of C3) from the "worst" supply bridge.

Stability investigations were done in combination with impedance calculations due to the fact that a high phase margin will certainly result in a relatively poor set impedance (low BRL) at low audio frequencies. Fig. 14 and Fig. 20 show in fact the results of both investigations. In chapter 2.2.2. is explained how to use these graphs.

Conditions.

- R9 = 20 Ω, C15 = 220 μF.
- C3 range: 0.47 μF to 4.7 μF.
- R15 range: 200 Ω to 600 Ω.
- R16: Ratio R15/R16 is the resulting parameter of this investigation.
- Rc range: 800 Ω to infinite.
- Set impedance Z1 (LN - VCC1) is varied from 400 Ω to 1200 Ω.
- Line load (LN - VEE): a) open circuit and b) line impedances corresponding with line length between zero and 10 km, terminated by one of the supply bridges according Fig. A3 and table A1.

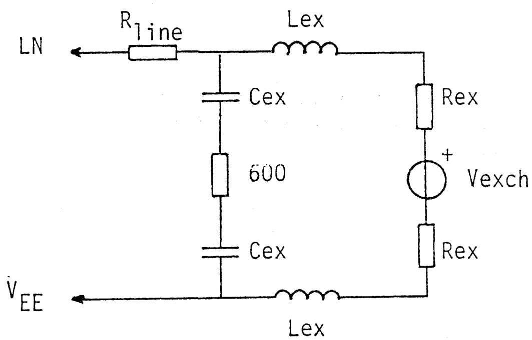


Fig. A3. Circuit diagram of the supply bridges according table A1 used for stability calculations.

Rex (Ω)	Lex (H)	Cex (μF)	PTT spec. documentation from:
200	1.5	2	UK BS6317 '82
200	10	100	UK BS6789 '84
500	10	47	Germany FTZ12R21 '88
500	1	1	Germany FTZ12R21 '88
150	5	10	Spec. for French/German tel. set'87
150	5	10	French CNET ST/CAA/ELR/305 '87
400	2	20	Netherlands BTR0148R/C/06 '88
800	3.5	100	Norway TVTST 8211A 124 '87

Table A1 Overview of supply bridges used for stability calculations.

A.3. MAXIMUM OUTPUT SENDING LEVEL.

As described in chapter 2.4. the maximum transmit line level depends on the available line current (current limiting) or on the DC voltage between LN and SLPE (voltage limiting). The maximum levels are calculated as follows:

Current limiting.

Fig. A1 is redrawn for AC into Fig. A4, assuming that for audio frequencies:

$$\frac{1}{p*C1} \ll Z1, \quad \frac{1}{p*C15} \ll (R15 + R16) \quad \text{and} \quad \frac{1}{p*C3} \ll Rp$$

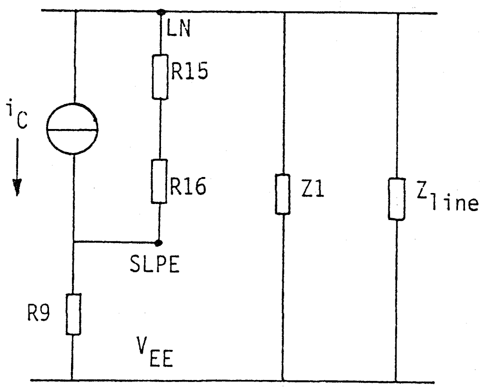


Fig. A4. Equivalent circuit diagram for AC (derived from Fig. A1)

The voltage between LN and SLPE is given by:

$$v_{LN-SLPE} = - i_C*Z_{LN-SLPE} \qquad [A13]$$

in which $Z_{LN-SLPE}$ is the total impedance between LN and SLPE.

$$Z_{LN-SLPE} = \frac{(R9 + Z1//Z_{line})*(R15 + R16)}{R9 + Z1//Z_{line} + R15 + R16} \qquad [A14]$$

The line voltage $$v_{LN} = \frac{Z1//Z_{line}}{Z1//Z_{line} + R9} * v_{LN-SLPE} \qquad [A15]$$

Substitution of [A14] in [A13] and in [A15] results in:

$$v_{LN} = - i_C * \frac{(Z1//Z_{line})*(R15 + R16)}{Z1//Z_{line} + R9 + R15 + R16} \qquad [A16]$$

The maximum AC collector peak current (Fig. A1) which can be generated equals the available DC current, thus:

$$i_C(p\text{-}p) = 2*I_C \quad [A17] \qquad \text{and} \quad I_C = I_{line} - I_{CC1} - Ip - Is \quad [A18]$$

where:
I_{CC1} is the internal current consumption; typ. 1.2 mA at V_{CC1} = 2.8 V.
Ip is the peripheral supply current.
Is is an internal current, which flows through R15 into pin V_{CC2} (Is typ. = 0.25 mA).

Substitution of [A18] in [A17] in [A16] gives the peak-peak level of the line signal as a result of the available line current.

$$v_{LN}(p\text{-}p) = 2*(I_{line} - I_{CC1} - Ip - Is) * \frac{(Z1//Z_{line})*(R15 + R16)}{Z1//Z_{line} + R9 + R15 + R16} \quad [A19]$$

Voltage limiting.

The voltage detector of the dynamic limiter (item 2.4.2) limits the maximum transmission level when the instantaneous voltage difference between LN and SLPE becomes less than 1.2 V. The sum of the voltages around the loop from V_{EE} to LN and back is:

$$V_{SLPE} + v_{SLPE}(p) + 1.2 + v_{LN}(p) - V_{LN} = 0 \quad [A20]$$

with $v_{LN}(p) = -\dfrac{Z1//Z_{line}}{R9} * v_{SLPE}(p)$ substituted in [A20] results in:

$$V_{LN\text{-}SLPE} - 1.2 - v_{LN}(p) * \left(1 + \frac{R9}{Z1//Z_{line}}\right) = 0 \quad [A21]$$

The maximum peak to peak line signal is:

$$v_{LN}(p\text{-}p) = 2 * \frac{V_{LN\text{-}SLPE} - 1.2}{1 + \dfrac{R9}{Z1//Z_{line}}} \quad [A21]$$

where:
$V_{LN\text{-}SLPE}$ is a constant DC voltage drop for the application with regulated line voltage. This voltage drop depends on Ip in case of regulated supply voltage.
Z1 is the set impedance between LN-VCC1 which can be replaced by R1 in case of 600 Ω set impedance.
Z_{line} is the line impedance between LN-VEE which has to be replaced by RL in case of 600 Ω test load.

APPENDIX B: TEA1060 FAMILY ANTI SIDETONE BRIDGE

This bridge type has been analyzed mathematically in Appendix A of Ref. 2. In this appendix it is shown that some simple rearrangements of the expressions found in Ref. 2 yields two bridge conditions that are more accurate than the ones derived in Ref. 2.
Fig. B1 shows the basic circuit diagram of TEA1060-family anti sidetone bridge. Fig. B2 shows the equivalent circuit under sending conditions.

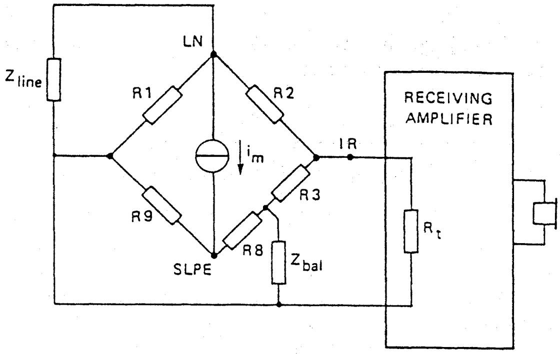


Fig. B1. TEA1060 family anti sidetone bridge

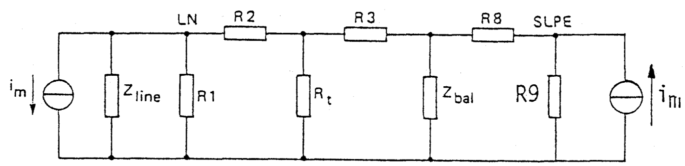


Fig. B2. Equivalent circuit to the TEA1060 bridge under sending conditions.

Using the same method as used in Ref. 2 the following expression can be derived [R1 has been substituted by Ri (the dynamic impedance of the TEA1064 between LN and VEE)]:

$$v_{IR} = \frac{im}{(R2//Rt)+R3+(R8//Zbal)} * \frac{Rt}{R2+Rt} * \left[R9 \frac{Zbal}{Zbal+R8} R2 - Ri \frac{Z_{line}}{Z_{line}+Ri} \{R3+(R8//Zbal)\}\right] \quad [B1]$$

where $Ri = R1//Req$ is the dynamic impedance of the circuit

$$Req = Rp*(1+\frac{R15}{R16}) \quad \text{and } Rp = 15.5k\Omega$$

The signal at the input of the receiving amplifier is completely cancelled when:

$$R9\frac{Zbal}{Zbal+R8+R9} * R2 - Ri*\frac{Z_{line}}{Z_{line}+Ri} * \{R3+(R8//Zbal)\}] = 0 \qquad [B2]$$

Rearranging [B2] yields:

$$R9ZbalR2(Ri+Z_{line}) - RiZ_{line}R3Zbal - RiZ_{line}R3R8 - RiZ_{line}R8Zbal = 0 \qquad [B3]$$

$$= ZbalZ_{line}[R9R2 - Ri(R3+R8)] + (R9R2ZbalRi - R3(R8+R9)RiZ_{line}) = 0 \qquad [B4]$$

This expression is fulfilled when:

(a) $R9R2 = Ri(R3 + R8)$ [B5]

and

(b) $$Zbal = \frac{R3(R8+R9)}{R9R2} Z_{line} = k\,Z_{line} \qquad [B6]$$

where k is a scale factor and with [B5] substituted in [B6]:

$$k = \frac{R3(R8+R9)}{R9R2} = \frac{R3(R8+R9)}{Ri(R3+R8)}$$

APPENDIX C: APPLICATION CIRCUIT EXAMPLES

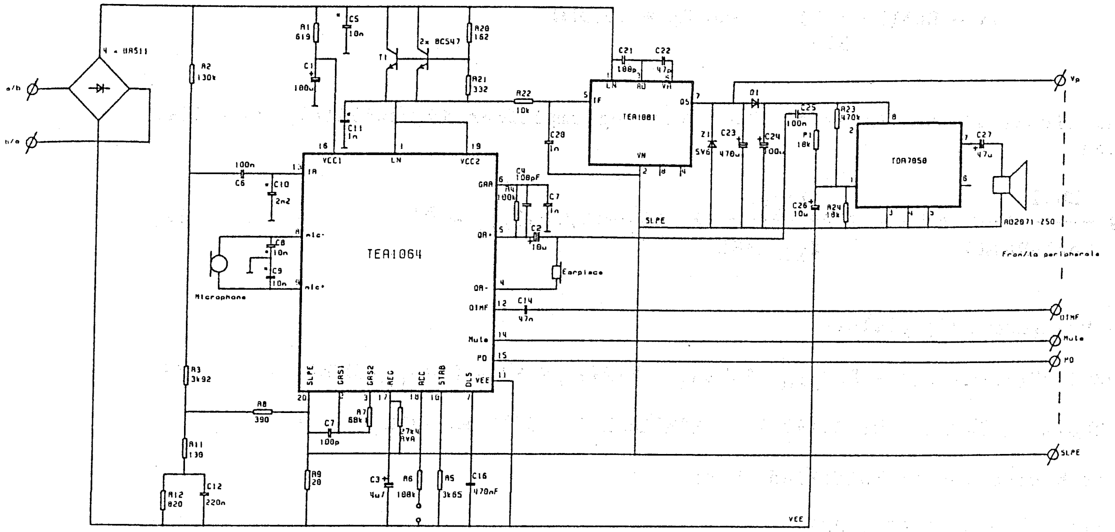


Fig. C1. Circuit diagram for listening-in application with TEA1064, TEA1081 and TDA7050 (Ref.7).

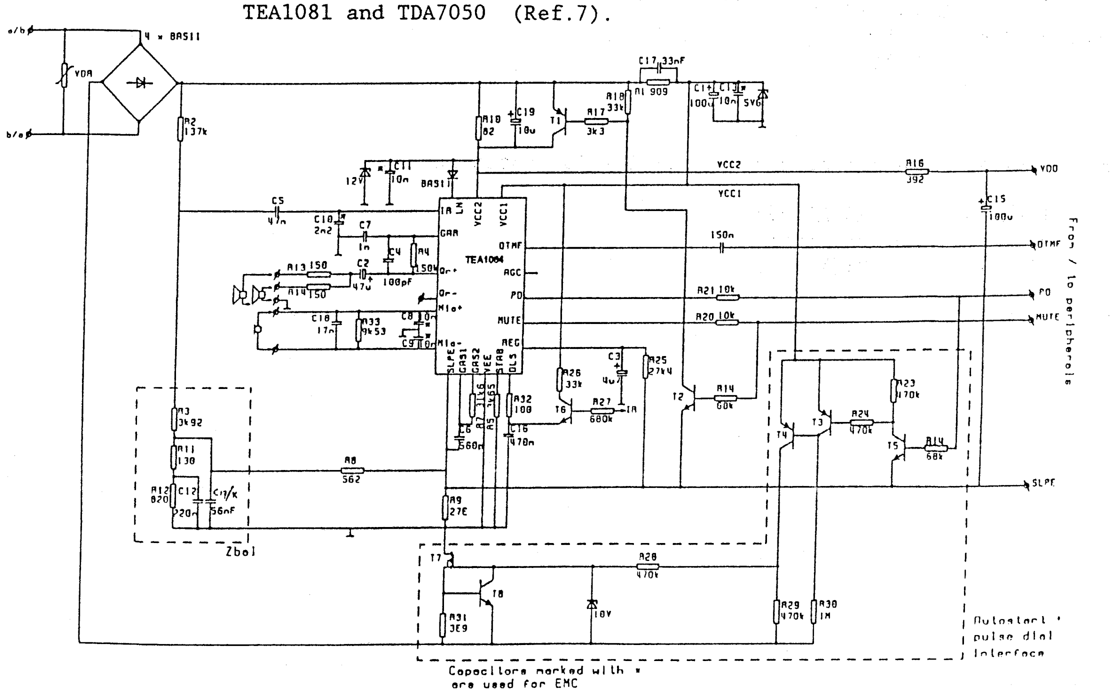
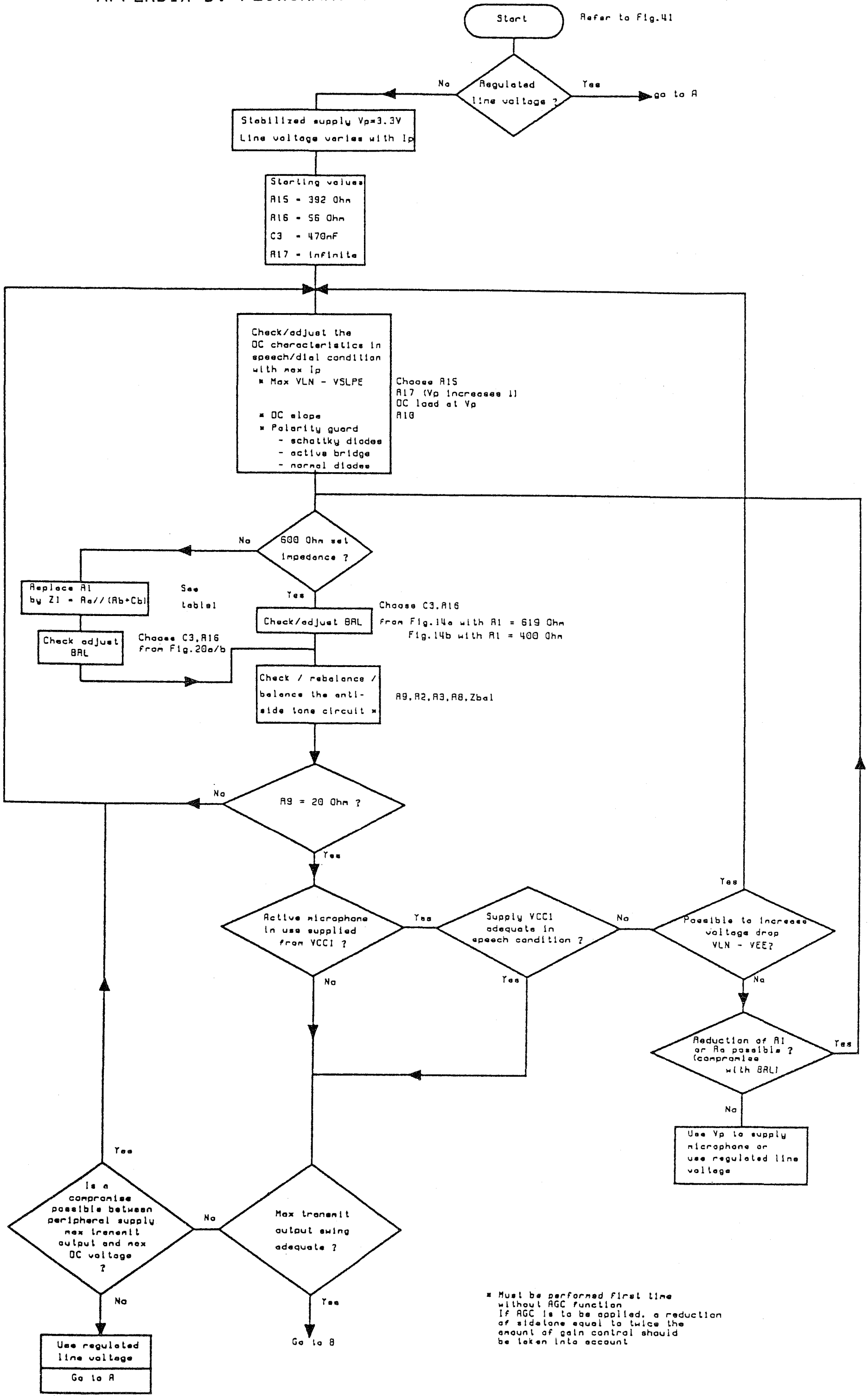
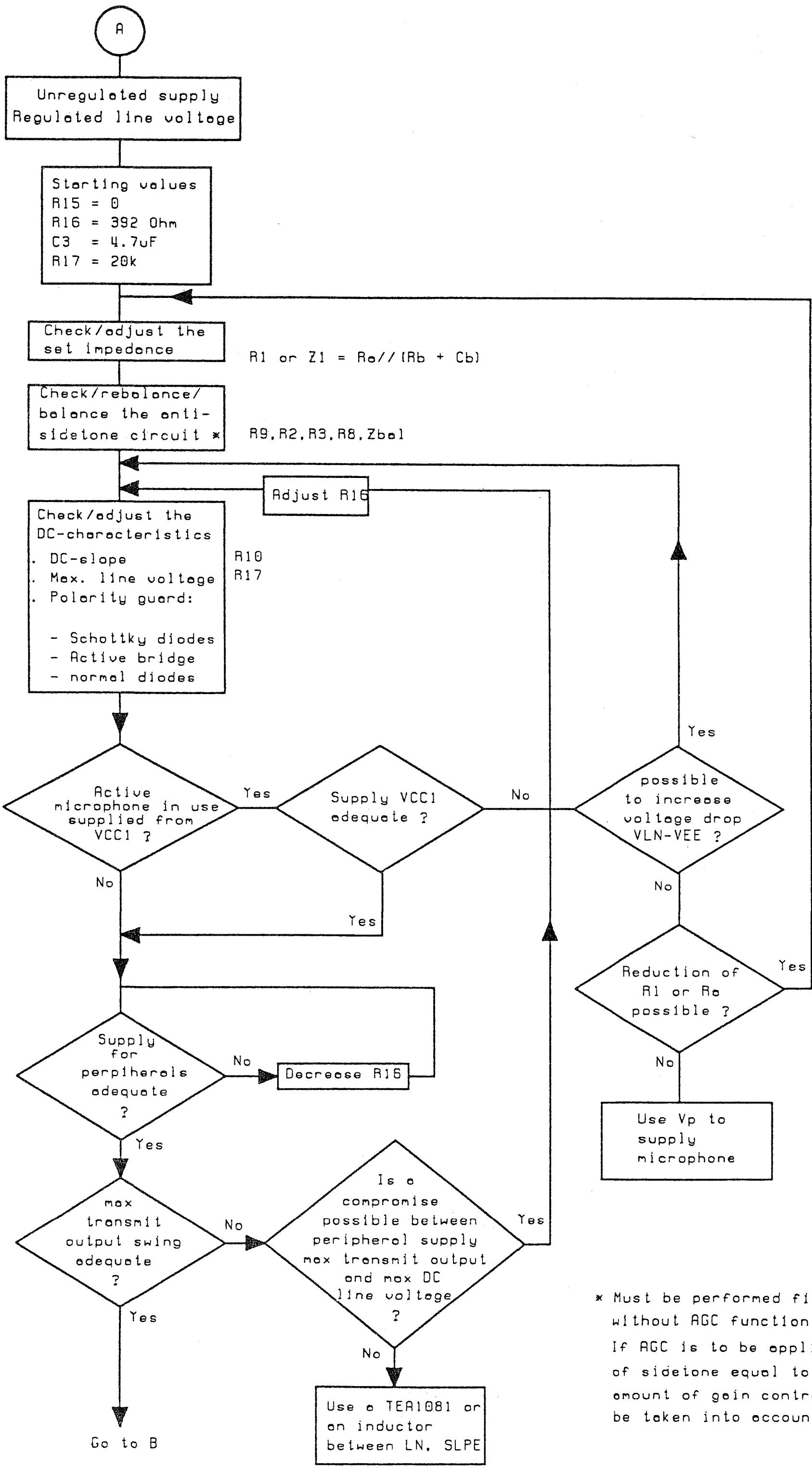


Fig. C2. Application proposal (Ref.8) of the transmission part for an electronic telephone set according to the German requirements as described in "FTZ Pflichtenheft: FTZ121TR8-8 (jan. '88)

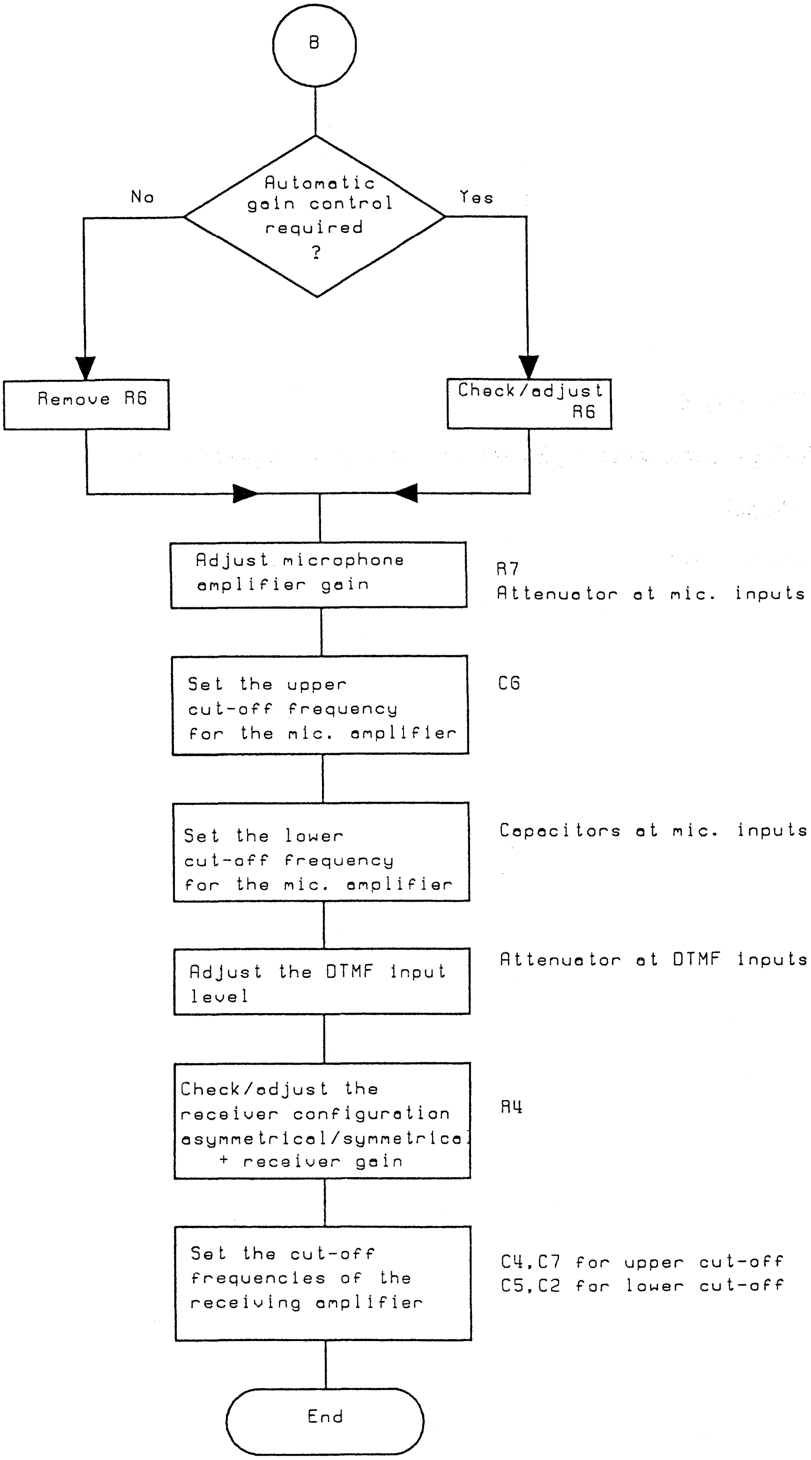
APPENDIX D: FLOWCHART TO ADJUST PARAMETERS OF THE TEA1064



* Must be performed first time without AGC function
If AGC is to be applied, a reduction of sidetone equal to twice the amount of gain control should be taken into account



* Must be performed first time without AGC function
If AGC is to be applied, a reduction of sidetone equal to twice the amount of gain control should be taken into account


B
Automatic gain control required ?
No
Yes
Remove R6
Check/adjust R6
Adjust microphone amplifier gain
R7
Attenuator at mic. inputs
Set the upper cut-off frequency for the mic. amplifier
C6
Set the lower cut-off frequency for the mic. amplifier
Capacitors at mic. inputs
Adjust the DTMF input level
Attenuator at DTMF inputs
Check/adjust the receiver configuration asymmetrical/symmetrical + receiver gain
R4
Set the cut-off frequencies of the receiving amplifier
C4,C7 for upper cut-off
C5,C2 for lower cut-off
End

APPLICATION NOTE Nr	ETT/AN90005
TITLE	The TEA1064A with complex set impedance and complex line termination
AUTHOR	K. Wortel
DATE	February 1990

Table of Contents:

1. Introduction

In the early days of electronic telephone sets PTT requirements were based upon the classical telephone sets. The AC impedance of such a set was 600Ω, a figure based upon the average characteristic impedance of the lines used (air cables). The allowed variation on this impedance is expressed as Balance Return Loss (BRL):

$$BRL = 20*\log\left|\frac{Zset + Zref}{Zset - Zref}\right| \quad \text{with } Zref = 600\Omega$$

Requirements nowadays vary from 6dB (USA) up to 18dB (Germany)

Over the last couple of years more and more PTT's switch over to the requirement of complex impedances for telephone sets, because of better line termination (ground cables).

Examples:

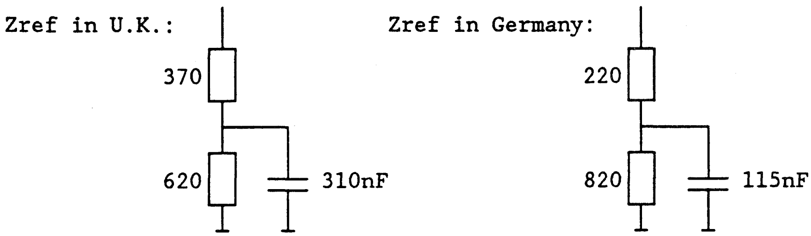


Figure 1: Complex reference impedances for UK and Germany

When using the TEA1064A with complex set impedance high frequency roll off (above ≈1500Hz) in sending direction occurs. This is caused by the sending amplifier configuration. Up to now however, most countries measure the sending gain curves with a 600Ω line termination. In this way the influence of the complex set impedance on the sending characteristics is smoothed (600Ω//Zset). In practice therefore, in most cases a compromise has to be made between set impedance and sending frequency curves in order to fulfill requirements.

Recently however, for instance the German PTT decided to measure the sending characteristics with the line terminated with Zref. Besides that, BRL figures have been increased up to 18dB in the frequency range of 500-2500Hz. These two new requirements make it impossible to compromise between BRL and frequency roll-off (Zset//Zref). Total roll-off within the audio band (300-3400Hz) will be up to 6dB! It can be expected that other PTT's will follow the German requirements in future.

This report describes an application hint for the TEA1064A to obtain a high balance return loss (with complex Zref) and a nearly flat frequency response (with complex termination).

2. Balance return loss

As explained in the introduction a high balance return loss (BRL) will be obtained if the telephone set impedance approximates the reference impedance. For the TEA1064A the set impedance is mainly determined by the impedance connected between LN and VCC1. In figure 2 the basic application of the TEA1064A is shown:

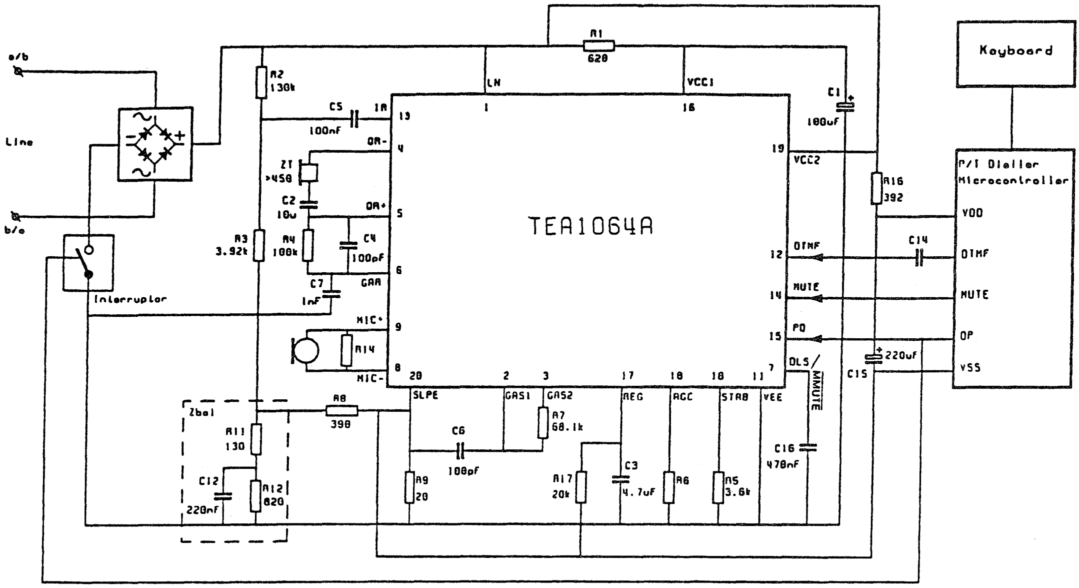


Figure 2 "Basic TEA1064A application"

The total equivalent impedance of the circuit of figure 2 (see also reference 1) is shown below:

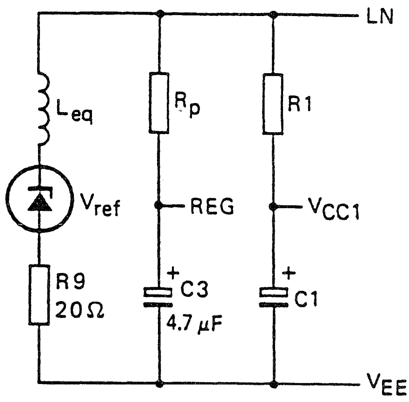


Figure 3 "Equivalent impedance of TEA1064A"

For R1 a complex impedance can be installed which equals the reference impedance. If fairly high BRL figures are required at low frequencies (e.g. >18dB at 500Hz) the absolute impedance of the inductance can influence the set impedance too much. In such a case the value of Leq (Leq = Rp*R1*C3) can be increased by increasing C3 (at pin REG) or R9 (at pin SLPE). It should be noted however, that changing R9 will also change other transmission parameters (e.g. sending gain, sidetone, AGC, etc.) while the DC start-up time is proportional to the value of C3.

SIDETONE:
The impedance between LN and VCC1 is a part of the anti sidetone bridge (2 to 4 wire conversion). If R1 is changed from the original value (figure 2) sidetone attenuation changes. For calculating the new bridge balance please refer to reference 1.

SUPPLY POINT VCC1:
The supply point VCC1 is used to supply the internal circuitry. This voltage is obtained from the line via the impedance between LN and VCC1 (≃ set impedance) and smoothed by a big capacitor (figure 2). Since a complex impedance usually has a higher DC resistance (≃1kΩ) compared to the standard 600Ω resistor, voltage VCC1 will be lower. For correct operation it can be necessary to increase the reference voltage of the TEA1064A by means of resistor R17 (see also figure 2). For more details refer to reference 1.

3. Frequency compensation in sending direction

The gain of the microphone amplifier, between LN and the microphone inputs of the TEA1064A, is given by the following equation (see also reference 1):

$$Am = 1.356 * \frac{RGS}{R5*R9} * RLN \qquad (1)$$

In which:
RGS = impedance between GAS1 and SLPE
R5 = 3.65kΩ resistor connected to STAB
R9 = resistor connected to SLPE
RLN = impedance seen at pin LN

In case of a complex set impedance and a complex line termination the load at LN will not be constant with frequency. The absolute impedance decreases above approximately 1500Hz, thus within the speech band (300-3400Hz). For equation (1) this means that the microphone gain will decrease!
In order to compensate for this effect the network between GAS1 and GAS2, as shown in figure 4, can be used:

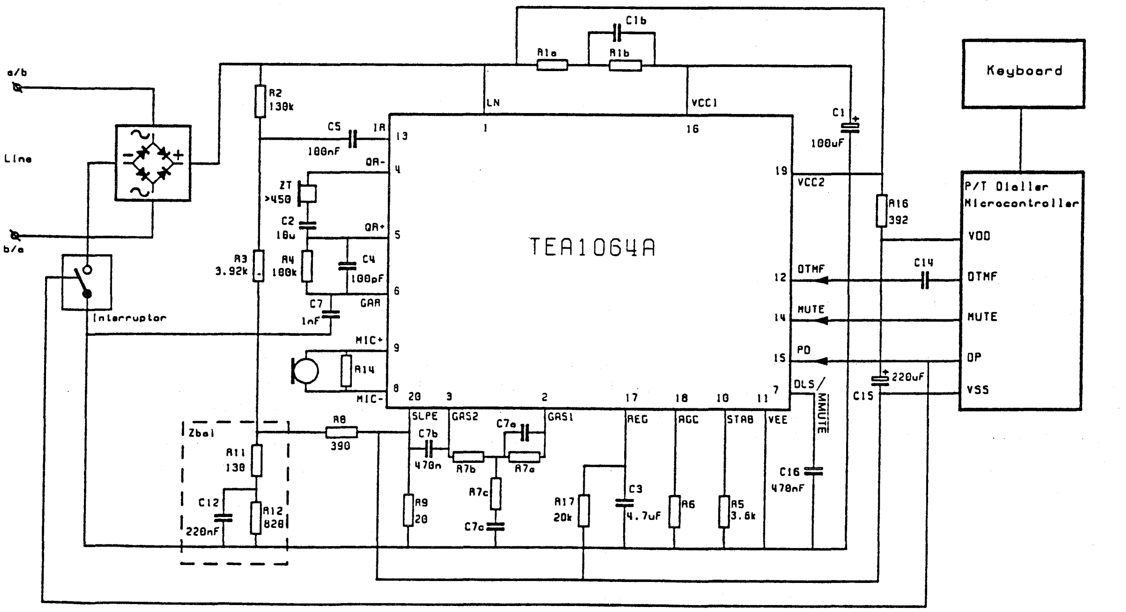


Figure 4 "TEA1064A application with complex set impedance"

With this network a pre-emphasis is created in the sending amplifier:

ZERO: $fz = 1/[2*\pi*C7c\{(R7a//R7b) + R7c\}]$ (2)
POLE: $fp = 1/[2*\pi*C7c*R7c]$ (3)

The pole and the zero should eliminate respectively the zero and the pole occurring at the line due to the complex load.

In order to calculate the optimum values for the components of this network the exact load at LN needs to be determined. For correct calculation of this load the internal impedance of the TEA1064A (15.5kΩ) has to be included. If also other components are connected to the line (e.g. EMC capacitor) they should be included in the calculations as well.

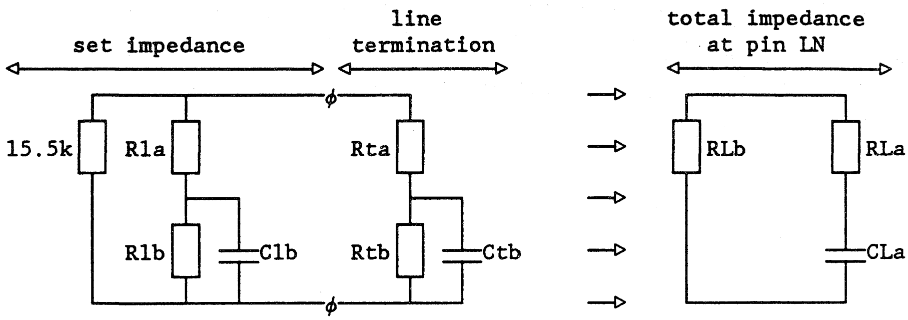


Figure 5: "Schematic representation of line load calculations"

$$\text{Impedance at LN: } Zline = \frac{RLb(1 + j*w*RLa*CLa)}{(1 + j*w*CLa(RLa + RLb)}$$

POLE: $fp = 1/[2*\pi*CLa(RLa + RLb)]$ (4)
ZERO: $fz = 1/[2*\pi*RLa*CLa]$ (5)

Equation 1 can be used to determine a value for RGS. For RLN the maximum value should be filled in (the low frequency value ≈300Hz). For low sending distortion and accurate gain setting the internal impedance of the TEA1064A between pins SLPE and GAS2 must be short circuited. In figure 4 this is done by means of the parallel capacitor C7b of 470nF. Now the following equation is valid:

$R7a + R7b = RGS$ (6)

The outcome of equation (4) and (5) (discrete frequencies) can be substituted in (2) and (3) respectively. Together with equation (6), 3 equations are left with 4 unknown parameters. By picking one parameter (e.g. R7a=R7b or a value from the E12 series for capacitor C7c) the others can be calculated.

For stability reasons a capacitor must be connected in parallel with R7a, which can also be used to create a first order roll-off. The minimum value of this capacitor is determined by the maximum cut-off frequency of about 20kHz:

$$1/[2*\pi*C7aR7a] < 20kHz$$

Example:

In Germany the complex set impedance and complex line termination are equal: 220 + 820//115nF. For calculating the total load at the line the 15.5kΩ internal TEA1064A resistor needs to be incorporated. We obtain (see also figure 5):

RLb = 500Ω
RLa = 140Ω
CLa = 143nF

with equation (4): fp = 1740Hz
with equation (5): fz = 7950Hz

Filling in these values in equations (2) and (3) respectively:

$$1/[2*\pi*C7c\{(R7a//R7b) + R7c\}] = 1740 \quad \text{(I)}$$
$$1/[2*\pi*C7c*R7c] = 7950 \quad \text{(II)}$$

For the microphone gain 44.5dB (168x) can be chosen as an example, with R9 = 39Ω and R5 = 3.65kΩ we obtain from equation (1):

$$Am = (1.356 * RGS * RLN) / (3.65k\Omega * 39\Omega)$$

For RLN the low frequency value of the line load has to be taken which is 500Ω, together with equation (6) we now obtain:

$$RGS = R7a + R7b = 35.3k\Omega \quad \text{(III)}$$

Choosing	C7c = 10nF
with (II) we obtain	R7c = 2kΩ
combined with (I) and (III):	R7a = 25.4kΩ, practical R7a = 25.5kΩ
	R7b = 9.9kΩ, practical R7b = 10kΩ

A practical value for C7a could be 470pF resulting in high frequency roll-off during sending at f-3dB = 13.3kHz.

Ripple

Ripple can be caused by mismatch between the pre-emphasis curve (pole and zero) and the frequency roll-off caused by the complex load at LN. The allowable sending ripple in the audio band determines the allowable tolerances of the resistors and capacitors for the GAS1-GAS2 network. Since in practice 1% resistors can be used the ripple is mainly determined by capacitor tolerances (in practice 5% or 10%). At 10% capacitor tolerance 0.8dB maximum ripple can be expected, at 5% tolerance 0.4dB ripple can occur.

4. Conclusions

In countries were a complex set impedance is required and sending characteristics are measured with a complex line termination, the standard TEA1064A application does not satisfy. This is mainly due to high frequency roll-off within the speech band (300Hz - 3400Hz).

The TEA1064A application hint, mentioned in this report, makes it possible to obtain high balance return loss figures with complex set impedances. Besides, the sending frequency characteristic remains flat even with complex line terminations. For this purpose the number of components for the standard application extends with only 2 ceramic capacitors and 2 resistors.

5. References:

1. Philips Laboratory Report ETT89009; Application of the Versatile Speech/Transmission Circuit TEA1064 in full Electronic Telephone Sets; by F. van Dongen and P.J.M. Sijbers.

Relevant literature:

2. Philips Laboratory Report ETT8802; Application Proposal for the German Market of the TEA1064 Electronic Speech Transmission Circuit; by K. Wortel

3. Philips Laboratory report ETT/AN90001; Documentation for TEA1064A printed circuit board CAB3458; by P. Biermans, P. Sijbers and K. Wortel

4. TEA1064A datasheet

APPLICATION NOTE Nr	AN95050
TITLE	Application of the TEA1112 and TEA1112A transmission circuits
AUTHOR	Fernand Courtois and Fred van Dongen
DATE	November 1995

CONTENTS

Tables and figures

1. INTRODUCTION

The TEA1112/A offer all the speech and line interface functions required in electronic telephone sets. They perform the interface between the telephone line and transducers such as microphone capsule(s), earpiece, loudspeaker (in case of LI or HF functions) as well as dialler circuit for DTMF and pulse dialling.

Moreover, they offer a hook-status indicator by means of LED output. Both ICs have a MUTE function to switch between conversation and dialling as well as a MMUTE function to disable the microphone channel to give some privacy (furthermore, this MMUTE function enables the sending DTMF channel, if needed for some specific applications).

The difference between the TEA1112 and the TEA1112A concerns the MUTE and MMUTE inputs. For TEA1112, the MUTE and MMUTE functions are active for a high level at the inputs; while for TEA1112A, the $\overline{\text{MUTE}}$ and $\overline{\text{MMUTE}}$ functions are active for a low level on these inputs. TABLE 1 shows the enabled channels depending on the levels on these two inputs. It can be seen that the MUTE function acts on both sending and receiving channels, while the MMUTE function only acts on the sending channel.

TABLE 1 Channel selection

TEA1112				
MUTE	'LOW'		'HIGH'	
MMUTE	'LOW'	'HIGH'	'LOW'	'HIGH'
Microphone	ON	OFF	OFF	OFF
DTMF	OFF	ON	ON	ON
Earpiece	ON	ON	OFF	OFF
Confidence Tone	OFF	OFF	ON	ON
TEA1112A				
$\overline{\text{MUTE}}$	'LOW'		'HIGH'	
$\overline{\text{MMUTE}}$	'LOW'	'HIGH'	'LOW'	'HIGH'
Microphone	OFF	OFF	OFF	ON
DTMF	ON	ON	ON	OFF
Earpiece	OFF	OFF	ON	ON
Confidence Tone	ON	ON	OFF	OFF

The report is divided into two parts. The first part, up to chapter 3, gives a detailed description of the different circuit blocks of the TEA1112/A consisting of operating principles, settings of DC and transmission characteristics and performances of the different functions.

The second part describes two application examples of the TEA1112 by means of descriptions, settings, measurement results and performances. The consecutive steps to design or to adjust the basic application of the TEA1112/A are handled.

An evaluation board with the basic application of the TEA1112/A is available [9]. The results of the RF immunity tests of this board are shown in this report extended with a brief description of the board and the layout of the board wiring.

2. BLOCK DIAGRAM AND PINNING

The block diagram of TEA1112/A is shown by means of Fig.1. The pinning is shown in Fig.2 and Fig.3.

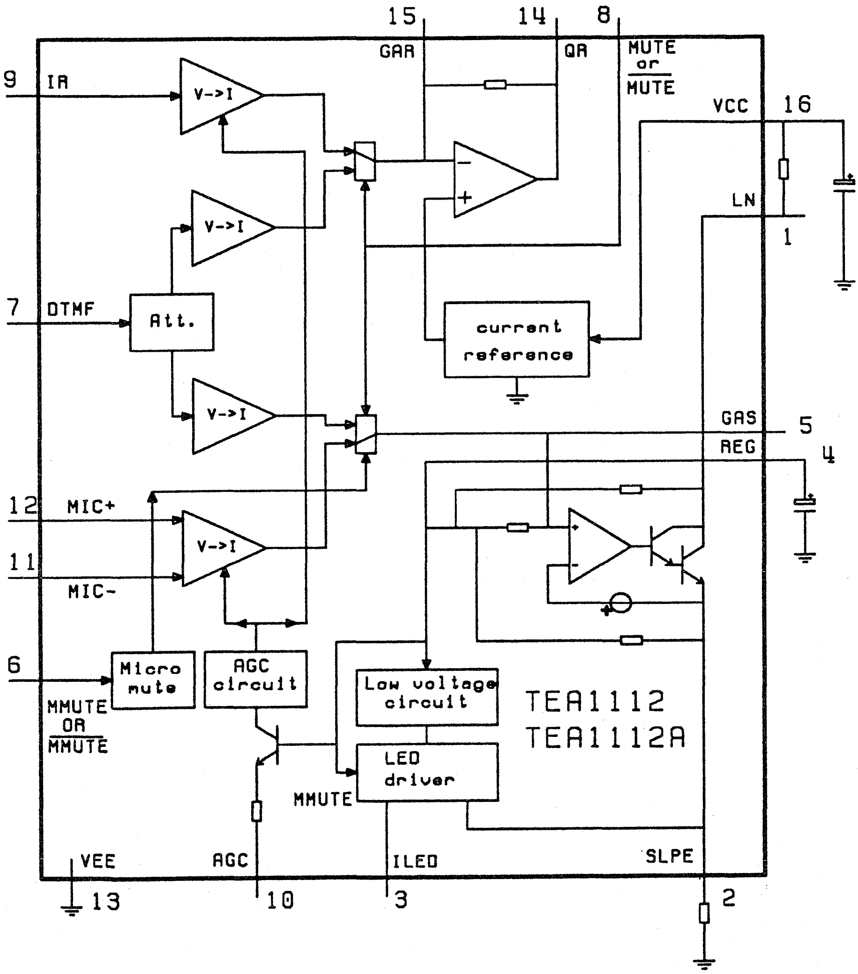


Fig.1 TEA1112/A Block Diagram

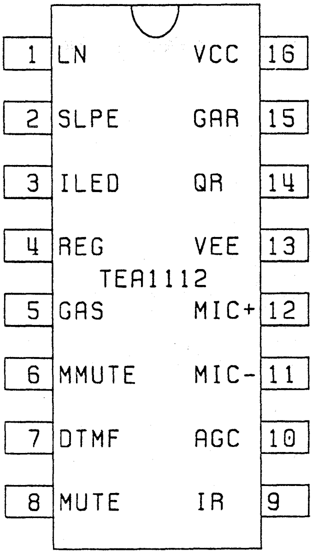


Fig.2 TEA1112 pinning

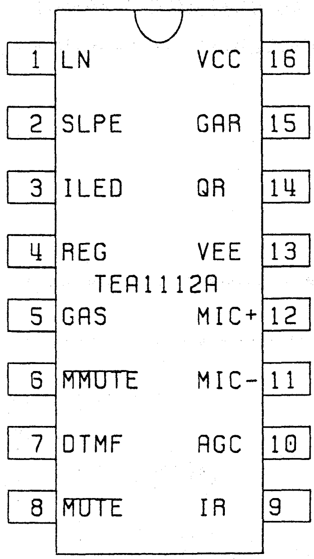


Fig.3 TEA1112A pinning

Pin	Name	Description
1	LN	Positive line terminal
2	SLPE	Slope adjustment
3	ILED	Current available to drive a LED
4	REG	Line voltage regulator decoupling
5	GAS	Sending gain adjustment
6	MMUTE	Microphone mute input
7	DTMF	Dual-Tone Multi Frequency input
8	MUTE	Mute input
9	IR	Receiving amplifier input
10	AGC	Automatic gain control
11	MIC-	Non-inverting microphone input
12	MIC+	Inverting microphone input
13	VEE	Negative line terminal
14	QR	Receiving amplifier output
15	GAR	Receive gain adjustment
16	VCC	Supply voltage for speech and peripherals

Pin	Name	Description
1	LN	Positive line terminal
2	SLPE	Slope adjustment
3	ILED	Current available to drive a LED
4	REG	Line voltage regulator decoupling
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6	$\overline{\text{MMUTE}}$	Microphone mute input
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8	$\overline{\text{MUTE}}$	Mute input
9	IR	Receiving amplifier input
10	AGC	Automatic gain control
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12	MIC+	Inverting microphone input
13	VEE	Negative line terminal
14	QR	Receiving amplifier output
15	GAR	Receive gain adjustment
16	VCC	Supply voltage for speech and peripherals

3. DESCRIPTION OF THE IC

All the curves shown in this section result from the measurement of a typical sample. All the component names refer to the basic application of the IC shown in Fig.4.

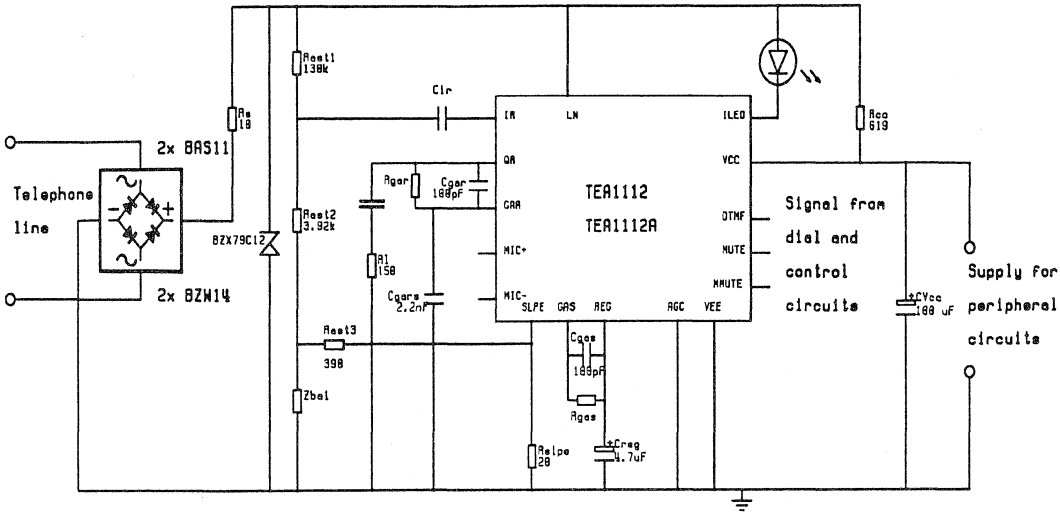


Fig.4 Basic application used for measurement

3.1 Supply; pins LN, SLPE, VCC, REG.

3.1.1 TEA1112/A Supply

Principle of operation

The supply for the TEA1112/A is obtained from the telephone line. The ICs generate a stabilized voltage (called VREF) between pins LN and SLPE. This reference voltage, typically 3.35 V, is temperature compensated. The voltage at pin REG is used by the internal regulator to generate the stabilized VREF voltage and is decoupled by a capacitor Creg connected to VEE. For effective operation of the telephone set, the TEA1112/A must have a low resistance to DC and a high impedance to speech signals. The Creg capacitor, converted into an equivalent inductance (as mentioned in the set impedance section), realizes this set impedance conversion from its DC value (Rslpe) to its AC value (Rcc in the audio frequency range). The DC voltage at pin SLPE is proportional to the line current.

The general supply configuration is shown in Fig.5

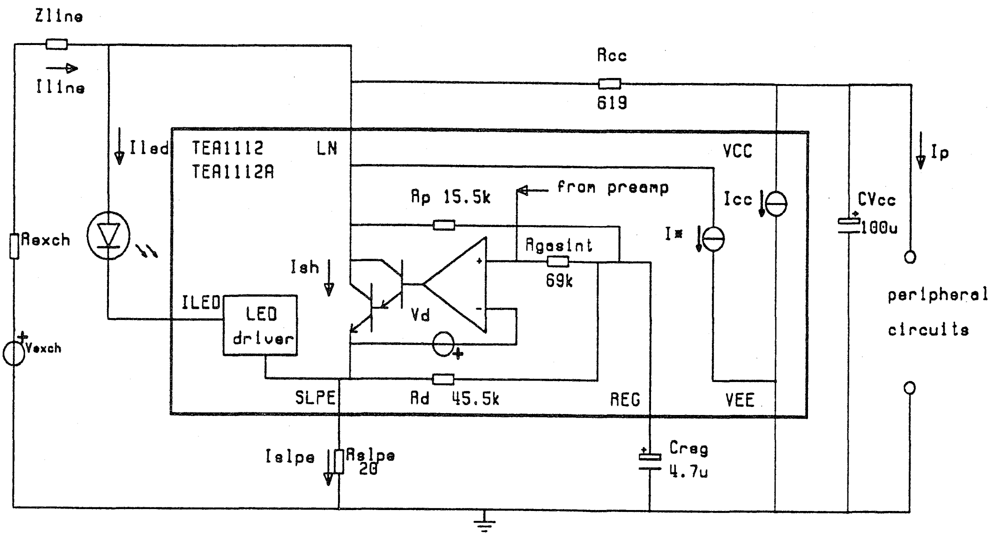


Fig.5 Supply configuration

The ICs regulate the line voltage at the pin LN. The voltage on pin LN can be calculated as:

$$VLN = VREF + Rslpe \times Islpe \quad (1)$$

$$Islpe = Iline - Icc - Ip - I^* = Iled + Ish \quad (2)$$

- Iline: Line current
- Icc: Current consumption of the IC
- Ip: Supply current for peripherals
- I*: Current consumed between LN and VEE
- Iled: Supply current for a LED component
- Ish: Excess line current shunted to SLPE (and VEE) from LN

The DC line current Iline flowing into the set is determined by the exchange supply voltage Vexch, the feeding bridge resistance Rexch, the DC resistance of the telephone line Rline and the voltage across the set including diode bridge.

Below a threshold line current (Ith is typically equal to 7.5 mA) the internal reference voltage (generating VREF) is automatically adjusted to a lower value. This means that more sets can operate in parallel with DC voltage down to an absolute minimum voltage of 1.6 V excluding the diode bridge. For line currents below this threshold current, the circuit has reduced sending and receiving performances. This is called the low voltage area.

The internal circuitry of the TEA1112/A is supplied from pin VCC. This supply voltage is derived from the line voltage by means of a resistor (Rcc) and must be decoupled by a capacitor (Cvcc). Fig.6 shows the IC current consumption (Icc) as a function of the VCC supply voltage.

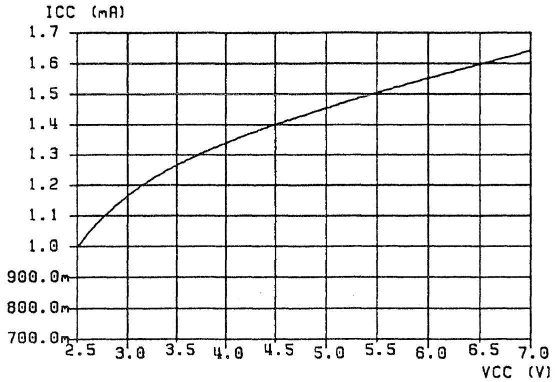


Fig.6 ICC versus VCC

Fig.7 shows the main DC voltages as a function of the line current, while Fig.8 shows the behaviour in the low voltage area.

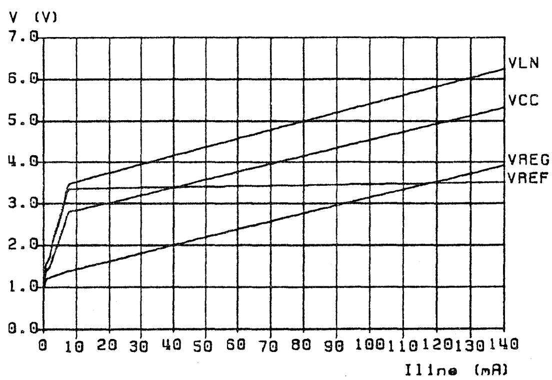


Fig.7 Main voltages versus line current

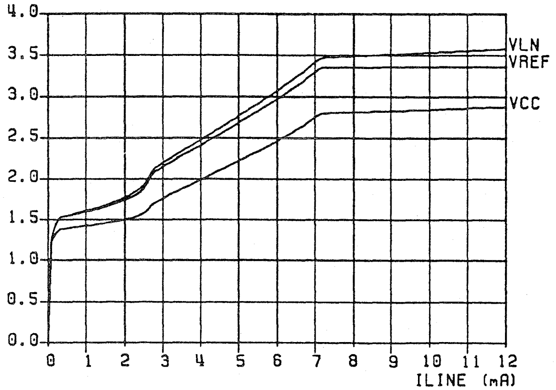


Fig.8 Low voltage behaviour

Adjustment

The reference voltage, VREF, can be adjusted by means of an external resistor Rva. It can be increased by connecting the Rva resistor between pins REG and SLPE, or decreased by connecting the Rva resistor between pins REG and LN. However this voltage reduction is possible, it is not recommended to use it, because it reduces the peripheral supply capability. Fig.9 shows the reference voltage, VREF, as a function of an Rva resistor. To ensure correct operation, the reference voltage is preferably not adjusted to a value lower than 3 V or higher than 7 V. These adjustments will slightly affect a few parameters: there will be a small change in the temperature coefficient of VREF and a slight increase in the spread of this voltage reference. Furthermore, the Rva resistor connected between REG and LN will slightly affect the set impedance (See section: 'Set impedance' 3.2).

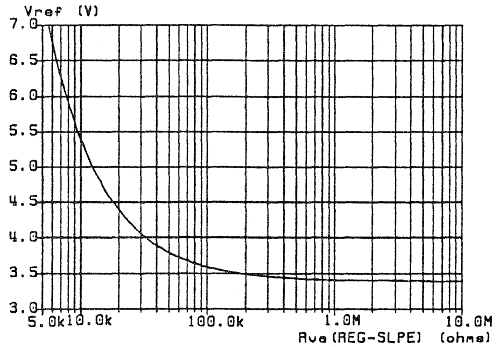


Fig.9 Influence of an Rva resistor between REG and SLPE on VREF

The DC slope of the voltage on pin LN is influenced by the Rslpe resistor as shown in Fig.10. The preferred value for Rslpe is 20 Ω. Changing this value will affect more than the DC characteristics. It also influences the microphone and DTMF gains, the LED supply current characteristic, the gain control characteristics, the sidetone level, the maximum output swing on the line and the low voltage current threshold Ith.

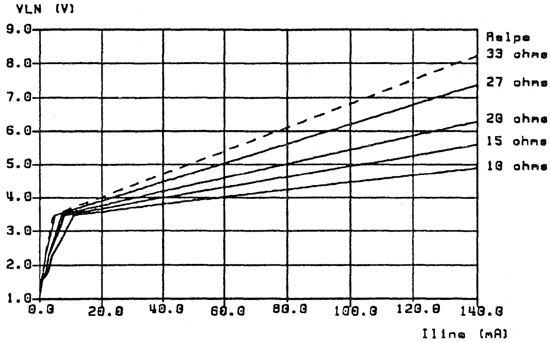


Fig.10 Influence of Rslpe on the DC slope of the line voltage

3.1.2 Supply for peripheral circuits

Principle of operation

The supply voltage at pin VCC is normally used to supply the internal circuitry of the TEA1112/A. However, a small current can be drawn to supply peripheral circuits having VEE as a ground reference. The VCC supply voltage depends on the current consumed by the IC and the peripheral circuits as shown by formula (3) (See also curves at Fig.11 and equivalent schematic of this supply point at Fig.12).

Rccint is the output impedance of the voltage supply point. As can be seen from Fig.6, the internal supply current Icc depends on the voltage on the pin VCC; it means that the impedance of the internal circuitry connected between VCC and VEE is not infinite. While supplying a peripheral circuit on VCC, the Ip supply current flows through the Rcc resistor, decreases the value of the voltage on the pin VCC. This voltage reduction affects the Icc consumption and than the voltage drop across the Rcc resistor. So to calculate the voltage drop across this resistor, both effects must be taken into account. The impedance to use in combination with Ip is not Rcc but Rcc in parallel with the impedance of the internal circuitry connected between VCC and VEE. That is what is called Rccint. For a line current equal to 15 mA and Rcc equal to 620Ω, this Rccint impedance is equal to 550Ω. The worst case for Rccint is Rcc.

$$VCC = VCCo - Rccint * (Irec + Ip) \qquad (3)$$

$$VCCo = VLN - Rcc * Icc$$

Irec = internal current necessary to supply the earpiece amplifier to realize an AC peak voltage Vq across the earpiece impedance Rl

$$Irec = \frac{Vq}{\pi \times Rl}$$

Rccint = is due to the fact that Icc slightly varies with the voltage on VCC. A worst case value for Rccint is Rcc

Rccint = Rcc // (internal impedance between VCC and VEE)

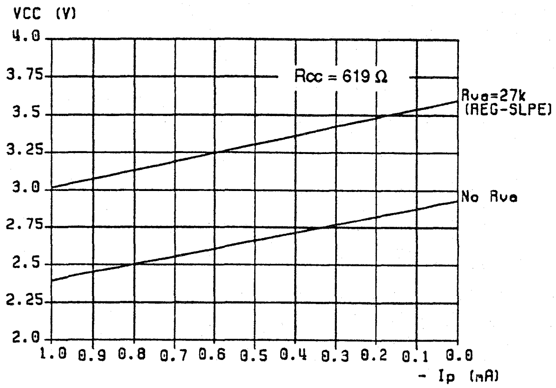


Fig.11 VCC supply voltage versus Ip consumed current for Irec = 0

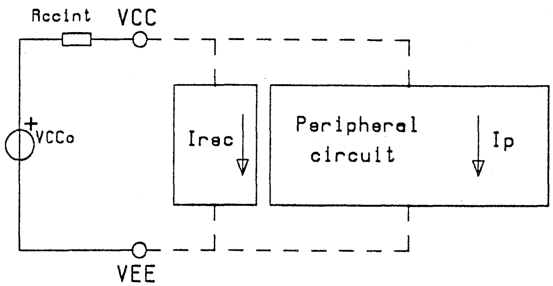


Fig.12 VCC supply point: equivalent schematic

As VCC is limited to a minimum value to ensure correct functioning, Ip will be limited to a maximum value. The limit is imposed by the requirement to maintain a minimum permitted voltage between VCC and SLPE which is called Vmin. So the maximum current available depends on the DC settings of the IC: VREF, Rcc, Rslpe and the required AC signal level at the line and receiver outputs. To simplify the calculation, we will use the worst case for Rccint which is Rcc. It gives:

$$VCC = VLN - Rcc(Icc + Irec)$$

$$VCC = VREF + Rslpe(Iline - Icc - irec) - Rcc(Icc + Irec)$$

$$VCCmin = Vmin + Rslpe(Iline - Icc - Irec - Ip)$$

$$Ipmax = \frac{VCC - VCCmin}{Rcc}$$

$$Ipmax = \frac{VREF - Vmin}{Rcc - Rslpe} - \frac{Rcc}{Rcc - Rslpe}(Icc + Irec)$$

$$Vmin = 1.7V + v\ln\left(\frac{Rslpe}{Zline || Rcc}\right)$$

Adjustment

As the impedance connected between LN and VCC also determines the set impedance, the easiest way to increase the current capability of the supply point VCC is to increase the reference voltage VREF by connecting an Rva resistor between REG and SLPE (see Fig.11). The maximum preferable value of VREF = 7 V; see *Adjustment* in section 3.1.1.

3.2 Set Impedance

Principle of operation

The ICs behave like an equivalent inductance that present a low impedance to DC (Rslpe) and a high impedance (Rp) to speech signals. Rp is an integrated resistance in the order of 15.5 kΩ +/- 15%. It is in parallel with the external RC filter realized by Rcc and Cvcc. Thus in the audio frequency range the set impedance is mainly determined by the Rcc resistor. Fig.13 shows an equivalent schematic for the set impedance, while Fig.14 shows measurement results of the set impedance and the Balance Return Loss (BRL). BRL measures the matching of the set impedance to a reference impedance of 600Ω (in this case) according to the formula:

$$BRL = 20\left(\log\frac{|Zset + 600\Omega|}{|Zset - 600\Omega|}\right)$$

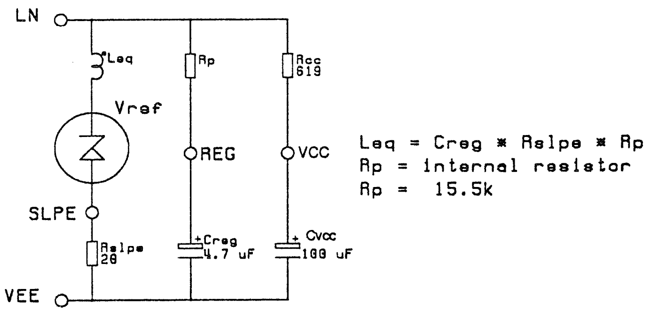


Fig.13 Equivalent set impedance

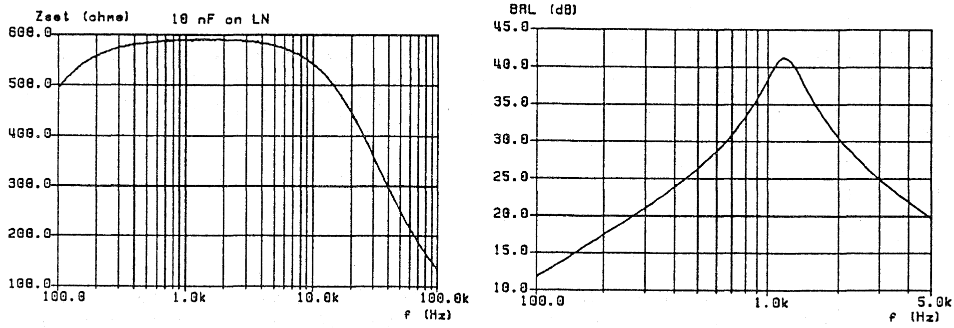


Fig.14 Set impedance and Balance Return Loss at 600 ohms reference impedance

Adjustment

When decreasing the reference voltage VREF, a resistor is connected between LN and REG in parallel of Rp (See Fig.13) so, slightly modifying the set impedance.

If complex set impedance is required, the Rcc resistor must be replaced by an equivalent complex network. Keep in mind that the DC resistance of this network influences the VCC voltage and current supply capability. (See section 3.1.2 'Supply for peripheral circuits')

3.3 Supply for a LED; pin ILED

Principle of operation

The TEA1112/A give an on-hook / off-hook status indication. This is done by a current available to drive a LED connected between pins ILED and LN. In the low voltage area, which corresponds to low line current condition, no current is available for this LED. For line currents higher than a threshold, Iled starts at 18mA typically, the Iled current increases proportionally to the line current (with a ratio of approximately one third). The Iled current is internally limited to 19.5 mA (typical value). See curves shown in Fig.15.

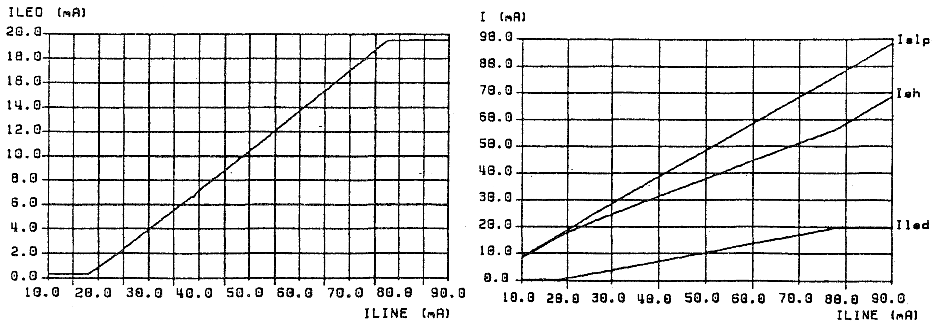


Fig.15 LED supply current versus the available line current

As the LED driver is connected to SLPE, all the Iled supply current will flow through the Rslpe resistor. Consequently, the AGC characteristics are not disturbed.

Adjustment

The ICs have been designed for use with all kind of LED's as long as the voltage across this device, at 20 mA current flowing through it, is lower than VREF - 0.8 V. The start and stop line currents as well as the maximum Iled current are internally fixed.

If no LED is required, the ILED output can be shorted to SLPE to avoid a floating pin.

3.4 Microphone amplifier; pins MIC+, MIC-, GAS

Principle of operation

In Fig.16 the block diagram of the microphone amplifier of the TEA1112/A is depicted.The microphone amplifier has symmetrical very high input impedances. The input impedance between pins MIC+ and MIC- is typically 64 kΩ (2 x 32 kΩ) with maximum tolerances of +/- 15%. Thanks to this high input impedance, the ICs are suitable for several kinds of microphones: dynamic, piezoelectric or electret microphones with symmetrical or a-symmetrical drives (See Fig.17 for some examples).

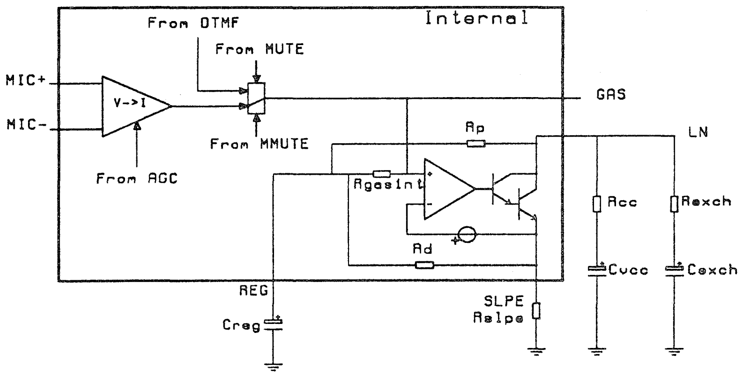


Fig.16 Microphone channel

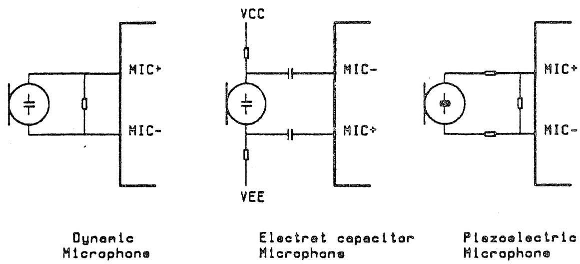


Fig.17 Microphone arrangements examples

As can be seen in Fig.16, the microphone amplifier itself is built up out of two parts: a pre-amplifier which realizes a voltage to current conversion, and an end-amplifier which realizes the current to voltage conversion. The overall gain (Gvtx) of the microphone amplifier from inputs MIC+/MIC- to output LN is given by the following equation

$$\mathrm{Gvtx} = 20 \times \log \mathrm{Avtx}$$

$$\mathrm{Avtx} = 1.31 \times \frac{\mathrm{Rgasint}}{\mathrm{Rrefint}} \times \frac{\mathrm{Ri}||\mathrm{Zline}}{\mathrm{Rslpe}} \times \alpha$$

with:

Ri = the dynamic set impedance, Rcc||Rp, typically 619 Ω || 15.5 kΩ

Rgasint = internal resistor realizing the current to voltage conversion, typically 69 kΩ with a spread of +/- 15%

Rrefint = internal resistor determining the current of an internal current stabilizer, typically 3.4 kΩ with a spread of +/- 15%(correlated to the spread of Rgasint)

Zline = load impedance of the line during the measurement

α = gain control factor varying from 1 at Iline = 15 mA to 0.5 at Iline = 75 mA when AGC function is applied; see chapter 3.8.

Using these typical values in the equation, we find a gain equal to:

Gvtx = 20 x log Avtx = 52 dB

at Iline = 15 mA

The different gain controls (AGC, MUTE, MMUTE) act on the microphone pre-amplifier stage, modifying its transconductance.

Adjustment and performance

The microphone gain can be decreased by connecting a resistor Rgas between pins GAS and REG. It can be adjusted from 52 dB down to 39 dB to suit application specific requirements. The gain dependency to this external resistor is calculated in equation (4) and shown in Fig.18 at 1 kHz and for a typical sample. The gain adjustment by an external Rgas resistor connected between pins GAS and REG may slightly change the gain spread.

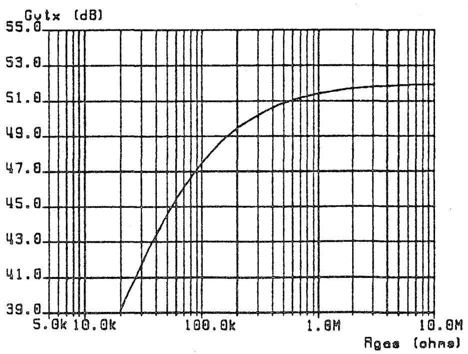


Fig.18 Microphone gain function of the Rgas resistor connected between GAS and REG

$$Gvtx = 20 \times \log\left(1.31 \times \frac{Rgasint||Rgas}{Rrefint} \times \frac{Ri||Zline}{Rslpe} \times \alpha\right) \quad (4)$$

A capacitor Cgas is generally connected between pins GAS and REG to provide a first order low pass filter, which cut-off frequency is determined by the product Cgas x (Rgasint || Rgas). Fig.19 shows the frequency response of the microphone amplifier at different temperatures (Cgas = 100 pF, Rgasint = 69 kΩ, no external Rgas).

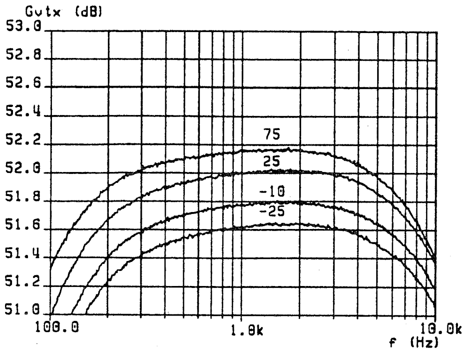


Fig.19 Microphone gain versus frequency: influence of temperature

Fig.20 shows the distortion of the signal on the line as a function of the microphone input signal for two different gains, at nominal DC settings. The inputs of the microphone amplifier can handle signals up to 18 mVrms with less than 2% Total Harmonic Distortion (THD). For overall gains (Gvtx) larger than 40 dB, the distortion will be determined by the output stage (clipping of the line signal). So Fig.20 a) shows a saturation due to the output stage, while Fig.20 b) shows a saturation due to the input stage.

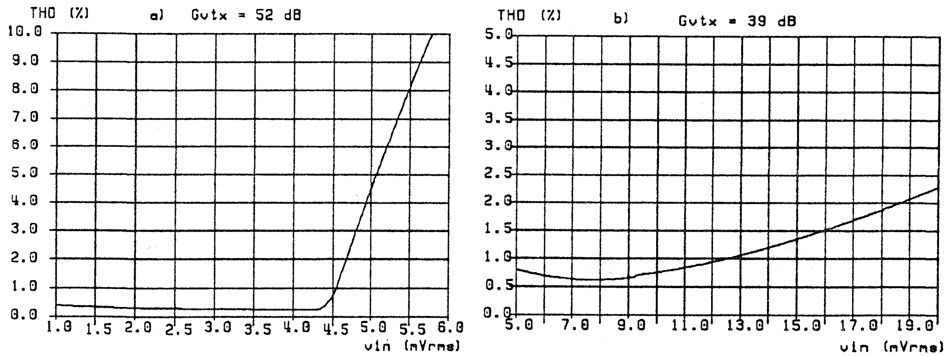


Fig.20 Distortion on the line as a function of the input signal for two microphone gains

Fig.21 shows the distortion of the line signal versus the rms voltage on the line at line currents equal to 4 mA and 15 mA at the nominal gain of 52 dB.

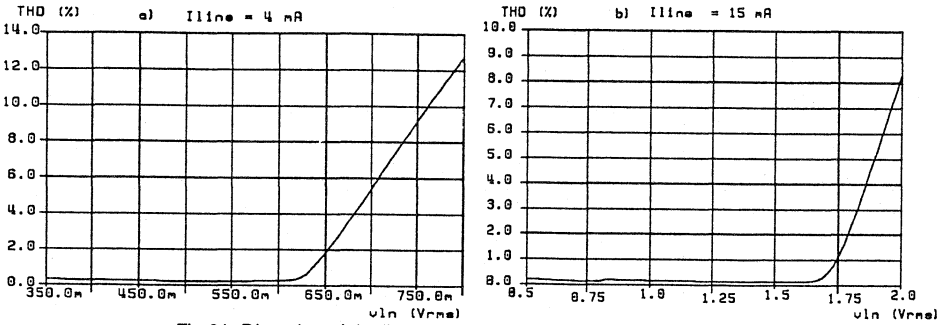


Fig.21 Distortion of the line signal versus the rms voltage on the line

To obtain optimum noise performance on the line, the microphone inputs must be loaded. Fig.22 shows the noise on the line (psophometrically weighted; P53 curve) as a function of the line current and the microphone gain with a 200Ω connected between the microphone inputs (typical application). These curves show the sensitivity of the noise to the microphone gain. The noise measures -79.5 dBmp at minimum send gain.

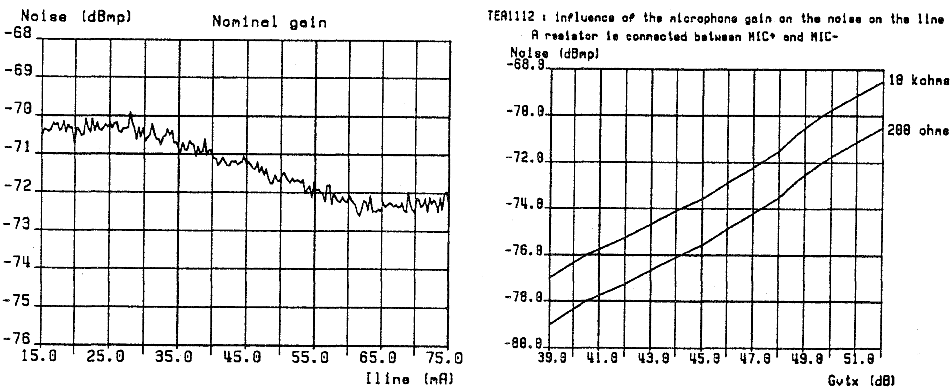


Fig.22 Noise on the line versus the line current and the microphone gain

The amplifier gain is temperature compensated. The gain adjustment by an external Rgas resistor connected between pins GAS and REG may slightly change the temperature coefficient; see reference [2].

Fig.23 shows the common mode rejection ratio at 15 mA and at nominal microphone gain. Two curves are present on this figure. The first one is the spectrum of the signal on pin LN when a sending signal is applied on pin MIC-, pin MIC+ being shorted to VEE by a decoupling capacitor. The second curve is the spectrum of the signal on pin LN when an sending signal is applied on the microphone inputs, MIC+ and MIC- being shorted. Both signals are at a frequency of 1 kHz. The difference between the two curves at this frequency gives the CMRR.

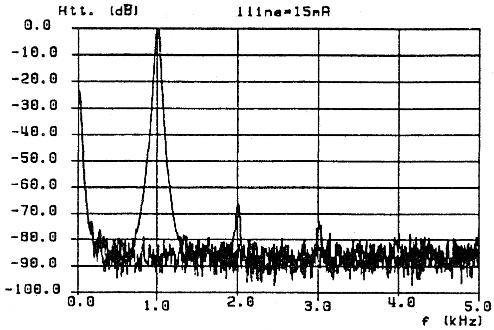


Fig.23 Common mode rejection ratio

3.5 MMUTE function (TEA1112 only); pin MMUTE

Principle of operation

The microphone mute function realizes an electronic switching between the microphone amplifier and the sending DTMF amplifier. This function disables the microphone channel to provide such kind of privacy and in the same time enables the DTMF channel if needed for some specific applications. If a high level is applied to the MMUTE input, the sending DTMF channel is activated, while the microphone amplifier is disabled. The microphone amplifier can be enabled (depending on the MUTE level; see TABLE 1) by either applying a low level (< 0.3 V typically) at the MMUTE input or leaving it open. Fig.24 shows the microphone amplifier gain reduction and the input current as a function of the input voltage on MMUTE. The threshold voltage level is 0.68 V typically (base-emitter junction) with a temperature coefficient of -2 mV/°C.

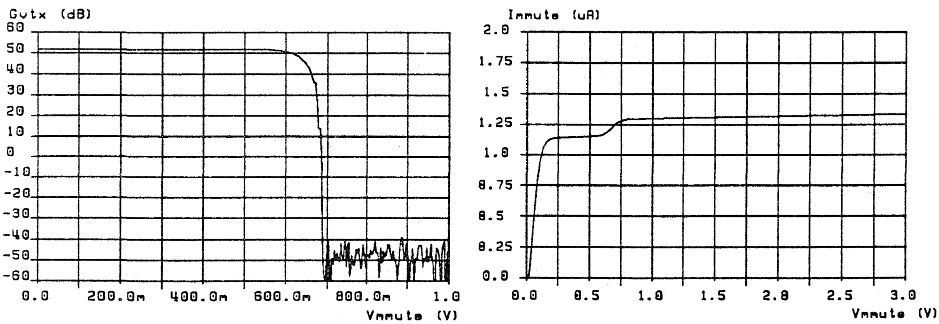


Fig.24 Microphone gain and MMUTE input current vs Vmmute

The microphone mute function has no effect on the receiving channel which is fully determined by the MUTE level.

Performance

Fig.25 shows the microphone amplifier gain reduction at Iline = 15 mA for an input signal at 1 kHz. Two curves are drawn in this figure. The first one shows the spectrum of the signal on the line in speech condition when a signal is applied on the microphone inputs. The second curve shows the same signal in DTMF condition. Both signals are at a frequency of 1 kHz. The difference between the two curves at this frequency gives the gain reduction.

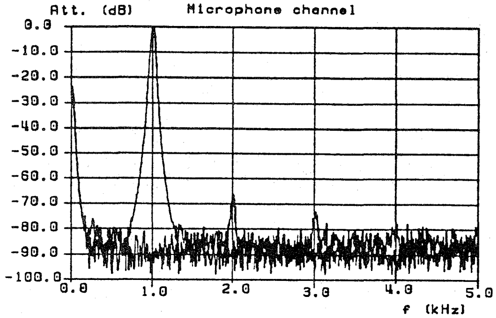


Fig.25 Microphone gain reduction in MMUTE condition

The MMUTE function works down to a voltage on VCC equal to 1.6V (Iline = 2.5 mA in the basic application). Below this threshold, the microphone amplifier stays always enabled independently of the MMUTE input level.

The maximum voltage allowed at the MMUTE input is VCC + 0.4 V.

3.6 MMUTE function (TEA1112A only); pin MMUTE

Principle of operation

The MMUTE function realizes an electronic switching between the microphone amplifier and the sending DTMF amplifier. This function disables the microphone channel to provide such kind of privacy and in the same time enables the DTMF channel if needed for some specific applications. If a high level is applied to the MMUTE input, the microphone amplifier can be activated (depending on the MUTE level, See TABLE 1), while the DTMF channel is disabled. The DTMF channel is enabled by either applying a low level (< 0.3 V typically) at the MMUTE input or leaving it open. Fig.26 shows the microphone amplifier gain reduction and the input current as a function of the input voltage on MMUTE. The threshold voltage level is 0.68 V typically (base-emitter junction) with a temperature coefficient of -2 mV/°C.

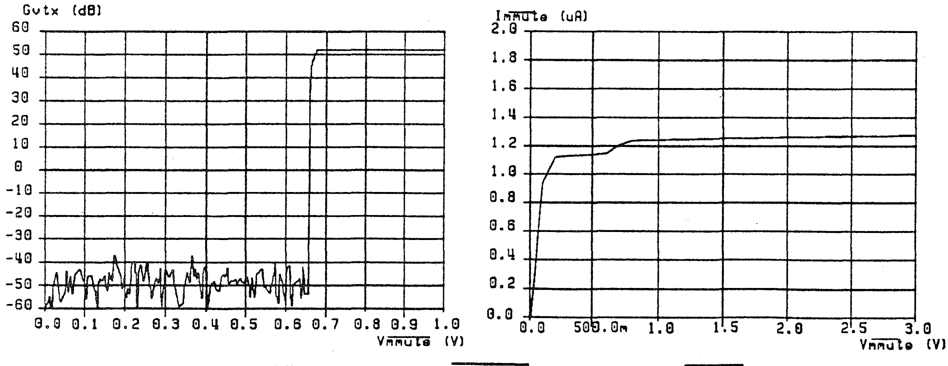


Fig.26 Microphone gain and MMUTE input current vs Vmmute

The $\overline{\text{MMUTE}}$ function has no effect on the receiving channel which is fully determined by the $\overline{\text{MUTE}}$ level.

Performance

Fig.27 shows the microphone amplifier gain reduction at Iline = 15 mA for an input signal at 1 kHz.Two curves are drawn in this figure. The first one shows the spectrum of the signal on the line in speech condition when a signal is applied on the microphone inputs. The second curve shows the same signal in DTMF condition. Both signals are at a frequency of 1 kHz. The difference between the two curves at this frequency gives the gain reduction.

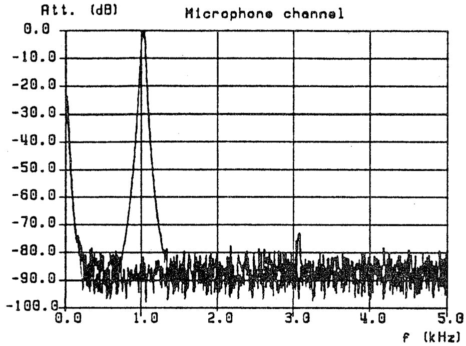


Fig.27 Microphone gain reduction in $\overline{\text{MMUTE}}$ condition

The $\overline{\text{MMUTE}}$ function works down to a voltage on VCC equal to 1.6V (Iline = 2.5 mA in the basic application). Below this threshold, the microphone amplifier stays always enabled independently to the $\overline{\text{MMUTE}}$ input level.

The maximum voltage allowed at the $\overline{\text{MMUTE}}$ input is VCC + 0.4 V

3.7 Receiving amplifier; pins IR, GAR, QR

Principle of operation

In Fig.28, the block diagram of the receiving amplifier of the TEA1112/A is depicted.

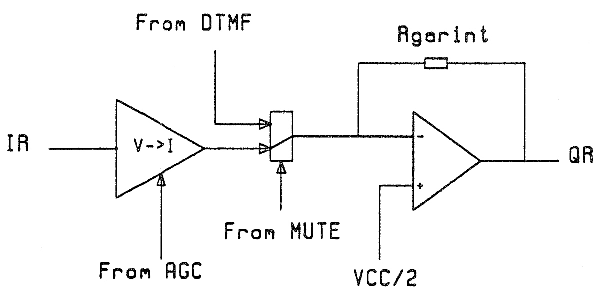
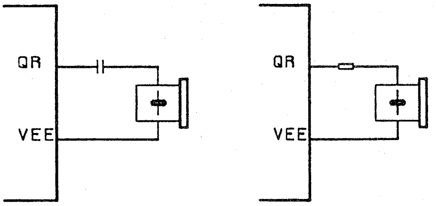


Fig.28 Receiving channel

The receiving amplifier has an a-symmetrical high input impedance between pins IR and VEE. It is equal to 20 kΩ with maximum tolerances of +/- 15%. The ICs are suitable for several kind of earpieces and can drive either dynamic, magnetic or piezo-electric earpieces (See Fig.29 for some arrangements examples).



Dynamic Earpiece   Piezoelectric Earpiece

Fig.29 Earpieces arrangements examples

As can be seen in Fig.28, the receiving amplifier itself is built up out of two parts: a pre-amplifier which realizes a voltage to current conversion, and an end-amplifier which realizes the current to voltage conversion. The overall gain Gvrx of the receiving amplifier from input IR to output QR is given by the equation:

$$\text{Gvrx} = 20 \times \log \text{Avrx}$$

$$\text{Avrx} = \alpha \times 1.21 \times \frac{\text{Rgarint}}{\text{Rrefint}}$$

with:

Rgarint = internal resistor realizing the current to voltage conversion, typically 100 kΩ with a spread of +/- 15%

Rrefint = internal resistor determining the current of an internal current stabilizer, typically 3.4 kΩ with a spread of +/- 15% (correlated to the spread of Rgarint)

α = gain control factor varying from 1 at Iline = 15 mA to 0.5 at Iline = 75 mA, when AGC function is applied.

Using these typical values in the equation, we find a gain equal to:

Gvrx = 20 x log Avrx = 31 dB at Iline = 15 mA.

The gain controls, AGC and MUTE, act on the receiving pre-amplifier stage, modifying its transconductance.

Adjustment and performance

The receiving gain can be decreased by connecting a resistor Rgar between pins GAR and QR. It can be adjusted from 31 dB down to 19 dB to suit application specific requirements. The gain dependency to this external resistor is calculated in equation (5) and shown in Fig.30. The gain adjustment by an external Rgar resistor connected between pins GAS and REG may slightly change the gain spread.

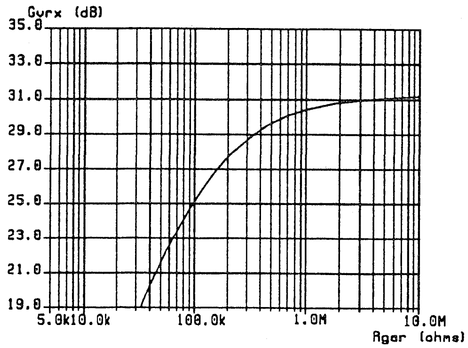


Fig.30 Receiving gain function of the Rgar resistor connected between GAR and QR

$$Gvrx = 20 \times \log\left(1.21 \times \frac{Rgarint||Rgar}{Rrefint}\right) \tag{5}$$

Two external capacitors Cgar (connected between GAR and QR) and Cgars (connected between GAR and VEE) ensure stability. The relationship Cgars >= 20 x Cgar should be fulfilled to ensure stability. The Cgar capacitor provides a first order filter, which cut-off frequency is determined by the relation Cgar x (Rgarint || Rgar). Fig.31 shows the frequency response of the receiving amplifier at different temperature (Cgar = 100pF, Cgars = 2.2 nF).

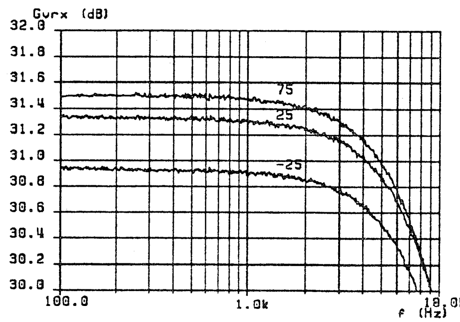


Fig.31 Receiving gain versus frequency: influence of temperature

The maximum output swing on QR depends on the DC line voltage, the Rcc resistor, the Icc current consumption of the circuit, the Ip current consumption of the peripheral circuits and the load impedance on QR. The receiving input IR can handle signals up to 18 mVrms with less than 2% THD. Fig.32 shows the distortion on QR as a function of the input voltage for a line current equal to 75 mA. The two curves correspond to a measurement with and without the AGC function which results in a difference of 6 dB in the receiving gain. With AGC, the gain is only 25 dB and the distortion is due to the input, while without AGC, the gain is 31 dB and the distortion comes from the output.

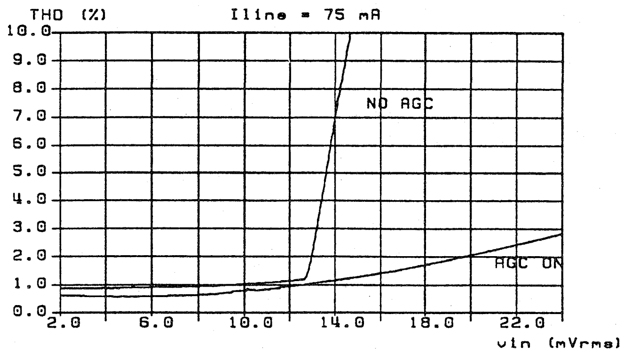


Fig.32 Distortion on QR versus the input signal on IR

The maximum level on QR for 2% THD increases with line current due to the increase of VCC and then is limited to a maximum value due to the input limitation.

Fig.33 shows the distortion of the signal on QR as a function of the rms voltage on QR at Iline = 15 mA for two different loads: 150 Ω and 450 Ω .

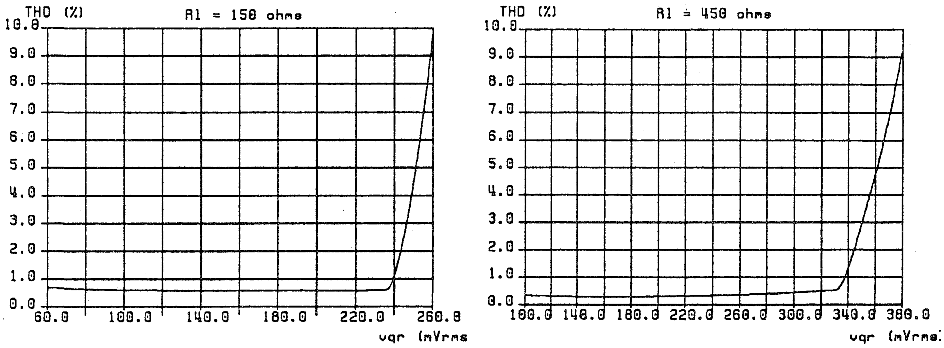


Fig.33 Distortion of the receiving signal for two loads

Fig.34 shows the noise on QR loaded with 150 Ω (psophometrically weighted; P53 curve) as a function of the line current. This curve has been done with an open input IR. With the anti-sidetone connected to the input, the noise generated on the line will add via the anti-sidetone circuitry to the equivalent noise at the input IR. The total noise generated at the earpiece output depends on the microphone amplifier gain that has been set, the sidetone suppression and the receiving amplifier gain. The influence of the AGC on the noise appears clearly in Fig.34.

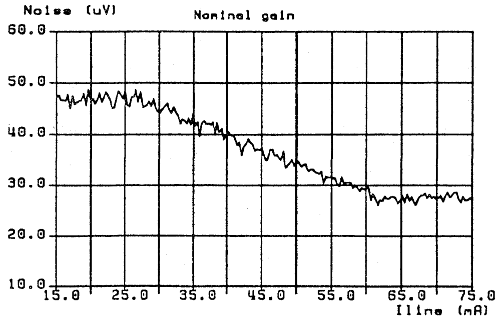


Fig.34 Noise on the earpiece

The amplifier gain is temperature compensated. The gain adjustment by an external Rgar resistor connected between pins GAR and QR may slightly change the temperature coefficient.

3.8 Automatic Gain Control; pin AGC

Principle of operation

The TEA1112/A perform automatic line loss compensation. The automatic gain control varies the gain of the microphone and receiving amplifiers in accordance with the DC line current. To enable the AGC function, the pin AGC must be connected to the pin VEE. For line currents below a current threshold, Istart (typical 26 mA), the gain control factor α is equal to 1, giving the maximum value for the gains Gvtx and Gvrx. If this threshold current is exceeded, the gain control factor α and the gain of both controlled amplifiers are decreased. When the line current reaches a second threshold current, Istop (typical 61 mA), the gain control factor α is limited to its minimum value equal to 0.5, giving the minimum value for the gains Gvtx and Gvrx. The gain control range of both amplifiers is typically 5.85 dB. This corresponds to a line length of 5 km for a 0.5 mm diameter twisted pair copper cable with a DC resistance of 176Ω/km and an average attenuation of 1.2 dB/km.

Adjustment and performance

The ICs have been optimized for use with an exchange supply voltage of 48V, a feeding bridge resistance of 2 times 300 Ω and the previously described line. To fit with other configurations, a resistor, Ragc, can be connected between pins AGC and VEE. This allows to increase the threshold currents Istart and Istop. Fig.35 shows the control of the microphone gain versus the line current for different values of Ragc.

If no AGC function is required, the AGC pin must be open circuit. So no gain control is applied, the gain control factor α stays at 1 and both controlled gains have their maximum value.

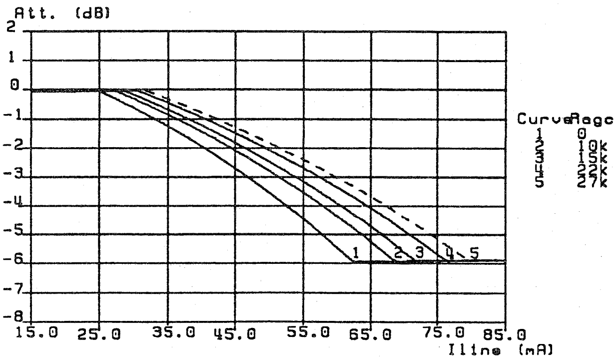


Fig.35 Automatic gain control on the microphone amplifier

3.9 DTMF amplifier; pin DTMF

Principle of operation

In Fig.36 the block diagram of the DTMF channel of the TEA1112/A is depicted.

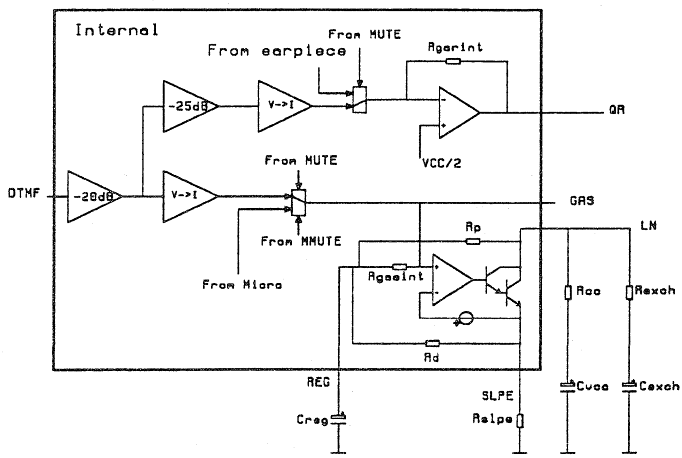


Fig.36 DTMF channel

The DTMF amplifier has an a-symmetrical high input impedance. The impedance between DTMF and VEE is typically 20 kΩ with maximum tolerances of +/- 15%. The DTMF amplifier is built up out of three parts: an attenuator by a factor 10, a pre-amplifier which realizes the voltage to current conversion and the same end-amplifier as the microphone amplifier. No AGC is applied on the DTMF channel.

Fig.37 shows the frequency response of the DTMF amplifier at 15 mA at different temperatures (Cgas = 100 pF).

Adjustment and performance

When a resistor Rgas is connected between GAS and REG to decrease the microphone gain, the DTMF gain varies in the same way: the DTMF gain is 26.5 dB lower than the microphone gain without control of AGC.

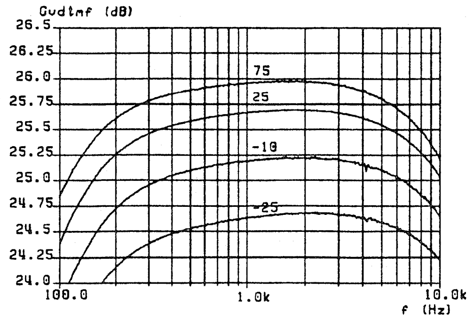


Fig.37 DMTF gain versus frequency: influence of temperature

The input of the DTMF amplifier can handle signals up to180 mVrms with less than 2% THD. Fig.38 shows the distortion of the line signal versus the rms input voltage for two different gains.

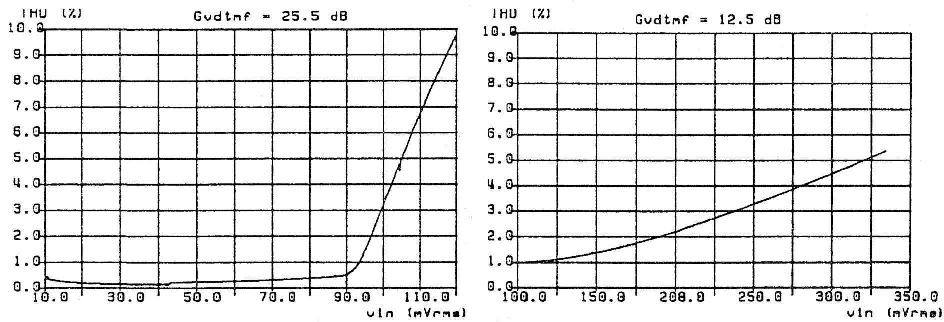


Fig.38 Distortion on the line function of the DTMF input signal for two different gains

3.10 MUTE function (TEA1112 only); pin MUTE

Principle of operation

The mute function realizes an electronic switching between the speech mode and the dialling mode. If a high level is applied to the MUTE input, the DTMF input is enabled and both microphone and receiving amplifiers are disabled. In this mode a confidence tone is provided in the earpiece. The microphone and receiving amplifiers are enabled by either applying a low level (< 0.3 V typically) at the MUTE input or leaving it open; (keep in mind that the microphone channel depends on the MMUTE level; See TABLE 1). In this case, the DTMF input is disabled.

Fig.39 shows the microphone amplifier gain reduction and the input current as a function of the voltage on MUTE. The threshold voltage is 0.68 V typically (base-emitter junction) with a temperature coefficient of -2 mV/ °C.

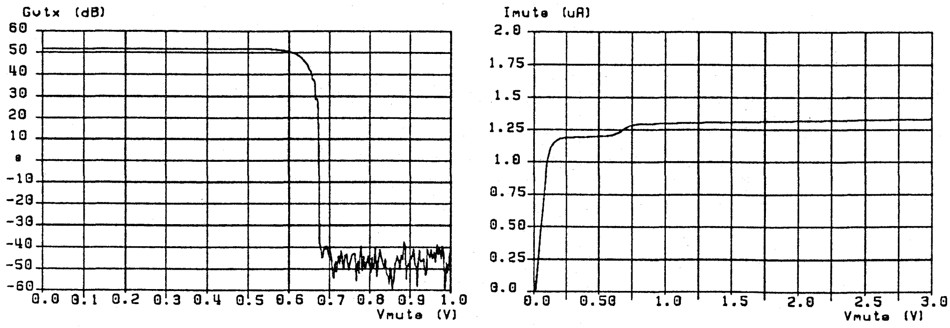


Fig.39 Microphone gain attenuation and MUTE input current vs Vmute

Adjustment and performance

Fig.40 shows the microphone and receiving gains reduction at Iline = 15 mA for an input signal at 1 kHz.Two curves are drawn on each graphic. The first one shows the spectrum of the signal on the line (QR) in speech condition when a signal is applied on the microphone inputs (IR input). The second curve shows the same signal in DTMF condition. Both signals are at a frequency of 1 kHz. The difference between the two curves at this frequency gives the gain reduction.

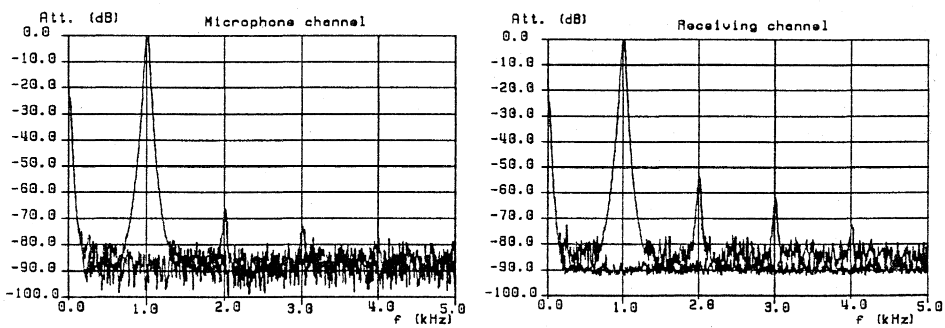


Fig.40 Microphone gain and earpiece gain reduction in MUTE condition

The MUTE function works down to a voltage on VCC equal to 1.6V (Iline = 2.5 mA in the basic application). Below this threshold, the microphone amplifier stays always enabled independently of the MUTE input level.

The maximum voltage allowed at the MUTE input is VCC + 0.4V.

3.11 $\overline{\text{MUTE}}$ function (TEA1112A only); pin $\overline{\text{MUTE}}$

Principle of operation

The $\overline{\text{MUTE}}$ function realizes an electronic switching between the speech mode and the dialling mode. If a high level is applied to the $\overline{\text{MUTE}}$ input, the microphone and receiving amplifiers are enabled and the DTMF input is disabled; (keep in mind that the microphone channel depends on the $\overline{\text{MMUTE}}$ level; See TABLE 1). The DTMF input is enabled by either applying a low level (< 0.3 V typically) at the $\overline{\text{MUTE}}$ input or leaving it open. In this mode a confidence tone is provided in the earpiece and the microphone and receiving amplifiers are disabled. Fig.41 shows the microphone amplifier gain reduction and the input current as a function of the voltage on $\overline{\text{MUTE}}$. The threshold voltage is 0.68 V typically (base-emitter junction) with a temperature coefficient of -2 mV/ °C.

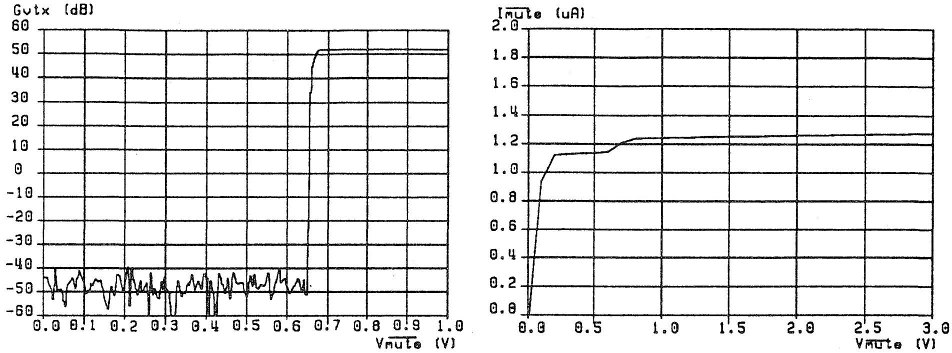


Fig.41 Microphone gain attenuation and $\overline{\text{MUTE}}$ input current vs Vmute

Adjustment and performance

Fig.42 shows the microphone and receiving gains reduction at Iline = 15 mA for an input signal at 1 kHz. Two curves are drawn on each graphic. The first one shows the spectrum of the signal on the line (QR) in speech condition when a signal is applied on the microphone inputs (IR input). The second curve shows the same signal in DTMF condition. Both signals are at a frequency of 1 kHz. The difference between the two curves at this frequency gives the gain reduction.

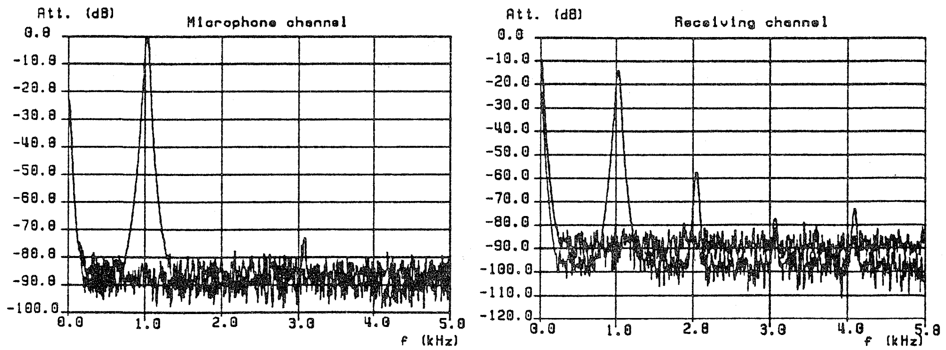


Fig.42 Microphone gain and earpiece gain reduction in $\overline{\text{MUTE}}$ condition

The $\overline{\text{MUTE}}$ function works down to a voltage on VCC equal to 1.6V (Iline=2.5 mA in the basic application). Below this threshold, the microphone amplifier stays always enabled independently of the $\overline{\text{MUTE}}$ input level.

The maximum voltage allowed at the $\overline{\text{MUTE}}$ input is VCC + 0.4V.

3.12 Anti-sidetone circuitry

Principle of operation

To avoid the reproduction of microphone signals in the earpiece, the anti-sidetone circuit uses the microphone signal from pin SLPE to cancel the microphone signal at the input IR of the receiving amplifier. The anti-sidetone bridge already used for the TEA106x family or a conventional Wheatstone bridge as shown in Fig.43 may be used as the basis for the design of the anti-sidetone circuit.

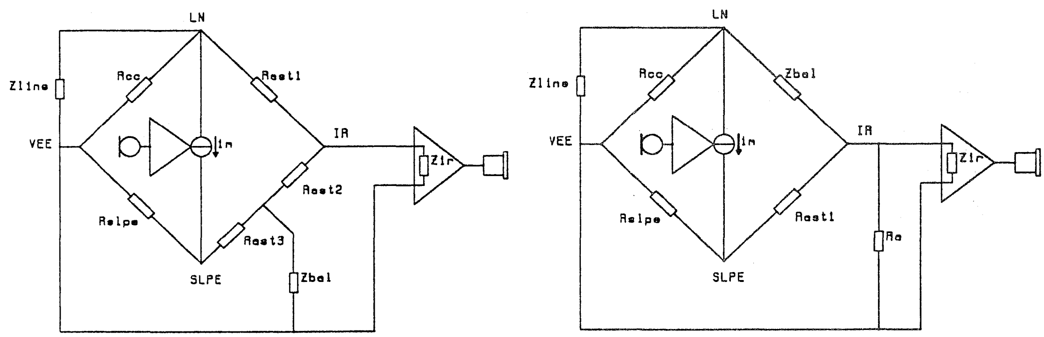


Fig.43 TEA106X family anti-sidetone bridge (left) and Wheatstone bridge (right)

The TEA106x family anti-sidetone bridge has the advantage of a relatively flat transfer function in the audio frequency range between pins LN and IR, both with real and complex set impedances. Furthermore, the attenuation of the bridge for the received signal (between pins LN and IR) is independent of the value chosen for Zbal after the set impedance has been fixed and the condition shown in equation (6) is fulfilled. Therefore, re*adjustment* of the overall receive gain is not necessary in many cases.

The Wheatstone bridge has the advantages of needing one resistor fewer than the TEA106x family bridge and a smaller capacitor for Zbal. But the disadvantages include the dependence of the attenuation of the bridge on the value chosen for Zbal and the frequency dependence of that attenuation. This necessitates some readjustment of the overall receive gain.

3.12.1 TEA106x family bridge

The anti-sidetone circuit is composed of: Rcc//Zline, Rast1, Rast2, Rast3, Rslpe and Zbal. Maximum compensation is obtained when the following conditions are fulfilled:

$$\text{Rslpe} \times \text{Rast1} = \text{Rcc} \times (\text{Rast2} + \text{Rast3}) \tag{6}$$

$$k = (\text{Rast2} \times (\text{Rast3} + \text{Rslpe})) / (\text{Rast1} \times \text{Rslpe})$$

$$\text{Zbal} = k \times \text{Zline}$$

The scale factor k is chosen to meet the compatibility with a standard capacitor from the E6 or E12 range for Zbal.

In practice, Zline varies strongly with the line length and line type. Consequently the value for Zbal has to be chosen to fit with an average line length giving satisfactory sidetone suppression with short and long lines. The suppression further depends on the accuracy with which Zbal equals this average line impedance.

Example

Let's optimize for a line length of 5 km 0.5 mm diameter copper twisted pair with an average attenuation of 1.2 dB / km, a DC resistance of 176Ω / km and a capacitance of 38 nF / km. The approximate equivalent line impedance is shown in Fig.44

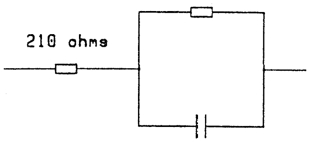


Fig.44 Equivalent average line impedance

For compatibility of the capacitor value in Zbal with a standard capacitor from the E6 series (220nF):

$$k = \frac{140\text{nF}}{220\text{nF}} = 0.636$$

For Rast3, a value of 3.92 kΩ has been chosen. So using the previous equations, we can calculate Zbal, Rast1, Rast2. We find Rast1= 130 kΩ, Rast2 = 390 Ω, and for Zbal 130Ω in series with (220nF//820Ω).

The attenuation of the received line signal between LN and IR can be derived from equation (7)

$$\frac{Vir}{Vln} = \frac{Zir||Rast2}{Rast1 + (Zir||Rast2)} \qquad (7)$$

if Rast2 >> (Rast3//Zbal).

With the values used in the example, it gives 32 dB at 1 kHz.

Zir is the receiving amplifier input impedance, typically 20 kΩ.

3.12.2 Wheatstone bridge

The conditions for optimum suppression are given by:

$$Zbal = \frac{Rast1}{Rslpe} \times \frac{Rcc \times Zline}{Rcc + Zline}$$

Also for this bridge type, a value for Zbal has to be chosen that corresponds with an average line length.

The attenuation of the received line signal between LN and IR is given by:

$$\frac{Vir}{Vln} = \frac{Rast1||Zir||Ra}{Zbal + (Rast1||Zir||Ra)}$$

Ra is used to adjust the bridge attenuation; its value has no influence on the balance of the bridge.

4. APPLICATION EXAMPLE 1 - LOW VOLTAGE BASIC SET -

Two application examples are described in this report; a 'low voltage basic set' in this chapter and a 'handsfree application with on-hook dialling' in chapter 5. Both examples are general purpose applications for exchanges with voltage regulation. Fine tuning is required to fulfil specific country requirements. Both applications have been build and tested on their functionality.

4.1 Description of the application

An application example for a low voltage basic telephone set is shown in Fig.45 and Fig.46. It is build up with the TEA1112 transmission IC and a discrete ringer circuit as shown in Fig.45, and the PCD3332-3 pulse/tone repertory dialler/ringer IC PCD3332-3 according Fig.46. The interconnections between both figures are indicated. The application offers the following features:

- Transmission functions with adjustable parameters as described for the TEA1112 in chapter 3.
- Microphone mute function
- Pulse, DTMF and mixed mode dialling, redial, 13-number repertory dialling as specified in [1]
- Ringer signal detection and melody generation

The application is build up around the TEA1112. The individual settings of the TEA1112 are for 600 Ω set impedance and 2.5 V minimum supply voltage for the PCD3332-3 at dialling. The several blocks of the application are briefly described in this chapter; details concerning the performances are given in chapter 4.2.

The **TEA1112** in this application cannot be replaced by the **TEA1112A** version because of the inverted MUTE and MMUTE ($\overline{\text{MUTE}}$ and $\overline{\text{MMUTE}}$) of the TEA1112A.

Polarity guard and protection

One diode bridge is applied for the transmission circuit part as well as for the ringer stage to ensure proper functioning independent of the polarity of the line voltage respectively to rectify the ringer signal. Protection is achieved by a break-over diode D18 between the A-B/B-A terminals, the current limiting components R20 and TR3, the 11 V zenerdiode Z5 between LN and VEE of the transmission IC and by a 5.6 V zener diode Z13 between VDD and VSS of the PCD3332-3.

The current limiter R20 and TR3 provides protection against current surges exceeding 150 mA. It is not designed for continuous limitation of the line current.

The voltage across the ringer output stage is limited at 24 V by means of the zener diodes Z11 and Z13 and diode D12.

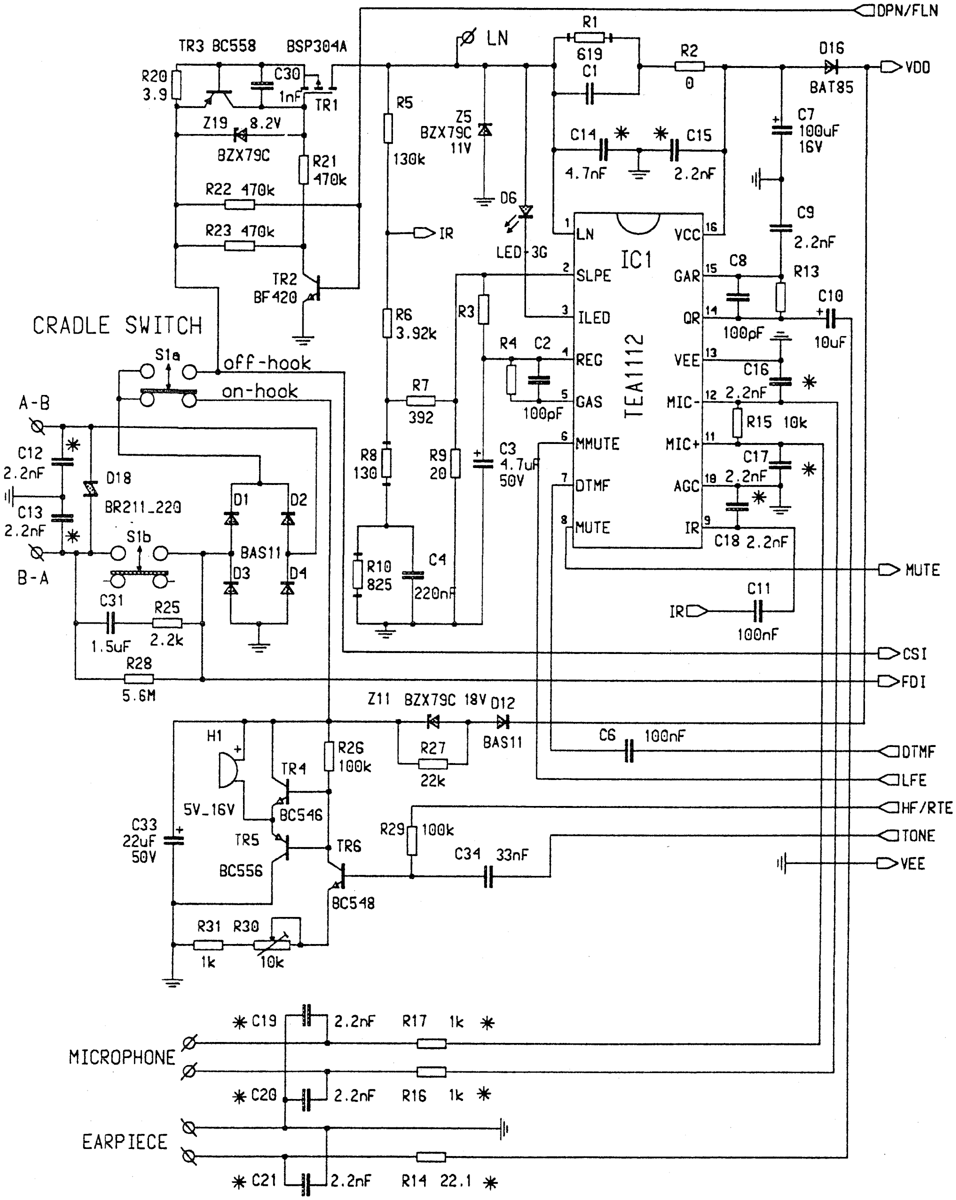
Interrupter

The interrupter consists of TR1, P-channel enhancement D-MOS BSP304A, and inverter TR2 controlled by the DPN/FLN open drain output of the PCD3332-3. When the handset is lifted, cradle switch S1 changes from ringer state (on-hook) to transmission state (off-hook); DPN/FLN is high resulting in a conducting TR1.

Interruption of the line current is achieved by a low DPN/FLN level.

Speech / transmission

The TEA1112 stabilizes the DC voltage between LN and SLPE. It delivers the supply voltage VCC for internal use and for the PCD3332-3 via diode D16. VCC is buffered by C7 while VDD is buffered by C35. The set impedance will be mainly determined by the impedance of the network between LN and VCC. This application has a set impedance of about 600 Ω realised by R1 = 619 Ω and R2 = 0 Ω.



*: EMC components

Fig.45 Application example 1; line interface TEA1112, discrete ringer

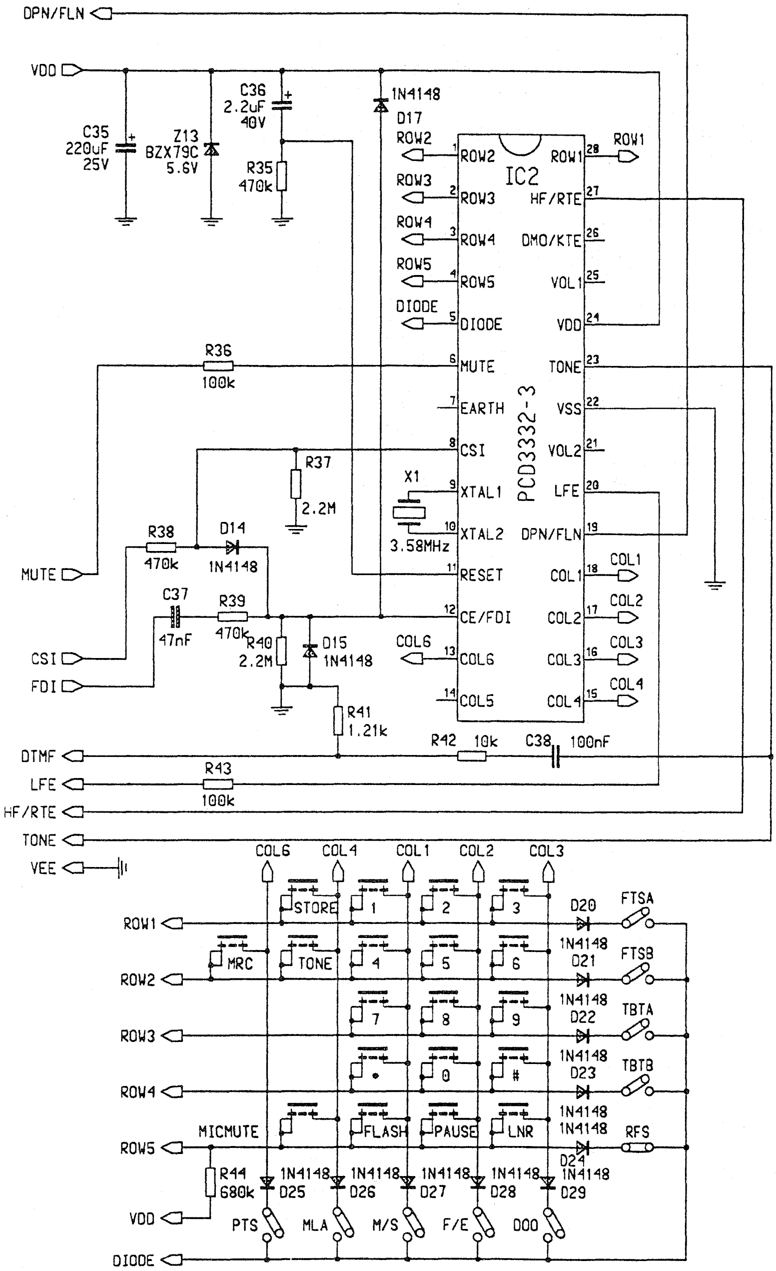


Fig.46 Application example 1; dialler/ringer PCD3332-3

Complex set impedance can be realised by means of the network R1, R2 and C1.

The application is intended for use with a dynamic microphone and dynamic earpiece. Use of an electret microphone requires a modification of the application. A supply has to be made from VCC while the gain has to be adapted by resistor R4 which has no value in Fig.45; see application example 2, chapter 5.

Buffer capacitor C7 is discharged heavily during the break periods at pulse dialing with the result that the VDD buffer capacitor C35 will not be charged during the whole dialling digit.

The supply voltage VCC has to be increased for pulse dialling applications, see chapter 4.2.

The microphone amplifier can be disabled by a high level at the MMUTE input pin. In this example is the MMUTE input coupled with output LFE of the PCD3332-3. LFE can be toggled by the 'MICMUTE-key' to disable or enable the handset microphone.

PCD3332-3 dialler/ringer

The dial parameters of the PCD3332-3 can be set by diode options to specific-country requirements. A single contact keypad matrix is connected with the corresponding COL and ROW I/O's. This simple keypad offers no direct access of the stored numbers as proposed for application example 2. The 'STORE-key' and 'MRC-key' has to be used to store and recall telephone numbers. Diode switch MLA has to be open.

As explained before is the 'MICMUTE-key' applied to toggle the microphone amplifier during conversation by means of the LFE output. However, use of the 'MICMUTE-key' during ringing, toggles also the ringing melody.

Reset is performed by the internal reset of the PCD3332-3 mainly. Reset components C36-R35 compensates the spread of the internal reset voltage. Output DPN/FLN drives the interrupter to perform pulse dialling (PTS switch 'closed') and flash function (F/E switch 'open"). The position of cradle switch S1 determines the CSI level during stand-by (CSI = low) and conversation mode (CSI = high).

Input CE/FDI is connected to the positive line wire and the diode-bridge to detect the operation mode of the PCD3332-3 in combination with CSI.

Resistor R36 in the MUTE wire is required to prevent discharging of the VDD capacitor during the break-periods at pulse dialling or flash when VCC is reduced below the VDD voltage level at MUTE is high.

Output TONE delivers the melody for the ringer circuit (at HF/RTE = high) and the DTMF dialling signal to the DTMF input of the TEA1112 via attenuator R41-R42.

ROW 5 of the PCD3332-3 is an open drain output which is pulled-up by R44.

Ringer circuit

The VDD capacitor is kept charged during stand-by to speed-up initialization of the PCD3332-3 at incoming calls; see '*Start-up*' in this chapter.

Supply of the ringer is delivered by the ringer signal from the exchange via the bridge and the series network C31-R25.

When CE becomes high and CSI is kept low the PCD3332-3 enters the ringer mode at frequencies of the ringer signal between 20 Hz and 57 Hz (RFS switch is 'open') or between 14 Hz and 75 Hz (RFS switch is 'closed'). Output HF/RTE will be high during ringing to select the ringer circuit.

Volume control is performed by potentiometer R30. In application example 2 the volume of the ringer sound is controlled by means of the VOL1 and VOL2 outputs and the 'VOL1/VOL2-keys'. This principle can be applied for this example also.

The ringer melody can be changed by means of the key-board buttons 1, 2 and 3.

4.2 Settings and performance of the application

4.2.1 DC behaviour

DC settings

The DC voltage at the A-B/B-A terminals is a result of the voltage drop across the TEA1112, line interrupter and diode bridge. The voltage drop across the TEA1112 depends on the setting of the reference voltage (VREF) between LN and SLPE and the voltage drop across R9 which depends of the line current.

Important for the minimum line voltage (A-B/B-A) is the minimum supply voltage VDD required by the PCD3332-3. VDD is supplied by VCC which depends on the resistance value of R1, or network between LN and VCC in case of complex impedance, and the total current consumption from VCC.

To guarantee a minimum VDD supply voltage of 2.5 V during DTMF as well as pulse dialling VCC has to be increased by an enlarged reference voltage of the TEA1112 by means of R3 between REG and SLPE.

In case of 600 Ω set impedance the A-B/B-A voltage measures 6.0 V at 20 mA line current to get a minimum VDD of 2.5 V at DTMF as well as pulse dialling; R3 = 40 kΩ. The minimum VDD level is reached at pulse dialling (long digits) when VCC decreases below the VDD level.

Fig.47 shows the line voltage VA-B across the A-B/B-A terminals as a function of line current Iline at nominal and increased line voltage by means of R3 = 40 kΩ.

Supply possibilities

VCC can be applied to supply peripherals such as the PCD3332-3 and an electret microphone. The possibilities are rather limited and depend in general of the LN-SLPE setting, the DC resistance of the network between LN and VCC and the total current consumption from VCC.

Take in account that the minimum VCC level, to keep the TEA1112 functioning, is about 2.0 V at 20 mA line current. Furthermore, the voltage difference between VCC and SLPE has to be more than 1.6 V, over the whole line current range (!), to keep the send stage fully functional.

Start-up

After connecting the application with the line supply the very first time, the handset has to be lifted to charge the VCC and VDD supply capacitors. The set is operational within 200 ms at 20 mA line current.

During on-hook the VDD capacitor C35 is kept charged by R28. The DC current in this stand-by mode has to be more than 6 µA.

Start-up after off-hook (t = 0, VDD capacitor has been charged) is given in Fig.48 by means of the voltage VA-B across the set and the supply voltages VCC and VDD versus time. The set is supplied from an exchange voltage of 48 V while the line current is 20 mA during off-hook. R3 = 40 kΩ.

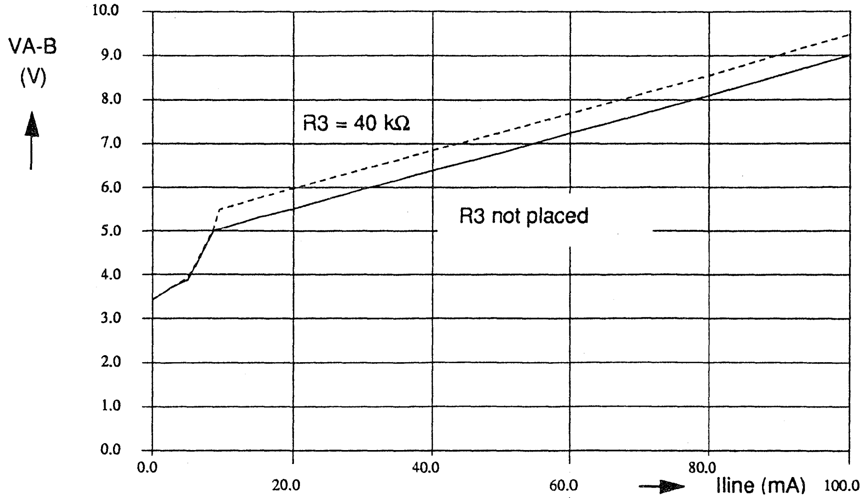


Fig.47 Line voltage across the set as a function of line current

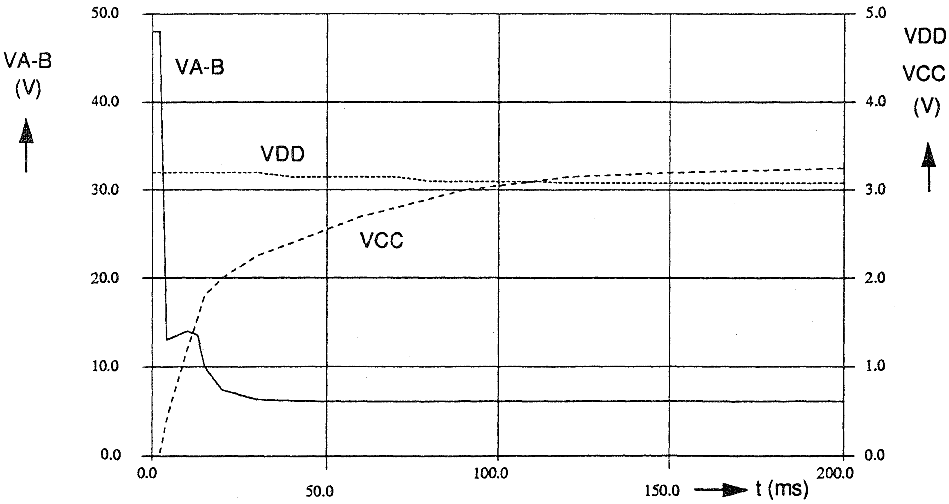


Fig.48 Start-up after off-hook

4.2.2 Transmission

Set impedance and BRL

A set impedance of 600 Ω can be realised with R1 = 619 Ω, while for complex set impedance the network between LN and VCC has to be defined.

Fig.49 shows the BRL (dB) of a '600 Ω set' measured with 600 Ω reference. In the same graph is given the BRL (dB) of a 'complex set' consisting of R1 = 825 Ω, C1 = 115 nF and R2 = 220 Ω measured with a reference impedance of 825 Ω // 115 nF + 220 Ω.

In case of complex set impedance the value of capacitor C3 has to be increased to meet BRL requirements.

In this example, at R1 = 825 Ω, C1 = 115 nF and R2 = 220 Ω, the value of C3 = 6.8 µF. To eliminate the influence of the transducers in the handset, they have been replaced by 200 Ω resistors during the measurement. The line current is 20 mA.

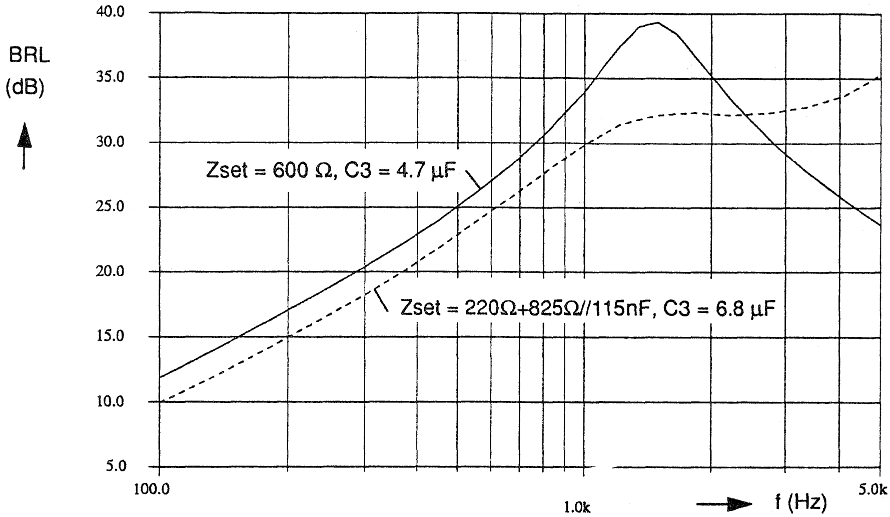


Fig.49 BRL of application example 1 at 'real' and 'complex' termination

Send and receive

This application is intended for use with a dynamic microphone. The total gain from microphone terminals to the line measures 50dB at 600 Ω set impedance and 600 Ω line load without AGC function. The internal setting of the TEA1112 is 52 dB typical, while about 2 dB is lost due the EMC components R16 and R17 (both 1 kΩ), in series with the MIC inputs, and the termination resistor R15 (10 kΩ) across the MIC inputs.

The send gain of the TEA1112 may be decreased to a minimum of 39dB according [2]. The overall send gain results in 37 dB with R4 = 20 kΩ.

The maximum swing of the line signal measures 5.5 dBm over the frequency range 300 Hz - 3400 Hz, at 600 Ω set impedance, 600 Ω line load and 20 mA line current. Capacitor C30 between the gate-source of TR1 keeps TR1 conducting at negative swings of the line signal; it improves the maximum swing of the line signal at lower frequencies.

The overall receive gain from line to earpiece is about –2.5 dB at an earpiece impedance of 150 Ω. This is due to the attenuation of 32 dB from line to IR input, the internally determined gain of the receive stage of 31 dB typically and the 1.5 dB attenuation due to EMC component R14 (22 Ω). The gain values are given without activated AGC function.

The receive gain can be reduced from 31 dB to 19 dB minimum by means of resistor R13. At R13 = 100 kΩ the receive gain is reduced by 6 dB which result in an overall receive gain of –8.5 dB typically.

Send and receive gains are internally defined by on-chip resistors. Reduction of these gains by external resistors (R4, R13) result in matching inaccuracies.

Side tone / AGC

Reproduction of the (electrical) microphone signal in the earpiece is reduced by the anti-sidetone circuit consisting of the components R5, R6, R7 and Zbal with R8, R10 and C4. The principle of the applied TEA1060-family bridge is given in chapter 3.12 and fully described in [8].

In case AGC is not applied (pin AGC open) the anti-sidetone circuit has to be re-calculated for a mean cable length of < 5 km. Readjustment of the balance circuit is necessary for other cable types, different line length, etc.

4.2.3 Dialling

DTMF dialling

The DTMF signal from the TONE output of the PCD3332-3 is attenuated by the network R41 and R42 and applied to the DTMF input of the TEA1112. Resistor R41 is in parallel to the input impedance of the DTMF amplifier (20 kΩ typ.). During dialling, MUTE is high, the signal is amplified by the DTMF stage and transferred to the line resulting in a total level of –6 dBm at 600 Ω set impedance and 600 Ω line load. The gain of the DTMF stage is 25.5 dB typical.

A reduction of the microphone gain by means of external resistor R4 reduces also the DTMF gain and transmitted signal levels. The attenuation network R41-R42 has to be redefined to correct the reduced signal transfer.

Take in account that VDD decreases during DTMF dialling because of the enlarged current consumption of the PCD3332-3 in this mode.

Pulse dialling / Flash

The line current will be interrupted by the electronic interrupter (TR1) under control of the DPN/FLN signal. During progress of the dialled digit (or flash) the PCD3332-3 has to be supplied by the stored energy of C35 because the level of VCC is too low to charge-up this capacitor. The value of buffer capacitor C35 has a value of 220 µF to keep the VDD supply level at >2.5 V.

Fig.50 shows the voltage at the A-B/B-A terminals, supply voltages VCC and VDD and the line current during dialling of a 'zero' at R3 = 40 kΩ, Vexchange = 48 V and 20 mA line current. The VDD voltage is reduced to about 2.5 V at the end of dialling phase as a test result for this application example. Take into account that VDD could be < 2.5 V at worst case conditions.

The selectable maximum FLASH time of the PCD3332-3 is 600 ms. At flash-times of 600 ms the VDD voltage remains > 2.5V.

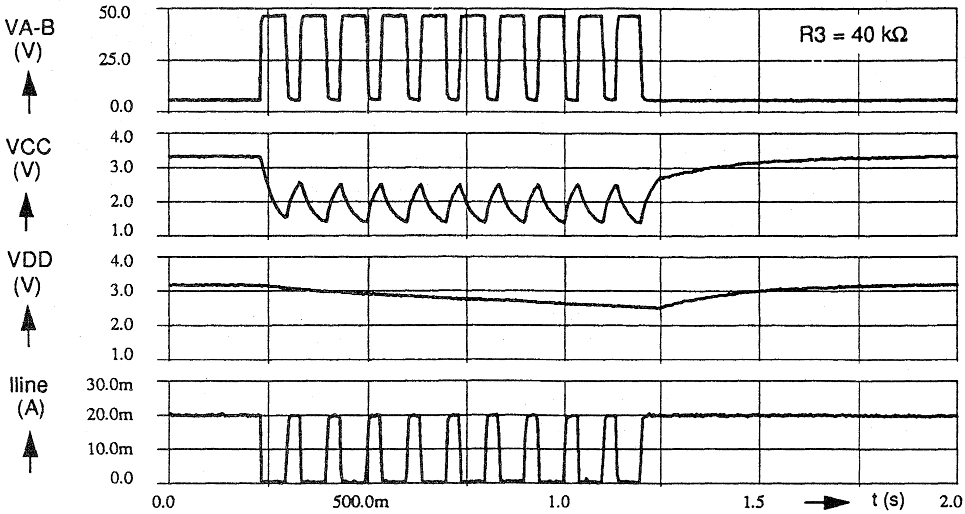


Fig.50 Behaviour of application example 1 during pulse dialling at 20 mA

5. APPLICATION EXAMPLE 2 - HANDSFREE SET -

5.1 Description of the application

The circuit diagram of the handsfree application is split-up in three figures. It consists of an electronic hook switch, TEA1112 transmission IC and discrete ringer circuit as shown in Fig.51, the TEA1093 handsfree application Fig.52 and the PCD3332-3 pulse/tone repertory dialler/ringer IC according Fig.53. Interconnections between Fig.51 and Fig.52 are indicated by means of net-in/net-out symbols while interconnections between Fig.51 and Fig.53 are given by offpage-symbols.

The application offers:

- Pulse, DTMF and mixed mode dialling, redial, 13-number repertory dialling with the PCD3332-3 [1]
- Transmission functions with adjustable settings as described for the TEA1112; chapter 3
- Handset operation
- Handsfree operation
- Ringer signal detection, melody generation and volume control.

Line connection / electronic hook switch / interrupter

The transmission circuitry and the ringer stage are connected with the line by two separate diode bridges to ensure proper functioning of the application independent of the polarity of line voltage and to rectify the ringer signal. The two zener-diodes Z14 and Z15 in series with the ringer bridge reduce the line-load from the ringer stage during transmission.

The application is protected against over-voltages at the line input by break-over diode D18. Components R20 and TR3 limit the current through TR1 when the line current exceeds about 150mA. This current limiter is not designed for continuous limitation of the line current. It only provides protection against current surges.

The electronic hook-switch / interrupter TR1 is controlled by inverter TR2 via the DPN/FLN open drain output of the PCD3332-3. During off-hook when the handset is lifted, or when the 'HOOK-key' is activated, DPN/FLN is high resulting in a conducting TR1. Conducting of TR2 is initiated by the high ohmic resistor R22 and is taken over by R24. Interruption of the line current is achieved by DPN/FLN is low.

When the application is connected with the line supply the very first time, the electronic hook-switch is switched-on for a short time resulting in a quick charge-up of the supply capacitors of VCC and VDD.

During stand-by the VDD capacitor is kept charged by means of R28.

Handset / Handsfree application

The handsfree circuit TEA1093 is connected between the positive line wire, which is connected to LN of the TEA1112 via R11 and R12, and SLPE of the TEA1112. The current into LN of the TEA1112 is as low as 3mA to have most of the line current available for the loudspeaker function of the TEA1093. Resistor R12 keeps the TEA1093 operational at saturation the line signal at large negative amplitudes.

The base microphone (HF-mic) or the handset microphone (HS-mic) are switched to the MIC input of the TEA1093 by means of TR7 respectively TR8 depending on the HF/RTE level of the PCD3332-3. Handsfree is switched-on when HF/RTE = high resulting in transfer of the HF-mic signal to the transmit input of the TEA1093 (MIC). Handset mode is achieved at HF/RTE is low; the HS-mic is operational while the TEA1093 is forced into the transmit mode by a low level of MUTER generated by TR9. Transmit or receive state is under control of the duplex controller of the TEA1093.

The discrete switching circuitry for the microphones could be replaced by the 74HC4053 multiplexer/demultiplexer IC as applied in [7].

The transmit output signal between MOUT and MICGND (in HS or HF mode) is transferred to the MIC inputs of the TEA1112 via attenuator R42, R43 and R15. The signal between the MIC inputs is amplified to the line.

The receive signal is transferred to the QR output of the TEA1112 and offered to the HS earpiece and RIN1 of the TEA1093. RIN2 is connected to VEE which is the ground reference of the receive signal of the TEA1112.

The signal between RIN1 and RIN2 is amplified by means of the loudspeaker amplifier and supplied to the loudspeaker. Volume control is performed by a simple potentiometer R41.

Transmit and receive gains of the transmit and receive channels of the TEA1093 and TEA1112 are in conformity with the sensitivities of the applied microphones, earpiece and loudspeaker; see 'Settings and performances of the application'; chapter 5.2. The TEA1112 is described in chapter 3 of this report while the settings of the duplex controller of the TEA1093 are according the Application Note of the TEA1093 [3] and the demonstration model of the TEA1093 [4].

PCD3332-3 dialler/ringer

A single contact 6 x 5 matrix keypad is connected with the corresponding COL and ROW I/O's. The keypad includes 10 memory keys, M0 to M9, for direct access of the stored numbers in case the MLA diode switch is closed. PCD3332-3 output DPN/FLN drives the electronic hook-switch to perform pulse dialling and flash function (F/E diode option not applied).

Reset is performed by the internal reset of the PCD3332-3 mainly. Reset components C71-R71 compensates the spread of the internal reset voltage. Input CE/FDI is connected to the positive line wire and the ringer bridge to detect the operation mode of the PCD3332-3 in combination with CSI. Series diode D17 in the positive line wire is applied to get a fast trailing edge of the CE pulse after on-hook or at line breaks.

Output MUTE is wired to the TEA1112 via R16 and to the TEA1093 via R52 and D20. This diode prevents levels at MUTET below the GND reference of the TEA1093.

Output TONE delivers the melody for the ringer circuit (at HF/RTE is high) and DTMF dialling signal to the DTMF input of the TEA1112 via the attenuator R41-R42.

The different modes of the PCD3332-3 are:

Stand-by mode: CE, CSI and HF/RTE are low during a specific time. The stand-by mode is left when CE goes high. It changes over to the ringer mode when an incoming ringer signal is detected, or changes over to the handset mode when CSI goes high or comes in the on-hook dialling or handsfree mode when the 'HOOK-key' is activated.

Ringer mode: CE = high, CSI = low resulting in HF/RTE = high. The ringer mode is left when CE goes low for time out (stand-by mode), when the handset is lifted (handset mode) or when the 'HOOK-key' is pressed (handsfree mode).

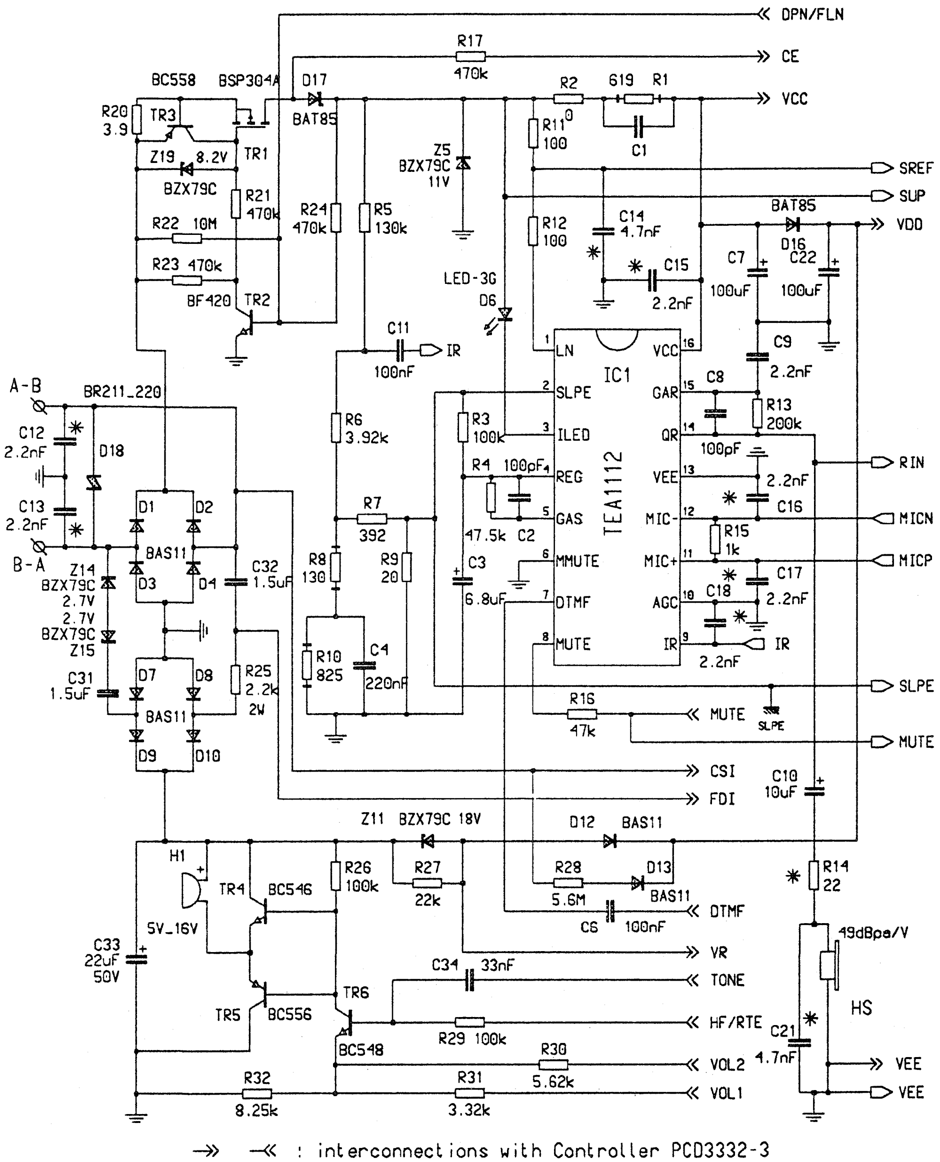
Handset mode: CE = high, CSI = high resulting in HF/RTE = low. The handset mode is left when the handset is put back on the cradle (stand-by mode) or when the 'HOOK-key' is pressed while the handset is put back (handsfree mode).

Handsfree mode: HF/RTE = high, CSI = low. This mode can be entered by pressing the 'HOOK-key'. The handsfree mode can be left by pressing the 'HOOK-key' (stand-by mode) or by lifting the handset (handset mode).

Dialling operations are possible in the handset and handsfree mode. Pulse or DTMF dialling can be selected by diode switch PTS [1].

Ringer circuit

The discrete ringer stage from example 1 is extended in this application with volume control by keypad via the PCD3332-3 outputs VOL1 and VOL2. The sound pressure from the PXE (Murata PKM34EW-1224) can be changed by 4 steps. Maximum volume is obtained when both VOL1 and VOL2 are low.



→» —« : interconnections with Controller PCD3332-3

—▷ —◁ : interconnections with Handsfree TEA1093 + local

∗ : EMC components

Fig.51 Application example 2; line interface TEA1112, electronic hook-switch and discrete ringer

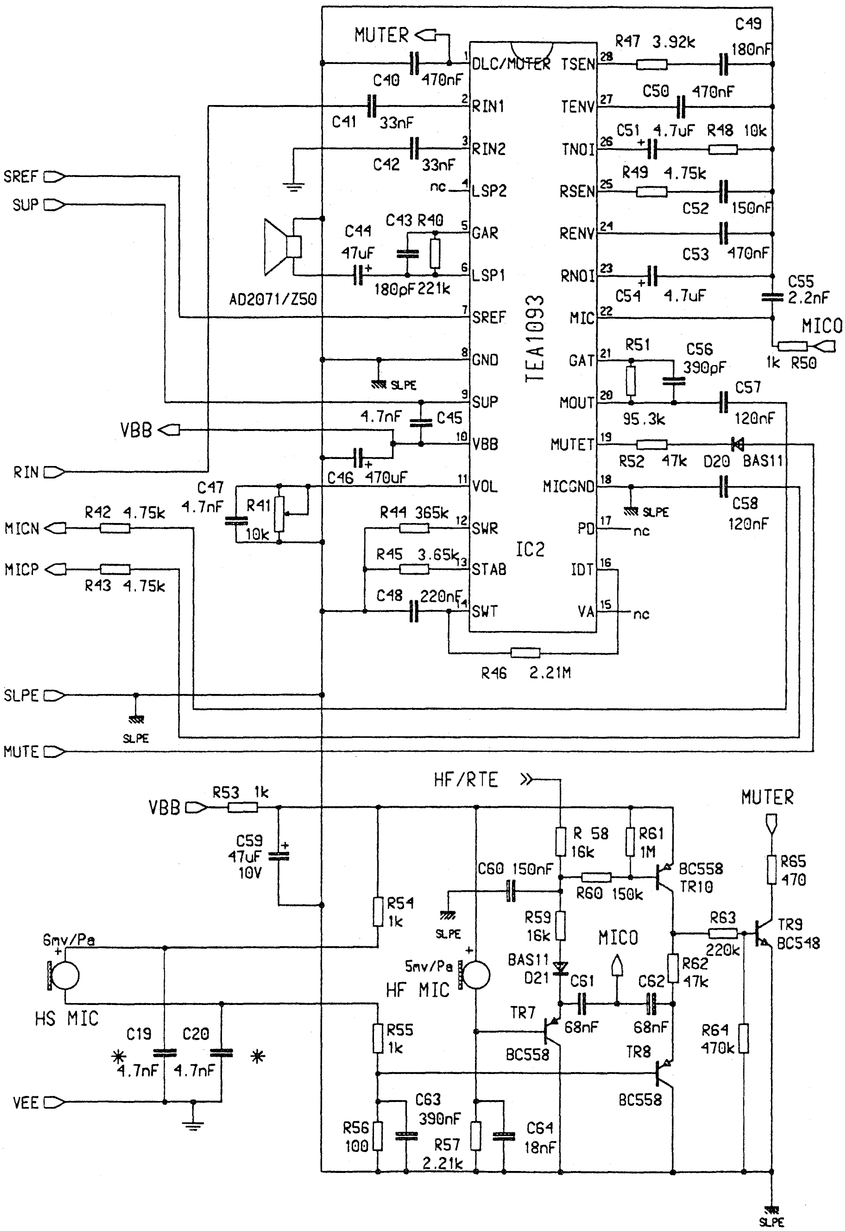


Fig.52 Application example 2; handsfree application TEA1093

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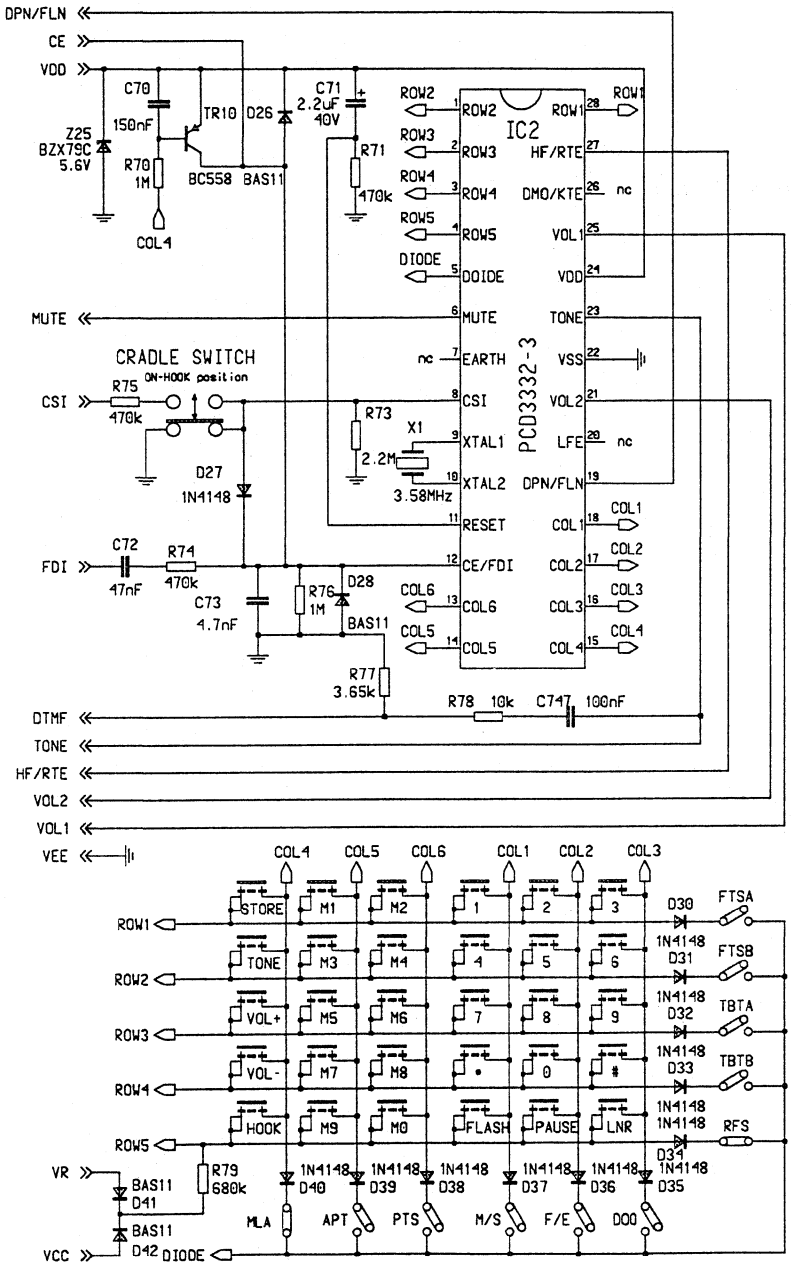


Fig.53 Application example 2; dialler/ringer PCD3332-3

5.2 Settings and performance of the application

DC settings

The stabilized voltage of the TEA1112 (VREF between LN and SLPE) is increased by means of R3 (100 kΩ) to adjust the voltage difference between SUP and VBB of the TEA1093 to 600mV. The voltage at the A-B/B-A line terminals measures 6.6 V at 20 mA. The DC slope of the Vline/Iline characteristic is about 45 Ω due to R20, R9 and the channel-resistance of TR1.

The stabilized voltage VBB (TEA1093) can be adjusted [3]. Take into account that VSUP–VBB has to be at least 600 mV to maintain maximum efficiency of the current switch of the TEA1093 at mean speech levels.

The line current can be split-up in:

- **Isup** flowing into SUP of the TEA1093 to supply the internal circuitry including loudspeaker amplifier and microphones
- **Iled** through D6 which is a function of the line current. Refer to chapter 3.3
- **Itr** flowing into LN of the TEA1112; realised by (VSUP–VSREF)/R11 = 0.32/100 = 3.2 mA typical
- **Ivcc** which includes the current consumption of the TEA1112, see chapter 3.1, and the current consumption of the PCD3332-3 in conversation mode [1]

Fig.54 shows these currents as a function of Iline in the conversation mode while Fig.55 gives the line voltage VA-B, supply voltages VCC and VDD both with respect to VEE, and the stabilized voltage VBB with respect to SLPE versus Iline.

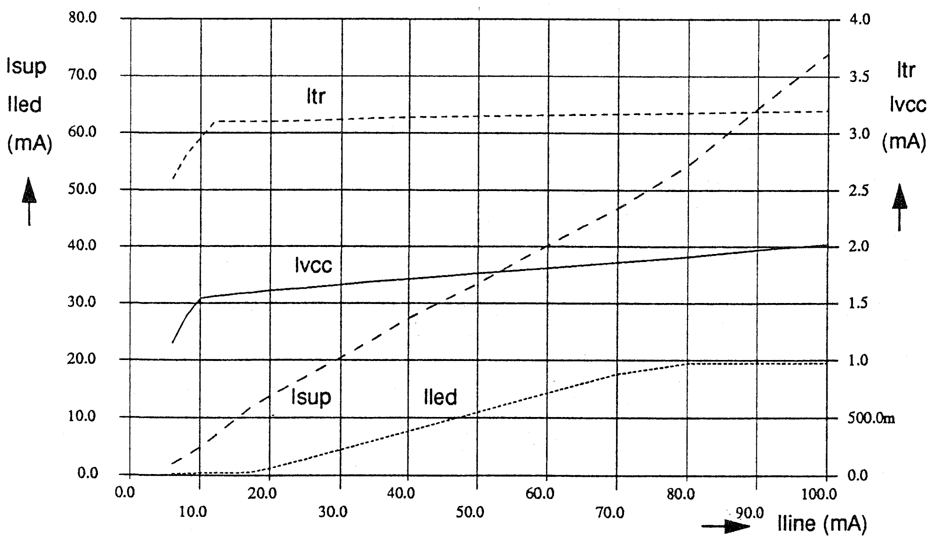


Fig.54 Currents Isup, Iled, Itr and Ivcc as a function of Iline

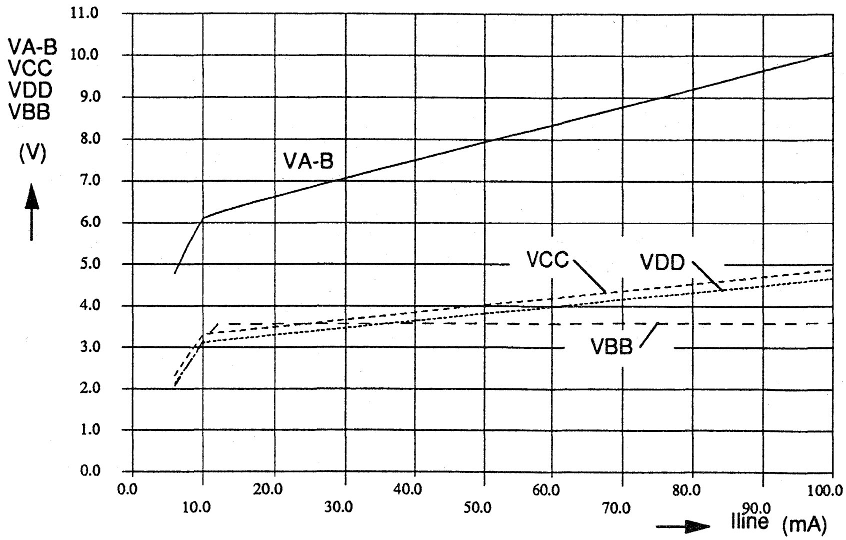


Fig.55 Voltages VA-B, VCC, VDD and VBB (with respect to SLPE) versus Iline

At 20 mA line current, the current into SUP measures 13.8 mA from which 5.5 mA (typ) is consumed by the internal circuitry of the TEA1093 and 250µA by the external circuitry connected to VBB. The remaining supply current to generate the loudspeaker signal (at 20 mA line current) is thus about 8 mA which gives a maximum output power of 15.8 mW theoretically into a 50 Ω loudspeaker. Measured is 12.5 mW at 20 mA; see also Fig.56.

Transmission

Transmit and receive gains are in conformity with the sensitivities of the proposed microphones, earpiece and loudspeaker and the performance of the application used as handset or handsfree set. The applied handset (Ericsson RLGN40201/8B6) contains an electret microphone with a sensitivity of –44.5 dBV/Pa (1 kHz, 2 kΩ load) and a dynamic earpiece of 150 Ω and 49 dBPa/V. The base contains the HF microphone with a sensitivity of –46 dBV/Pa (1 kHz, 2 kΩ load) and a 50 Ω loudspeaker (Philips type AD2071/Z50).

The overall transmit gain, at 600 Ω line load, from R56 or R57 to the line measures 48 dB as a result of 24 dB gain from R56 (or R57) to MOUT, 20 dB attenuation from MOUT to the MIC inputs of the TEA1112 and 44 dB gain from MIC inputs to the line. The default microphone gain of the TEA1112 is reduced to 44 dB by means of R4 (47.5 kΩ).

The receive gain from line to earpiece is –6.5 dB; from line to loudspeaker about 24 dB. The receive gain of the TEA1112 application is reduced from –1 dB default to –4.5 dB by means of R13 (200 kΩ). Volume control is achieved by potentiometer R41. A proposal for volume control by the PCD3332-3 can be found in [3], while a circuit realisation is offered in [6].

The BRL is more than 18 dB at 300 Hz to 3400 Hz, complex set impedance (R2 = 220 Ω, R1 = 825 Ω, C1 = 115 nF) and same reference impedance while measured without handset. C3 has to be at least 6.8 µF for complex set impedances but can be 4.7 µF for 600 Ω set impedance to meet BRL requirements.

Fig.56 shows the maximum power generated into a loudspeaker of 100 Ω, 50 Ω and respectively 25 Ω as a function of the available line current, with and without connected LED. The nominal VBB supply voltage measures 3.55 V. At the 'rising edges' of the curves the power is limited by the available supply current. The power in the 'flat area' of the curves is limited by the supply voltage VBB. The power in this area can be increased by an enlarged VBB voltage (VBB > 3.55 V) by means of a resistor between pin VA and pin GND of the TEA1093 [3]. Adjust in this case also the voltage at SUP, by means of R3, to get a minimum DC level of 600 mV between SUP and VBB.

The current consumed by the LED, see Fig.54, is not available for the handsfree loudspeaker function; it reduces the maximum power in the loudspeaker at lower line currents, as shown in Fig.56.

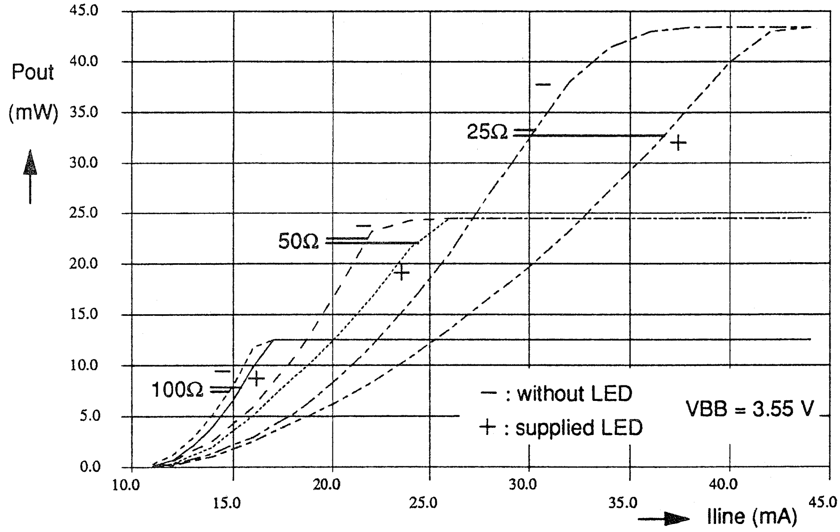


Fig.56 Maximum power into 100Ω, 50Ω respectively 25Ω loudspeaker versus Iline

Dialling

DTMF: The signal from the TONE output is attenuated by R77 and R78 (12.5 dB) and amplified by the TEA1112 (17.5 dB) to get a DTMF level of –6.5 dBm at 600 Ω line load.

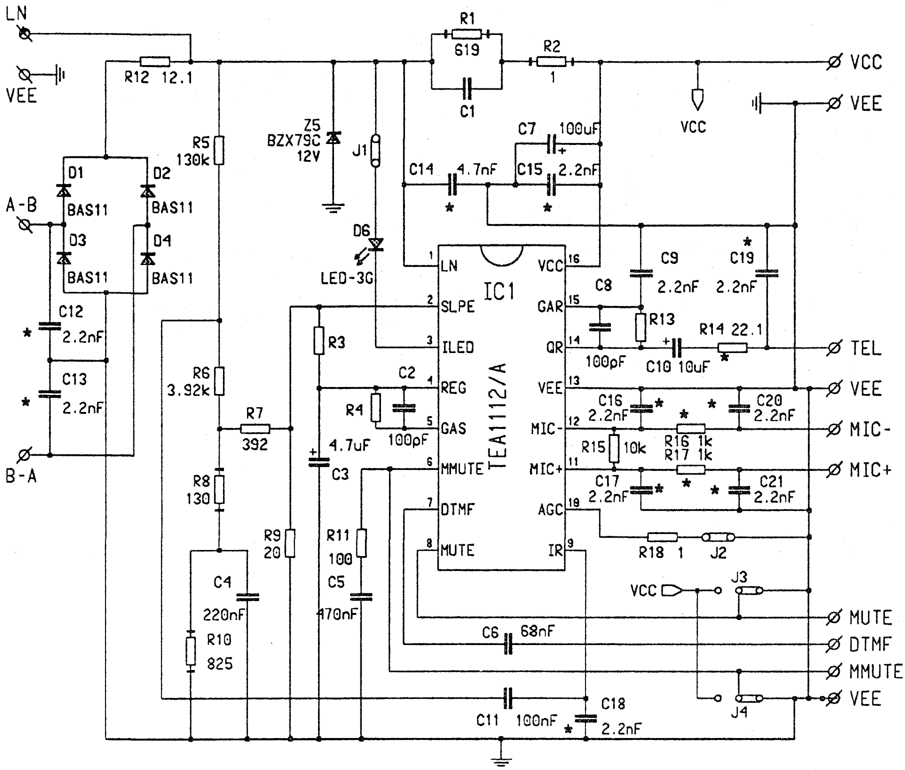
Pulse dialling: The line voltage of this application has been increased to create a voltage space of 600 mV between SUP and VBB of the TEA1093. This results in a VDD back-up level of more than 2.5 V during pulse dialling and flash times up to 600 ms (maximum selectable flash time).

6. DESIGN / ADJUSTMENT STEPS TEA1112/A APPLICATION

This chapter gives a number of adjustment steps which should be made to design or to adjust the basic application of the TEA1112/A. For every 'Adjustment' the 'Components' are given. The influence on the characteristics of the application and the considerations which have to be taken into account are added as 'Remarks'. The components refer to circuit diagram Fig.57 which is the application of evaluation board OM4776 as described in chapter 7.

Adjustment	Component(s)	Remark(s)
Set impedance	R1 or Z1	Zset depends mainly on R1 or network Z1 (R2 + R1//C1) for frequencies from 300 Hz up to 3400 Hz; R12 is in series with R1 or Z1. VCC supply depends on DC resistance of R1 or Z1.
BRL	R1 (Z1), C3	BRL depends on 'Set impedance' with respect to reference impedance (PTT requirement). Value of Leq (depends on the values of C3, R9 and resistor between LN and REG if applied) is important at the lower frequencies (300 Hz). Adapt C3 to improve BRL at 300 Hz if necessary. Value of C3 has also influence on the start-up time!
Side tone	Zbal (R8+R10//C4), R5, R6, R7	Depends on cable type, mean cable length, AGC function and Zset
DC slope	R12 (R9)	R12 is the best choice. Modification of R9 means also an adaption of Leq, VLN, low voltage threshold current, microphone gain, AGC function and side tone balancing.
VLN increase	R3 (REG-SLPE)	Refer to local PTT requirements. Increases VCC supply possibilities.
VLN decrease	R(LN-REG)	Reduces Leq; reduces the BRL at lower frequencies; see BRL. Reduces VCC supply voltage level; take in account the minimum operating level of VCC (2 V at 20 mA) and the minimum permitted voltage space between VCC and SLPE (1.6 V).
VCC supply	R1 (Z1), C1	VCC supply level depends on VLN, resistance of R1 or network (Z1) between LN and VCC and current consumption from VCC. Take in account the minimum operating level of VCC and the minimum voltage space between VCC and SLPE.
AGC	R18	Internally defined when AGC pin is connected to VEE. Adjustable by R18 to **increase** 'start and stop' currents in relation with Vexch and Rexch. See chapter 3.8. AGC function can be disabled by leaving pin AGC open.

Microphone gain	R4	Internally defined at 52 dB by internal resistance Rgasint. Can be reduced by R4. No matching of R4 with Rgasint. Take into account the attenuation from capsule to MIC inputs due to R16 and R17 with respect to R15 (in parallel with Zmic = 64 kΩ typ).
- High pass		Value of couple capacitors of microphone with respect to input impedance of external microphone network.
- Low pass	C2	Value of C2 in combination with R4//Rgasint.
- Supply		Electret microphone supplied from VCC via extra RC filter
DTMF gain		DTMF gain is microphone gain – 26.5 dB. Total DTMF gain has to be set by means of the attenuation network between DTMF generator and TEA1112/A DTMF input.
Receive gain	R13	Internally defined at 31 dB (from IR to QR) by internal resistance Rgarint. Can be reduced by R13. No matching of R13 with Rgarint. Take into account the attenuation from QR output to earpiece due to R14. Overall receive gain from line to earpiece depends on attenuation from line to IR input.
- High pass		C10 in combination with earpiece impedance. C11 in combination with source impedance and Zir (20 kΩ typ).
- Low pass	C8	Value of C8 in combination with R13//Rgarint
- Stability	C9	C9 has to be 20 x C8
MUTE	TEA1112	MUTE is active high. MUTE from dialler has to be high during dialling. Apply a series resistance in the MUTE wire from dialler to TEA1112 (ca. 50 kΩ) to prevent discharge of the VDD capacitor during break periods at pulse dialling (or flash) at MUTE = high.
$\overline{\text{MUTE}}$	TEA1112A	$\overline{\text{MUTE}}$ is active low. MUTE from the dialler has to be low during dialing.
LED		Current consumed by the LED is not available for an added HF application. The ILED-pin can be connected with SLPE when the LED function is not used.



*** : EMC components**

Not placed: R3, R4, R13 and C1

R11 and C5 are mounted on the PCB to demonstrate the TEA1113

Fig.57 Circuit diagram of the OM4776 evaluation board with the basic application of the TEA1112/A

7. RF IMMUNITY OF THE TEA1112 /A

The TEA1112 and TEA1112A have been designed with on-chip measures to keep RF disturbances away from sensitive circuit parts at higher RF frequencies (> 80 MHz). For the lower frequency range (from 150kHz upwards) the coupling into the IC occurs mainly via the A/B-lines and the handset cord. Improvement of the immunity at those frequencies can be realised by filtering at the PCB connectors and IC pins and a PCB layout which is designed with respect to EMC.

An evaluation board OM4776 [9] has been made for the TEA1112/A with the basic application according Fig.57. The components side is shown in Fig.58 and the board layout in Fig.59. The dimensions of the board are 6.5 x 8 cm. It is provided with connection terminals at the PCB edge and jumpers (J3, J4) to define the state of the logic inputs of the TEA1112 as well as the TEA1112A. Jumpers J1 and J2 are for the LED and AGC function respectively. Some of the components are mounted on soldering pins to simplify modification of the application.

Components R3, R4, R13 and C1 are not placed while R11 and C5 are intended for use of the board with the TEA1113. The TEA1113 is not described in this report; refer to [10]. See Fig.57 for components values.

The OM4776 has a single-sided wiring with filled ground plane between the interconnections. The EMC measures on the PCB are:

- Filtering from A-B/B-A terminals to line input LN of the TEA1112/A by means of C12 and C13 at the line terminals, R12 and C14 from pin LN to VEE.
- Filtering from the PCB terminals MIC-/MIC+ to the MIC inputs of the IC by C20 and C21 at the PCB terminals, series resistors R16 and R17 and decoupling at the MIC pins by means of C16 and C17. The bandwidth of microphone amplifier is limited by C2.
- Filtering of the receiver channel at input IR by C18 and from output QR to the earpiece terminals by means of R14 and C19. Furthermore is the bandwidth of the receiver amplifier limited by C8 and stability guaranteed by means of the combination of C8 and C9.
- Decoupling at VCC pin by means of C15.

General recommendations of EMC measures to design the PCB are:

- Use a filled ground between the wires in case of a single-sided PCB or a ground plane when a double-sided PCB is applied.
- Place line and handset connectors close to each other on the same side of the PCB and decouple the connections by means of EMC capacitors.
- Place EMC capacitors as close as possible to the corresponding IC pins. Use small size ceramic capacitors.
- Make interconnection-wires as short as possible. Use wire-bridges instead of a clever design with long wires.
- Design a symmetrical microphone entry from connector to MIC inputs of the IC.

Test method and results

The RF immunity test is split up in two test methods. The conducting test [11], in the frequency range of 150 kHz to 150 MHz, is carried out with a RF disturbance signal coupled into the A-B/B-A cable via coupling/decoupling networks. The RF signal with an amplitude of 3 V for f < 30 MHz and 0.5 V for f > 30 MHZ is modulated with an AM signal of 1 kHz sinewave and 80% modulation depth.The results of the measurements are given in Fig.60 by means of detected levels at the A-B lines and receiver output with respect to 1 Vrms (0 dBV) reference level.

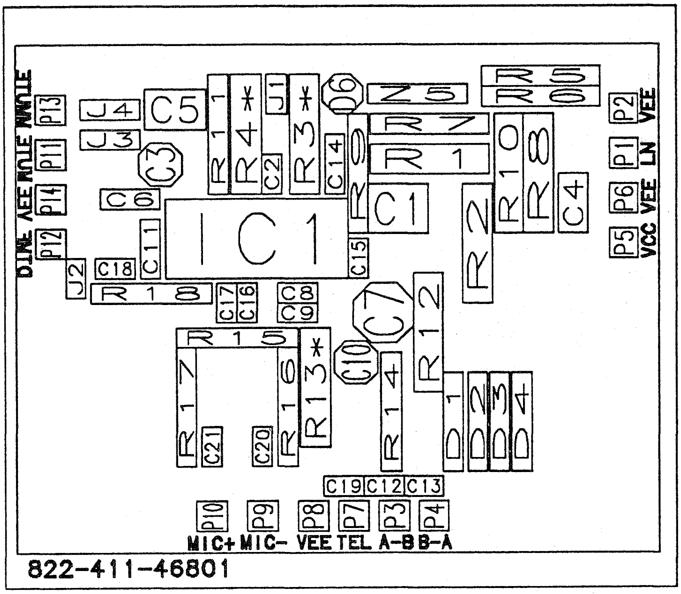


Fig.58 Components side of the OM4776 evaluation board

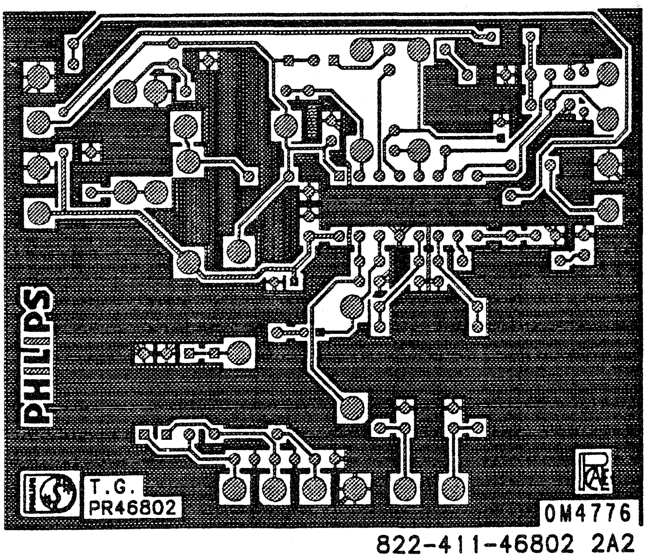


Fig.59 Layout of the wiring of the OM4776

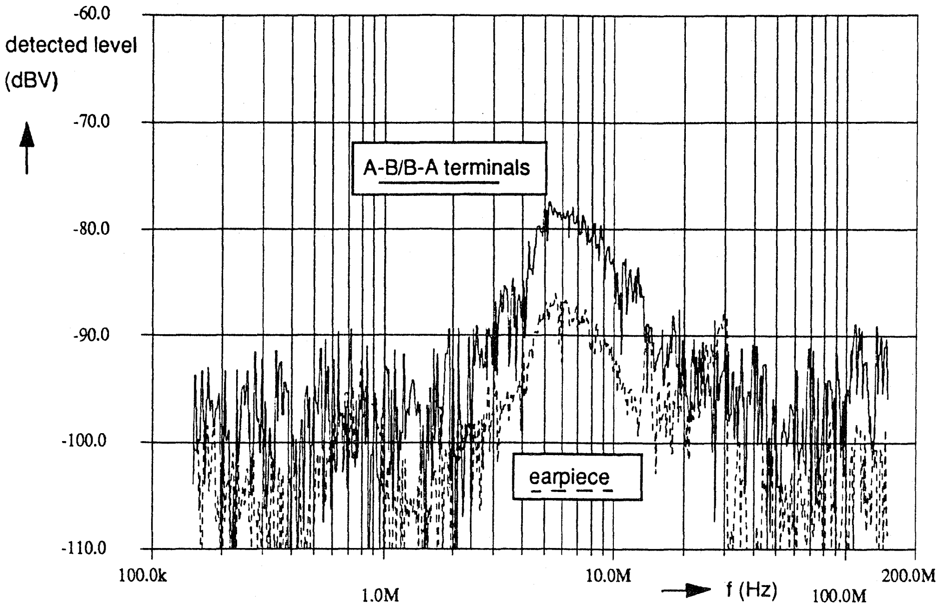


Fig.60 EMC behaviour of the OM4776; conducting test

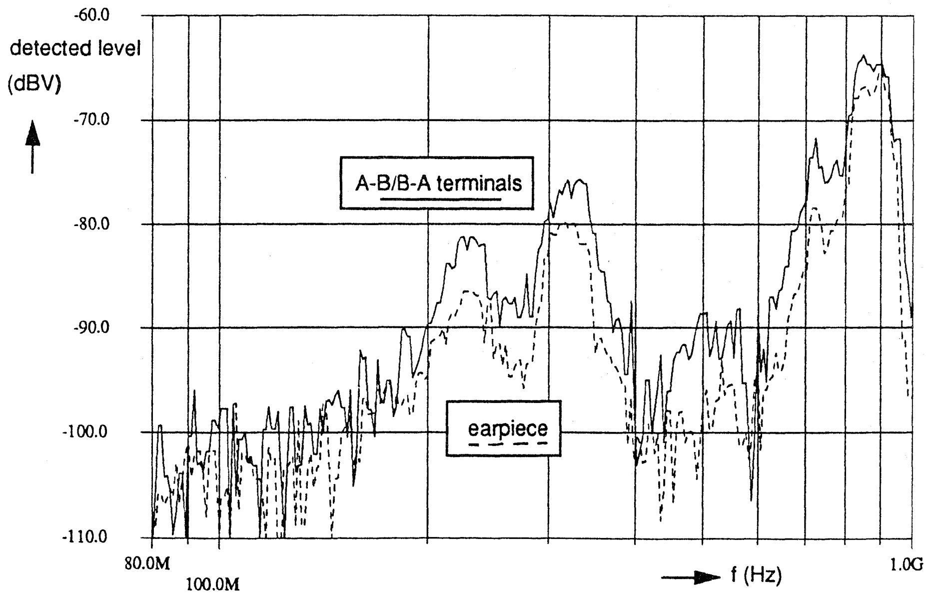


Fig.61 EMC behaviour of the OM4776; radiation test

The second test [12] of the OM4776 is carried out in an electro-magnetic field. The field strength is 3 V/m over the frequency range 80 MHz to 1 GHz while the signal is modulated with an AM signal of 1 kHz and 80% modulation depth. The results of the measurements are shown in Fig.61 by means of detected levels at the A-B lines and receiver output with respect to 1 Vrms (0 dBV) reference level.

The OM4776 evaluation board meets the requirements according [11] and [12]. The detected signal levels, as a result of the measurements, are in both cases less than the -60 dBV demands.

Note: The logic inputs MUTE, MMUTE (TEA1112) and $\overline{\text{MUTE}}$, $\overline{\text{MMUTE}}$ (TEA1112A) are sensitive because of the rather low internal pull-down currents. When they are not used connect them to VEE, in case of the TEA1112, or to VCC, in case of the TEA1112A.

8. REFERENCES

[1] Philips Semiconductors DATA DHEET PCD3332-3 Multi-standard pulse/tone repertory dialler/ringer.

[2] Philips Semiconductors Tentative Device Specification TEA1112 / TEA1112A. Low voltage versatile telephone transmission circuits with dialler interface.

[3] Philips Semiconductors Application Note ETT/AN93015. 'Application of the TEA1093 handsfree circuit', by C.H.Voorwinden & K.Wortel.

[4] Philips Semiconductors Application Note ETT/AN94001. 'User Manual for OM4750: Demonstration board TEA1093 and TEA1094', by R.v.Leeuwen & C.Voorwinden.

[5] DATA HANDBOOK IC03 'Semiconductors For Telecom Systems'.

[6] Philips Semiconductors Application Note ETT/AN94002. 'Design considerations for a high_end telephone set with PCA1070, TEA1093 and PCD335X', by K.Wortel.

[7] Philips Semiconductors Application Note ETT95007. 'OM4757 Demonstration Board PCD3332-3/ TEA1064B-1062/TEA1093-1094', by F.v.Dongen.

[8] Philips Components Laboratory Report ETT89009. 'Application of the versatile speech/transmission circuit TEA1064 in full electronic telephone sets', by F.v.Dongen & P.J.M.Sijbers.

[9] Philips Semiconductors User Manual ETT/UM95011. 'Evaluation board for the TEA1112/A and TEA1113', by E.Bosma.

[10] Philips Semiconductors Tentative Device Specification TEA1113. Low voltage versatile telephone transmission circuit with dialler interface.

[11] IEC Publication DIS 1000-4-6 (formerly 801-6). Electromagnetic compatibility for electrical and electronic equipment. Part6: Immunity to conducted disturbances, induced by radio frequency fields above 9 kHz.

[12] IEC 1000-4-3 Draft International Standard (Annex B). Immunity to radiated, radio frequency, electromagnetic fields (formerly IEC 801-3).

APPENDIX 1 List of abbreviations and definitions

A-B/B-A	Line terminals of application examples
AGC	Automatic Gain Control; line loss compensation facility
APT	Access Pause Time selection PCD3332-3
BRL	Balance Return Loss
CE/FDI	Chip Enable / Frequency Discriminator Input PCD3332-3
COL	Column keyboard input PCD3332-3
CSI	Cradle switch input PCD3332-3
DIODE	Diode option input PCD3332-3
DOO	DTMF output selection PCD3332-3
DPN/FLN	(Inverted) Dial Pulse / FLash output PCD3332-3
DTMF	Dual Tone Multi Frequency
EMC	Electro Magnetic Compatibility
Electret	Electret microphone with amplifier
F/E	Flash Earth selection PCD3332-3
GND	Ground reference TEA1093
GNDMIC	Ground reference microphone amplifier TEA1093
Gvrx	Gain factor of receive stage TEA1112/A
Gvtx	Gain factor of transmit stage TEA1112/A
HC4053	Philips IC with 3, 2-channel analogue switches
HF	Handsfree
HF-mic	Handsfree microphone
HF/RTE	Handsfree / Ringer Tone Enable output PCD3332-3
HOOK	HOOK-key PCD3332-3
HP	High Pass
HS-mic	Handset microphone
ICC, Icc	Current consumption of the TEA1112/A (from VCC)
Iled	Current through the LED connected between LN and ILED
Iline	Line current
Ip	Current consumption of the peripheral devices connected to VCC
Irec	Internal current consumption (from VCC) of the receiver amplifier of the TEA1112/A
Ish	Excess of line current from LN to SLPE
Istart, Istop	Start and stop currents of the AGC function
Ith	Threshold current of low voltage function
Itr	Current in transmission circuit of HF application
k	Scale factor of balance network
LED	Light Emitting Diode
LFE	Enable output PCD3332-3
LI	Listening-in
Leq	Artificial inductor of voltage stabilizer TEA1112/A; Leq = R9. C3. Rp
M/S	Mark Space ratio selection PCD3332-3

M0-M9	Memory location keys PCD3332-3
MIC	Microphone input TEA1093
MLA	Memory Location Access selection PCD3332-3
MOUT	Microphone amplifier output TEA1093
MRC	Memory Recall-key PCD3332-3
MUTE	MUTE output PCD3332-3 / MUTE input TEA1112
MUTER	Receive channel MUTE input TEA1093
MUTET	Transmit channel MUTE input TEA1093
OM4776	Evaluation board for the TEA1112/A
PCB	Printed Circuit Board
PCD3332-3	Multi standard pulse/tone repertory dialler/ringer IC
PTS	Pulse Tone Selection PCD3332-3
PXE	Piezo Ceramic Buzzer Element
Power Down/PD	Reduced current consumption mode during pulse dialling or flash
Ra	Resistor to adjust the sidetone bridge attenuation
Rast	Anti sidetone resistor
RESET	Reset input PCD3332-3
RF	Radio Frequency
RFS	Ringer Frequency Selection PCD3332-3
RINn	Receiver amplifier inputs TEA1093
ROW	Row keyboard input PCD3332-3
Rexch	Bridge resistance of exchange
Rgarint	Internal resistance (100 kΩ) to define receive gain TEA1112/A
Rgar	External resistance to reduce receive gain TEA1112/A
Rgasint	Internal resistance (69 kΩ) to define microphone gain TEA1112/A
Rgas	External resistance to reduce microphone gain TEA1112/A
Rp	Internal resistance of TEA1112/A between LN and REG
SREF	Supply reference input TEA1093
STORE	Store-key programming mode PCD3332-3
SUP	Supply input TEA1093
TEA1093	Handsfree IC
TEA1112	Transmission IC, MUTE and MMUTE active high
TEA1112A	Transmission IC, $\overline{\text{MUTE}}$ and $\overline{\text{MMUTE}}$ active low
TEA1112/A	General notation of the TEA1112 as well as the TEA1112A
TEA1113	Transmission IC of the TEA111X-family with dynamic limiter
THD	Total Harmonic Distortion (%)
TONE	Tone generator output PCD3332-3
VA-B	Voltage across the A-B/B-A line terminals
VBB	Supply output TEA1093
VCC	Supply pin / supply voltage of TEA1112/A
VDD	Positive supply PCD3332-3
VEE	Ground reference TEA1112/A

VLN	DC level at LN of the TEA1112/A (with respect to VEE)
VOL	Receiver volume adjustment TEA1093
VOLn	Volume control outputs PCD3332-3
VREF	Stabilized reference voltage between LN and SLPE of the TEA1112/A
VSLPE	DC level at SLPE TEA1112/A, DC level at GND TEA1093 of HF application
VSS	Negative Supply PCD3332-3
Vexch	Exchange voltage
XTALn	Oscillator inputs PCD3332-3
Zir	Input impedance of receive amplifier TEA1112/A
Zmic	Symmetrical Input impedance of microphone amplifier TEA1112/A
Z1	Complex network between LN and VCC TEA1112/A
Zbal	Balance network to reduce side tone
Zset	Set impedance between A-B/B-A terminals
α	Gain control factor of AGC function; $0.5 < \alpha <= 1$
[x]	Reference to REFERENCE chapter
(x)	Reference to equation (x)

APPLICATION NOTE Nr	AN94016
TITLE	Basics of PCA1070: a Programmable Analog Transmission IC (PACT)
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DATE	March 1994

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1 Introduction

The PTT's all over the world have different requirements for the electrical and acoustical characteristics of telephone sets and other equipment (like cordless telephones, fax, modems, personal communicators etc.) that must be connected to the public telephone network or to private network (PABX). In order to meet these requirements the electronics of modern telephone sets are different between countries. Especially the line interface part needs adaptation to meet the various requirements.
This means that virtually for each country where a telephone set must be approved, a different printed circuit board version is needed. Sometimes a number of components must be mounted or not mounted, or the value of some components must be changed.
The process to get approval from several PTTs for a new set design is very time consuming and costly. Because of the complex requirements of the various countries it may happen therefore that during the real type approval surprises may pop-up which require then fine-tuning of some parameters of the telephone set (e.g. changing some component values).
The number of different pcb versions for different countries and the time in which approval can be reached can be reduced enormously when a programmable line interface is used. This saves costs in the logistics.

A new IC that opens the possibility to manufacture a programmable transmission part for use in the line interface of all electronic devices that must be connected to the telephone line, is the PCA1070.
The PCA1070 is a CMOS integrated circuit performing all speech and line interface functions required in fully electronic telephone sets. In its basic application it needs a minimum number of external components. The transmission parameters are programmable via I²C-bus. This makes the IC adaptable to nearly all country requirements in the world, and to a various range of speech transducers, without changing the (few) external components of the basic application. For some countries a 12kHz or 16kHz tax pulse filter must be switched in series with line connection of the basic application.
The transmission parameters are stored in the EEPROM of a µC and are loaded into the PCA1070 via the I²Cbus during the start-up phase of the transmission IC (hook-off).
PCA1070 also allows adaptation to the connected telephone line, by reading the line current via I²C-bus and processing it in the µC.

Main application areas for PCA1070 are: wired telephony (basic till feature phones), combination-terminals (e.g. telephone + answering machine or fax), modems and base units of cordless telephones. With PCA1070 the number of country versions of the electronics in a design is minimized and the number of components that are needed can be reduced drastically.

This report gives a detailed description of the PCA1070 and its basic application in electronic telephone sets. Also EMC aspects, protection and tax pulse filtering are discussed. Furthermore an application example of PCA1070 together with a preprogrammed µC PCD3353A/008 is given and some measurement results of this application are shown.

For product details is referred to the Device Specification PCA1070 in Ref. [1]
For specification of I²C-bus see Ref. [2]

The Philips PCD335x family of µC is recommended for use with PCA1070. This µC family is supported with a One Time Programmable µC (PCD3755A). For details see Ref. [3] and Ref. [4]. In Appendix A an overview of the features of this µC family is given. Appendix B gives a list of abbreviations used in this report.

2 Block diagram

This chapter gives the block diagram, the pinning and a concise description of the block diagram including the function of the external components. Also an overview of the programmability of the circuit is given and a brief description of the I²C-programming protocols.

2.1 Block diagram and pinning

The block diagram of the PCA1070 together with the most essential external components is shown in Fig. 1. Parameters that can be programmed via I²C-interface are indicated below each block.
Fig. 2 gives the pinning of the circuit and Table 1 gives the pin description.

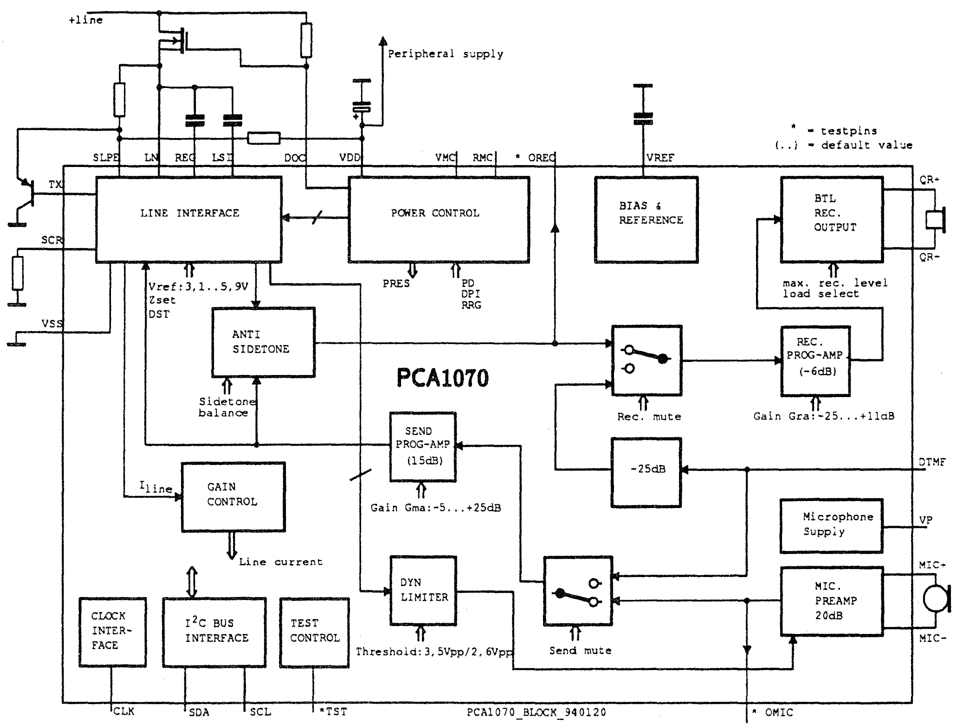


Figure 1 Block diagram PCA1070

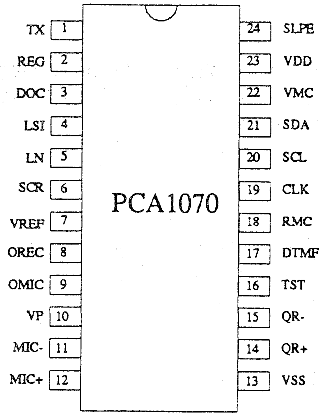


Figure 2 Pinning PCA1070

Pin	Name	Description
1	TX	drive output
2	REG	voltage regulator decoupling
3	DOC	dial output connection
4	LSI	line signal input
5	LN	positive line terminal
6	SCR	sending current resistor
7	VREF	voltage reference decoupling
8	OREC	output receive preamplifier
9	OMIC	output microphone preamplifier
10	VP	supply for electret microphone
11	MIC-	inverting input of microphone preamplifier
12	MIC+	non-inverting input of microphone preamplifier
13	VSS	negative line terminal
14	QR+	non-inverting output of BTL receive output amplifier
15	QR-	inverting output of BTL receive output amplifier
16	TST	testpin; to be connected to VSS in application
17	DTMF	dual tone multi-frequency input
18	RMC	reset output for micro controller
19	CLK	clock signal input
20	SCL	serial clock line of I^2C-bus
21	SDA	serial data line of I^2C-bus
22	VMC	input to sense supply voltage micro controller
23	VDD	positive supply decoupling
24	SLPE	slope (DC resistance) adjustment

TABLE 1 Pin description PCA1070

2.2 Concise description of the blocks

Fig. 1 shows that the IC consists basically out of 17 blocks. Looking to functionality the circuit can be split up in 5 functions: the line interface and supply, the microphone channel, the receive channel, the DTMF channel, the control and interface part.
These functions and blocks are discussed briefly here. Programmability is discussed in paragraph 2.3.

Line interface and supply
The line interface consists of an AC part and a DC part. The IC is supplied from the telephone line via the polarity guard (not shown in Fig. 1), the N-MOST interrupter, the slope resistor R_{SLPE} (fixed value 20Ω) and the resistor R_{SUP} (250Ω) between SLPE and VDD.

The DC line interface consists of a voltage stabilizer which stabilizes the DC voltage between SLPE and VSS and sinks the excess line current from pin SLPE via the external pnp transistor (driven from pin TX). An artificial coil function between SLPE and VSS is created by C_{REG}, R_{SLPE} and an on-chip resistor Rp (between REG and VSS). This ensures high impedance for audio signals in order to minimize influence on the set impedance generated by the AC line interface.
The SLPE voltage is passed through a low pass filter (R_{SUP}, C_{VDD}) to provide a smoothed supply voltage at VDD for all blocks of the IC that are not supplied from the line interface terminals (LN, SLPE). This supply voltage is also intended to power peripheral circuitry (e.g. the μC).

The AC line interface synthesises the set impedance and modulates the line via output LN. Therefore it takes a (constant) bias current from the line. The resistor at pin SCR (R_{SCR}=100Ω fixed value) is a reference resistor for the generation of the set impedance and it determines the DC bias current I_{LN} of the AC line interface. Pin LSI is the (voltage) feedback input of the active output stage and serves also as an input for the receive channel. The capacitor C_{LSI} is meant for DC blocking.

A low line current supervisor reduces the DC voltage drop at SLPE and the bias current into LN at low line currents to enable operation with relaxed performance under poor feeding conditions (e.g. parallel operation of sets).

The microphone channel
Microphone signals are (pre-)amplified with 20dB in the mic. pre-amp and are then passed via the I^2C-controlled send-mute switch and a programmable amplifier (the send prog-amp) to the AC line interface part which modulates the line by modulation of the bias current in the reference resistor at pin SCR (R_{SCR}).
A dynamic limiter prevents distortion of the transmitted microphone signals by reducing the gain (speech controlled) of the mic. pre-amp.
Pin VP provides a stabilized supply voltage for electret microphones.

The receive channel
The line signals on LN, consisting of receive signals superimposed with sending signals, enter the LSI input via a DC blocking capacitor C_{LSI} and are then attenuated in the AC line interface with 6dB. The output signal of the AC line interface is then passed via the anti sidetone block to remove (a large portion of) the sending signal. Receive signals are not attenuated in the anti-sidetone block. The output signal from the anti sidetone block is the receive signal (plus the remainder of the sending signal: the sidetone) and is then passed via the I^2C-controlled receive mute switch and a programmable amplifier (the receive prog-amp) into the receive output stage which drives an earpiece in either BTL mode (6dB gain) or SEL mode (0dB gain).
BTL = Bridge Tied Load or symmetrical drive. SEL = Single Ended Load or asymmetrical drive.

The DTMF channel

DTMF signals are fed directly into the send mute switch. The DTMF channel is activated by setting the send mute switch via I^2C-bus. After the mute switch the DTMF signals follow the same path as the microphone channel.
Via a 25dB attenuator the DTMF signals are fed into the receive mute switch and if this switch is activated, the signals are then processed further in the receive path to drive the earpiece with low level DTMF signals. This is done to obtain a confidence tone during DTMF dialling .

Control and interface blocks

Power control

This block ensures correct start-up and switch-off of the PCA1070 under several working conditions (speech and ringing). It supervises the supply voltage V_{DD} of the PCA1070 and VMC of the µC and takes care of internal power on reset (POR) and provides a reset signal for the µC at pin RMC.
A power down (PD) function can be activated to switch the IC in low power mode to survive line breaks and to facilitate pulse dialling. The circuit can also be switched in standby mode. In this mode the power consumption of the IC is extremely low and it is meant for applications where power backup from the telephone line in on-hook situation is needed (e.g. electronic hook-switch applications).
The block also controls the DOC output which drives an external N-MOST (or P-MOST) interrupter for pulse dialling and/or flash.
The internal ring/speech detector supervises if the IC is supplied from the line (in speech condition) or from an external source (e.g. during ringing). This is needed when only 1 supply capacitor for both the PCA1070 and the µC is used. The circuit is switched in normal operating condition during supply from the line and in standby condition during supply from an external supply voltage. This prevents reverse supply of the IC during ringing.

Gain control

The DC current flowing in the resistor between LN and SLPE (R_{SLPE}=20Ω) is measured and via an A/D converter a bit-code is written into a register. The contents of this register can be read via I^2C-bus to provide information about the DC line current.

Clock interface

An external clock signal of 3.58MHz must be applied to the CLK input. This block then provides all necessary internal clock signals to other blocks where needed. If no clock is applied, the set impedance and the sidetone balance impedance are both switched automatically to 600Ω and cannot be programmed to other values.

I^2C-interface

This block is a standard slave device which works fully according to the standard I^2C-specifications. The SDA and SCL pins must be connected to the µC supply voltage VMC via 2 pull-up resistors.

Bias & reference

This block contains an on-chip bandgap reference which is available at pin VREF. An external capacitor at pin VREF (C_{VREF}=100nF) ensures optimum noise performance of the PCA1070.
The block also generates the necessary bias voltages and bias currents for the PCA1070.

Test control

This block is used only for test purposes during production of the PCA1070. It can set-up a number of test-conditions which are used for easy access of separate blocks.

2.3 Overview of programmability

The following parameters (see also block diagram in Fig. 1) can be programmed by means of a bit-code via the I^2C-bus:

Block	Parameter	Symbol	Bits	Description
Line Interface	Vref Zset DST	VDCx ZSAx ZSBx ZSPx DST	3 3 3 4 1	DC voltage SLPE-VSS AC set impedance Ra AC set impedance Rb AC set impedance fp DC Start Time
Power Control	PD DPI RRG	PDx DPI RRG	2 1 1	Power Down Dial Pulse Input Reset RinG detector
BTL Rec. Output	Max. rec. level Load select	HPL RFC	1 1	Hearing Protection Level Resistive/Capacitive load
Anti Sidetone	Sidetone balance	ZOSAx ZOSBx ZOSPx	4 4 4	Z Optimum Sidetone Rsa Z Optimum sidetone Rsb Z Optimum Sidetone Cs
Rec. Mute	Rec. mute	RM	1	Receive Mute
Rec. prog-amp	Gain Gra	GRAx	6	Gain Receive prog-Amp
Send prog-amp	Gain Gma	GMAx	6	Gain send prog-amp
Send Mute	Send mute	SM	1	Send Mute
Dynamic Limiter	Threshold	DLT	1	Dynamic Limiter Threshold

TABLE 2 Programmable parameters PCA1070

The following parameters (see also block diagram) can be read as a bit-code via the I^2C-bus:

Block	Parameter	Symbol	Bits	Description
Power Control	PRES	PRES	1	Pact RESet
Gain Control	Line current	LCx	5	Line Current

TABLE 3 Readable parameters PCA1070

For details about the setting ranges and the step resolution see Ref. [1] (Data sheet).

2.4 Programming via I²C-interface

The I²C-interface is used to program the transmission parameters and control functions.
For details about the I²C-bus protocol and standard specifications see Ref. [2]. A concise hardware description of the I²C-interface block of the PCA1070 plus some programming examples are given in paragraph 3.13.

	A6	A5	A4	A3	A2	A1	A0	R/Wn
The device address is	0	1	0	0	0	1	0	x

All functions can be accessed by writing an 8 bit word to the PCA1070.
In order to set up the PCA1070, a control message comprising the device address, a R/Wn bit, a subaddress byte and one or more data bytes must be written to the PCA1070.
If more than one data byte follows the subaddress, these bytes are stored in the successive registers by the automatic increment feature.

The control word format for the **slave receiver**:

```
   Device address        Subaddress               Data/control byte
|S|0 1 0 0 0 1 0 0|A|I7 I6 I5 I4 I3 I2 I1 I0|A|D7 D6 D5 D4 D3 D2 D1 D0|A|P|
                 |
                 └─ R/Wn
```

Table 4 shows the bit arrangement of each data byte used in the control word. The bits that are not indicated must be set to "0".

Function	Sub-address	D7	D6	D5	D4	D3	D2	D1	D0
DC voltage	H00		VDC2	VDC1	VDC0				DST
Sidetone and Set impedance	H01	ZOSB3	ZOSB2	ZOSB1		ZOSA3	ZOSA2	ZOSA1	ZOSA0
	H02	ZOSP3	ZOSP2	ZOSP1			ZSA2	ZSA1	ZSA0
	H03		ZSB2	ZSB1	ZSB0	ZSP3	ZSP2	ZSP1	ZSP0
Send channel	H04	DLT		GMA5	GMA4	GMA3	GMA2	GMA1	GMA0
Receive channel	H05	RFC	HPL	GRA5	GRA4	GRA3	GRA2	GRA1	GRA0
Control	H06	PD1	PD0		RRG	RM	SM		DPI

TABLE 4 Bit arrangement PCA1070 (write)

The control word format for the **slave transmitter**:
Note: change in direction of read/write bit

```
   Device address       Data/status byte
┌─┬───────────────┬─┬───────────────────────┬─┬─┐
│S│0 1 0 0 0 1 0 1│A│D7 D6 D5 D4 D3 D2 D1 D0│A│P│
└─┴───────────────┴─┴───────────────────────┴─┴─┘
                 │
                 └─ R/Wn
```

Function	Sub-address	D7	D6	D5	D4	D3	D2	D1	D0
Status PCA1070		PRES			LC4	LC3	LC2	LC1	LC0

TABLE 5 Bit arrangement PCA1070 (read)

PRES : Indicates if PCA1070 has received an internal reset (when V_{DD} has dropped below 1.2V+/-0.2V); PRES will be set to "1" with internal reset and is set to "0" after reading the register via I^2C-bus.

LC4 - LC0 : Information about value of line current.

An example of a SW specification for a pre-programmed µC suitable for use with PCA1070 (the PCD3353A/008) is given in Ref. [5].

3 Description of the circuit

3.1 Basic application of the PCA1070

The basic application diagram of PCA1070 with the simplest supply structure is shown in Fig. 3. Only 1 supply decoupling capacitor is used here for the supply pins of both the PCA1070 and the µC.
In Fig. 4 the basic application diagram with a supply system using 2 decoupling capacitors is given. Here a diode is used between the 2 supply decoupling capacitors to prevent discharge of the µC supply capacitor into the PCA1070 in the case of a (long) line break or to prevent reverse supply when the µC is supplied from another source (e.g. during ringing condition). More complex supply structures are of course possible but are not discussed in this report.

A suitable family of µC for use with PCA1070 is the PCD335x (Ref. [3]). This µC family has on-chip EEPROM and DTMF generator and its oscillator (if set to 3.58MHz) can be used to deliver a clock signal to the PCA1070 CLK input without the need for external components. An overview of the PCD335x family is given in Appendix A. Also the PCD3755A, an 8-bit µC with on-chip DTMF generator, 8k OTP and 128-bytes EEPROM is a suitable µC for use with PCA1070. The instruction set is based on that of the MAB8048 and is software compatible with the PCD335x family. For details please see Ref. [4].

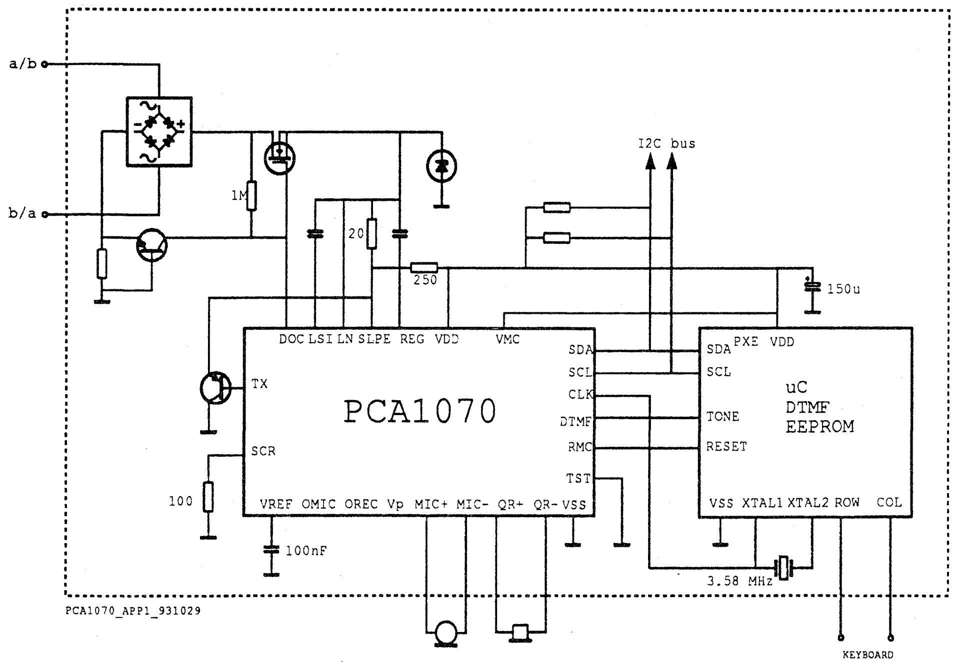


Figure 3 Basic application diagram PCA1070 (supply system with 1 capacitor)

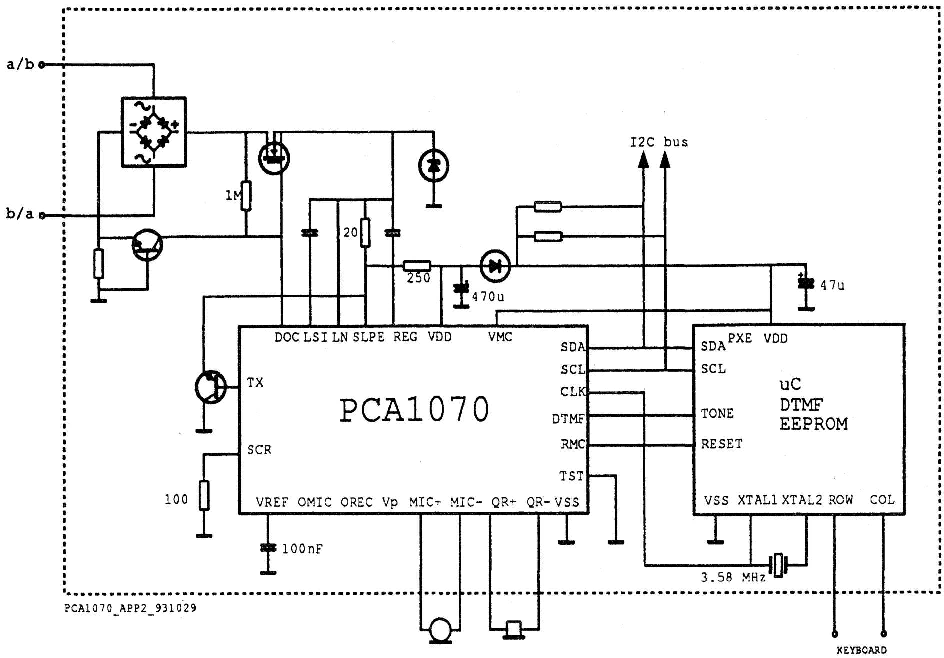


Figure 4 Basic application diagram PCA1070 (supply system with 2 capacitors)

3.2 Test circuit

The measurement results given in the following paragraphs, have been performed in the test schematic of Fig. 5. The settings of the programmable parameters are according to Table 6 unless otherwise noted. Test conditions and component values are as follows (unless otherwise noted):
I_{LINE}=20mA, V_{SS}=0V, f=1000Hz, I_P=0mA, I_{VP}=0mA, f_{CLK}=3.58MHz, Tamb=25°C, Z_{LINE}=220Ω+(820Ω//115nF), Rm=150Ω, Rt=150Ω, PNP_{TX}=BC328.

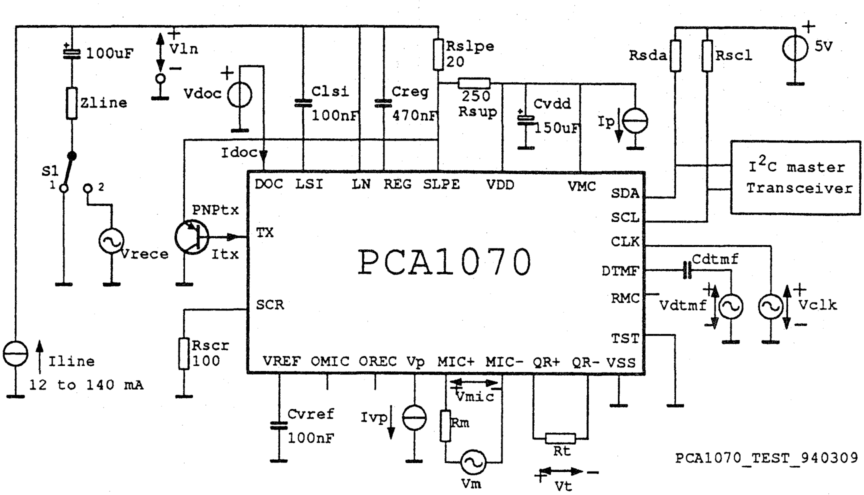


Figure 5 Test circuit PCA1070

Function	Sub-address	D7	D6	D5	D4	D3	D2	D1	D0
DC voltage	H00		VDCx=4.7V						DST =0
Sidetone and Set impedance	H01	ZOSBx=1259Ω				ZOSAx =492Ω			
	H02	ZOSPx=134nF					ZSAx =200Ω		
	H03		ZSBx=800Ω			ZSPx =1915Hz			
Send channel	H04	DLT=0		GMAx =15dB					
Receive channel	H05	RFC=0		GRAx =-6dB					
Control	H06	PDx=00			RRG=0	RM =0	SM =0		DPI =0

TABLE 6 Programmed settings used for measurements

3.3 Regulated line voltage and supply

The IC is supplied with current from the telephone line. For effective operation of the telephone circuitry it must have a low resistance to DC and a high impedance to speech signals (300 to 3400Hz). This is done by incorporating a voltage regulator in the IC in series with an artificial inductor. The principle diagram of the voltage regulator and current management is shown in Fig. 6. This configuration regulates the DC voltage drop between pins SLPE and VSS to a constant value. So the AC sending and/or receiving signals being present on the line pin LN are also superimposed on the DC voltage between SLPE and VSS. Therefore this voltage is passed via a first order lowpass (R_{SUP}, C_{VDD}) to provide a smoothed supply voltage V_{DD} for the IC (typically I_{DD}=2.3mA at V_{DD}=2.5V) and for the peripheral circuits in such a way that the line termination impedance is not deteriorated.
Typical internal current consumption I_{DD} as a function of V_{DD} is shown in Fig. 7.

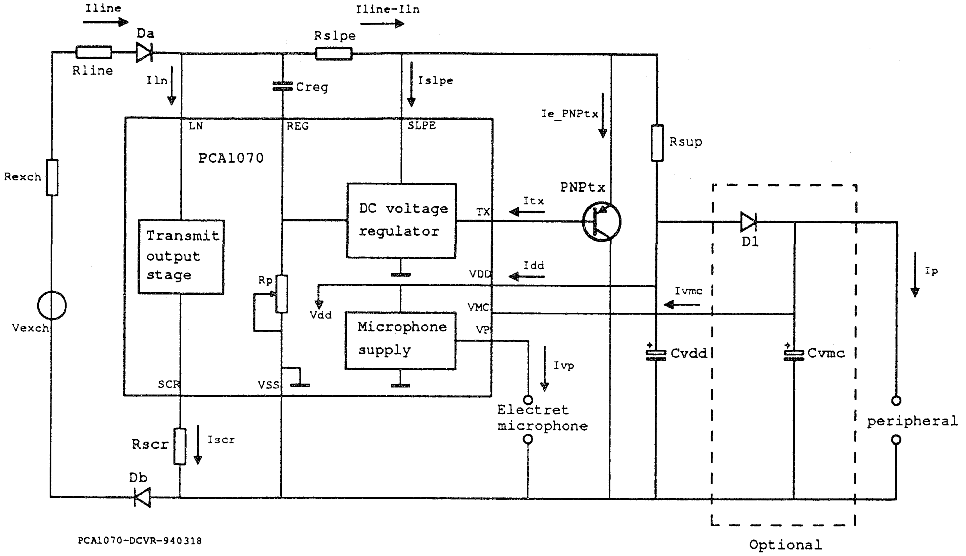


Figure 6 Principle of DC regulator and current management

The direct current I_{LINE} which flows into the set is determined by:
* the exchange supply voltage (V_{EXCH}).
* the resistance of the feeding bridge (R_{EXCH}).
* the resistance of the subscriber line (R_{LINE}).
* the DC voltage across the subscriber set, including the polarity guard.

If the line currents exceeds $I_{LN} + I_{SLPE} + I_{DD} + I_{VMC} + I_P$ then the voltage regulator diverts the excess current through the pnp transistor which is driven from pin TX.

Where:

I_{LN}	=	Bias current of the AC sending output stage (≈3mA if I_{LINE}>17mA)
I_{SLPE}	=	Bias current of the DC line interface (I_{SLPE}≈0.35mA)
I_{DD}	=	Internal supply current (Fig. 7)
I_{VMC}	=	Internal supply current sense input µC supply (I_{VMC}≈4µA)
I_P	=	Supply current to peripheral circuits (See paragraph 3.3.2)

Because the receiving output stage of the PCA1070 is of the class-AB type, the internal supply current I_{DD} will increase when an earpiece is driven. Also when a current (I_{VP}) is drawn from the supply pin VP for electret microphones, the internal supply current will increase. The currents I_{SLPE} and I_{VMC} are negligible during normal operation and will not be taken into account in the following chapters.

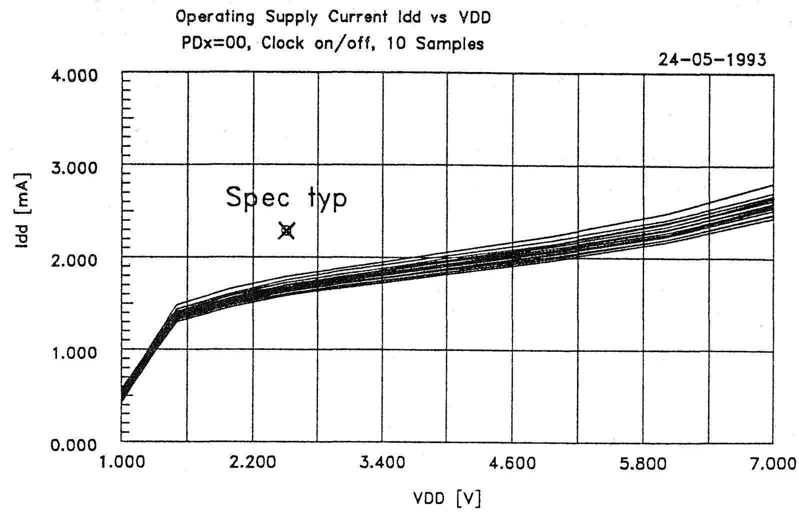


Figure 7 Internal current consumption I_{DD} as a function of V_{DD} in operating condition (I_{VP}=0, V_M=0, Vrec=0)

3.3.1 DC characteristics

The DC voltage V_{SLPE} can be programmed via I^2C-interface to suit the DC requirements in loop condition and during pulse dialling (NSA function). The programming range is 3.1 to 5.9V in 0.4V steps. Its default value is 4.7V.
The DC line voltage at pin LN is :

$$V_{LN} = V_{SLPE} + (I_{LINE} - I_{LN}) \bullet R_{SLPE}$$

where

I_{LN}	=	dc bias current flowing into pin LN (≈3mA if I_{LINE}>17mA).
R_{SLPE}	=	external 20Ω resistor between LN and SLPE

The typical DC-voltages V_{LN} and V_{SLPE} as a function of line current are shown in Fig. 8. The slope of the V_{LN} DC characteristic is determined by R_{SLPE}. This resistor has a fixed value of 20Ω.
At line currents below typ. 6mA the DC voltage V_{SLPE} is automatically adjusted to a lower value (Fig. 9). This means that the operation of more sets in parallel is possible with reduced sending and receiving levels and relaxed performance.
At line currents below typ. 16mA (max. 17mA) the dc bias current I_{LN} (≈I_{SCR}) is reduced from ≈3mA to a lower value (Fig. 10) to ensure under all line current conditions maximum possible transmit level capability.

Increasing the DC slope

The gradient of the DC voltage drop V_{LN} as a function of line current is normally fixed to 20Ω (=R_{SLPE}). It may of course be increased by connecting a resistor in series with the line.
However with PCA1070 it can be influenced also under software control without the need to change external components. This is possible by reading the line current via I^2C-bus (valid range 15 to 91mA) into the µC and then reprogramming the DC voltage V_{SLPE} via I^2C-bus to a value that depends on the actual line current. This may be done once after start-up of the telephone set or regularly (e.g. every 200ms or every 10s).

NSA function

In some countries the DC requirements are different for loop (= normal speech) condition and dialling condition. In practice this means that during pulse dialling the DC voltage drop of the set must be switched to a lower voltage. This function is known as NSA in Germany or MUTE2 in other countries. With PCA1070 this can be realized under software control by reprogramming the DC voltage V_{SLPE} during the dialling period. It is recommended to program the DST control bit to "1" to ensure short settling time (see paragraph 3.15.1).

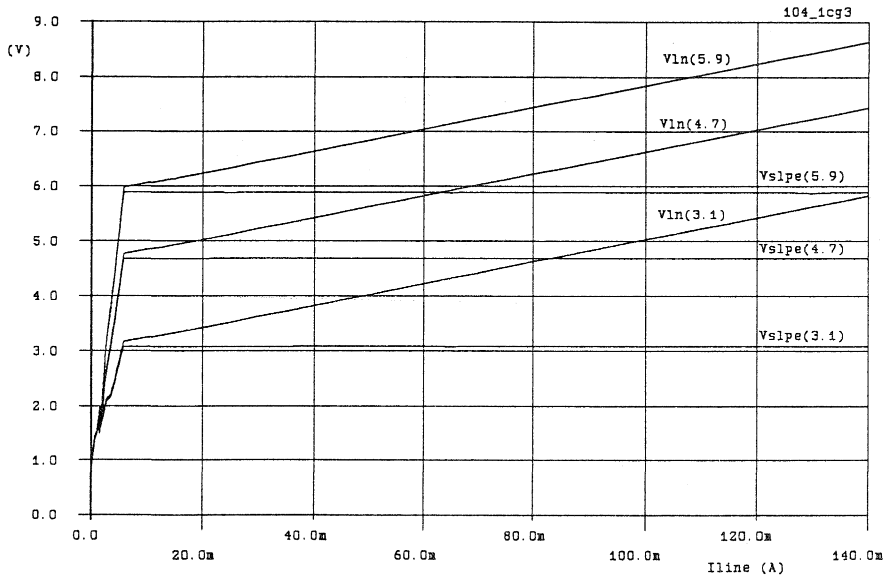


Figure 8 DC characteristics V_{LN} and V_{SLPE} = f(I_{LINE}) at V_{SLPE}=3.1, 4.7 and 5.9V

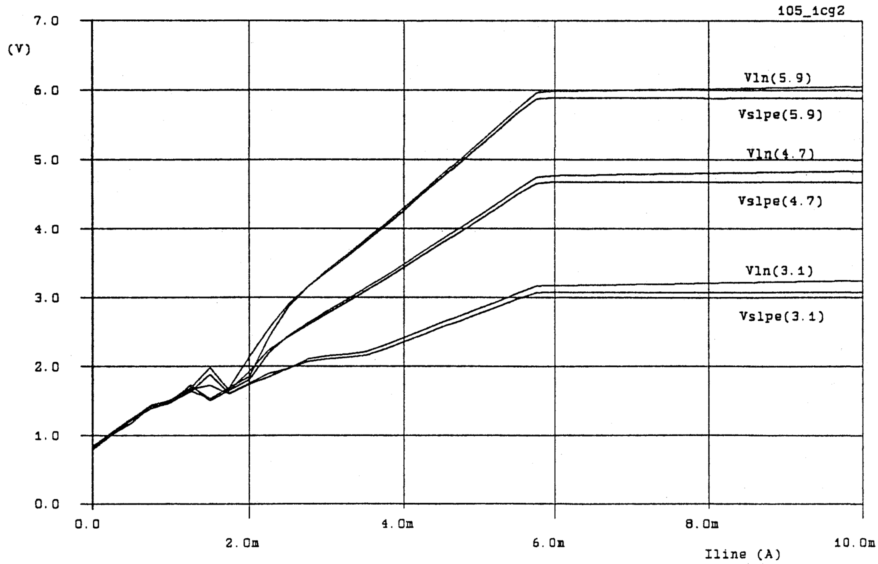


Figure 9 DC characteristics V_{LN} and $V_{SLPE} = f(I_{LINE})$ in low line current range

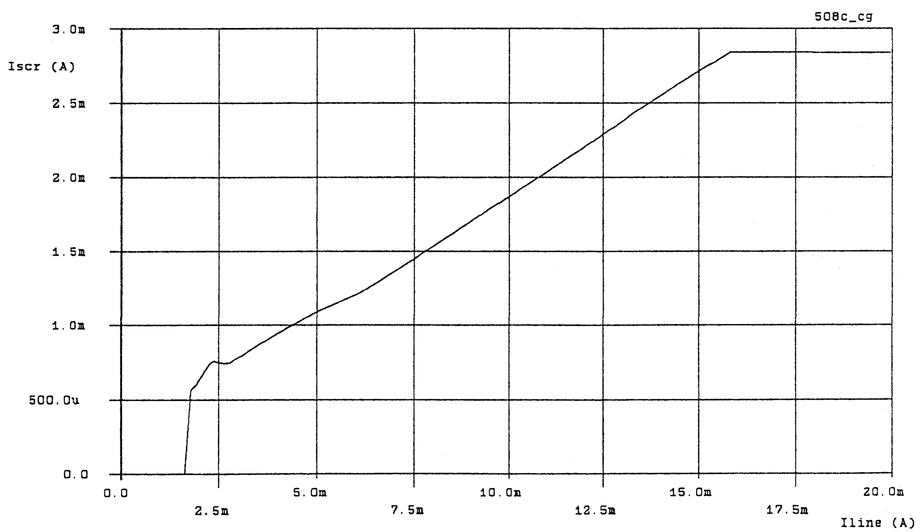


Figure 10 DC bias current AC line interface $I_{LN} \approx I_{SCR} = f(I_{LINE})$ in low line current range

3.3.2 Supply for peripheral circuits

The voltage available between VDD and VSS can be used to power peripheral circuits (see Fig. 6). This can be done directly (see Fig. 3, system with 1 supply capacitor; pin VMC is connected to VDD so $V_{VDD}=V_{VMC}$) or via a diode (see Fig. 4, system with 2 supply capacitors; pins VMC and VDD are now separated via a diode; the voltage between VMC and VSS is now the peripheral supply voltage).
The supply capabilities depend on the DC voltage V_{SLPE} (which is programmable via I^2C-bus) and the value of the resistor between SLPE and VDD and the voltage drop across the diode D1 (in the case this is used).

3.3.2.1 System with 1 supply capacitor

Fig. 11 shows the supply voltage V_{DD} (=V_{VMC}) as a function of peripheral supply current in the case of a supply system with 1 capacitor.
For Philips microcontrollers PCD335x family, the minimum supply voltage is 1.8V (2.5V for DTMF and EEPROM writing). For a lowest supply voltage of 2.5V Fig. 11 shows that typically I_P=4mA at V_{DD}=2.5V is available for V_{SLPE}=3.9V.

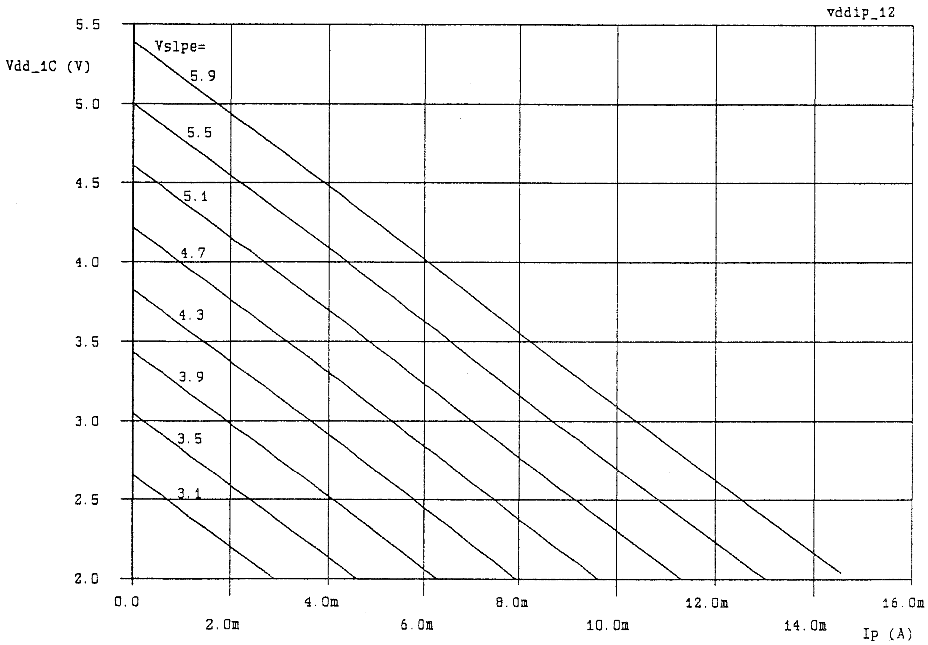


Figure 11 Peripheral supply: V_{DD} versus I_P (system with 1 supply capacitor); parameter is V_{SLPE}

3.3.2.2 System with 2 supply capacitors

The supply voltage V_{VMC} as a function of peripheral supply current in the case of a supply system with 2 capacitors (using diode type 1N4148 for D1) is shown in Fig. 12.

For Philips microcontrollers PCD335x family, the minimum supply voltage is 1.8V (2.5V for DTMF and EEPROM writing). For a lowest supply voltage of 2.5V Fig. 12 shows that with V_{SLPE}=3.9V a supply current of typically I_P=1.1mA is available in the case of the 2 supply capacitor system. A higher V_{SLPE} shows that with =3.9V a supply current of typically I_P=1.1mA is available in the case of the 2 supply capacitor system. V_{SLPE} results in a considerably higher capability. The supply capabilities can be increased also by using a Schottky diode (e.g. BAT85 has a forward voltage of less than 0.32V at 25°C at 1mA).

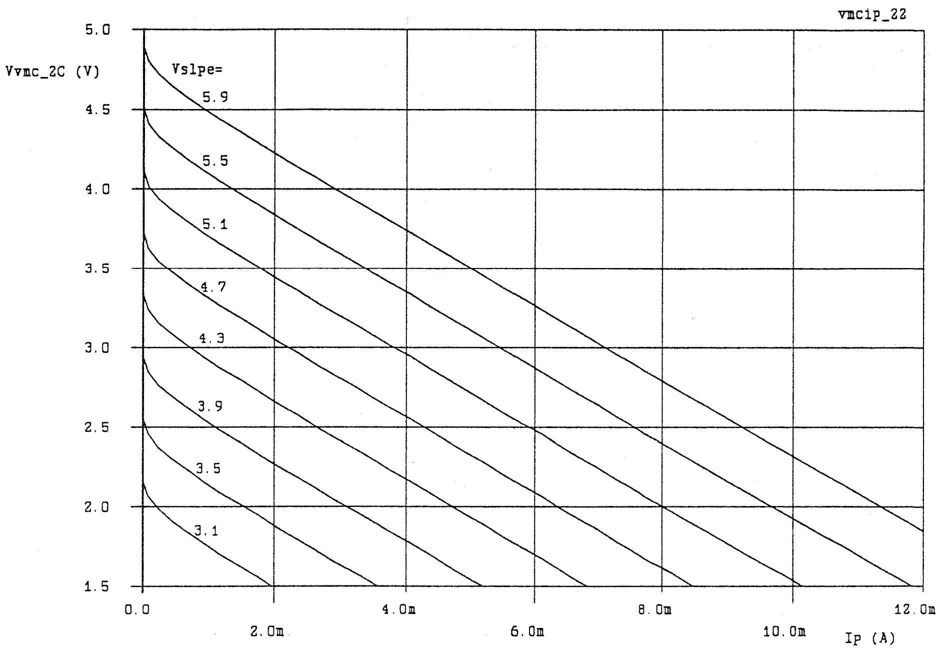


Figure 12 Peripheral supply: V_{VMC} versus I_P (system with 2 supply capacitors); parameter is V_{SLPE}

3.3.3 Microphone supply

Pin VP is available to supply electret microphones. The supply capabilities are depicted in Fig. 13 in unloaded condition I_{VP}=0 and with I_{VP}=500µA. The available supply voltage is stabilized at typically 1.9V for V_{DD}>=2.5V with maximum load of 500 µA. For symmetry reasons it is important that the output impedance of VP is low in the audio frequency range. This is shown in Fig. 14. To prevent unwanted feedback of the AC line signal via the first order low- pass filter between SLPE and VDD, into the sending amplifier, the power supply rejection (PSR) between VDD and VP must be rather high especially at low frequencies. Fig. 15 shows that the PSR between VDD and VP as a function of frequency has excellent performance.

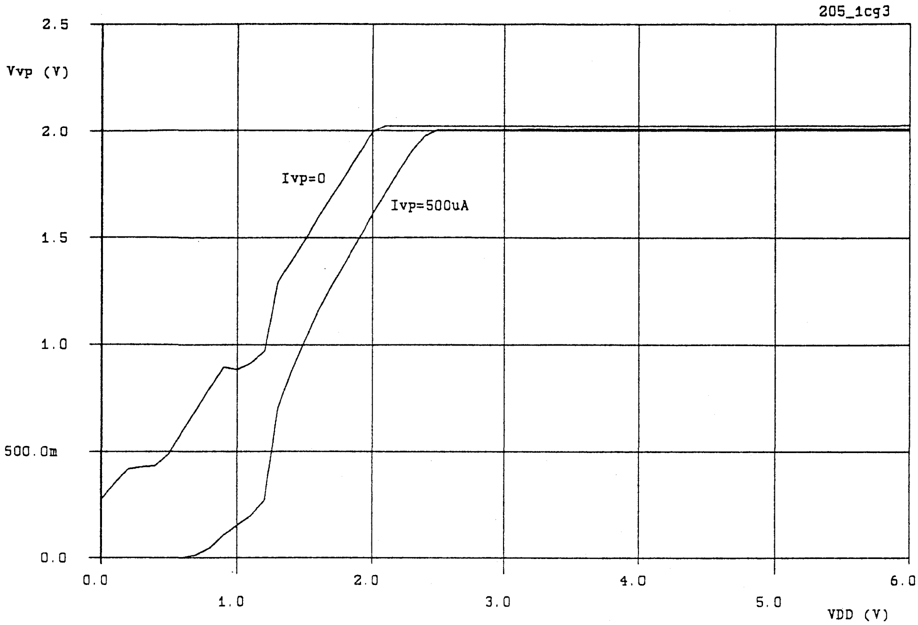


Figure 13 Supply for electret microphones: $V_{VP}=f(V_{DD})$ at $I_{VP}=0$, 500µA

Microphones needing a higher supply voltage than 1.9V

Although the typical supply voltage has been chosen such that the most frequently used types of electret microphones may be supplied from VP, it may happen that a certain microphone type needs a higher voltage. In this case an alternative supply configuration may be used taking VDD as a supply voltage. An extra low-pass filter is needed then to eliminate ripple from the VDD supply and assuring sufficient suppression of the residual AC line signals being present on VDD.

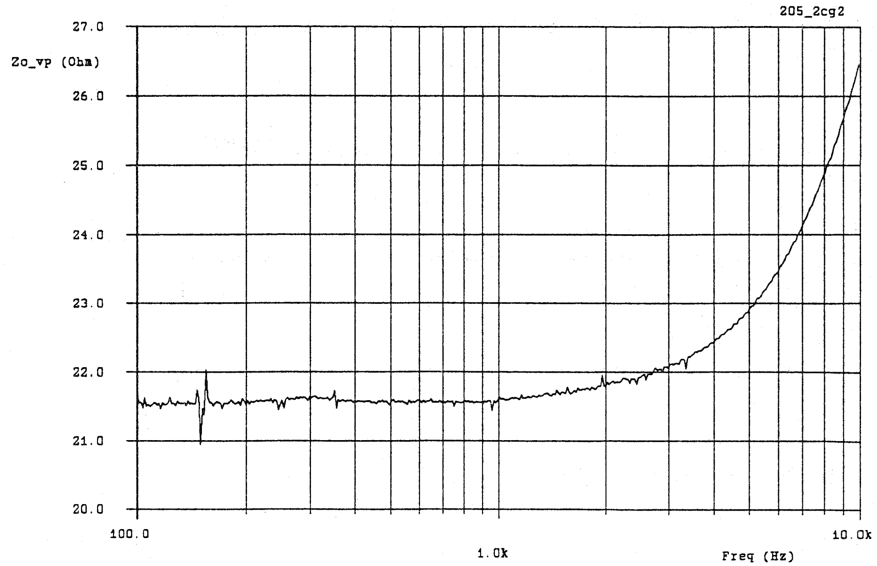


Figure 14 Supply for electret: output impedance Zo_{VP}=f(freq) at I_{VP}=500μA, V_{SLPE}=4.7V

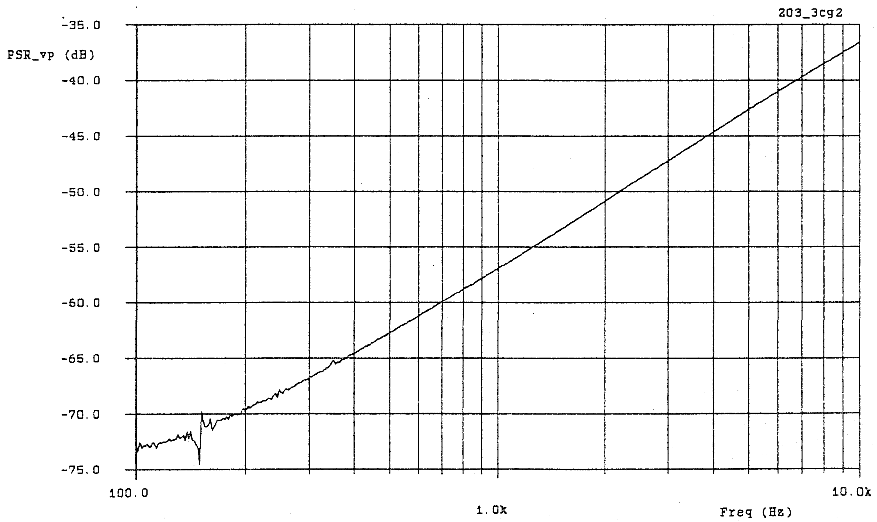


Figure 15 Supply for electret VP: power supply rejection PSR_{VP}=f(freq) at I_{VP}=500μA, V_{SLPE}=4.7V

3.3.4 Maximum sending level

The maximum possible peak to peak output swing on the line pin LN in sending direction as a function of the effectively available current (I_{LINE}-I_P-I_{VP}) is shown in Fig. 16. The effectively available current depends on the supply current I_P used for peripherals and the current I_{VP} sinked from the supply for electret microphone. Fig. 16 shows also that the programmed DC voltage of the circuit (V_{SLPE}) has influence on the sending output swing capabilities.
The on-chip dynamic limiter prevents distortion for most of the overdrive conditions of the microphone amplifier input that may occur in practice (see paragraph 3.6).

In principle two mechanisms limit the maximum possible sending level on the line: limitation by current or limitation by voltage. This is explained now in more detail.

Limitation on current

a. The bias current in the sending output stage I_{LN} ($\approx I_{SCR}$) is constant (typically $\approx$3mA) for I_{LINE} > 16mA typically. For I_{LINE} < 16mA this current decreases to ensure under all line current conditions maximum possible transmit level capability. Fig. 10 shows I_{LN} as a function of I_{LINE}. I_{LN} is about equal to the current in the (100Ω) resistor connected between SCR and VSS ($I_{LN} \approx I_{SCR}$).

The maximum possible swing on LN:

$$V_{LN}p\text{-}p = 2 \bullet (I_{LN} \bullet |Z_{LINE}|)$$

b. Because of the artificial inductor function of the DC voltage stabilizer, the AC signal on SLPE equals the AC signal on LN. This means that the available DC emitter current in the external pnp transistor PNP_{TX} (Fig. 6) equals the maximum possible AC current to modulate the current into the resistor between SLPE and VDD (R_{SUP}=250Ω in the basic application).
The maximum swing is:

$$V_{LN}p\text{-}p \approx V_{SLPE}p\text{-}p = 2 \bullet (I_{LINE} - I_{LN} - I_{DD} - I_P - I_{VP}) \bullet R_{SUP}$$

Limitation on voltage

c. The dynamic limiter threshold (programmable on either 3.5Vp-p or 2.3Vp-p) of course limits the sending signal level.

d. On SLPE the lower clipping level is determined by the DC voltage stabilizer. The lower clipping level (negative half-cycle of a 1kHz sinewave) of the AC signals at SLPE and TX are shown in Fig. 17 as a function of the DC current into pin TX. It is clear that the programmed DC voltage setting, the actual line current and the current gain of the pnp transistor therefore also have influence on the maximum possible swing on SLPE and because $V_{LN}p\text{-}p \approx V_{SLPE}p\text{-}p$ this has influence also on LN.

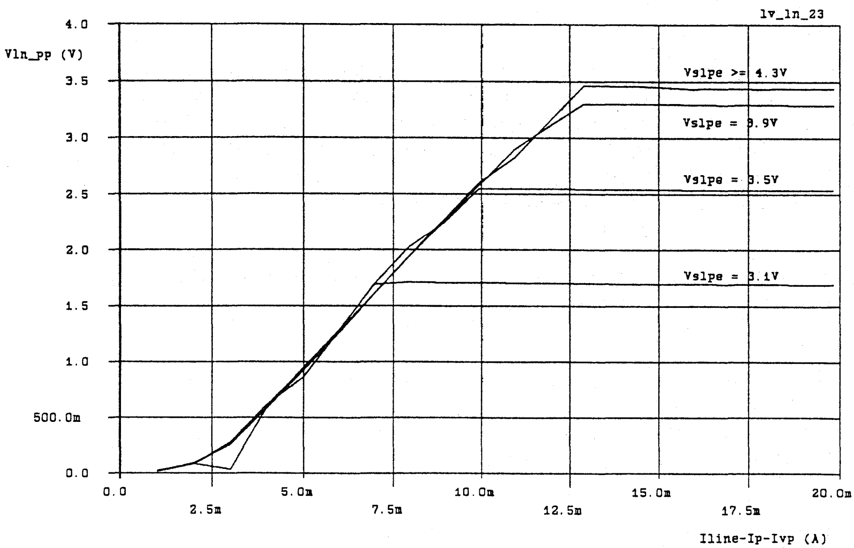


Figure 16 Maximum AC sending level on the line as a function of I_{LINE}-I_P-I_{VP}; parameter is the DC setting V_{SLPE}; DLT="0"

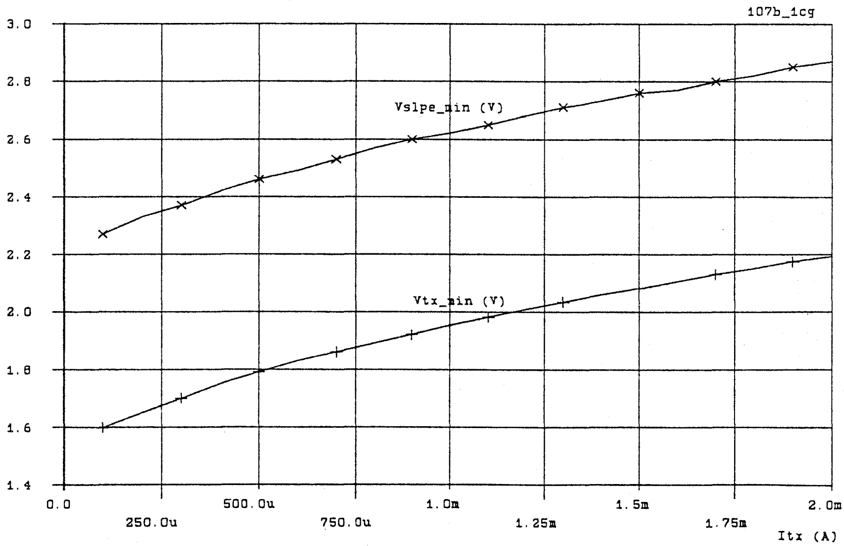


Figure 17 Lower clipping level for the AC signals on SLPE and TX as a function of DC current into pin TX

3.4 Set impedance

In normal conditions where $I_{LINE} >> I_{LN}+I_{DD}+I_P$ the static behaviour of the circuit (pin LN) is equivalent to a voltage regulator diode with a series resistor R_{SLPE} (fixed value 20Ω) (see paragraph 3.3). In the audio frequency range (300-3400Hz) the dynamic impedance Z_{LN} is determined mainly by internal components Zs=Ra+(Rb//C). The equivalent impedance Z_{LN} is shown in Fig. 18. The values of Ra, Rb and C can be programmed via I²C-interface. The value of C is not programmed directly but via the pole frequency: $C = 1/(2 \cdot \pi \cdot fp)$.

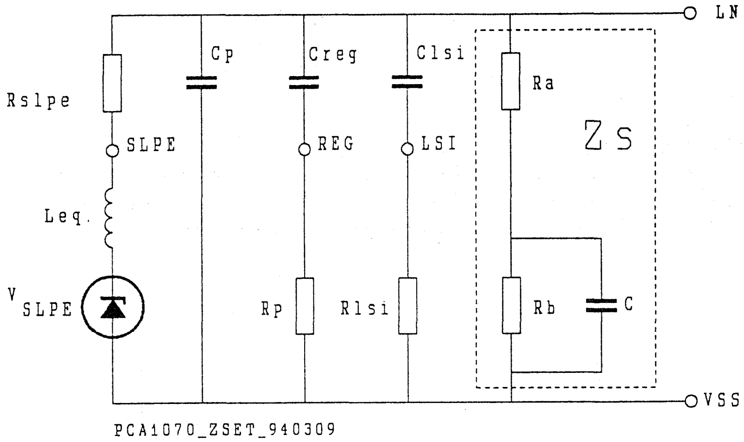


Figure 18 Equivalent impedance Z_{LN} between LN and VSS (f=300-3400Hz)

where

R_{SLPE}	=	resistor between LN and SLPE (20Ω)
Rp	=	internal resistor (typ. 1075k with V_{SLPE}=4.7V, DST=0)
R_{LSI}	=	internal resistor (typ. 240k)
C_{REG}	=	capacitor at pin REG (470nF)
Leq	=	$Rp \cdot R_{SLPE} \cdot C_{REG}$ = typ. 10.1H with V_{SLPE}=4.7V, DST=0 (artificial inductor)
C_{LSI}	=	capacitor at pin LSI (100nF)
Cp	=	internal capacitor (typ. 6nF)

Remarks:

- Rp changes when another DC voltage setting for V_{SLPE} is programmed. Rp increases/decreases with 100k for every step of 0.4V that V_{SLPE} increases/decreases.
- When the DST bit is set to 1 (e.g. for fast reprogramming of the DC voltage), the resistor Rp is switched to a low value (Rp= 26kΩ at V_{SLPE}=4.7V) and consequently the value of the artificial inductor decreases. This lowers the set impedance.

The set impedances as required by the PTTs of various countries differ a lot. As an example plots of the magnitude of the impedances of D (220Ω+[820Ω//115nF]), GB (370Ω+[620Ω//310nF]) are shown in Fig. 19 together with 600Ω which is required in many countries. An overview of the set impedance requirements of a number of European countries is given in Table 7 (Ref. [11]).

PCA1070 allows programming of an impedance which has values of Ra, Rb and C that come very close to the desired values as required by most PTTs of the world. This is sufficient to meet the BRL (Balance Return Loss) requirements. For example Fig. 20 shows the BRL of PCA1070 for the impedance setting according to the German Bundespost (D) (Ra=200Ω, Rb=800Ω, fp=1915Hz (C=104nF)). The required BRL is in this case:

300Hz-500Hz	BRL>14dB
500Hz-2500Hz	BRL>18dB
2500Hz-3400Hz	BRL>14dB

These requirements are met with quite a large margin.

For the basic application the country programming table is given in Table 8. The table is valid when the microphone and earpiece are replaced by their equivalent impedance. It must be kept in mind that when a handset is connected, the BRL may be influenced considerably because of the acoustic feedback path that exists between earpiece and microphone. This may result in slightly different settings for Zs to obtain optimum BRL.

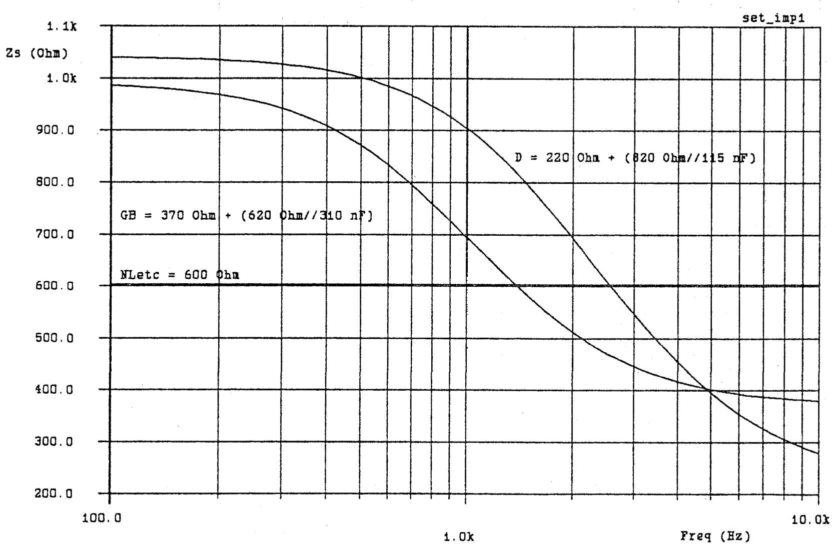


Figure 19 Impedances as required by some countries; D=Germany, GB=Great Britain, NL= Netherlands

Country	Ra (Ω)	Rb (Ω)	C (nF)
A,B,CY,E,IRL,IS, F,I,L,NL,P,SF	600	0	x)
B 1)	150	830	72
D,CH	220	820	115
GR	0	1000	100
N	120	820	110
GB	370	620	310
S	275	850	150
F (complex) 2)	210	880	150
DK	400	500	330

TABLE 7 Required set impedance (Ra+[Rb//C]) for a number of countries

Country	Ra (Ω)	Rb (Ω)	C (nF)	fp (Hz)
A,B,CY,E,IRL,IS, F,I,L,NL,P,SF	600	0	x)	x)
B 1)	100	800	79	2533
D,CH	200	800	104	1915
GR	0	1000	110	1448
N	100	800	104	1915
GB	400	600	320	828
S	300	800	182	1095
F (complex) 2)	200	900	161	1095
DK	400	600	320	828

TABLE 8 Set impedance: Country programming table PCA1070 in basic application (no tax pulse filter)

x) don't care
1) Only for PABX
2) According to CNET specification Ref.[14]

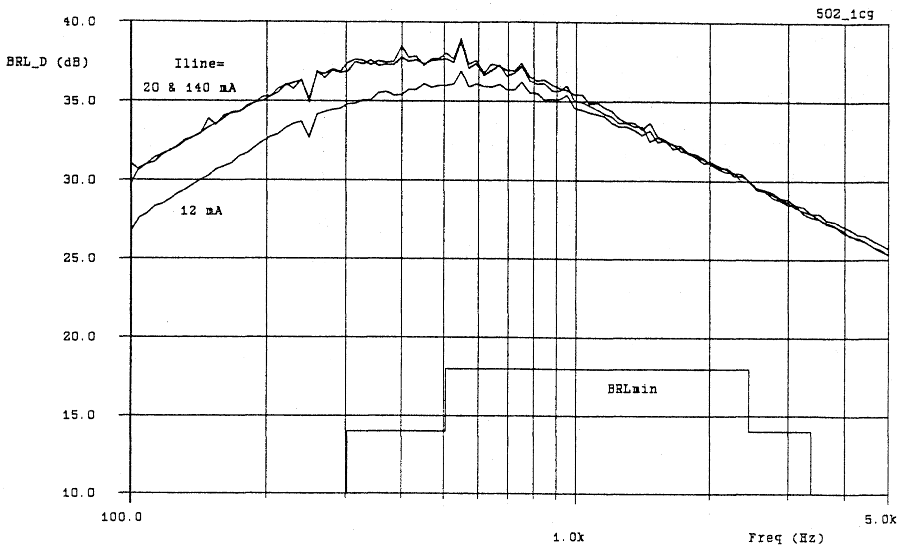


Figure 20 BRL of the impedance Z_{LN} against the German reference impedance Zref=220Ω+(820Ω//115nF) with programmed Ra=200Ω, Rb=800Ω, fp=1915Hz and I_{LINE} as a parameter

3.4.1 Tax pulse filter

In case a 12kHz or 16kHz tax pulse filter is connected between the line connections and pin LN, the total set impedance is of course affected. This is mainly caused by the capacitor used in the series LC filter that shunts the line. The overall set impedance can be corrected by reprogramming the impedance Zs of PCA1070. PCA1070 allows correction of the overall set impedance for capacitive loads between pins LN and VSS of up to about 33nF. It is recommended to use a capacitor value of <33nF in the series LC filter which is normally connected between the a/b lines.
More details about tax pulse filtering can be found in Chapter 6 and in Ref. [8].

3.4.2 EMC capacitor on LN

For EMC reasons it may be necessary to connect a capacitor between LN and VSS and/or between the a/b lines (practical values may be in the range 1nF - 4.7nF). Of course this also influences slightly the impedance of the telephone set at higher frequencies. Normally it is not necessary to correct the programmed Zs to meet the BRL requirements.

3.4.3 Set impedance without clock signal

In case no clock (3.58MHz) is applied to pin CLK the set impedance switches automatically to 600Ω. Programming of other impedances is not possible then.
In practice the oscillator of the μC may switch off under extremely low voltage conditions (e.g. parallel operation of sets). The PCA1070 will then still operate but with somewhat relaxed performance.

3.5 Microphone channel

The sending channels for microphone signals and DTMF signals are shown in Fig. 21. Microphone signals are (pre-)amplified with 20dB and are then passed via an I²C-controlled mute switch and a programmable amplifier (the send prog-amp) to the transmit output stage which is the AC part of the line interface block.
The transmit output stage (or AC line interface) is of the active type and takes care of the set impedance synthesis and modulation of the line. For transmit signals the equivalent circuit diagram consists of a voltage source (which has a 12dB higher signal level than the output signal of the send prog-amp) in series with the synthesised set impedance. DTMF signals are fed directly into the mute switch.
A dynamic limiter prevents distortion of the transmitted microphone signals by reducing the gain (speech controlled) of the mic. preamp.

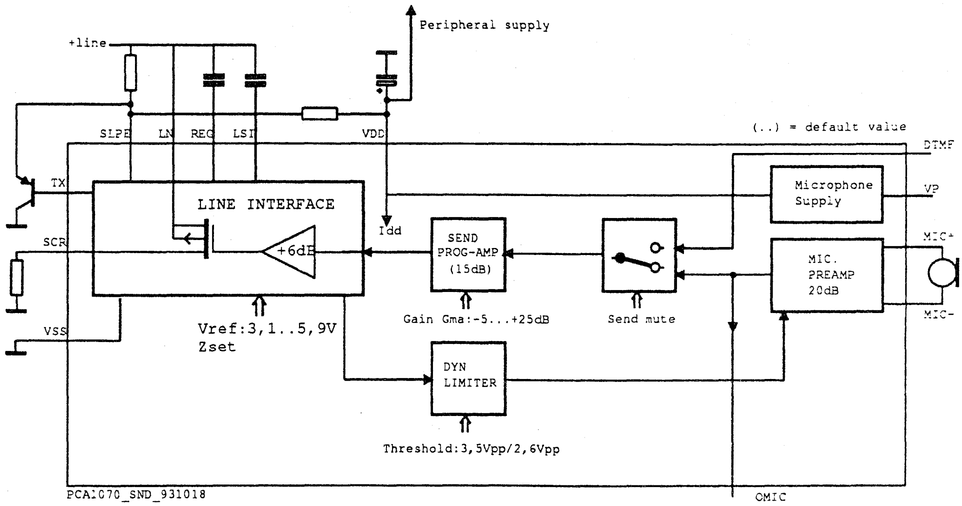


Figure 21 Sending Channel

3.5.1 Microphone inputs, connecting microphones

The PCA1070 has symmetrical microphone inputs. The input impedance is typically 100kΩ (2 x 50 KΩ) with a minimum of 60 kΩ. As the input impedance is rather high, this allows accurate matching of several microphone types by means of external components. The circuit is suitable for dynamic, magnetic, piezoelectric microphones with symmetrical drive. Electret microphones with built-in buffer of preamplifier may be connected in either asymmetrical mode or symmetrical mode. Some possible microphone arrangements are shown in Fig. 22.

Many electret microphones have a flat frequency characteristic. Therefore a bandpass filter may be needed to meet the PTT requirements w.r.t. frequency characteristics (acoustical in at microphone, electrical out at the telephone line). Fig. 23 shows such a bandpass filter suitable for use with electret microphones. A voltage limiting circuit at the microphone terminals may be used to prevent excessive overdrive (see paragraphs 3.5.2 and 3.6).

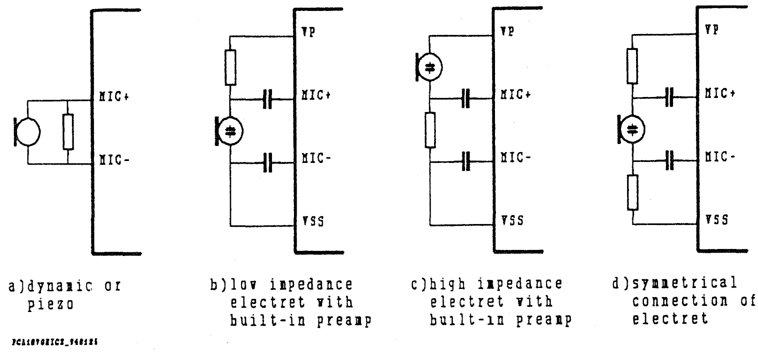


Figure 22 Some possible microphone arrangements

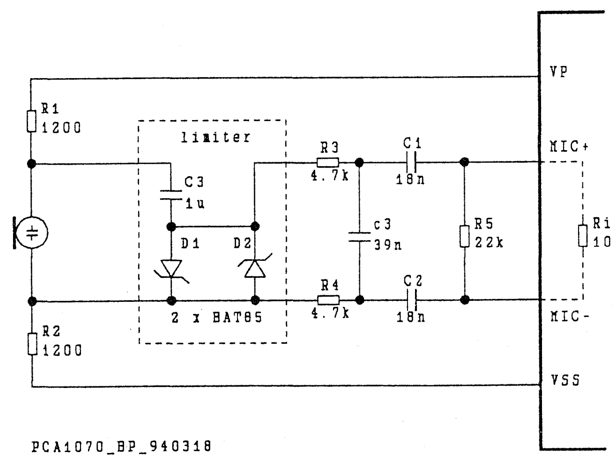


Figure 23 Example of a bandpass filter for electret microphones; the voltage limiting circuit is optional

Asymmetrical drive

In case asymmetrical drive of the microphone inputs is used, the MIC+ input should be used as a signal input. Care should be taken that both inputs MIC+ and MIC- see equal impedances to the common, otherwise residual line signals being present on the supply point VDD will cause inaccuracy in gain and sometimes (with a large DC-blocking capacitor connected to MIC+) even low-frequency hicking may occur.

For measuring purposes, in practice a low-ohmic (e.g. 50 Ohm) sinewave generator can be connected via a DC blocking capacitor to pin MIC+; a capacitor with the same value has to be connected between pin MIC- and VSS.

3.5.2 Maximum input signal

To prevent overdrive of the internal circuitry of PCA1070 after the microphone preamp the maximum allowed microphone input signal is limited.

The microphone preamplifiers accepts input signal up to 70mVpeak for THD=2% (with V_{DD}>=2.5V). The distortion versus signal level of the output signal of the microphone preamplifier (measured at pin OMIC) is shown in Fig. 24.

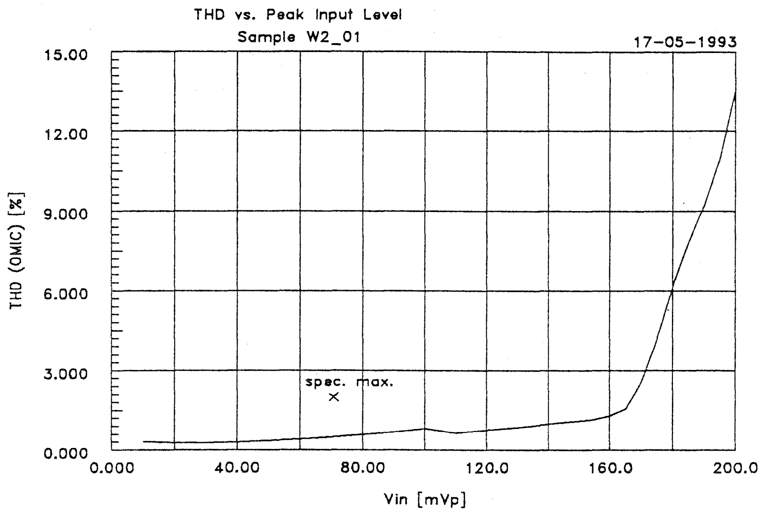


Figure 24 Distortion at the output OMIC of the microphone preamp: THD=f(Vin)

To prevent overdrive of the microphone channel, the minimum allowed gain setting is 30dB. At 30dB gain an input signal of 62mVp is needed to reach the maximum specified threshold voltage of the dynamic limiter at the LN (3.9Vp-p). An external attenuation network between microphone capsule and microphone inputs is required if a very sensitive type of microphone is used needing less than 30dB of gain.

Such an attenuation network may be combined with a bandpass filter which is normally needed in case an electret microphone with flat frequency response is used (Fig. 23)

In case the transmit output stage is overdriven due to lack of line current or due to voltage clipping, the dynamic limiter prevents distortion of the transmitted line signal.
The gain control range of this dynamic limiter is 12dB. This is sufficient for most microphone types. However in case a microphone with a high dynamic range is used, it may happen that peaks of the microphone signal overdrive the microphone channel in such a way that the 12dB gain control range of the dynamic limiter is exceeded. This means that the transmitted line signal will distort and this will lead to an annoying distortion of the sidetone signal.
For some types of electret microphones which are capable of generating extremely high output signals which exceed 400mVp-p, overdriving the dynamic limiter may be prevented by limiting the maximum possible input signal of the microphone amplifier. This can be done by using a simple limiting circuit consisting of 2 Schottky diodes and a capacitor in front of the bandpass filter (Fig. 23)
This solution may require some fine tuning w.r.t. attenuation of the bandpass filter and gain setting of PCA1070.

3.5.3 Output OMIC

The output of the microphone preamplifier (pin OMIC) is used as a testpin for production testing of PCA1070. In practice this pin may be used for applications with voice switched listening-in (e.g. with the Philips handsfree IC TEA1093; see Ref. [9]) and for answering machines.
Pin OMIC can deliver a swing of up to typically 2.4Vp-p (THD=2%; V_{SLPE}=4.7V) into a 20kΩ load.
Its output impedance is approximately 400Ω.

3.5.4 Gain setting

The total gain can be programmed between 30dB and 51dB in 1dB steps by means of a programmable amplifier ("send prog-amp") which can be controlled via the I^2C-bus.
The "send prog-amp" has an allowed gain range between 4 and 25dB.

Total gain of the microphone channel is

$$G_M = 32 + Gma + 20\log(Z_{LINE}/(Z_{LINE}+Zs)) \qquad [dB]$$

where Gma = Gain "send prog-amp"
Z_{LINE} = load impedance presented to LN
Zs = programmed set impedance

With the assumption that the load presented to the LN terminal of the IC is equal to the programmed set impedance:

$$G_M = 26 + Gma \qquad [dB]$$

Default $G_M = 26 + 15 = 41$ [dB]

In Fig. 25 the effect of gain settings between 30 and 51dB is shown. Fig. 26 shows the frequency characteristic with nominal settings.

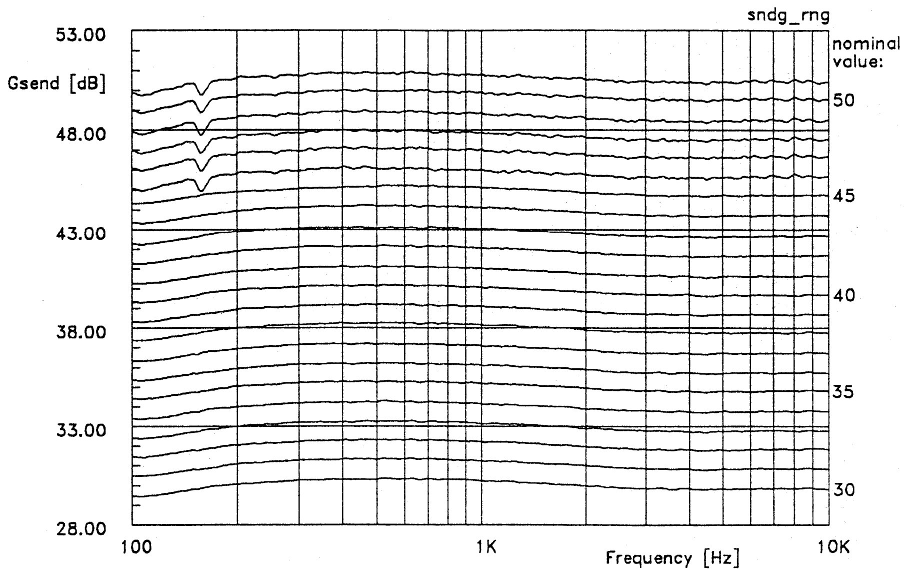


Figure 25 Gain setting range of microphone channel versus frequency; G_M=30-51dB

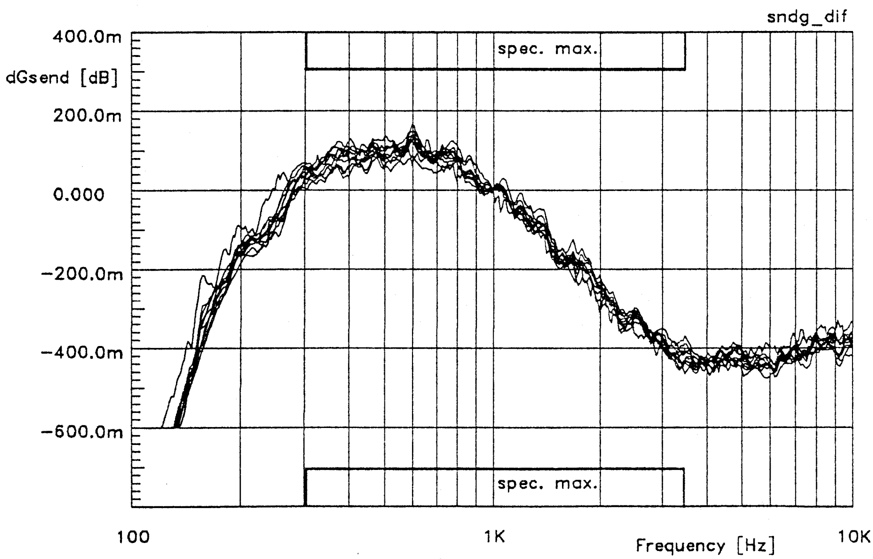


Figure 26 Frequency characteristic of the microphone channel; G_M=41dB

3.5.5 Noise

For optimum noise performance the microphone inputs must be loaded. Fig. 27 shows the psophometrically weighted (according to CCITT recommendation O41) noise voltage on the LN output as a function of overall gain of the microphone channel. The microphone inputs are loaded with 150Ω.

In the case of +3dB SLR and -9dB RLR (German requirement; Ref. [12]), and taking as a reference the Reference vocal level at 25mm (lips) which equals 89.3dBSPL (Ref. [7]), this means that in sending direction the line signal will be about 110mVrms.
From Fig. 27 it can be seen that to fulfil the German requirements for noise level on the line (<-75dBmp), the electrical gain must be set to a value lower than 40dB. This means that a microphone type must be selected that has a sensitivity of >= 1.1mVrms/89.3dBSPL or 1.9mVrms/Pa. With 40dB electrical gain a noise level of -75dBmp can be expected and with +3dB SLR a S/N of 58dB at reference vocal level is obtained.
A more sensitive microphone type will allow a lower gain setting resulting in less electrical noise on the line and an even better S/N.

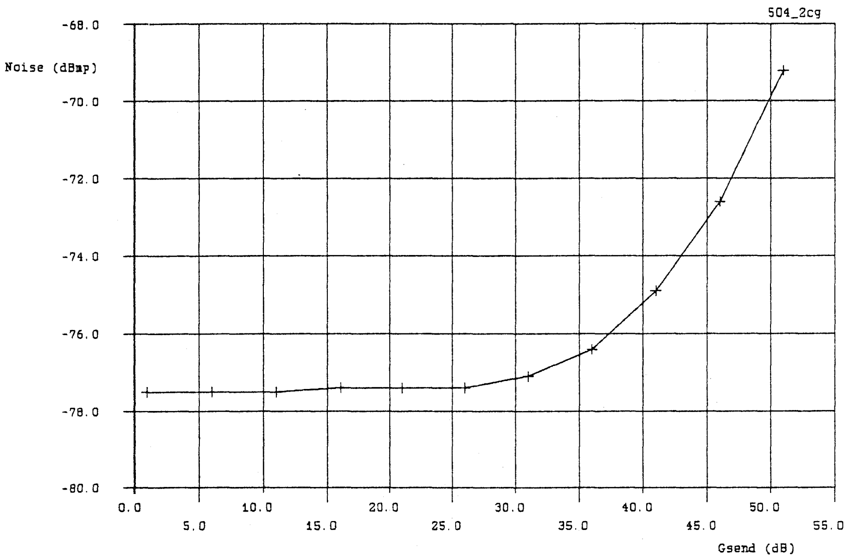


Figure 27 Psophometrically weighted noise on the transmitter output LN as a function of microphone channel gain

3.5.6 Parallel operation

In the case of parallel operation of sets, the parallel set takes a large portion of the line current such that the PCA1070 may enter the low current range (<6mA typically; see paragraph 3.3.1) and the DC voltage is automatically adjusted to a lower value. Of course this will influence the performance of the circuit.

Fig. 28 shows the output swing capabilities of the circuit as a function of line current in the case that a telephone set with impedance 220Ω+(820Ω//115nF) is connected in parallel with the circuit in Fig. 5. At I_{LINE}<16mA typically, the transmit output swing is limited by current (I_{LN}) when the DC setting is programmed >= 3.5V. The flat part of the curve with V_{SLPE}=3.1V shows voltage limiting in the DC line interface.

Fig 29 shows the transmit gain as a function of DC voltages V_{LN}, V_{DD} in the low line current range.

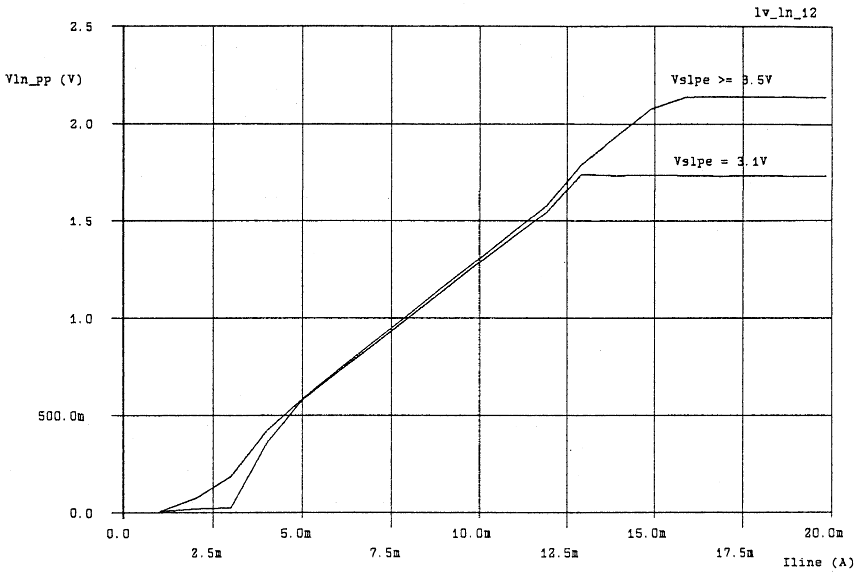


Figure 28 Maximum output swing of the transmitting output stage versus I_{LINE} in the low line current range (a set with impedance 220Ω+[820Ω//115nF] connected in parallel); I_P=I_{VP}=0

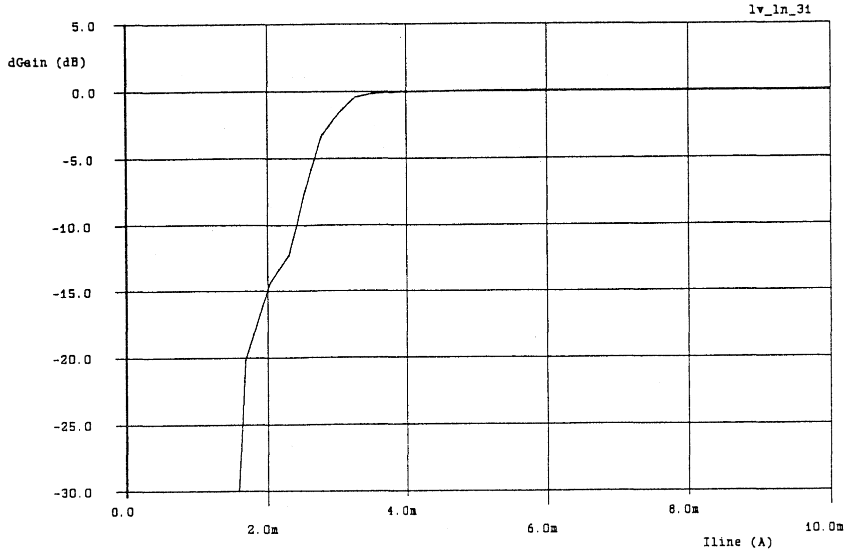


Figure 29 Transmit gain versus I_{LINE} in low line current range (set with impedance 220Ω+[820Ω//115nF] connected in parallel)

3.6 Dynamic limiter

The maximum signal level at the line output LN is limited by the DC voltage drop across the IC and/or by the current in the transmit output stage (AC line interface) or the current in the DC voltage stabilizer (DC line interface). For a description of the mechanisms see paragraph 3.3.4.
To prevent distortion (clipping) of the transmitted signal due to these mechanisms a dynamic limiter is incorporated. Another reason for using a dynamic limiter is that some PTTs (Ref. [11]) impose limitations w.r.t. the highest possible transmit level e.g. in Germany V_{LINE}<6.3dBm(600Ω) (=1.6Vrms=4.5Vp-p). In France the maximum allowed line level is +3dBm[600Ω] (=1.1Vrms=3.1Vp-p).
Dynamic limiting also improves the sidetone performance in overdrive conditions (less distortion and limited sidetone level).

3.6.1 Attack and release times

When peaks of the transmitted signal on the line exceed an internally determined threshold, the gain of the microphone preamplifier is reduced rapidly. The time in which gain reduction is affected (the attack time) is rather short (See Fig. 30). The circuit stays in gain-reduced condition until the peaks of the transmitted signal remain below the threshold level. The sending gain then returns to normal after a time determined on the chip (the release time; see Fig. 31).

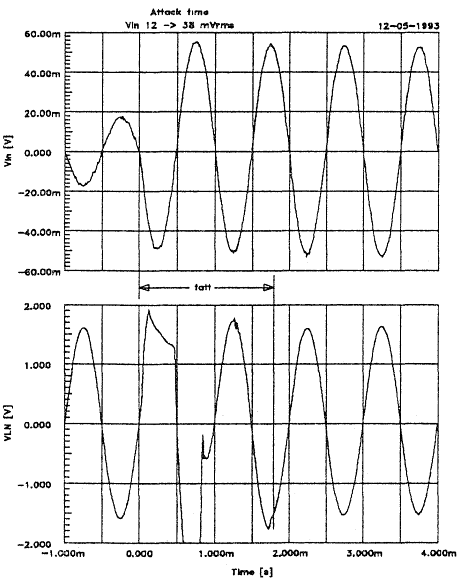


Figure 30 Dynamic behaviour of the dynamic limiter: Attack Time (Vin=12 to 38mVrms).

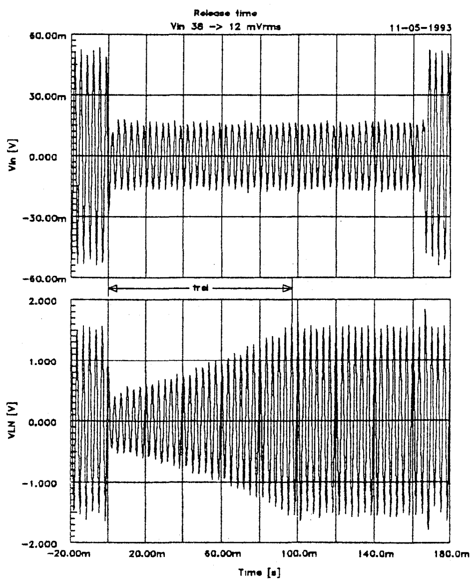


Figure 31 Dynamic behaviour of the dynamic limiter: Release Time (Vin=38 to 12mVrms)

3.6.2 Threshold, maximum sending level

The threshold levels can be selected via I^2C-bus by bitcode DLT; if DLT = "0" the AC peak-to-peak line voltage on pin LN is default typ. 3.5V(p-p) and if DLT = "1" the level is typ. 2.6V(p-p).
Fig. 32 shows the sending signal on LN as a function of the signal level at the microphone inputs.

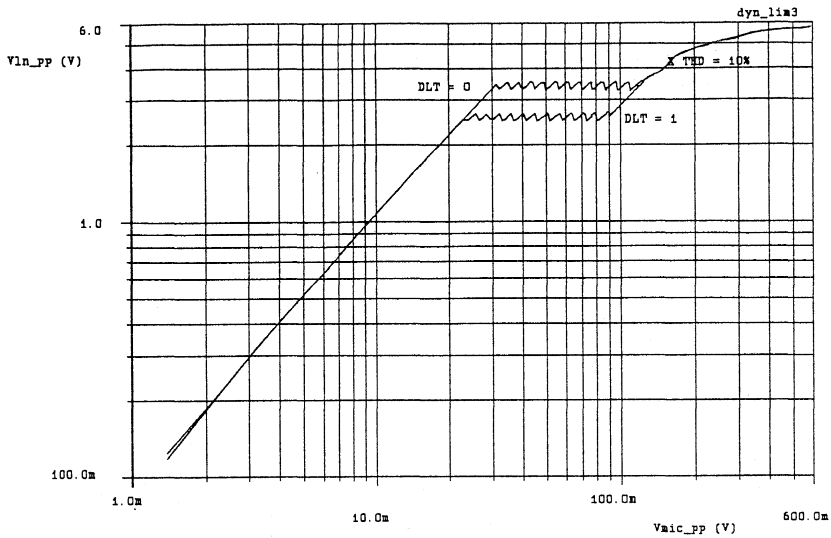


Figure 32 Static behaviour of the dynamic limiter; V_{LN}p-p=f(V_{MIC}); DLT="0", "1"

In case the programmed DC voltage of the circuit is not high enough to reach the programmed threshold level, the internal detector threshold adapts automatically to a lower value.
Also in the case that the bias current in the transmit output stage (I_{LN}) is not sufficient to modulate the line load or when the current in the DC voltage stabilizer is not sufficient to modulate the current in the supply resistor between SLPE and VDD, the threshold is adapted automatically to a lower value.

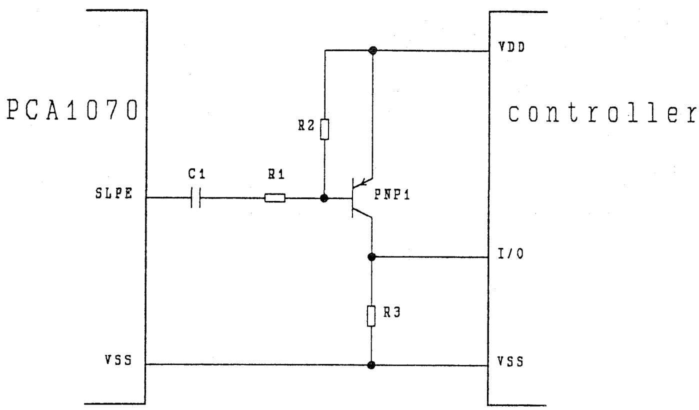
Fig. 16 in paragraph 3.3.4 shows the maximum output swing versus the effectively available line current (I_{LINE}-I_P-I_{VP}) with the programmed DC voltage V_{SLPE} as a parameter and DLT="0".

The gain control is done in 1dB steps. Because of the fast attack time and slow release time, the mean peak to peak output level on the line is slightly below the specified threshold value. The gain control range is typically 12dB. During extreme overdrive conditions where the microphone input signal is increased such that the 12dB control range is exceeded, the line signal will increase again and voltage or current clipping will occur according to the mechanisms as described in paragraph 3.3.4 or due to overdrive of internal circuitry.
This may happen when a microphone is used with an extremely high dynamic range. This means that the transmitted line signal will distort and this will lead to an annoying distortion of the sidetone signal (see paragraph 3.9).

One way of preventing this is to limit the maximum possible input signal of the microphone amplifier. This is described in paragraph 3.5.2. A second possibility is to extend the gain control range of the dynamic limiter under SW control as is described in the next paragraph (3.6.3).

3.6.3 Software controlled dynamic limiter

A µC may be used to extend the control range of the dynamic limiter considerably. A small external circuit is needed then to detect overdrive of the transmit output LN. An example of such a detector is shown in Fig. 33. Typical component values may be R1=11kΩ to 35kΩ, R2=18kΩ, R3=100kΩ, C1=100nF, PNP1=BC558.



PCA1070SVDL_940318

Figure 33 Example of a detector for SW controlled dynamic limiter

The threshold of the detector may be adjusted with the resistive divider R1, R2 and is determined by the junction voltage Vbe of the pnp transistor.
In formula:

$$V_{LN}p\text{-}p = 2 \bullet (R1+R2)/R2 \bullet Vbe$$

The mean level of the transmit signal on LN will be slightly lower than the threshold because the gain control is done stepwise. For example with R2=18kΩ, R3=100kΩ, C1=100nF, PNP1=BC558 and R1=35kΩ the mean transmit output level on LN would be V_{LN}p-p=3.4V. With R1=24kΩ this will result in V_{LN}p-p=2.9V.

Note that the threshold is independent of the programmed DC voltage V_{SLPE} of the PCA1070. However at low DC voltage settings, voltage clipping may occur in the line interface. Therefore the threshold must be chosen to be sufficiently low to prevent clipping in the line interface (see paragraph 3.3.4 Fig. 17).
The output of the detector gives a '1' to one of the I/O ports of the µC during the times when the adjusted threshold is exceeded. The µC detects this and by means of SW the gain of the send channel can be reduced by reprogramming the gain of the send prog-amp via I^2C-bus. The attack and release times can be set by SW. Of course the I^2C-bus imposes restrictions w.r.t. the control speed. In practice when the µC uses a SW controlled I^2C-bus protocol (10kHz), an attack time of 3dB/3ms and a release time of 1dB/20ms is feasible.
The gain control range is in principle limited by the setting range of the send prog-amp, the nominal gain setting needed for the specific microphone type and the maximum allowed input signal of the microphone preamp, the send mute switch and the send prog-amp.

An extra advantage of the SW controlled dynamic limiter is that is works also for signals applied via the DTMF input.

3.7 DTMF channel

The channels for DTMF signals are shown in Fig. 34. DTMF signals are supplied via an on-chip DC blocking capacitor into the I²C-controlled send mute switch and a programmable amplifier (the send prog-amp) to the transmit output stage which is the AC part of the line interface block.
The DTMF signals are also fed into the receiving channel via a 25dB attenuator. They then enter via the I²C-controlled receive mute switch and a programmable amplifier (the receive prog-amp) into the receive output stage which drives the earpiece. In this way the DTMF signal can heard in the earpiece (confidence tone).

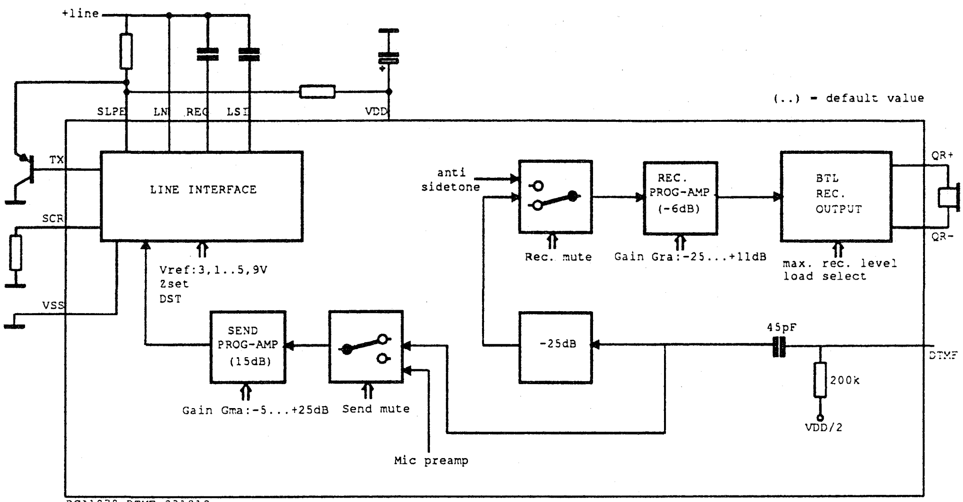


Figure 34 DTMF Channel

3.7.1 DTMF input

The PCA1070 has an asymmetrical DTMF input. Its input impedance is typ. 200 kΩ//45pF. A DC blocking capacitor is integrated onto PCA1070. For EMC reasons, the DTMF input is connected internally via a high-ohmic resistor (typ. 200kOhm) to the bias voltage $V_{DD}/2$. The DTMF input can sink/source currents up to typ. 10 µA.
In applications where the DTMF generator or the µC with an on-chip DTMF generator is supplied directly or via a (Schottky) diode) from VDD of the PCA1070, the DTMF input can be coupled directly to the TONE output of the DTMF generator without using a DC blocking capacitor.

The DTMF input is enabled by setting the Send Mute control bit (SM) to "1" via I²C-bus.

3.7.2 Maximum input signal

To prevent overdrive of the internal circuitry of PCA1070 the maximum input signal is limited.
The DTMF input accepts signals up to 700mVp for THD=2% (with V_{DD}>=2.5V).
To prevent overdrive of the DTMF channel, the minimum allowed gain setting is +1dB. This means that the maximum allowed line signal is then about 1.6Vp-p (=-3dBm[600 Ohm]) which is sufficient to meet the PTT requirements for normal DTMF levels on the line.

In case the DTMF input is used for other purposes (e.g. to amplify the speech signal from the base microphone of a handsfree telephone set or to amplify a recorded message from an answering machine), the maximum line signal that may occur will be higher than for DTMF. This limits the minimum allowed gain setting. For an output swing of 3.5Vp-p at LN and a maximum input signal of 700mVp, a minimum gain setting of +8dB is recommended.

The dynamic limiter is inoperative for the DTMF channel and therefore the maximum sending level is determined by the normal clipping mechanisms as described in paragraph 3.3.4 (a, b and d).

If a dynamic limiter is needed for the DTMF channel, the SW controlled dynamic limiter may be used. See paragraph 3.6.3 for details.

3.7.3 Gain setting

The total gain can be programmed in 1dB steps by means of a programmable amplifier ("send prog-amp") which can be controlled via the I²C-bus.
In normal DTMF mode the normal gain setting range of the "send prog-amp" is between -5 and +15dB. This results in an overall DTMF gain setting range of +1 to +21dB.
Maximum allowed line signal is in this case -3dBm[600 Ohm].

In the case that the DTMF channel is used to amplify speech signals (e.g. in combination with the handsfree circuit TEA1093), the recommended gain setting range is +8 to +31dB (Gma=+2 to +25dB).

Total gain of the DTMF channel is

$$G_{DTMF} = 12 + Gma + 20\log(Z_{LINE}/(Z_{LINE}+Zs)) \quad [dB]$$

where Gma = Gain "send prog-amp"
Z_{LINE} = Load impedance presented to LN
Zs = Programmed set impedance

With the assumption that the load presented to the LN terminal of the IC is equal to the programmed set impedance:

$$G_{DTMF} = 6 + Gma \quad [dB]$$

Default $G_{DTMF} = 6 + 15 = 21$ [dB]

In Fig. 35 the frequency characteristic with a gain setting of 7dB (Gma=+1dB) is shown.

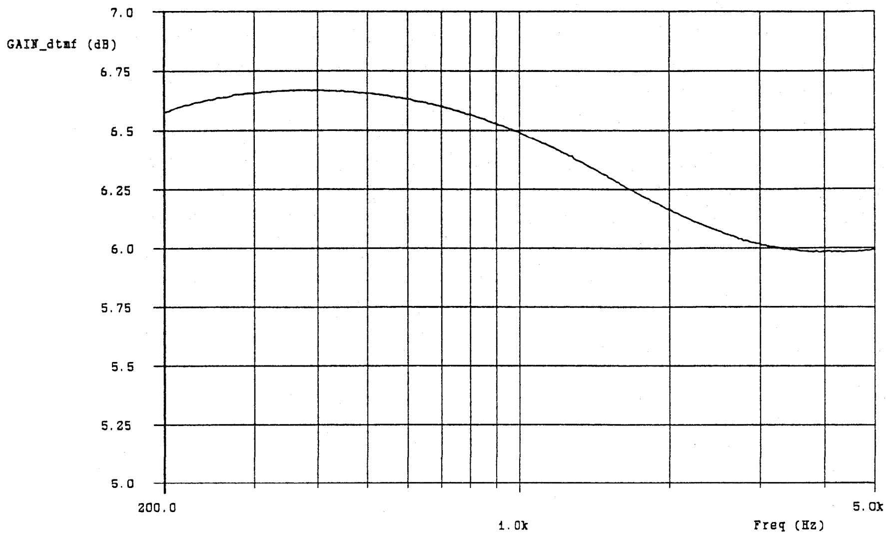


Figure 35 Frequency characteristic DTMF channel (DTMF to LN), Gma=+1dB

3.7.4 Confidence tone

The confidence tone gain (between DTMF input and earpiece outputs QR) can be programmed between approximately -40dB and -19dB (symmetrical drive of earpiece) by setting the gain (Gra) of the "rec-prog-amp" (recommended setting range Gra:-25 to 0dB).

The confidence tone gain (DTMF to QR outputs) is:

With symmetrical drive (=BTL) of earpiece

	G_{CTs}	= Gra	-	19		[dB]
Default	G_{CTs}	= -6	-	19	= -25	[dB]

Where Gra = Gain of the "rec. prog-amp".

Fig. 36 shows the frequency characteristic with a gain setting of Gra=-25dB.
Fig. 37 shows the confidence tone gain versus the programmed gain of the receive prog-amp. At low gain settings (Gra<-10dB), the confidence tone gain will be slightly higher than the calculated value. This is caused by a residual signal.

The confidence tone channel is enabled by setting the Receive Mute control bit (RM) to "1" via I^2C-bus.

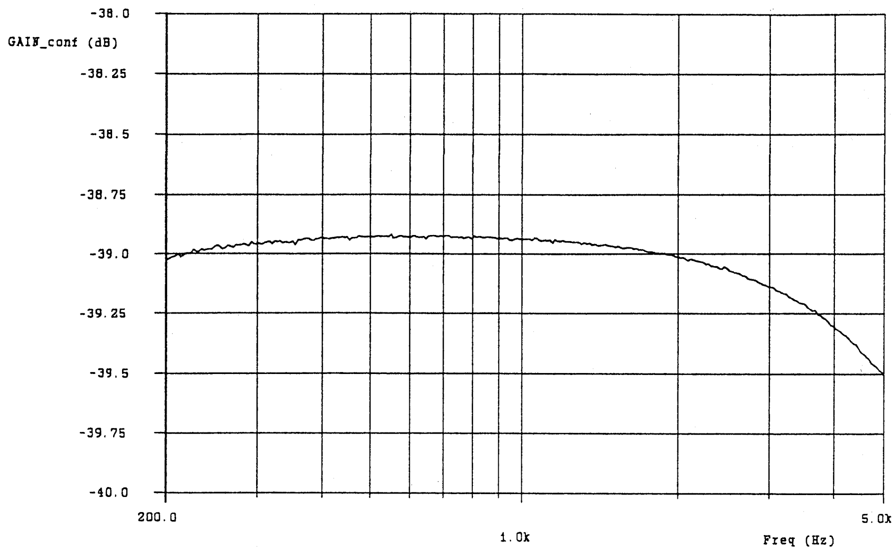


Figure 36 Frequency characteristic confidence tone (DTMF to QR+,QR-), Gra=-25dB

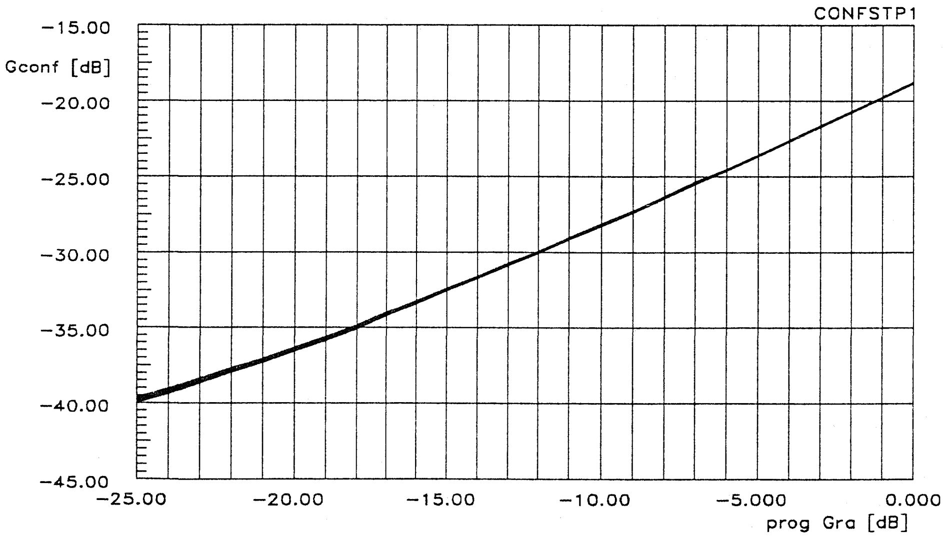


Figure 37 Gain adjustment range confidence tone (DTMF to QR+,QR-)

3.8 Receive channel

The receive channel is shown in Fig. 38. The transmit output stage (AC line interface) which is a part of the line interface block synthesises the correct set impedance. The line signals consisting of receive signals superimposed with sending signals, enter the LSI input via a DC blocking capacitor and are then attenuated in the AC line interface with 6dB. The output signal of the AC line interface is then passed via the anti sidetone block to remove (a large portion of) the sending signal. The output signal from the anti sidetone block is the receive signal (plus the remainder of the sending signal: the sidetone) and is then passed via the I²C-controlled receive mute switch and a programmable amplifier (the receive prog-amp) into the receive output stage which drives an earpiece in either BTL or SEL mode.

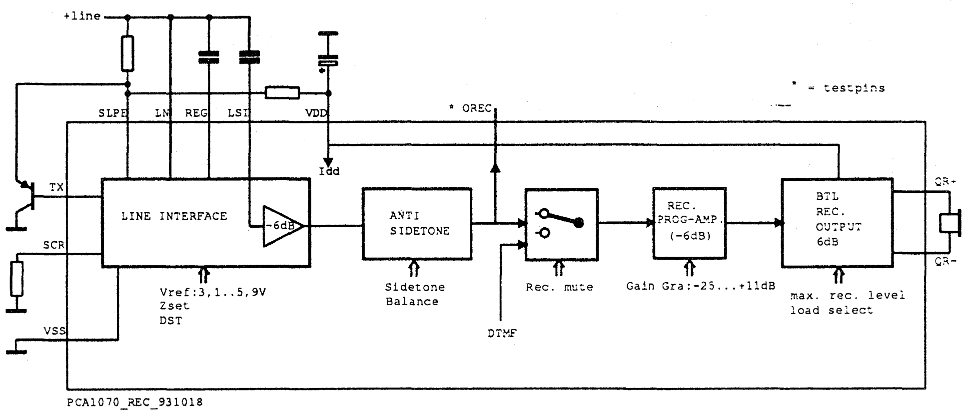


Figure 38 Receive Channel

3.8.1 Line input, max input signal

The AC line interface takes care of correct set impedance in sending directing as well as in receiving direction. This means that also for receiving signals on LN the transmit output stage is modulated and therefore the current in the 100Ω reference resistor at pin SCR is modulated.
The DC line interface (voltage stabilizer) takes care of correct DC setting and keeps the DC current in R_{SLPE}, the 20Ω resistor between LN and SLPE, constant (artificial inductor). So the AC signal being present on LN is also on SLPE and therefore the current flowing in the supply resistor between SLPE and VDD (R_{SUP}) must be modulated by the pnp transistor which sinks the excess current from pin SLPE and is driven by output TX.

From the previously mentioned it may be clear that the maximum acceptable line signal is determined by several mechanisms which are in principle the same as for the maximum sending level as described in paragraph 3.3.4 (mechanisms a,b,d). Of course the dynamic limiter is ineffective for receiving signals.

The AC line signal is coupled via a DC blocking capacitor into input LSI. The input structure of LSI consists of a 240kΩ resistor which is connected to the inverting input of an amplifier Pin LSI is biased at a DC voltage of $V_{DD}/2$. A further limitation of the maximum line signal is therefore imposed by the LSI input which gives clipping on voltage in case of extremely low V_{DD} because of its on-chip ESD protection diode which is connected between LSI and VSS (anode to VSS). This limits the maximum swing to $V_{LN}p <= ½V_{DD} + 0.7$ V.

The maximum receive signal on the line pin LN (for THD=2% at QR+,QR-) as a function of V_{DD} and with the DC voltage V_{SLPE} as a parameter is shown in Fig. 39.

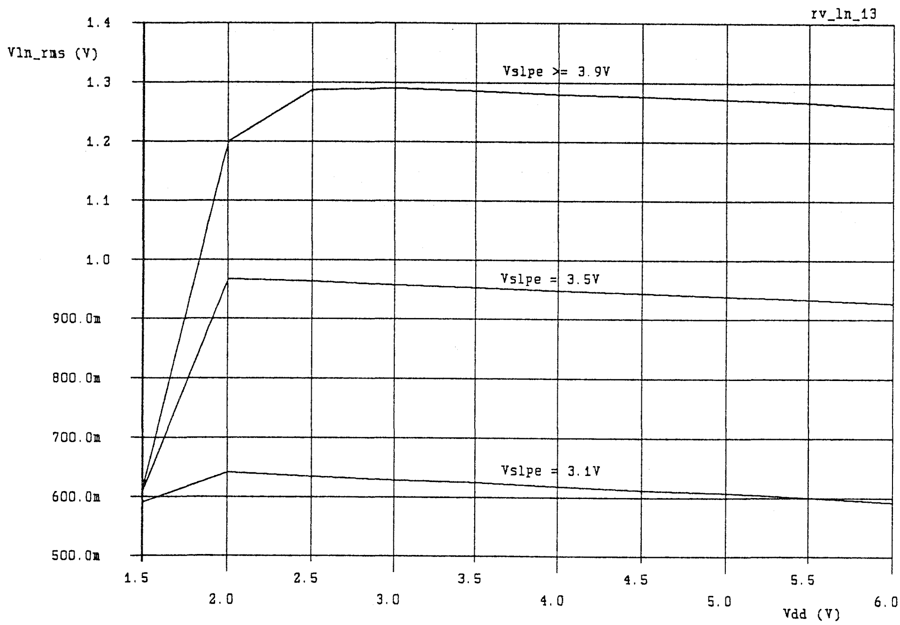


Figure 39 Maximum receive input signal on the line input LN versus DC setting V_{SLPE}; THD=2% at QR+, QR-, Rt=∞, I_{LINE}=30mA, external V_{DD}

3.8.2 Receive output OREC

The output signal of the anti sidetone block is available at pin OREC and is used for production testing of PCA1070. In practice this pin may be used for applications with handsfree, (voice switched) listening-in (e.g. with the Philips handsfree IC TEA1093; see Ref. [9]) or line monitoring (e.g. with TEA1085) and for answering machines.
Pin OREC can deliver a swing of up to 1.4Vp-p (THD=2%; V_{SLPE}=4.7V) into a 20kΩ load.
Its output impedance is approximately 1kΩ.
The output signal is superimposed by the residual signals of the internal clock(s) (224kHz). These signals may be removed by passing the OREC signal through a suitable low pass filter before further processing.

3.8.3 Receive outputs, connecting earpieces

The outputs QR+, QR- may be used to drive dynamic, magnetic or piezoelectric earpieces with single ended load (SEL) or symmetrical load (BTL). Earpiece arrangements are shown in Fig. 40.
The load select bit RFC is default "1" to guarantee stable operation in case of a (heavy) capacitive load (piezoelectric earpiece) or in unloaded condition. With a resistive load (dynamic capsule) or a moderate capacitive load (piezoelectric earpiece up to 80nF) it is recommended to set RFC to "0" via I^{2}C-interface to obtain optimum performance w.r.t. distortion and bandwidth.

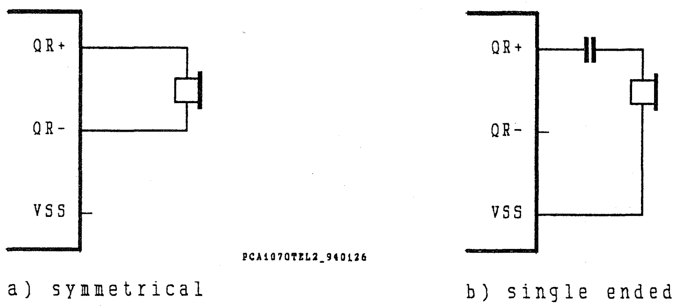


Figure 40 Earpiece arrangements

3.8.4 Gain setting

The gain of the receive channel is defined between the line connection LN and the earpiece outputs QR+, QR-.
The gain of the receive channel can be programmed between -19dB and +11dB (BTL or symmetrical drive) in 1dB steps by means of a programmable amplifier "rec prog-amp" which can be controlled via the I^2C-bus (recommended range Gra:-19dB to +11dB).

Total gain of the receiving channel is:
Symmetrical drive (is the same as BTL)

	G_{RS}	= Gra				[dB]
Default	G_{RS}	= -6				[dB]

Asymmetrical or single ended drive (SEL)

	G_{RA}	= Gra -	6			[dB]
Default	G_{RA}	= -6 -	6	=	-12	[dB]

Where Gra = Gain of the "rec prog-amp"

In Fig. 41 the effect of a number of gain settings between -19dB and +11dB is shown. Fig. 42 shows the frequency characteristics of 10 samples.

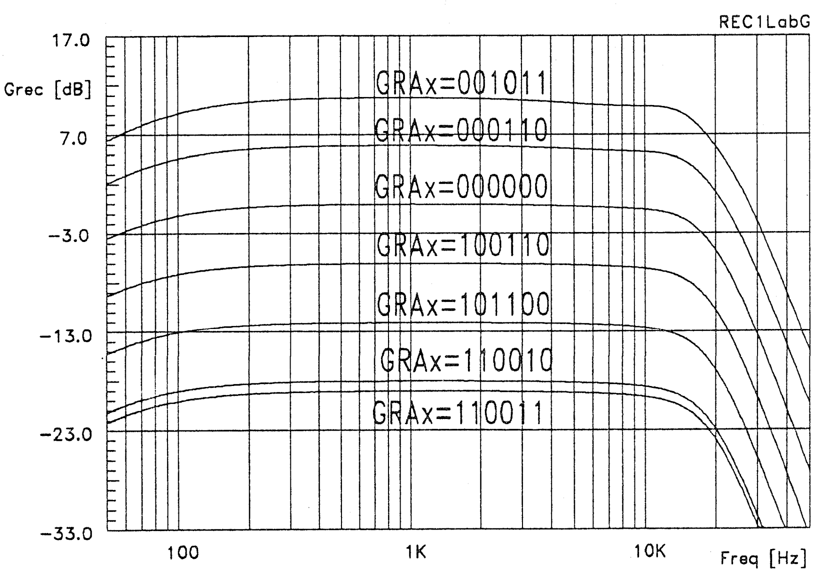


Figure 41 Gain adjustment range receive channel, at G_{RS}=-19- to +11dB (BTL)

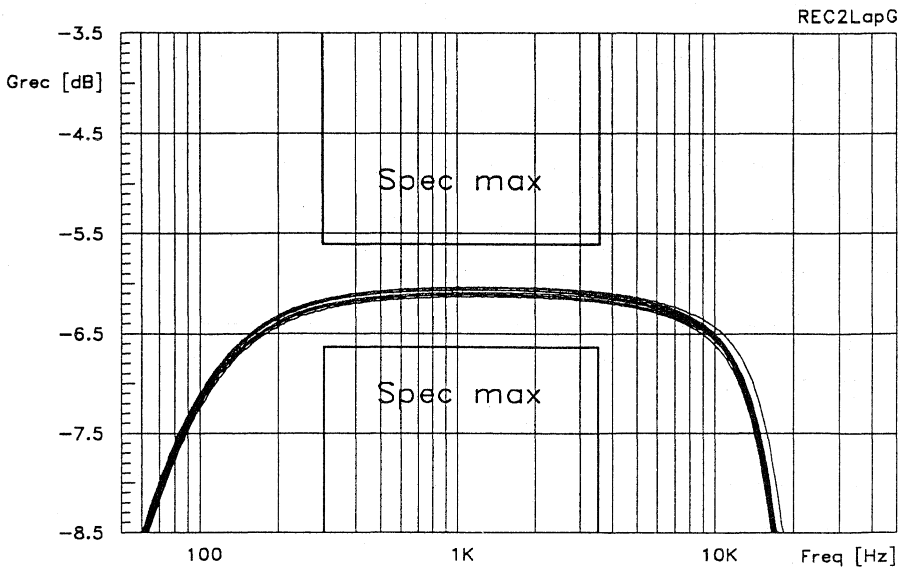


Figure 42 Frequency characteristic receive channel at G_{RS}=-6dB

3.8.5 Maximum output signal

The maximum output swing of the receive output stage depends on the impedance of the earpiece and the DC voltage drop across the circuit.
The maximum output swing versus the supply voltage V_{DD} in the case of 150Ω BTL load and with 450Ω BTL load is shown in Fig. 43. With a capacitive load of 80nF Fig. 44 applies.

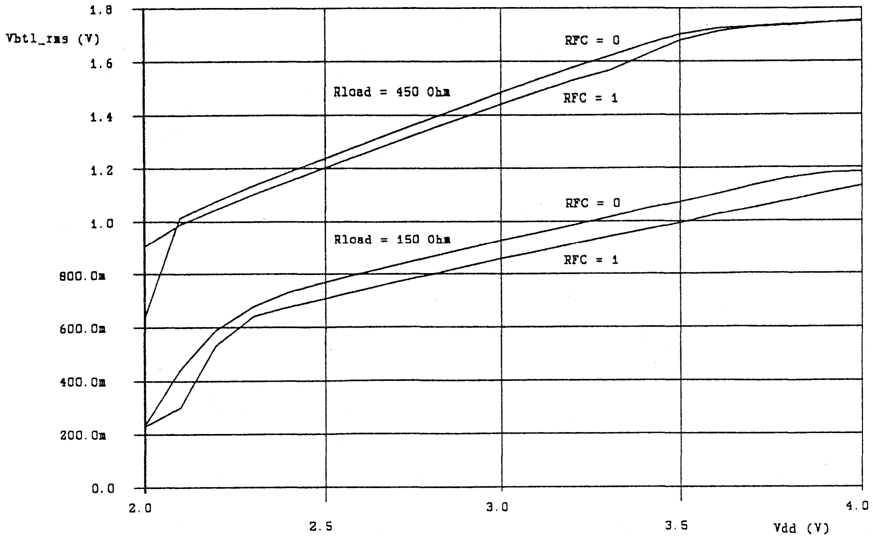


Figure 43 Maximum output swing of the receive amplifier for THD=2% as a function of V_{DD}; Rt=150Ω and 450Ω (BTL); G_{RS}=+3dB; RFC= 0,1; HPL=1

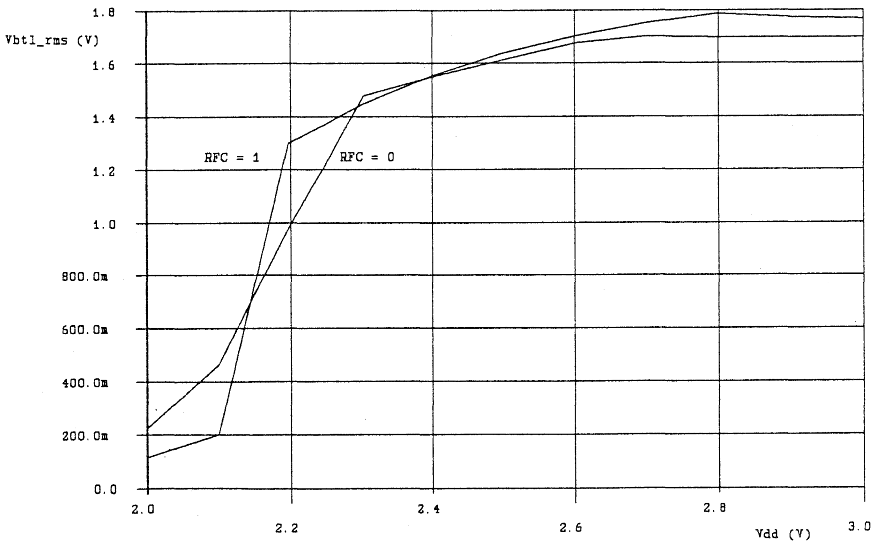


Figure 44 Maximum output swing receive amplifier for THD=2% versus V_{DD}; Ct=80nF (BTL); HPL=1; G_{RS}=+3dB; RFC= 0,1

3.8.6 Hearing protection

On-chip hearing protection circuitry is available. This eliminates the need for external protection devices and thus saves external components.
Two levels for hearing protection can be selected via I²C-interface with control bit HPL.
If HPL="0" the output level is limited to typ. 2.3 Vp-p. This is shown in Fig. 45. When HPL="1" the output level is limited to typ. 5.9 Vp-p (See Fig. 46).

The lower level is intended for sensitive earpieces such as dynamic or magnetic. The higher level is intended for use with insensitive types of capsules such as piezoelectric earpieces.

The electrical hearing protection levels can be adapted to the electrical/acoustical characteristics of a particular earpiece, by connecting a series resistor between the earpiece outputs and the earpiece. In this way the output signal at the QR+, QR- outputs can be attenuated as much as needed. Of course the overall gain between line and the earpiece must be kept constant and this means that the attenuation which is introduced by the series resistor must be corrected by increasing the gain of the receive prog-amp.

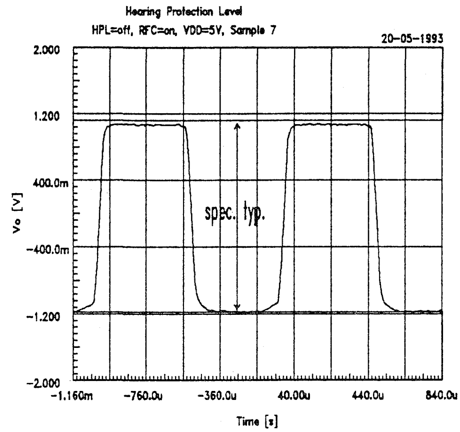


Figure 45 Hearing protection: output voltage for HPL=0

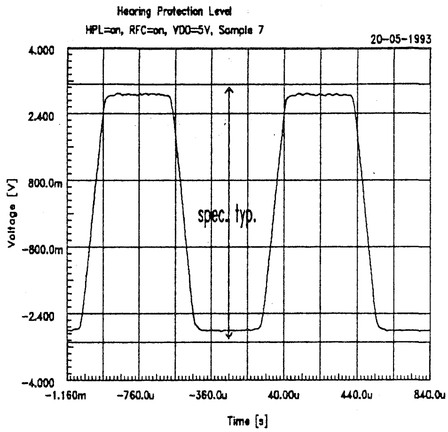


Figure 46 Hearing protection: output voltage for HPL=1

3.8.7 Noise

Fig. 47 shows the psophometrically weighted (according to CCITT recommendation O41) noise voltage across the earpiece (with microphone and earpiece replaced by Rm=Rt=150Ω and Rm=Rt=82nF) as a function of receive gain. The acoustical noise level produced by the earpiece of course depends on the sensitivity of the particular earpiece type which is used.

For gain settings above roughly -15dB the noise level at the output is linearly proportional to the gain and may be represented by one equivalent noise voltage source at the line of about -76dBmp[600Ω].

In the case of +3dB SLR and -9dB RLR (German requirement), and taking as a reference the Reference vocal level at 25mm (lips) which equals 89.3dBSPL (Ref. [7]), this means that in sending direction the line signal will be about 110mVrms.
With 110mVrms (=-17dBm[600Ω]) as a reference signal on the line and RLR=-9dB, the acoustical level produced by the earpiece will be about 95.3dBSPL.
So a noise level on the line of -76dBmp would result in 36.3dBSPL acoustical level produced by the earpiece. This is a S/N of about 59dB. Of course the acoustical level in dBA depends on the electrical frequency characteristics of the receiving channel and the electrical to acoustical transfer characteristic of the particular earpiece that is used.

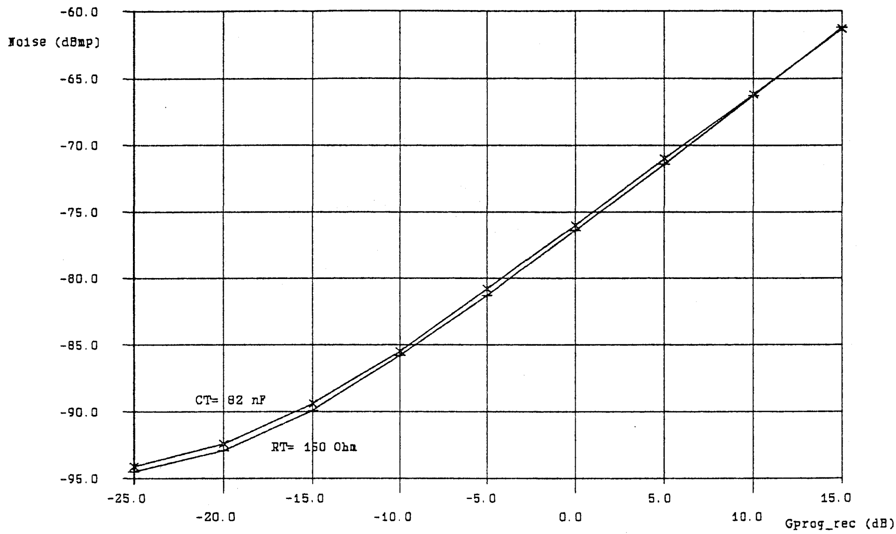


Figure 47 Psophometrically weighted noise at the receive outputs versus receive gain

3.8.8 Parallel operation

In the case of parallel operation of sets, the parallel set may take a large portion of the line current such that the PCA1070 may enter the low current range (<6mA typically ; see paragraph 3.3.1) and the DC voltage V_{SLPE} is automatically adjusted to a lower value. Consequently V_{DD} will follow and stabilize to a lower value which depends mainly on I_P, I_{VP}, I_{DD}.
Of course the performance of the receive channel will be affected by this.
Fig. 48 shows the output swing capabilities as a function of I_{LINE} which is valid in the low line current range I_{LINE}= 2 to 20mA.
Fig 49 shows the receive gain versus supply voltage V_{DD} in the low current range .

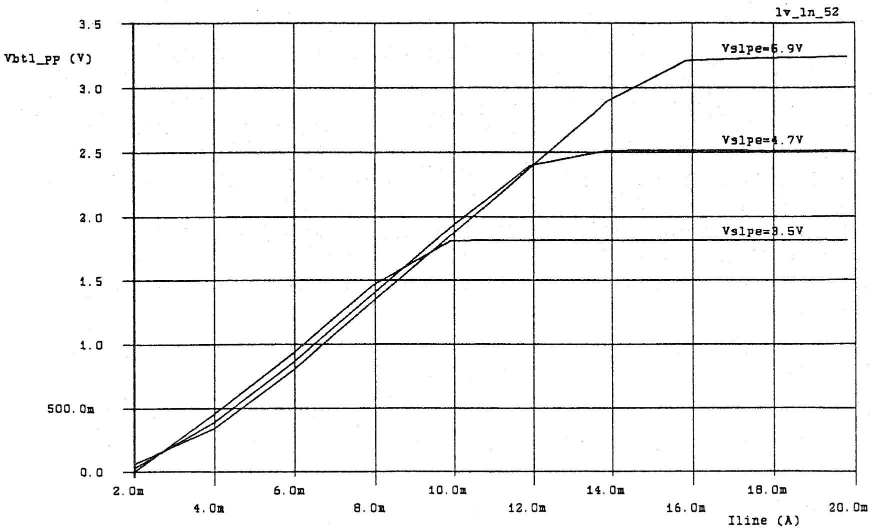


Figure 48 Maximum output swing of the receive amplifier into a 150Ω load (BTL) versus I_{LINE} in the low current range (parallel operation with a 220Ω+[820Ω//115nF] set)

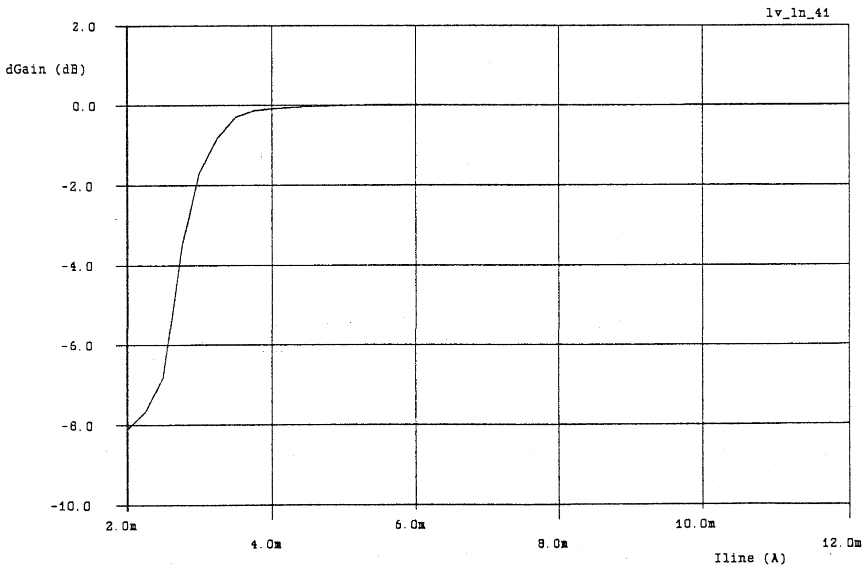


Figure 49 Receive gain versus I_{LINE} in the low current range (parallel operation with a 220Ω+[820Ω//115nF] set)

3.9 Anti sidetone

The function of the anti-sidetone block is to remove the sending signal on line for the receive path (see Fig. 50). Its basic structure is a Wheatstone bridge in which the sending channel structure is emulated by on-chip filters. The set impedance is emulated by a filter Zs' which is a replica from the AC-line interface block. The optimal sidetone impedance Zoss to match the external line impedance Z_{LINE}, is modelled by a resistor Rsa in series with a resistor Rsb in parallel to a capacitance Cs. It is emulated by a S.C.-filter which may be tuned in steps for a wide variety of possible cable types (11, 12 and 16 values for each of the components Rsa, Rsb and Cs respectively.).

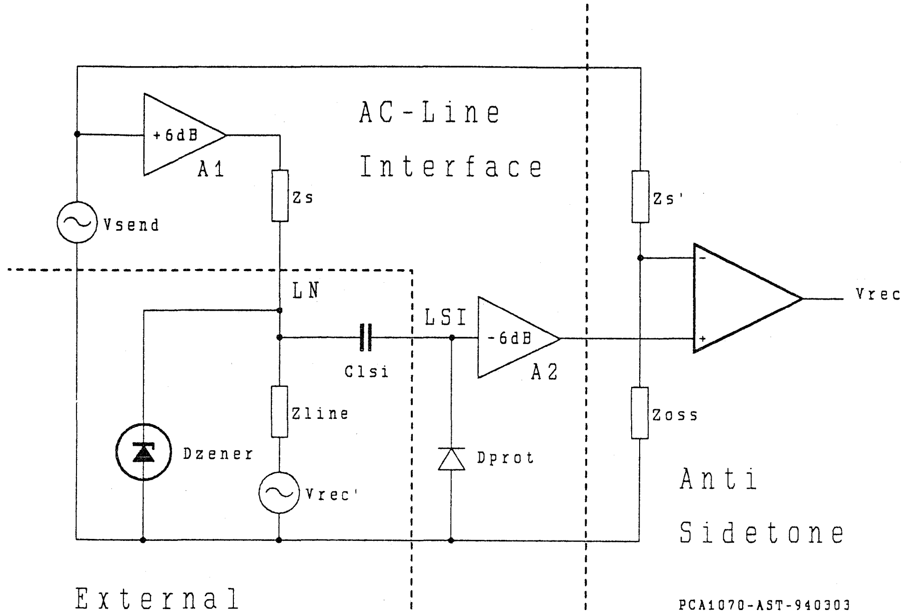


Figure 50 Principle structure of anti sidetone bridge

For the sidetone signal we can say

If Vrec'=0 then Vrec = Vsend (Zs/[Zs+Z_{LINE}] - Zs'/[Zs'+Zoss]

if Zs'=Zs and Zoss=Z_{LINE} then also Vrec=0

In Fig. 51 the side-tone gain (Gst) for the default values of the PCA1070 is measured with Z_{LINE}=Zoss. Sidetone suppression (Gsupp) is gain of the microphone channel (G_M) + gain of the receiving channel (G_{RS}) - Gst. Default values for G_M=41dB and G_{RS}=-6dB. So

$$Gsupp = G_M + G_{RS} - Gst = 41 - 6 - Gst = 35 - Gst$$

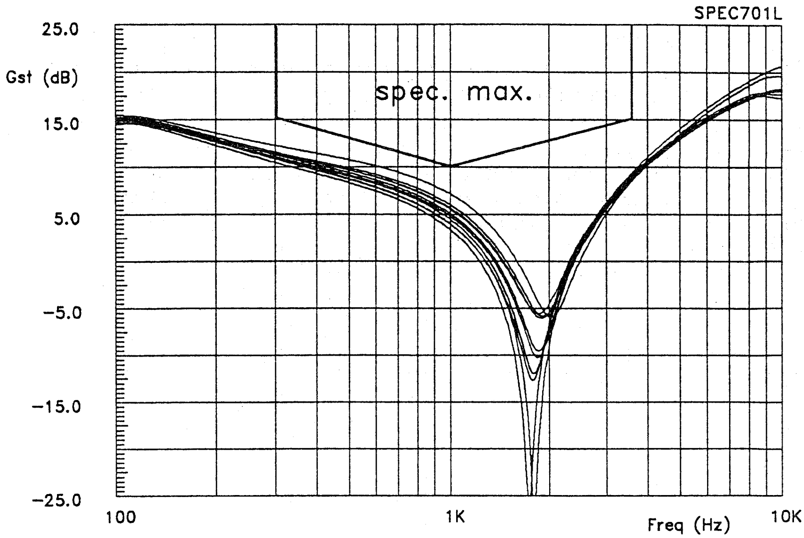


Figure 51 Sidetone Gain MIC+/MIC- to QR+/QR- for default values; Z_{LINE}=Zoss=492Ω+[1259Ω//134nF]

According to the specification is the Gsupp > 20dB at 300 and 3400Hz and Gsupp > 25dB at 1kHz.

For optimum sidetone reduction, the internal balance impedance Zoss must be equal to the external line impedance Z_{LINE}. In practice Z_{LINE} varies strongly with the line length and line type. Consequently a value for Zoss has to be chosen that corresponds with an average line length giving satisfactory sidetone suppression with short and long lines. The suppression further depends on the accuracy with which Zoss equals this average line impedance.

To fulfil the German PTT requirements for sidetone (tested with 35 different artificial lines), a practical method is to optimize for a cable 2.0km/0,4mm terminated with Zref (220Ω + 820Ω//115nF) and feeding-bridge type "A". A sufficiently good sidetone suppression is found with Rsa=192Ω, Rsb=1410Ω and Cs=105nF see Fig. 52.

3.9.1 Reprogramming set impedance

Reprogramming of the set impedance Zs has no influence on sidetone suppression. The sidetone bridge balance impedance Zoss can be chosen independent of Zs. The receive gain and the frequency characteristic of the receive channel is independent of the settings of Zs and Zoss.

3.9.2 Multiple anti sidetone by line current control

With the PCA1070 it is possible to measure the line current. The line current has mostly a strong relation with the line length. After reading this current, it is possible to change (under µC control) the programming value for Rsa, Rsb and Cs in order to get a better sidetone suppression.

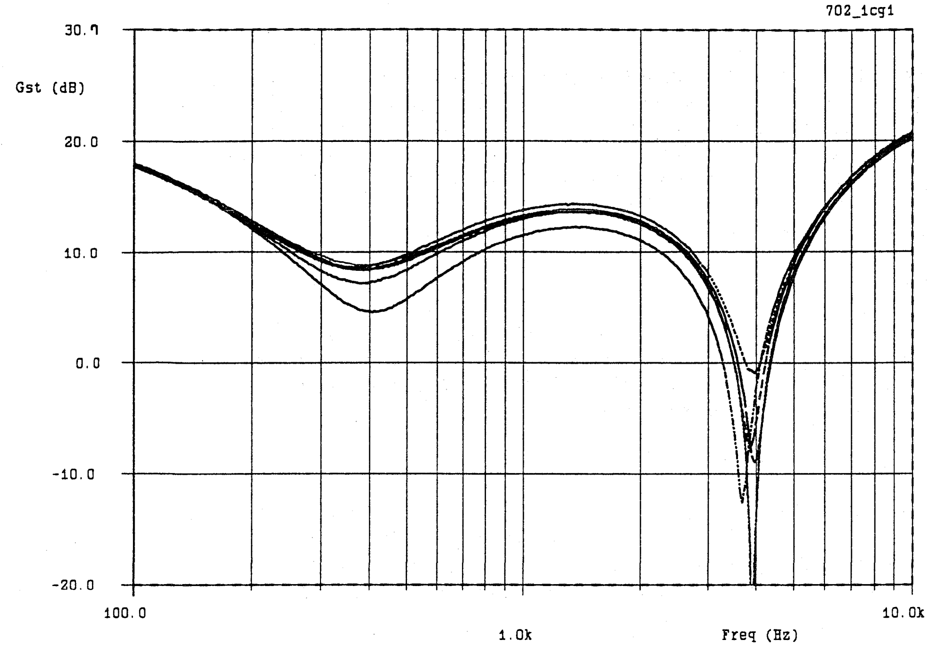


Figure 52 Sidetone Gain MIC+/MIC- to QR+/QR- with 2km/0.4mm cable terminated with 220Ω+[820Ω//115nF]

3.9.3 Considerations about sidetone distortion

Distortion in the 2 amplifiers A1 and A2 of Fig. 50 is not compensated in the anti sidetone block. The dynamic limiter of the PCA1070 will prevent distortion in amplifier A1. Amplifier A1 is limited by a zener-diode at pin LN. In normal operating amplifier A2 gives also no distortion. In case of very high sending signals that exceed the gain control range of the dynamic limiter, distortion at the line may occur. This can be due to the mechanisms a described in paragraph 3.3.4. A further limitation is imposed by input LSI which is the input of amplifier A2. This signal is limited to 0.7 V below V_{SS} (due to a protection diode at pin LSI). The output of amplifier A2 is limited between V_{DD} and V_{SS}.

3.9.4 Stability restrictions

To safeguard stable operation of the anti sidetone block in all situations, the following condition must be fulfilled: Rsa > Ra/2. In a practical situation can this may mean that the set impedance Zs has to be changed. For a telephone set with Zs=600Ω (Ra=600Ω, Rb=0, fp=12kHz) and optimum sidetone suppression for a cable 5km/0.5mm (176Ω/km, 38nF/km), the sidetone impedance must be Rsa=221Ω, Rsb=1259Ω and Cs=145nF. Now Rsa<300Ω, so change Zs to Ra=0Ω, Rb=600Ω and fp=12kHz.

3.9.5 Tax pulse filter

In case a 12kHz or 16kHz tax pulse filter is connected between the line connections and pin LN, the balance of the anti sidetone bridge is of course affected. This is mainly caused by the capacitor used in the series LC filter that shunts the line. The sidetone balance impedance can be corrected by reprogramming the Zoss of PCA1070. PCA1070 allows correction for capacitive loads between pins LN and VSS of up to about 33nF. It is recommended to use a capacitor value of <33nF in the series LC filter which is normally connected between the a/b lines.
More details about tax pulse filtering can be found in Chapter 6 and in Ref. [8].

3.9.6 EMC capacitor on LN

For EMC reasons it may be necessary to connect a capacitor between LN and VSS and/or between the a/b lines (practical values may be in the range 1nF - 4.7nF). Of course this also influences slightly the sidetone balance of the telephone set at higher frequencies. Normally it is not necessary to correct the programmed Zoss to meet the sidetone requirements.

3.9.7 Anti sidetone without clock signal

In case no clock (3.58MHz) is applied to pin CLK the set impedance and the optimal sidetone impedance switches automatically to 600Ω. Programming of other impedances is not possible then.
In practice the oscillator of the µC may switch off under extremely low voltage conditions (e.g. parallel operation of sets). The PCA1070 will then still operate but with somewhat relaxed performance.

3.10 Line current control (a.o. gain control; multiple anti sidetone)

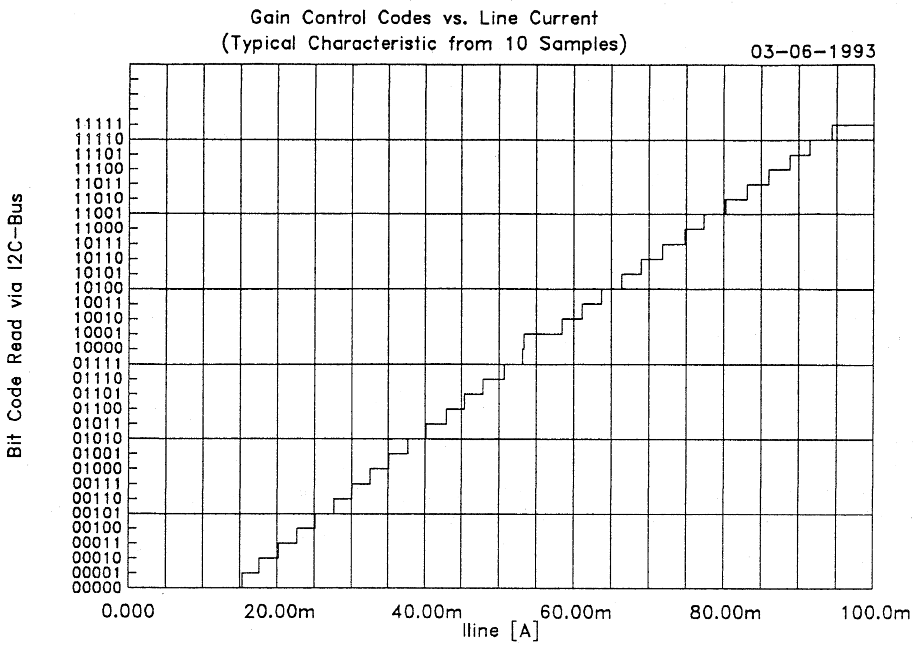


Figure 53 I²C-bus codes of line current register versus I_{LINE}

In many application the DC line current gives information about the line length. The DC line current can be read via I²C-bus interface. A typical relation between increasing I_{LINE} and the received bit-code can be seen in Fig. 53. To prevent unstable information a digital hysteresis is built in, so with decreasing I_{LINE} the bit-code is 1 higher with respect to Fig. 53. The PCA1070 is so designed that the bit-code is always monotone with the line current. The non-linearity at Iline≈53mA is a direct consequence of the circuit implementation.

The information about line-current can be used for gain control or multiple anti sidetone. Gain control to compensate line-losses by programming higher send and receive gains for long lines respectively. low line-currents. Multiple anti sidetone for better balancing the wheatstone-bridge by reprogramming Zoss respectively. Rsa, Rsb an Cs.

Line current control can be done once during start-up or repeated, depending on µC program. For test and measurements purposes it must be done repeated (e.g. every 200msec).

3.11 Send mute/Receive mute

The PCA1070 has 2 mutes, a send mute (SM) and a receive mute (RM), see Fig. 1. They can be programmed independent.
The function SM is to disconnect the microphone input and connect the DTMF input to the sending path. Gain programming is done with send prog-amp (Gma). In a telephone set it can be used for microphone mute.
The function RM is to disconnect the receive signal and connect the DTMF input via an attenuation (-25dB) to the receive outputs. Gain programming is done with receive prog-amp (Gra).
In case of tone dialling SM=1 is used for sending tones to the line and RM=1 is used for hearing a confidence tone.
To avoid clicks it is advised to program the new gain(s) first if this is lower than the previous one.

3.12 Clock Input

An external clock signal (f_{CLK} = 3.579545MHz; 200mVp-p < V_{CLK} < V_{VMC}-V_{SS}) must be applied to the CLK input. The clock interface block then provides all necessary internal clock signals to other blocks where needed. Internal clock signals are needed for the generation of set impedance (Zs) and optimal sidetone impedance (Zoss) as well as for operating of the dynamic limiter. If no clock is applied to pin CLK then Zs=600Ω and Zoss=600Ω.

The clock input consist of a dc blocking capacitor (≈4pF) followed by a buffer which is coupled back via a high-ohmic resistor. At the input pin an ESD protection diode is added with anode connected to VSS. The input structure is shown in Fig. 54. The equivalent impedance is typically 4pF in series with 800Ω.

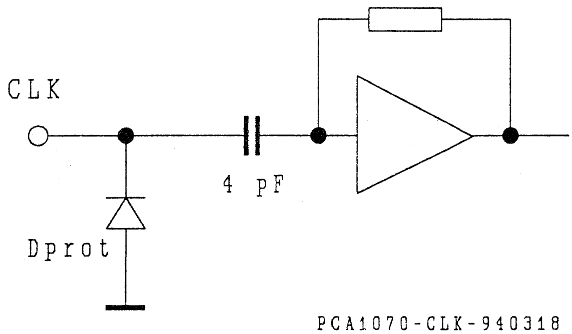


Figure 54 Internal structure clock input CLK

In a practical application the oscillator signal is taken from a µC (PCD335x) supplied from pin VMC, so the oscillator signal will be between V_{VMC} and V_{SS}. For test-purposes it is recommended to use a coupling capacitor. If no coupling capacitor is used, the applied clock signal must be superimposed on a DC voltage to prevent negative voltages on pin CLK.

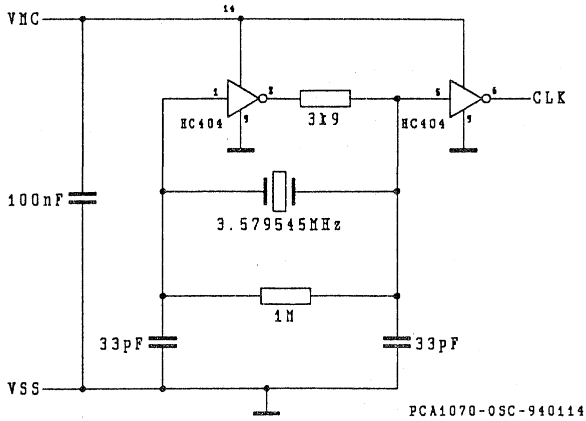


Figure 55 External oscillator

An example of an external oscillator is given in Fig. 55. It is important that the first inverter has a low gain so use a non-buffered device like HCU04. The second inverter is used as a buffer to drive the

PCA1070. If this oscillator is supplied from VMC of PCA1070, no coupling capacitor is needed.

Normally in a telephone set a micro-controller with a built-in oscillator is used. If this is a 3.579545MHz oscillator, it can be used to drive pin CLK of PCA1070. The supply of the µC is connected to pin VMC.

3.13 I²C-bus interface

The I²C-bus interface on the PCA1070 is a standard slave device. It works fully according to I²C-bus specification see Ref. [2]. For operating of the I²C-bus, SDA and SCL must be connected to VMC via 2 pull-up resistors. The value of these pull-up resistors depends on the clock-speed of the I²C-bus see Ref. [2]. In case of using a µC with software I²C-bus where $f_{SCL}\approx 10kHz$ pull-up resistors of 10kΩ are correct.

I²C-bus is used to program the PCA1070 with settings other than default.
An example of making a line-break (DPI=1) and also powerdown (PD0=1)

	Dev.addr.		subaddr.		data H06		
S	0100 0100	A	0000 0110	A	0100 0001	A	P

The I^2C-bus is provided with an automatic increment for the subaddressing. After writing data H06 the subaddress-counter is not returned to zero.
An example of changing send and receive gains to their lowest values (-25dB)

```
   Dev.addr.    subaddr.      data H04      data H05
|S|0100 0100|A|0000 0100|A|0011 1001|A|0011 1001|A|P|
```

In stead of making a Power-On-Reset all the defaults of the PCA1070 can be written with the following I^2C-bus command.

```
   Dev.addr.    subaddr.      data H00      data H01      data H02
|S|0100 0100|A|0000 0000|A|0100 0000|A|0110 1010|A|1001 0010|A| - -

                                 data H03      data H04      data H05      data H06
                         - - |0011 0011|A|0000 1111|A|1010 0110|A|0000 0000|A|P|
```

For more information see paragraph 2.4 and the PCA1070 specification Ref. [1].

3.14 Power control

The function of the power control block is generating reset signals, internal reset (RESET) and reset for µC (RMC), circuit controlling during start-up in speech mode or ring-mode (RG) and DC start-up (DST), power control during dial pulses and long line breaks. For functional block diagram see Fig. 56.

3.14.1 Reset conditions

The PCA1070 is in reset when V_{DD} is below the Power-On-Reset (POR) level 1.2V ± 0.2V. The PCA1070 goes out reset when V_{DD} is above the POR level and VMC is above the Micro-Controller-On-Reset (MCOR) level 2.0V ± 0.2V. When VMC goes below MCOR level the PCA1070 stays working. The PCA1070 goes in reset again when V_{DD} is below POR level. In reset mode all parts of the PCA1070 are conducting current from VDD except LI-block from LN, SLPE and VDD.

3.14.1.1 Internal reset (PRES-bit)
The PCA1070 is in reset when V_{DD} is below the POR level 1.2V ± 0.2V. In reset mode the default values are written into all registers. For each time the PCA1070 goes in reset the Power RESet (PRES) bit is set to "1". This PRES-bit can be seen by reading it via I^2C-bus. After reading the PRES-bit, it is set to "0". To check a power failure the PRES-bit is very useful.

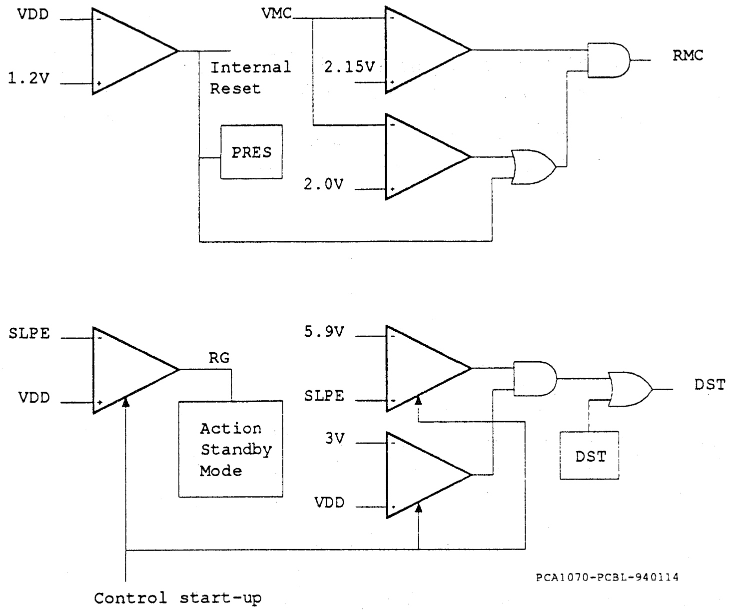


Figure 56 Circuit diagram Power Control block

3.14.1.2 Reset Micro Controller

The output Reset Micro Controller (RMC) depends on the voltage of VMC. When VMC is below the MCOR-level RMC is "1"(≈VMC). When VMC is above the MCOR-level RMC is "0". The MCOR-level depends on the mode of the PCA1070. In normal operation and powerdown mode the MCOR-level is 2.0 ± 0.2V. In stand-by and reset mode a P.O.R. circuit with low power consumption is used. This is important for applications with software controlled ringer see Ref. [15]. Now the spread is higher, but the minimum value remains 1.8V, so the MCOR-level is 2.15 ± 0.25V. In normal applications there is always a capacitor connected to VMC, so during start-up $\delta V_{VMC}/\delta t$ always slow enough to guarantee a good RMC pulse.

3.14.2 Ring/speech detector

The RinG/speech detector gives information to the PCA1070 whether it is supplied via VDD (ring mode) or via SLPE (speech mode). This is used in a system with 1 supply capacitor, VMC connected to VDD. When after an internal reset V_{DD} goes high and also V_{SLPE} via R_{SUP}(250Ω) the RG signal becomes "1" and the PCA1070 will go in stand-by mode. When after some time (ringing) the PCA1070 is supplied via SLPE (hook-off) the RG-signal becomes "0" and the PCA1070 returns automatically to normal operation mode. The RinG detector can be disabled by setting control bit RRG=1 (Reset RinG) via I²C-bus. In a system with 2 supply capacitors the RinG/speech detector has no function, but is still operational. To prevent unwanted actions make RRG=1.

3.14.3 Power down/Standby

For controlling current consumption in the PCA1070 during pulse dialling or long line breaks the PCA1070 can be switched in 2 low power modes: powerdown and stand-by (see Fig. 57). Stand-by can be reached by programming PD1=1 and PD0=1 via I²C-bus and is used in situations where VMC will stay above 2.5V (2.4V + [temp-25]•0.003V = 2.415V for 60°C). If VMC decreases below 2.5V but stays above 2.2V it is recommended to use powerdown mode by programming PD1=0 and PD0=1.

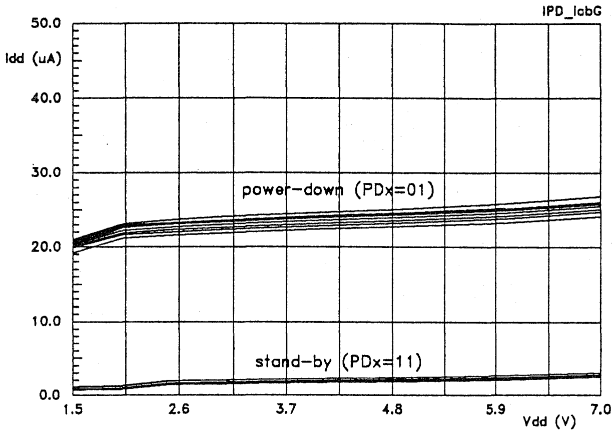


Figure 57 Power down and standby current consumption $I_{DD} = f(V_{DD})$.

3.15 Start-up and switch-off behaviour (timing diagrams)

3.15.1 DC starting time (DST)

The IC is equipped with circuitry for fast DC start-up. This circuit is automatically activated as soon as V_{DD} reaches typically 3V after hook-off, and is deactivated when V_{SLPE} drops below typical 5.9V. This ensures that only a relatively short time is needed to reach the default DC setting (V_{SLPE}) of the circuit and that V_{DD} will not exceed the maximum permitted voltage of 6V.
This start-up circuit can be also activated under software control by setting DST=1 via I²C-bus. The start-up time can be optimized by programming the DST=1 during the start-up procedure. In practice this is possible as soon as the µC has become operational. The DST bit can also be used to quickly restore of DC settings (V_{SLPE}) after long line breaks (e.g. flash) or during reprogramming of the DC voltage drop V_{SLPE} (e.g. during pulse dialling in case NSA or MUTE2 is required).
Remark: the AC-impedance into pin LN is reduced considerably when DST=1 (see paragraph 3.4).

3.15.2 System with 2 supply capacitors

The basic application of PCA1070 in a system with 2 supply capacitors is shown in Fig. 4.

3.15.2.1 Hook-off / Start-up (outgoing call)

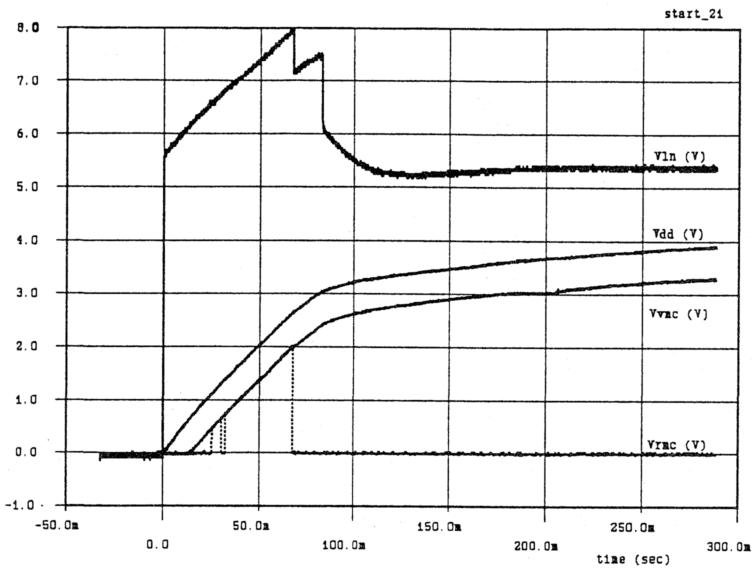


Figure 58 Hook-off 2 supply C system at V_{LINE}=60V and R_{LINE}=2520Ω.

Fig. 58 shows the start-up behaviour after hook-off. At time=0 there is no line current (I_{LINE}) and all voltages at capacitors are 0. When the I_{LINE} starts flowing, the voltage at LN (V_{LN}) will be V_{LN} = I_{LINE}•(R_{SUP}+R_{SLPE}) = I_{LINE}•(250+20)V. For high line currents (I_{LINE}>40mA) V_{LN} must be limited to 12V by the external zener. The voltage at SPEECH will be V_{LN} + the gate-source voltage of the external mostfet MNex. Now the capacitor at VDD (C_{VDD}) starts charging; the PCA1070 is in the RESET mode, I_{DD} rising with V_{DD} and I_{VMC}=2uA (max 5uA). As soon as V_{DD} reaches the Power-On-Reset level of 1.2V +/- 0.1 V, the PCA1070 goes out his reset mode; the bandgap-voltage (VBGAP) becomes available 1.2V +/- 4% ; the Power Control block becomes active; Line-Interface AC and DC (L.I.) part of the PCA1070 is still in Powerdown (PD_LI=1).

When V_{DD} reaches a diode voltage (external diode D1 between VDD and VMC), the capacitor at VMC (C_{VMC}) starts charging. The Micro-Controller is in the reset mode (RMC=1). As soon as VMC reaches the switching level for the Micro-Controller reset (MCreset) RMC goes to 0 and the L.I. goes out powerdown (PD_LI=0). MCreset = 1.67•VBGAP = 2.0V +/- 0.2V. The current now flowing into pin LN (≈3mA) gives a decreased current into R_{SUP} and also voltage drop at pin LN (see Fig. 58, time≈70msec). All blocks are active now but the most important part is the DC-Line-interface block. The capacitor between LN and REG (C_{REG}) starts charging. If V_{CREG} reaches I_{LINE}•R_{SLPE}+0.4V the external PNP transistor at TX starts sinking current (I_{PNP}). As soon as I_{PNP} is equal to I_{LINE}-I_{LN}, V_{SLPE} will switch to the default value of 4.7V. To prevent that the voltage at VDD becomes above 6V, C_{REG} must be charged faster than normal. This only happens when V_{DD} > 3V and V_{SLPE} > 5.9V by activating the DC-Start-Time (DST) bit. So if V_{SLPE} > 5.9V and V_{DD} reaches 3V, DST is automatically set to 1 and when V_{SLPE} becomes below 5.9V, DST is set to 0. Now the PCA1070 is fully operating.

Beside C_{REG} an other capacitor between LN and LSI (C_{LSI}) must be loaded for correct working of the AC-Line-interface. The charging of C_{LSI} starts as soon as V_{LN} becomes high. In normal operating the dc-voltage at pin LSI (V_{LSI}) must be the internal bias voltage (V_{DD}/2) which is normally V_{DD}/2. When V_{SLPE} and V_{LN} switches to the default values also V_{LSI} is switching. Three things can occur after switching: V_{LSI} above V_{DD}, between V_{DD} and V_{SS} or below V_{SS}. When V_{LSI} is not between V_{DD} and V_{SS} a current greater than expected can flow from LN to SCR. Current through R_{SCR} normal about 3mA. Due to this fact the voltage on V_{SLPE} can have a positive or negative overshoot. To prevent this overshoot the relation between C_{LSI} and C_{VDD} is important or make use of software DST during start-up (see Fig. 80).

3.15.2.2 Hook-off after ringing (incoming call)

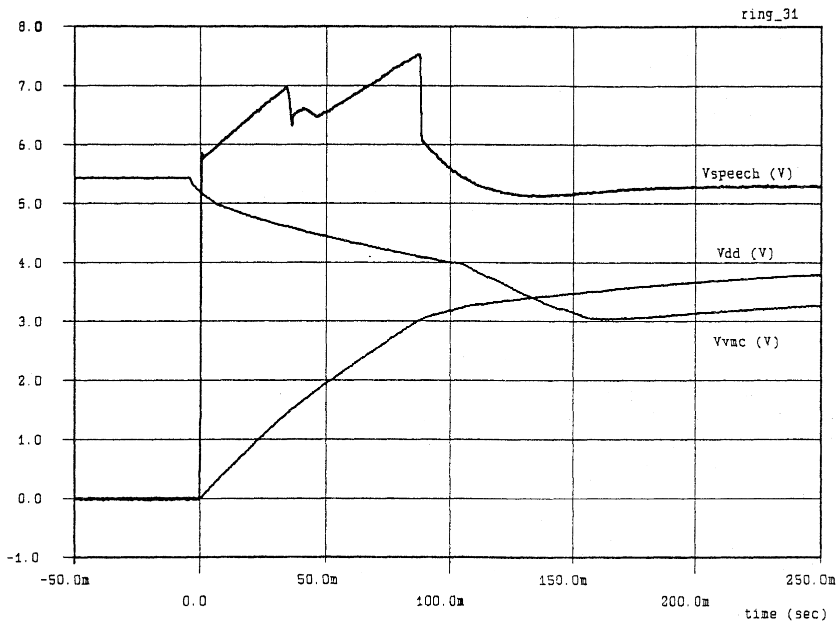


Figure 59 Hook-off after ringing, 2 supply C system at V_{LINE}=60V and R_{LINE}=2520Ω.

Fig. 59 shows the start-up behaviour after a ringing burst which already has charged the supply capacitor of the µC. At time=0 there is no line current (I_{LINE}) and all voltages at capacitors are 0. When the ringer voltage V_{RING} (see fig. 70) goes high, the capacitor at VMC (C_{VMC}) starts charging via diode D7 (see fig 71). As soon as VMC reaches the switching level for the Micro-Controller reset (MCreset) RMC goes to 0. MCreset = 2.15 +/- 0.35V. PCA1070 stays in RESET/stand-by mode while V_{DD} is still 0. Now the µC is working for the software controlled ringing function. System still on-hook.

When the system goes off-hook, I_{LINE} starts charging the capacitor at VDD (C_{VDD}). As soon as V_{DD} reaches the Power-On-Reset level of 1.2V +/- 0.1V, the L.I. goes immediately out powerdown. All blocks are active now. The capacitor between LN and REG (C_{REG}) starts charging. If V_{CREG} reaches $I_{LINE}*R_{SLPE}$+0.4V the external PNP transistor at TX starts sinking current (I_{PNP}). As soon as I_{PNP} is almost equal to I_{LINE}, V_{SLPE} will switch to the default value of 4.7V. To prevent that the voltage at VDD becomes above 6V C_{REG} must be charged faster than normal. This only happens when V_{DD} > 3V and V_{SLPE} > 5.9V by activating the DC-Start-Time (DST) bit. So if V_{SLPE} > 5.9V and V_{DD} reaches 3V, DST is automatic set to 1 and when V_{SLPE} becomes below 5.9V DST is set to 0. Now the PCA1070 is full operating.

For more information about software controlled ringer see Ref. [15].

3.15.2.3 Hook-on (Switch-off)

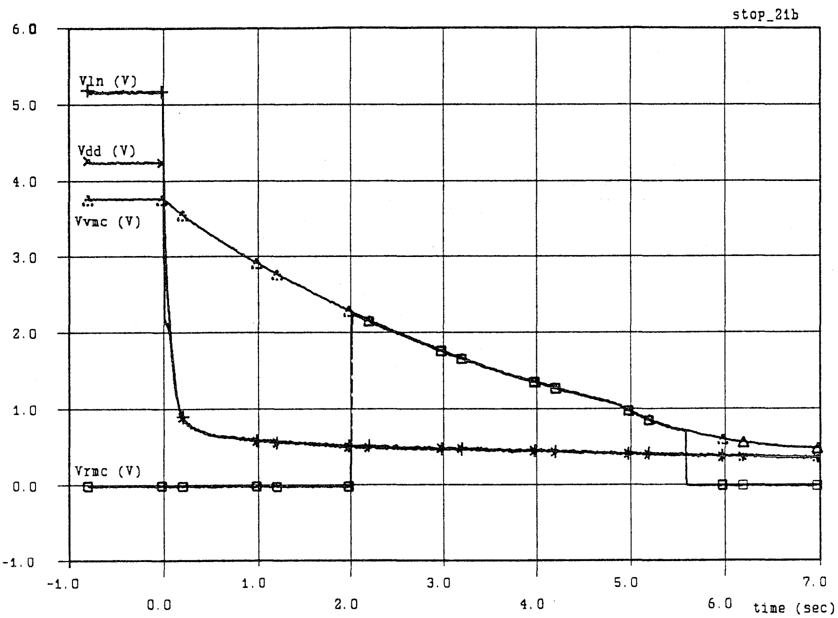


Figure 60 Hook-on 2 supply C system at V_{LINE}=60V and R_{LINE}=2520Ω.

When I_{LINE} is switched off and PDx is not set, all capacitors will be discharged. The PCA1070 is operating until V_{DD} has reached the POR level of 1.2V. When VMC has reached the MCreset level, the µC stops, but the clock for the PCA1070 is still there.

3.15.3 System with 1 supply capacitor

The basic application of PCA1070 in a system with 1 supply capacitor for both the µC and the PCA1070 is shown in Fig. 3.

3.15.3.1 Hook-off / Start-up (outgoing call)

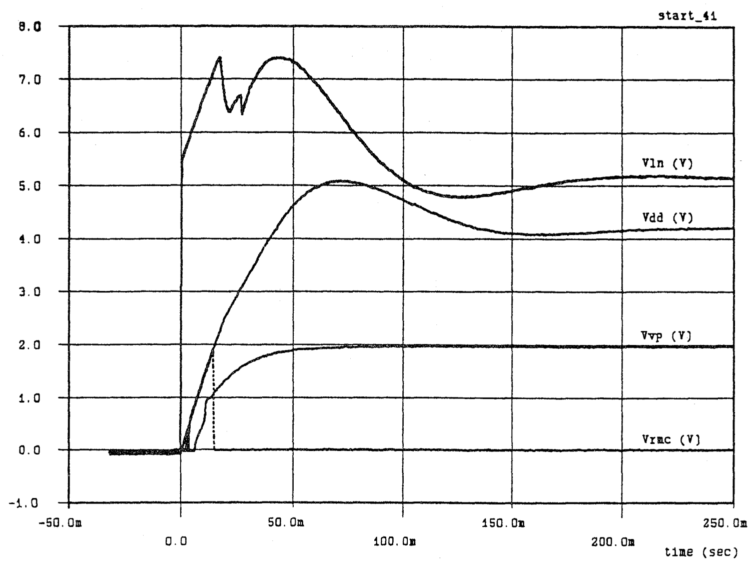


Figure 61 Hook-off 1 supply C system at V_{LINE}=60V and R_{LINE}=2520Ω.

Fig. 61 shows the start-up behaviour after hook-off. At time=0 there is no line current (I_{LINE}) and all voltages at capacitors are 0. When the I_{LINE} starts flowing, the voltage at LN (V_{LN}) goes immediately to the zener voltage maximum 12V if I_{LINE} is greater than 40mA. If I_{LINE} is lower than 40mA, the voltage on LN will be $V_{LN} = I_{LINE} \bullet (R_{SUP}+R_{SLPE}) = I_{LINE}*(250+20)V$. The voltage at SPEECH will be V_{LN} + the gate-source voltage of the external mostfet MNex. Now the capacitor at VDD/VMC (C_{VDD}) starts charging; the PCA1070 is in the RESET mode, I_{DD} rising with V_{DD} and I_{VMC}=2uA (max 5uA). As soon as V_{DD} reaches the Power-On-Reset level of 1.2V +/- 0.1 V, the PCA1070 goes out his reset mode; the bandgap-voltage (VBGAP) becomes available 1.2V +/- 4% ; the Power Control block becomes active; Line-Interface AC and DC (L.I.) part of the PCA1070 is still in Powerdown (PD_LI=1).

The Micro-Controller is in the reset mode (RMC=1). As soon as V_{DD} reaches the switching level for the Micro-Controller reset (MCreset) RMC goes to 0 and the L.I. goes out powerdown (PD_LI=0). MCreset = 1.67*VBGAP = 2.0V +/- 0.2V. The current now flowing into pin LN (≈3mA) gives a decreased current into R_{SUP} and also voltage drop at pin LN (see fig61, time≈18msec). All blocks are active now but the main part is the DC-Line-interface block. The capacitor between LN and REG (C_{REG}) starts charging. If V_{CREG} reaches $I_{LINE}*R_{SLPE}$+0.4V the external PNP transistor at TX starts sinking current (I_{PNP}). As soon as I_{PNP} is almost equal to I_{LINE}, V_{SLPE} will switch to the default value of 4.7V. To prevent that the voltage at V_{DD} becomes above 6V, C_{REG} must be charged faster than normal. This only happens when V_{DD} > 3V and V_{SLPE} > 5.9V by activating the DC-Start-Time (DST) bit. So if V_{SLPE} > 5.9V and V_{DD} reaches 3V, DST is automatically set to 1 and when V_{SLPE} becomes below 5.9V, DST is set to 0. Now the PCA1070 is full operating.

Beside C_{REG} an other capacitor between LN and LSI (C_{LSI}) must be loaded for correct working of the AC-Line-interface block. The charging of C_{LSI} starts as soon as V_{LN} becomes high. In normal operating the dc-voltage at pin LSI (V_{LSI}) must be the internal bias voltage ($V_{DD}/2$) which is normally $V_{DD}/2$. When V_{SLPE} and V_{LN} switches to the default values also V_{LSI} is switching. Three things can occur after switching: V_{LSI} above V_{DD}, between V_{DD} and V_{SS} or below V_{SS}. When V_{LSI} is not between V_{DD} and V_{SS} a current greater than expected can flow from LN to SCR. Current through R_{SCR} normal about 3mA. Due to this fact the voltage on V_{SLPE} swings more or less.

3.15.3.2 Hook-off after ringing (incoming call)

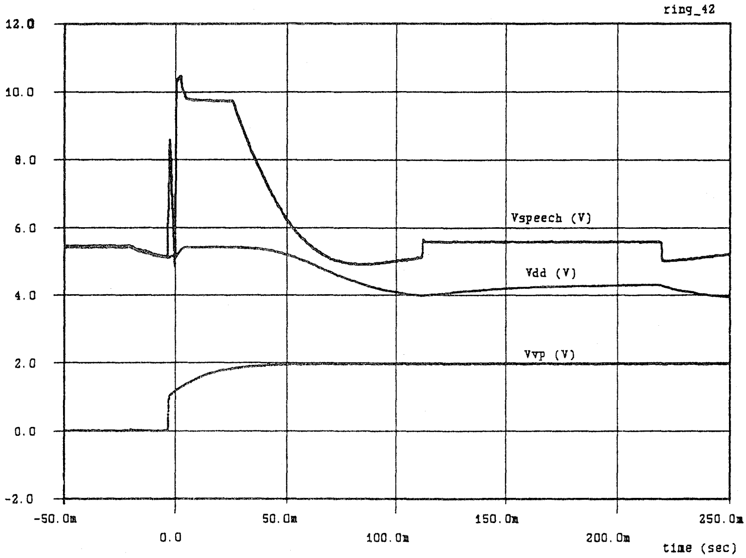


Figure 62 Hook-off after ringing, 1 supply C system at V_{LINE}=60V and R_{LINE}=2520Ω.

Fig. 62 shows the start-up behaviour after a ringing burst already has charged the supply capacitor of the µC and the PCA1070. At time=0 there is no line current (I_{LINE}) and all voltages at capacitors are 0. When the ringer voltage V_{RING} (see fig. 70) goes high, the capacitor at VDD/VMC (C_{VDD}) starts charging via diode D7 (see fig. 71); the PCA1070 is in the RESET mode, I_{DD} is rising with V_{DD} and I_{VMC}=2uA (max 5uA). As soon as V_{DD} reaches the Power-On-Reset level of 1.2V +/- 0.1 V, the PCA1070 goes out his reset mode; the ringer detector becomes active because $V_{SLPE} < V_{DD}$. Due to the internal RG=1 signal the PCA1070 will switch very fast to its STAND-BY mode. As soon as V_{DD} reaches the switching level for the Micro-Controller reset (MCreset) RMC goes to 0. MCreset = 2.15 +/- 0.35V. PCA1070 stays in STAND-BY mode. Now the MC is working for the software controlled ringing function. System still on-hook.

When the system goes off-hook, I_{LINE} starts flowing and the voltages at LN and SLPE will switches above V_{DD}. The ringer detector will be disabled because $V_{SLPE} > V_{DD}$ + 100mV and PCA1070 goes immediately out stand-by. All blocks become active. The capacitor between LN and REG (C_{REG}) starts charging. If V_{SLPE} > 5.9V (and V_{DD} > 3V) the DC-Start-Time (DST) bit will be activated. If V_{CREG} reaches I_{LINE}*R_{SLPE}+0.4V the external PNP transistor at TX starts sinking current (I_{PNP}). As soon as I_{PNP} is almost equal to I_{LINE}, V_{SLPE} will switch to the default value of 4.7V and the DST bit is disabled.

3.15.3.3 Hook-on (Switch-off)

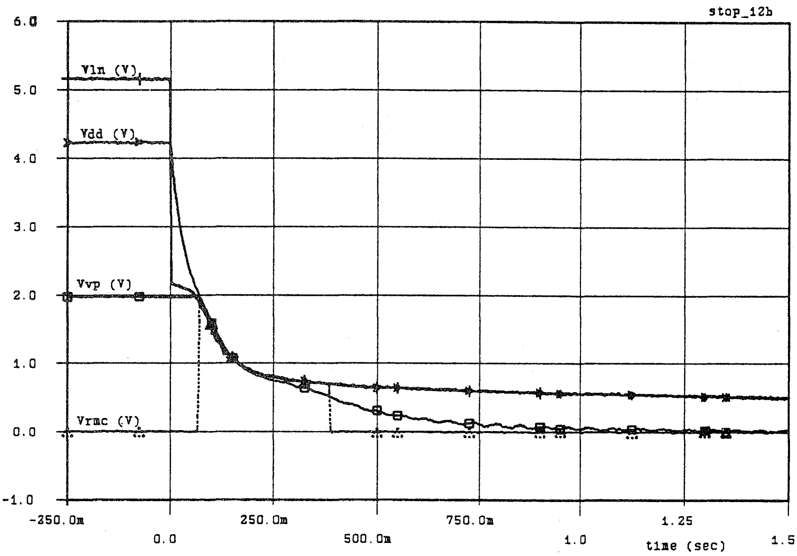


Figure 63 Hook-on 1 supply C system at V_{LINE}=60V and R_{LINE}=2520Ω.

When I_{LINE} is switched off (at t=0 in Fig. 63) and PDx is not set, all capacitors will be discharged. When V_{DD} has reached the MCreset level. The pin RMC goes high to reset the µC. The PCA1070 is operating until V_{DD} has reached the POR level of 1.2V. When the oscillator of the µC switches off, the clock for the PCA1070 is no longer there but PCA1070 will still work with relaxed performance.

3.15.4 Special requirements during start-up

In Germany are special requirements for dc-resistance of the telephone set (Rset-dc) during start-up (see Ref. [12]. We look now at the requirement that Rset-dc must be lower than 480 Ω after 60 msec when the set is supplied with 60V with 2 internal resistors of 500Ω and a series-resistor of 1530Ω. So the dc-voltage on the set must be lower than 480/(480+1000+1530)•60=9.6V. (Between 60 and 150 msec there may-be some spikes greater than 480 Ω if they are shorter than 3 msec).
With the basic application (Fig. 4) this requirement is met, but with little margin.

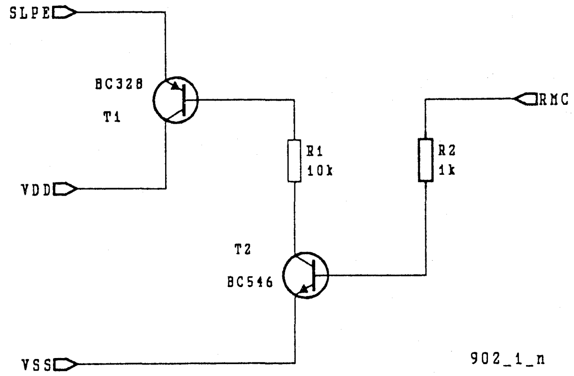


Figure 64 Add-on circuit for lower DC line voltage during start-up

This voltage depends on the voltage over the 2 bridge-diodes (Vdiode-bridge), the voltage over R_{SLPE}=20Ω ($V_{LN\text{-}SLPE}$), the voltage over R_{SUP} ($V_{SLPE\text{-}VDD}$) and the voltage at VDD (V_{DD}). So

$$V_{LINE} = \text{Vdiode-bridge} + V_{LN\text{-}SLPE} + V_{SLPE\text{-}VDD} + V_{DD}$$

where $V_{LN\text{-}SLPE} = R_{SLPE} \bullet (I_{LINE}\text{-}I_{LN})$
$V_{SLPE\text{-}VDD} = R_{SUP} \bullet (I_{LINE}\text{-}I_{LN}\text{-}I_{SLPE})$

In a 2 capacitor system $V_{DD} = V_{VMC}$ + Vdiode. When RMC=1 the current in pin LN (I_{LN}) and the current in slpe (I_{SLPE}) are 0. We look at V_{LINE} when V_{VMC}=MCreset level, so V_{VMC}= 2.0 ± 0.2 V. We assume Vdiode=0.7V, so V_{DD}=2.7V typical. We assume Vdiode-bridge = 1.4 V at I_{LINE} = 20mA. R_{SUP}=243Ω.

$$V_{LINE} = 1.4 + 20\bullet0.02 + 243\bullet0.02 + 2.7 = 9.36 \text{ V}$$

Due to spread this value can be higher than 9.6 V. To lower this value, the circuit of Fig. 64 can be used. This circuit is only active when RMC=1. When RMC=1 R_{SUP} is shorted by transistor T1. Measurements in Fig. 65 show the difference with and without this circuit.

When RMC switches to "0" the currents I_{LN} and I_{SLPE} are not 0, I_{LN}≈2.7mA and I_{SLPE}≈0.3mA, so V_{LINE} will therefore decrease about 20•0.0027 + 243•(0.0027+0.0003)=0.783V to a value of V_{LINE}=8.58V.
When V_{DD}=3V (V_{LINE}≈9V) DST is switched to "1" by the internal start circuit and V_{SLPE} goes fast to the default value. Now

$$V_{LINE} = \text{Vdiode-bridge} + V_{LN\text{-}SLPE} + V_{SLPE} = 1.4 + 20\bullet0.017 + 4.7 = 6.44 \text{ V}$$

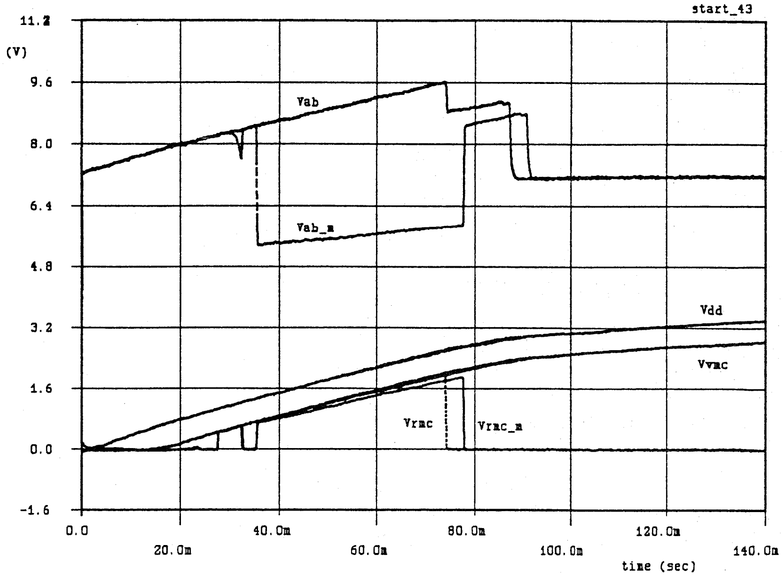


Figure 65 Lower DC line voltage during hook-off in a 2 supply C system

3.16 Dial pulse input / DOC-output

3.16.1 N-most interrupter

By setting control bit Dial Pulse Input (DPI) to "1" via I^2C-bus (see fig. 66), the Dial Output Connection (pin DOC) will become low-ohmic and the line will be disconnected (V_{SPEECH} goes high) with an external N-most (see Figs 3, 4). In the basic application this is a n-channel depletion type (e.g. Philips type BSD254A). This has several advantages: minimum number of components, low voltage drop and self-starting.
To prevent discharging of all the capacitors (C_{VDD}, C_{REG}, C_{LSI}) the powerdown (PDx) bits must be set. If PDx=01 the PCA1070 is in powerdown mode and the I_{DD} current will be < 100uA; if PDx=11 the PCA1070 is in stand-by mode and the I_{DD} current will be < 5uA. Pull down When the line current is disconnected, the PCA1070 is operating until the C_{VDD} is discharged and V_{DD} has reached the reset level 1.2V.

During line-break the V_{SLPE}=V_{DD} and C_{VDD} is discharged. During line-make C_{VDD} must be charged, so current through the external PNP-transistor must change in comparison with the beginning of the line-break. due to Z_{LINE}, C_{LSI} and large external C_{REG}=470nF and large internal resistor (Rp) typical 1075kΩ the current will change very slowly. Faster switching of the current can be achieved by making DST=1. It is recommended to make DST=1 for the complete dialling period.

When an N-most enhancement is used in stead of the N-most depletion, a considerably higher voltage drop across the interrupter must be taken into account.

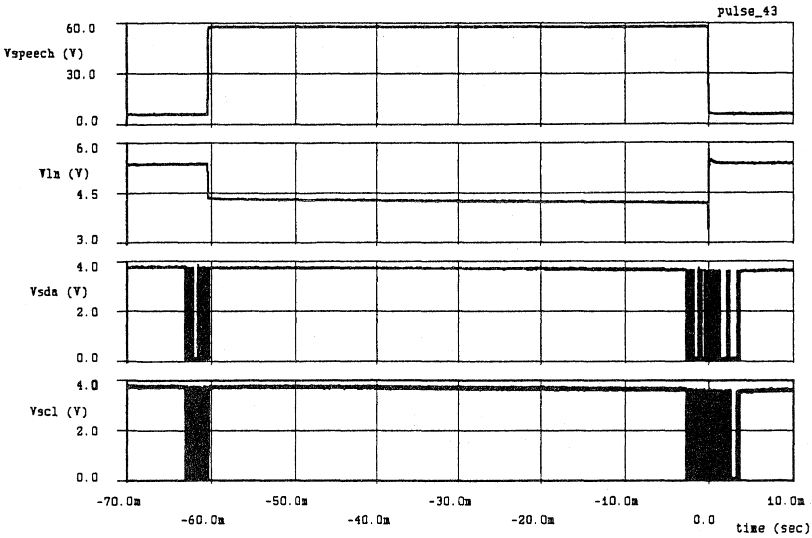


Figure 66 Pulse dialling in a 2 supply C system; waveforms on line (V_{SPEECH}), LN and I^2C-bus

3.16.2 P-most interrupter

Instead of an N-most interrupter, PCA1070 may also be used with a P-most interrupter transistor. suitable transistor types are BSP254A (250V) and BSP304A (300V). An example of such an interrupter is shown in Fig. 67. Transistor BC558 and the 3.9Ω resistor limit the current through the P-most when it exceeds Vbe/3.9 (about 150mA). This current limiter is only intended to provide protection against current surges and is not to be used (for reasons of power dissipation) for continuous limitation of line current.

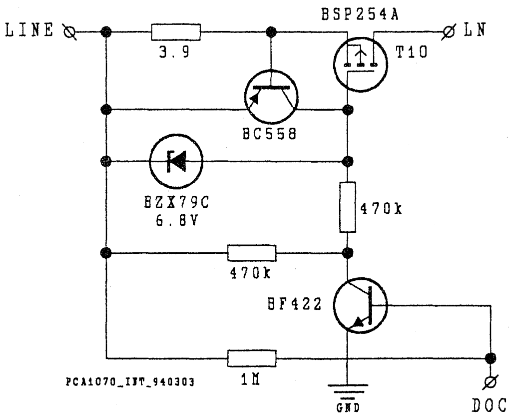


Figure 67 P-most interrupter

4 Immunity to R.F. signals (EMC)

The printed circuit boards on which the PCA1070 is used in his application are commonly small compared to the wavelengths of the applied interference signals (frequency < 300MHz /λ > 1 meter). The traces on the printed circuit board should be kept small (<0.1) compared to this wavelength to ensure that direct pick-up will be negligible.
In this situation the only part of the telephone set capable of acting as an antenna in order to receive a part of the interference signal will be the leads towards the public telephone network on one hand the leads towards the handset on the other.

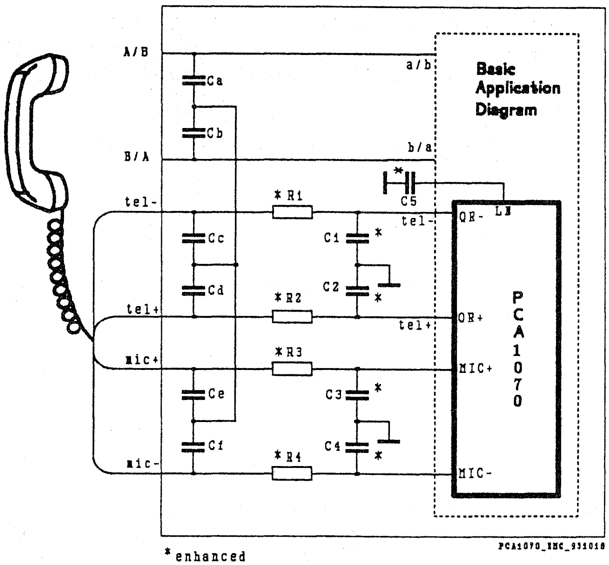


Figure 68 Circuit-diagram for EMC

4.1 Basic protection

To eliminate the working of the LINE-leads and the handset-leads as antenna, we can surround them with an "RF-guard", shown in Fig. 68. How the "RF-guard" works can be explained as follows:
The RF-signals are mainly picked up by the leads connected to the telephone base set (handset-cord and base-cord). The induced RF-current flows now through the conductor from Ca to Cf, instead of through the reference of the telephony circuitry.

Care must be taken that the re-radiation of the RF-currents through this extra conductor will not be picked up by the telephony circuitry again. Therefore the LINE and the Handset connectors should be placed close together. RFI-capacitors must be mounted as close as possible to these connectors. An open space must be realised on the printed circuit board between the circuitry and the extra conductor to reduce mutual coupling between the two.

Besides the RF-guard, sometimes a small capacitor at LN may be needed to fulfil the basic E.M.C. requirements. This depends strongly on print lay-out.

4.2 Enhanced protection

At designing the PCA1070, a lot of care has been taken that RF-signals on the pins have minimum influence on the circuit. In order to satisfy the most stringent E.M.C. requirements it may sometimes be necessary to add simple filters in the traces from the handset connector and the pins of the PCA1070. In Fig. 68 these RC-filters are marked with an *.

For immunity testing the current injection method is used according the German requirements, VDE 0878 part 200 (draft 1986), which can also be found in the draft IEC publication 801/6 "Conducted immunity requirements".

More info on design for EMC can be found in Ref. [13].

5 Protection

The transmission IC can be destroyed by excessive current surges on the telephone lines if no proper measures are taken.

A break over diode is needed to protect the circuit against spikes. This break over diode (e.g. BR211-240) is connected between the a,b lines in front of the polarity guard.

According to the PCA1070 specification Ref. [1] the voltage on most of the pins may not exceed the 7V. Some pins are allowed to have 12V.

In an application where the PCA1070 is supplied from the telephone line, the VDD and other related pins will never be higher than the programmed DC reference voltage on SLPE (maximum 5.9 + 0.5 V). During start-up the DST-bit assures that the voltage regulator of the PCA1070 is operating before V_{DD} reaches the maximum allowed level (see also paragraph 3.15.1).

The voltage on LN (also LSI, REG, SLPE, TX and DOC) may not exceed 12V. To prevent this a voltage regulator diode (e.g. BZD23C,11V) must be connected between LN and VSS. This zener has also a function during start-up at high line currents (>40mA). Special attention has to be taken that the voltage at pin DOC will not exceed 12V during start-up at high line currents.

In the basic application in which an N-most is used for pulse-dialling and line-interrupt (see Fig. 4), a resistor in the GROUND-line senses the line current. If this current exceeds a limit, an extra transistor pulls the voltage on DOC down. It is important that the collector-emitter leakage of this transistor is very low to prevent unwanted line interrupt.

In an application with a P-most used for pulse-dialling and line-interrupt, we can add a resistor in the SPEECH-line to measure the current (see Fig. 67). If this current exceeds a limit, an extra transistor pulls the voltage on the gate of the P-most up.

In an application with external supply (e.g. software controlled ringer) a zener of 5V6 must be applied to the VDD of the µC. This zener also limits the supply voltage of the PCA1070.

6 Tax pulse filtering

Fig. 69 shows schematically a tax pulse filter-transmission circuit combination.

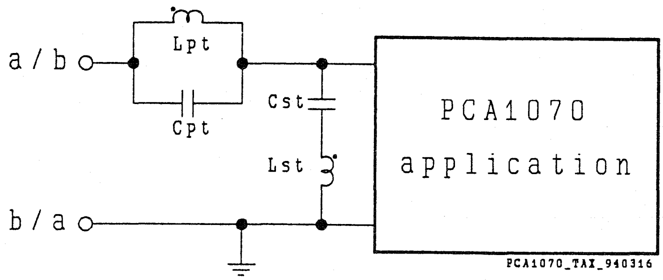


Figure 69 Tax pulse filter-transmission circuit combination.

The block "PCA1070 application" includes protection, polarity guard, hook-switch, interruptor and other peripheral components. The basic application diagrams (without hook-switch are shown in Fig. 3 and 4. The recommended component values are listed below.
12kHz: Lpt= 3050µH, Lst= 6100µH (2*3050), Cpt= 57nF (47//10), Cst= 28.8nF (27//1.8)
16kHz: Lpt= Lst= 3050µH, Cpt= Cst= 33nF.
N.B. The coils (See Note) must be mounted perpendicular to each other and at a sufficient distance (>30mm between their centres).
The programmable set- and sidetone impedances must be adapted to compensate the effect of the tax pulse filter. The adjustment for optimum sidetone suppression was done for a load impedance consisting of Zref terminating an artificial line of 2 km/0.4mm (R= 260 Ω/km, C= 44nF/km). Table 9 and 10 give the programmed values of set- and sidetone impedance for optimum BRL and sidetone gain (Gst) for the PCA1070 with the 12 and 16kHz filter respectively. Gst is the gain from microphone input to earpiece output for an adjusted send gain of +41 and a receive gain of -6dB.

Note: Manufacturer of tax pulse filter coils:
Fastron Bauelemente - Elektronik; Zum Kaiserblick 25; D-83620 Westerham; Germany
Tel (08063) 9935; Fax (08063) 6188

Country	Zref	Ra	Rb	fp	BRL	Rsa	Rsb	Cs	Gst
-	-	Ω	Ω	Hz	dB	Ω	Ω	nF	dB
A, F, IS IRL, P, E	600Ω	600	0	12k	15.9	295	803	145	23
F(complex) 1)	210Ω+880Ω//150nF	300	800	1448	26.3	193	1259	186	13
S	275Ω+850Ω//150nF	400	700	1448	27.9	221	1259	186	14
CH	220Ω+820Ω//115nF	300	700	2533	28.4	153	1259	145	13

1) According to CNET specification Ref. [14]

TABLE 9 Survey of Zref, programmed set- and sidetone impedances, minimum BRL and maximum Gst in the frequency range 300-3400Hz for PCA1070 with 12kHz tax pulse filter.

Country	Zref	Ra	Rb	fp	BRL	Rsa	Rsb	Cs	Gst
-	-	Ω	Ω	Hz	dB	Ω	Ω	nF	dB
B, CY, SF	600Ω	600	0	12k	14.5	295	803	166	23
B(PABX)	150Ω+830Ω//72nF	200	800	5859	26.2	153	1259	145	14
D	220Ω+820Ω//115nF	300	700	2533	27.2	153	1259	166	13
GR	1000Ω //100nF	0	1k	3350	21.8	153	1410	166	6
N	120Ω+820Ω//110nF	100	800	3350	32.4	153	1410	166	10

TABLE 10 Survey of Zref, programmed set- and sidetone impedances, minimum BRL and maximum Gst in the frequency range 300-3400Hz for PCA1070 with 16kHz tax pulse filter.

7 Hints for printed circuit board layout

Care must be taken to avoid that the large line current flows into common ground tracks to which sensitive points (such as amplifier inputs) are connected. For this reason special attention should be payed to the transistor PNP_{TX} which conducts almost the complete line current.
The copper tracks connecting the external components to the corresponding IC-pins should be kept as short as possible.

The ground connection of all RFI-capacitors should be realized by means of as large as possible copper planes or grids. RFI-capacitors must be connected as close as possible to the pins that have to be decoupled.

The ground plane on the circuit board must be kept as large as possible where every copper area must be connected to the ground-plane (or grid) on at least two points.

An 'RF-guard' as described in paragraph 4.1 can be realised as shown in Fig. 68 and furthermore the remarks related to this (end of 4.1) should be followed.

An example of a printed circuit board layout for the PCA1070 can be found in Ref. [6].

8 Application example of PCA1070+PCD3353A/008

Figs. 70 and 71 show an application example of the PCA1070 in a feature phone application with a preprogrammed µC (the PCD3353A/008).

Functionality of the feature phone:

- Speech/transmission with privacy switch.
- Ringer detection and generation.
- Dialling features including:
  - # Pulse, DTMF and Mixed mode dialling.
  - # Last number redial.
  - # Repertory dialling.
- Settings of programmable parameters are stored in an EEPROM.
- Changing of programmable parameters via keyboard.

The PCD3353A/008 preprogrammed µC has four functions:

- Control of the normal feature phone functions such as: pulse/tone dialling, redial/repertory dialling and software controlled ringer function.
- Setting of the transmission parameters of the PCA1070, which are stored in the on-chip EEPROM, via the I^2C-bus.
- Setting of the dialling and ringer parameters which are also stored in the EEPROM.
- Changing of all the programmable parameters via keyboard to show the flexibility of the total application.

For software program details is referred to Software Specification PCD3353A/008 in Ref. [5].
The complete application is described extensively in Ref. [6].

In the following paragraphs measurement results of this application are given. Settings of programmable values and test conditions are the same as in paragraph 3.2 unless otherwise noted.

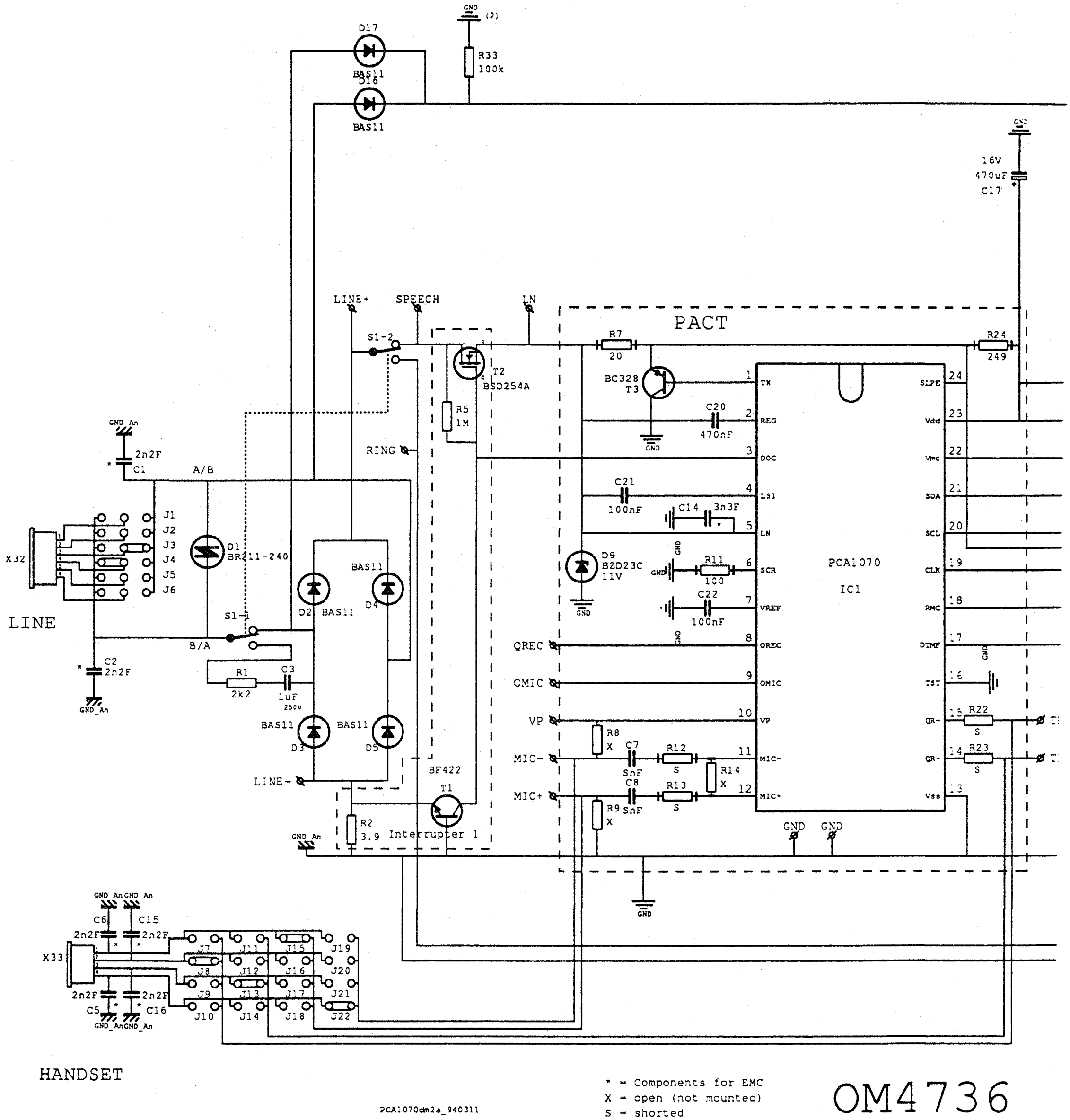


Figure 70 Circuit-diagram of feature phone application PCA1070 + PCD3353A/008 (left hand part)

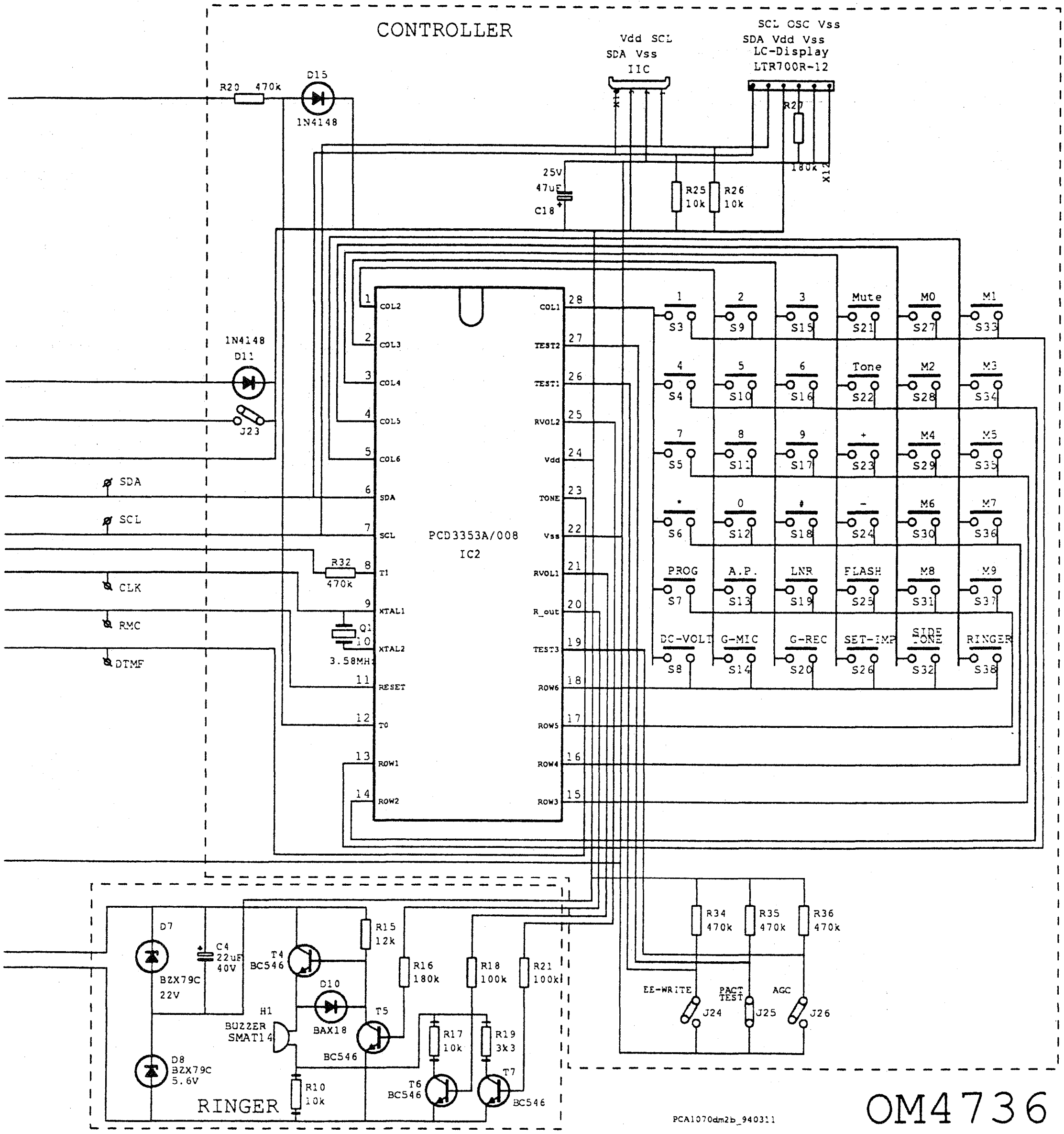


Figure 71 Circuit-diagram of feature phone application PCA1070 + PCD3353A/008 (right hand part)

8.1 DC characteristics

Fig. 72 shows the DC voltage at the telephone line (tip-ring or Vab) versus the line current with the programmed voltage V_{SLPE} as a parameter.

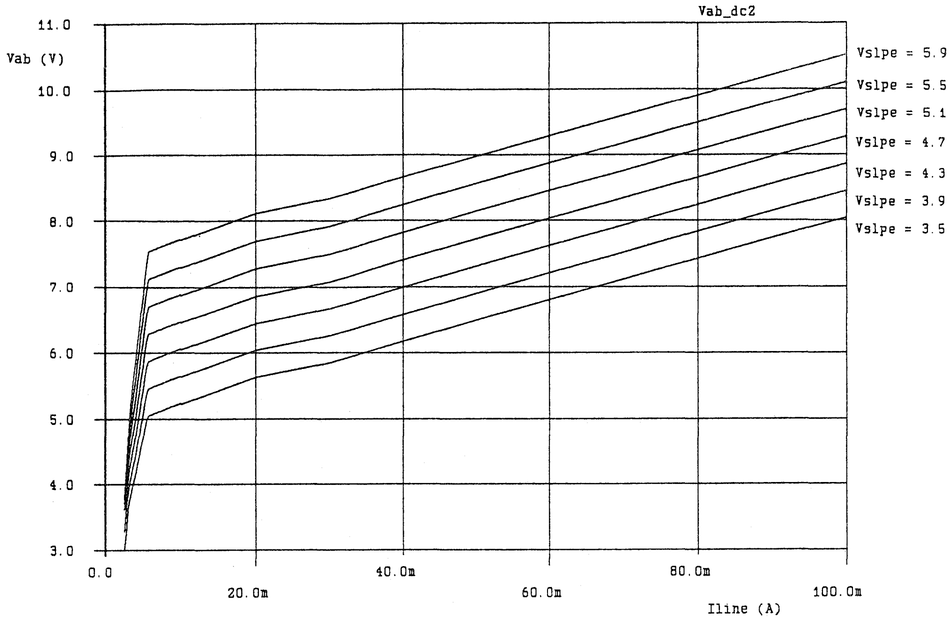


Figure 72 DC characteristics Vab versus I_{LINE} at V_{SLPE}=3.9-5.9V

8.2 Balance Return Loss

Fig. 73 gives the BRL with a programmed set impedance Zs=200Ω+(800Ω//104nF) according to the German requirements. The reference impedance is Zref=220Ω+(820Ω//115nF). Because EMC capacitors are used in this application at the a and b lines and between pins LN and VSS of the PCA1070, the BRL at high frequencies is somewhat lower than for the PCA1070 alone. However BRL requirements are met with quite a large margin.
Fig. 74 shows the BRL when the set impedance is reprogrammed according to the GB requirement and Fig. 75 shows the BRL for 600Ω impedance.

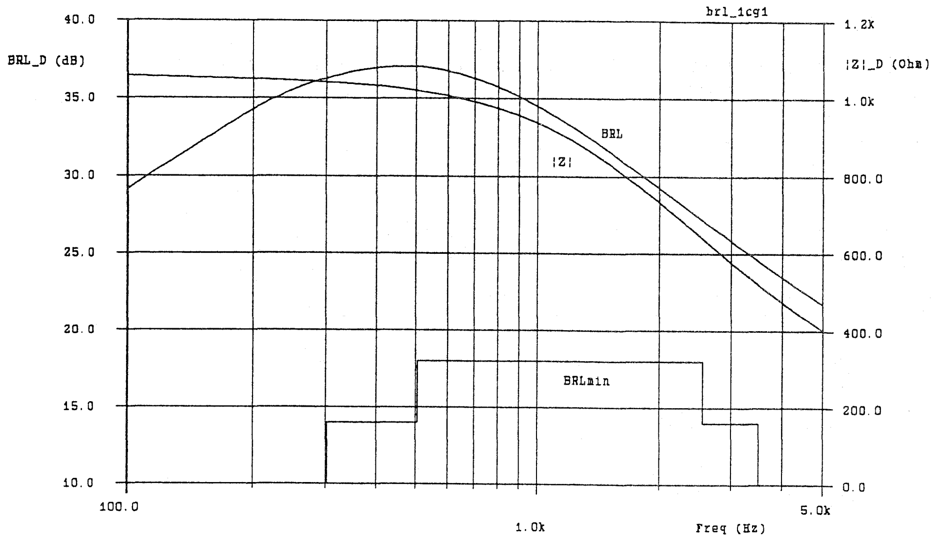


Figure 73 Balance return loss Application against Zref for D Z_D=(220Ω+[820Ω//115nF]; settings Ra=200Ω, Rb=800Ω, fp=1915Hz

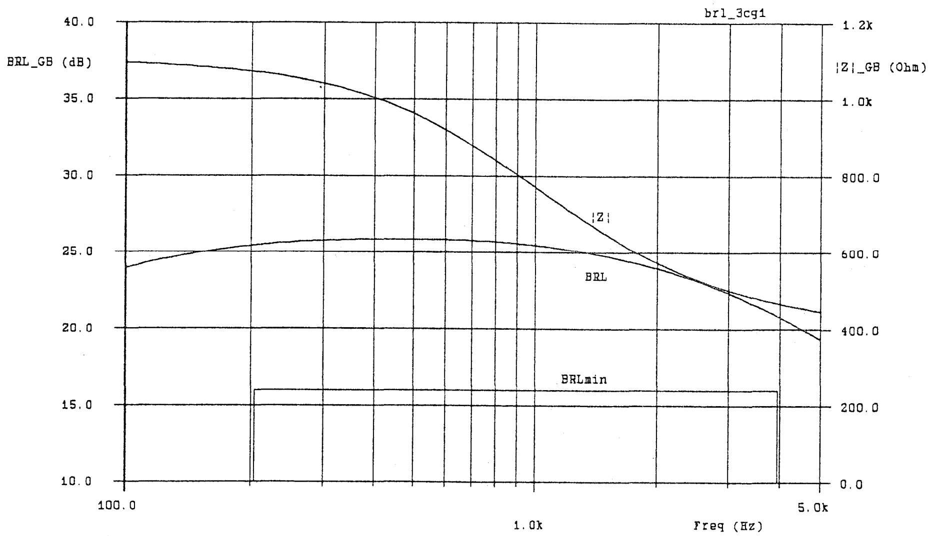


Figure 74 Balance return loss Application against Zref for GB Z_{GB}=(370Ω+[620Ω//310nF]; settings Ra=400Ω, Rb=600Ω, fp=828Hz

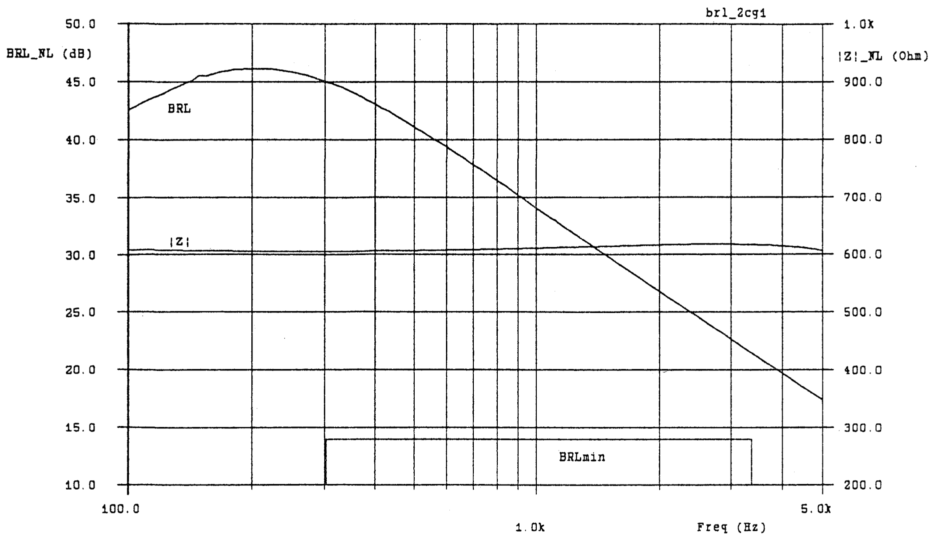


Figure 75 Balance return loss Application against Zref for NL (600Ω); settings Ra=600Ω, Rb=0Ω, fp=12kHz

8.3 Frequency characteristics

The frequency response between microphone inputs and the line (a/b) outputs is shown in Fig. 76. Programmed gain of the microphone channel is 41dB (Gma=15dB). The line load at the a/b is the German reference impedance.

In Fig. 77 the frequency characteristic between line and the earpiece outputs (150Ω symmetrical load) is given. The receive signal is injected onto the a/b terminals by means of a voltage source in series with the German reference impedance. The gain setting of the receive channel is G_{RS}=-6dB (Gra=-6dB).

The frequency characteristic between microphone inputs and earpiece outputs (150Ω symmetrical load) is shown in Fig. 78. This represents the electrical sidetone gain. Because the microphone gain has been set to 41dB and the receive gain to -6dB, the sidetone gain would be +35dB in case the sidetone suppression would be zero. During the measurement the line has been loaded with an artificial line (0km and 2km 0.4mm diameter) of 260Ω/km and 44nF/km terminated with 220Ω+(820Ω//115nF). The sidetone suppression has been optimized for 2km line length by programming the sidetone balance impedance of the PCA1070 to Zoss= 193Ω+ (1410Ω//121nF).

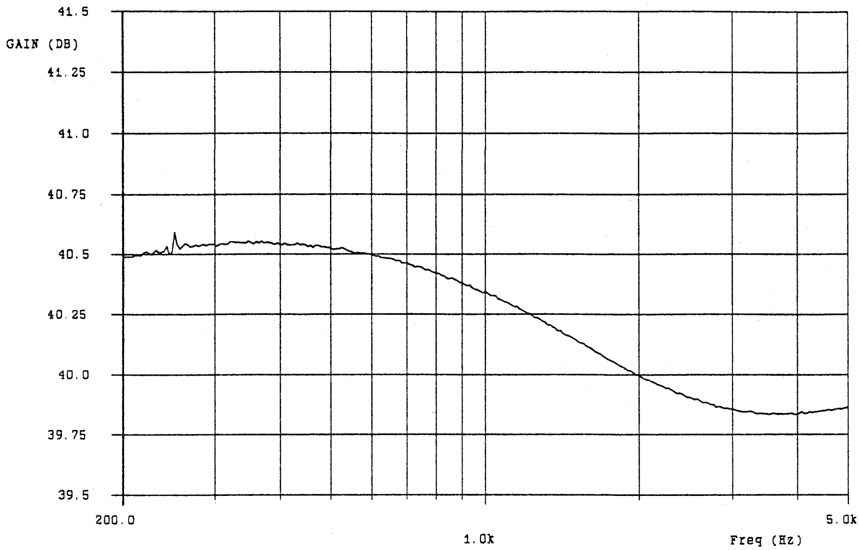


Figure 76 Frequency characteristic microphone amplifier; Z_{LINE}=(220Ω+[820Ω//115nF], G_M=41dB

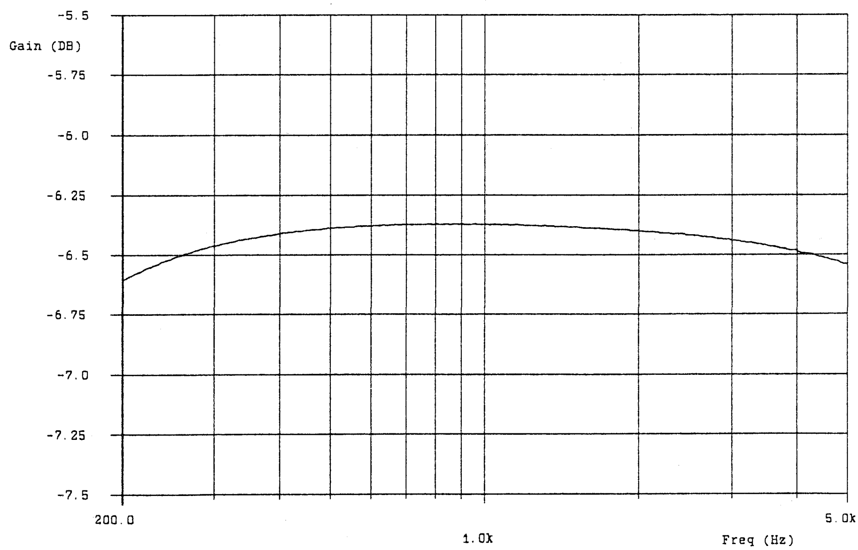


Figure 77 Frequency characteristic receive channel; G_{RS}=-6dB

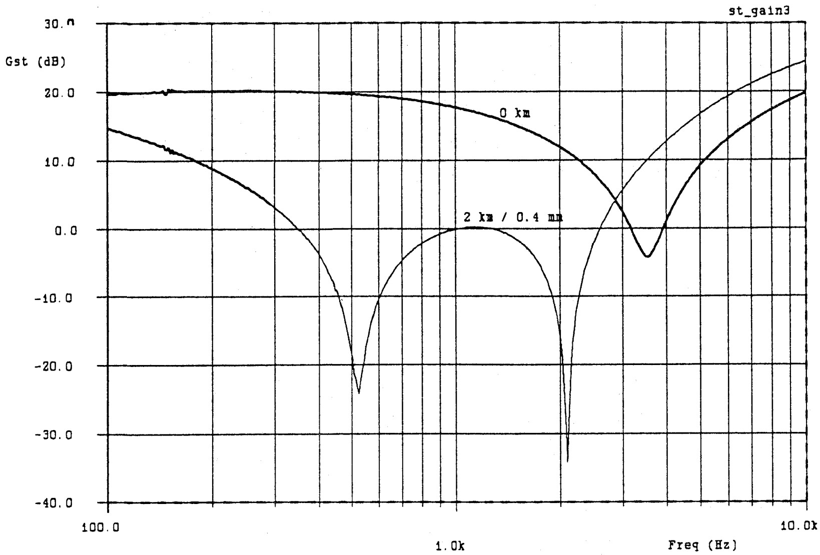


Figure 78 Frequency characteristic of the electrical sidetone (mic inputs to tel outputs) with 0km and 2km LINE (0.4mm diameter) terminated with 220Ω+[820Ω//115nF]; Settings Zoss=193Ω+[1410Ω//121nF]

8.4 Pulse dialling

In Fig. 79 the voltage before the interrupter (V_{SPEECH}) and the voltage at LN of the PCA1070 during pulse dialling are shown. Also the supply voltages V_{DD} and V_{VMC} are shown. Keys "0" and a "1" were pressed. At the start of the dialling procedure the µC switches the DST bit to "1" to improve the switching behaviour of the system. The dialling is controlled fully by the µC by sending the appropriate signals to the PCA1070 via the I^2C-bus. The SW protocol is described in Ref. [5].

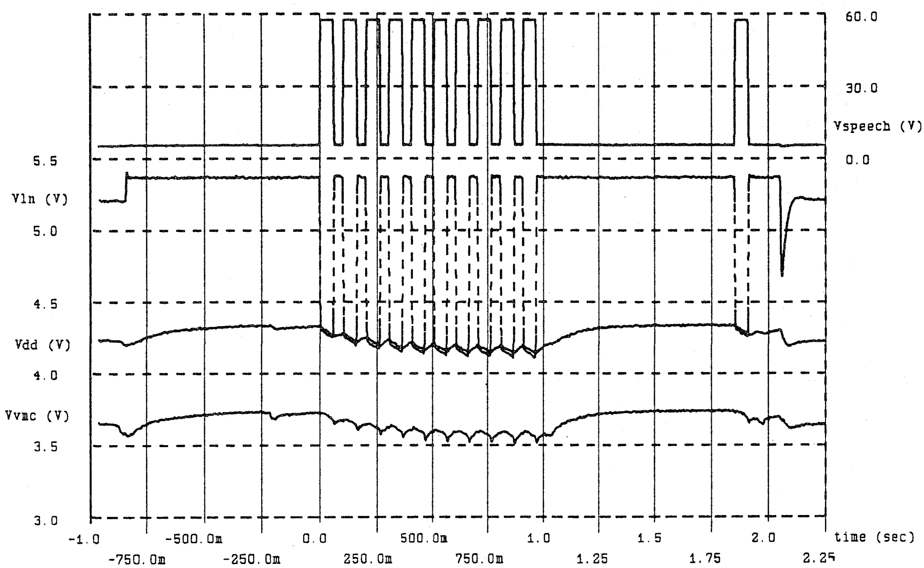


Figure 79 Line voltage V_{SPEECH}, V_{LN} and supply voltages V_{DD} and V_{VMC} during pulse dialling; Vexch=60V, Rexch=1000Ω, R_{LINE}=1520Ω

8.5 Start-up/Switch-off

8.5.1 Hook-off

The start-up behaviour after hook-off has been measured with a 60V supply voltage with a series resistor of 2520Ω. This results in a line current of approximately 20mA.
Fig. 80 shows the voltages on SPEECH, VDD, SDA and RMC on a timescale between 0 and 350ms and Fig. 81 shows an enlargement of the time frame between 74 and 86ms. This gives a good view on the activity on the I^2C-bus during the start-up phase. At roughly 76ms the I^2C-bus becomes active and the µC loads all programmable parameters into the PCA1070. This is done with the DST-bit set to "1" to activate the DC start circuitry in the PCA1070 to minimize the settling time. Then after about 100ms (t≈180ms) all programmable parameters are loaded once more into PCA1070 with DST="0". Then the line current is read via I^2C-bus and the gain of the microphone channel and the receive channel are adapted to match the connected line (line current dependent gain control see paragraph 8.6). The PCA1070 is then in normal operating condition.

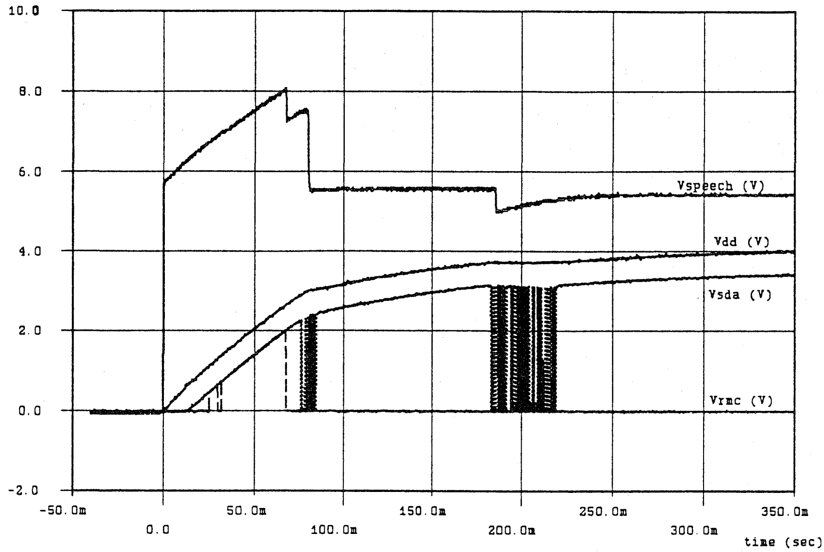


Figure 80 Start-up behaviour

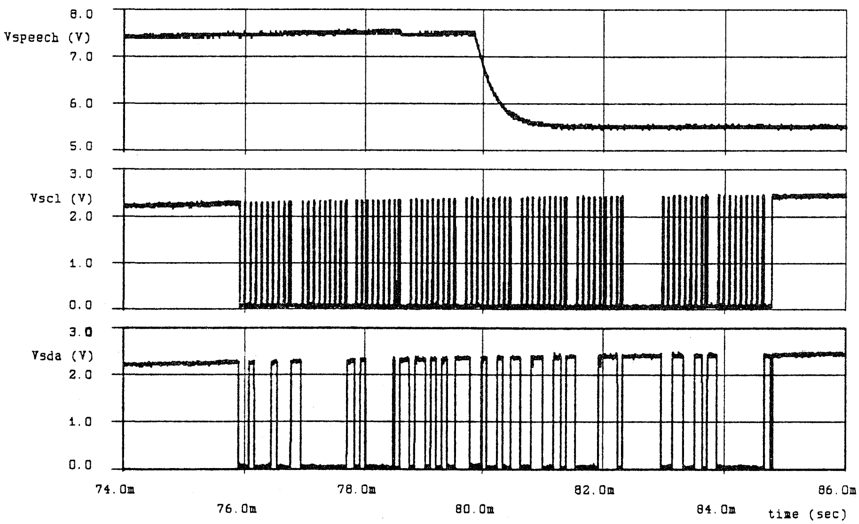


Figure 81 Start-up behaviour: enlarged view on I^2C bus voltages and line voltage

8.5.2 Hook-on

In Fig. 82 the switch-off behaviour is shown. When the hook-switch is switched from "SPEECH" into "RING" position at t=0, the µC will detect a line break and switches the PCA1070 in power down mode via I^2C-bus. V_{DD} will decrease then slowly. After a waiting time (the reset delay time) of 160ms the µC switches the PCA1070 into normal operating condition again and the µC goes into stand-by mode. The voltages V_{LN}, V_{DD} now decrease rapidly and the reset for the µC (pin RMC) becomes active at $V_{VMC}=V_{SDA}=2.15V$ typically.

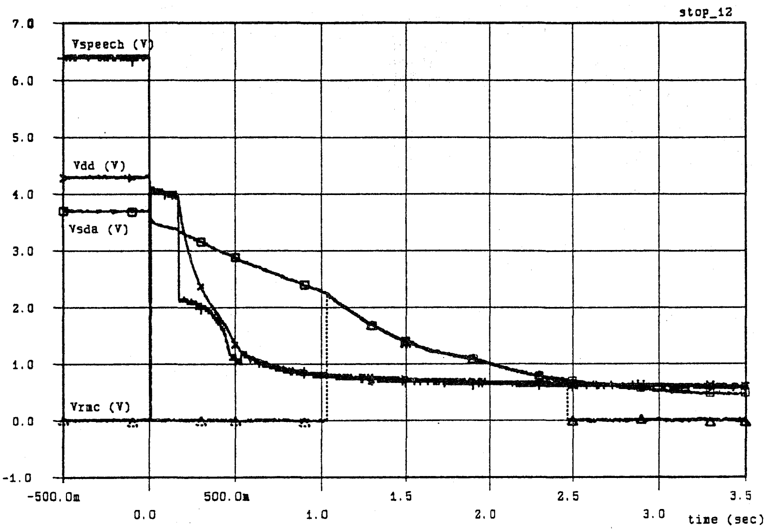


Figure 82 Switch-off behaviour

8.5.3 Ringer start-up

In Fig. 83 the start-up of the software controlled ringer system in ringing condition is shown.
The supply voltage at the µC (VMC) is built up by the ringer burst and the PCA1070 provides the reset signal. At V_{VMC}=2.15V, the reset signal at pin RMC changes from "1" to "0" and the µC starts the ringer procedure. The frequency of the incoming signal at pin T0 is tested and a melody is generated at the TONE output to drive the buzzer. The incoming signal at pin T0 is related to the rectified line signal V_{LINE_DC} in fig. 83.

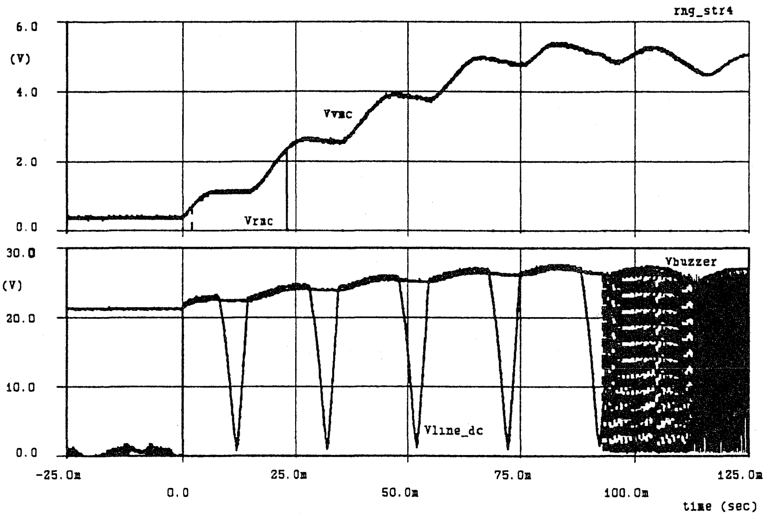


Figure 83 Ringer start-up

8.6 Line current dependent gain control

In this application line current dependent gain control is used for automatic line loss compensation (also knows as Automatic Gain Control or AGC). The characteristics are determined by software. The gain control characteristic of both the microphone and earpiece amplifiers can be programmed under keyboard control. As shown in Ref. [5] a start current (dc_i_0), 5 current values (dc_i_1 to dc_i_5), a stop current (dc_i_6) and the desired step resolution of the gain control (dc_i_d_G) can be programmed.
In Fig. 84 the gain control characteristic has been programmed to obtain optimum tracking with the line attenuation and the required amplifier gain for a system with a 2 x 300Ω feeding bridge and 48V exchange supply voltage. The gain control range of the microphone channel and the receive channel has been set to 6dB (6 steps of 1dB each). This corresponds to a line length of 5km of 0.5mm diameter copper twisted pair cable which has a resistance of 176Ω/km and an average AC attenuation of 1.2dB/km. Fig. 84 shows the ideal (desired) gain control curve, the measured curve (increasing I_{LINE}) and the difference between both representing the deviation.

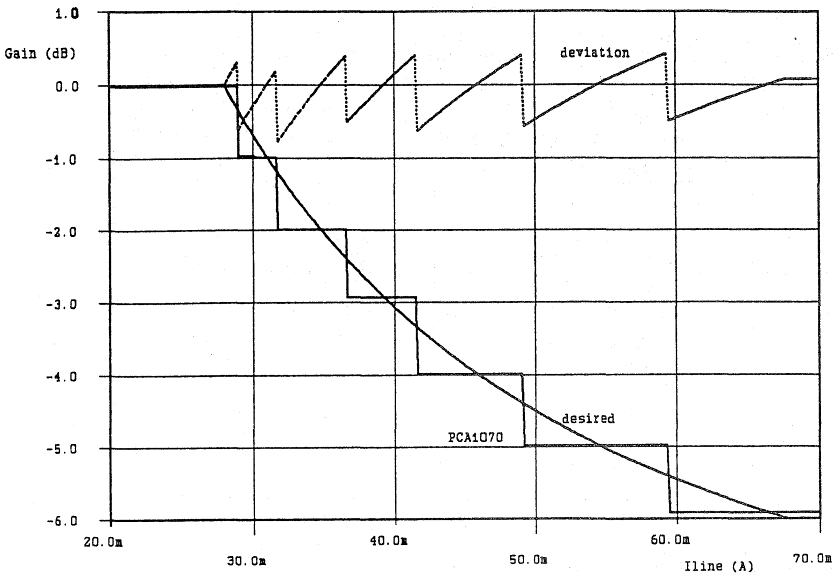


Figure 84 Gain control of microphone amplifier versus LINE current; Vexch=48V, Rexch=600Ω, LINE: 176Ω/km, 38nF/km, AC attenuation 1.2dB/km

8.7 Acoustical characteristics

The loudness ratings have been measured according to the German requirements of the Deutsche Telekom (Ref. [12]). A handset with dynamic 250Ω capsules has been used (see also Ref. [6]). The gain settings were Gma=+23dB (G_M=48dB), Gra=-3dB (G_{RS}=-3dB). The programmed set impedance Ra=200Ω, Rb=800Ω, fp=1915Hz. The sidetone balance impedance: Rsa=193Ω, Rsb=1410Ω, Cs=121nF.

The measurement results are shown in Table 11. The sidetone is measured with 35 different lines and the final result that indicates sidetone performance is a calculated figure (using weighing factors for the results of the 35 lines).

For clarity the results are summarized here:

The send loudness rating SLR=3.69dB (required +3dB ± 3dB).

The receive loudness rating RLR=-8.4dB (required -9dB ± 3dB).

The weighted and corrected sidetone mask rating STMR=17.33dB (required >=14.5dB)

Loudness Rating nach CCITT Rec. P79, P65		
Fernsprecher: PCALE OM4736 PACT Datum: 93.04.20 Speisung: 60V / 2 • 500Ω Kabelnachbildung: BOSSE LNB 200		
STMR-Berechnung nach FTZ **SLR: 3.69dB** XQ: 3.00dB **RLR:-8.40dB** XQ:-9.00dB KF:-1.29dB STMR-FeAp gesamt: 18.62dB **Korrigierter Wert:17.33dB** Sollwert: >=14.5dB	STMR_1 =10.38 dB STMR_2 =13.77 dB STMR-3 = 9.78 dB STMR-4 =15.94 dB STMR-5 =20.20 dB STMR-6 =13.47 dB STMR-7 =15.87 dB STMR-8 =13.08 dB STMR-9 =20.45 dB STMR-10=22.90 dB STMR-11=20.53 dB STMR-12=18.76 dB STMR-13=19.42 dB STMR-14=22.49 dB STMR-15=26.48 dB STMR-16=26.69 dB STMR-17=20.42 dB STMR-18=24.67 dB	STMR_19=21.05 dB STMR_20=24.08 dB STMR-21=22.22 dB STMR-22=20.39 dB STMR-23=23.80 dB STMR-24=19.76 dB STMR-25=21.64 dB STMR-26=18.50 dB STMR-27=14.16 dB STMR-28=14.45 dB STMR-29=17.76 dB STMR-30=15.01 dB STMR-31=13.80 dB STMR-32=12.59 dB STMR-33=13.54 dB STMR-34=13.35 dB STMR-35=12.93 dB ***SEALED***

TABLE 11 Loudness Ratings for Germany

8.8 Immunity to R.F. signals (EMC)

Programmed settings:

- DC voltage: V_{SLPE}=5.9V
- Set impedance 200Ω+(800Ω//104nF); fp=1915Hz
- Sidetone impedance: 492Ω+(1259Ω//134nF)
- Microphone channel: gain = 39dB
  dynamic limiter threshold: DLT=1
- Earpiece channel: gain = +3dB
  Load select bit: RFC=1
  Hearing Protection Level: HPL=1

PCB layout: Double layer; virtually homogeneous ground-plane on components side (see Ref. [6]).

EMC components: 6 capacitors of 2.2nF at all in and outgoing lines (handset cord, base cord) to the common ground plane. One capacitor of 3.3nF between LN and VSS of PCA1070.

Test set-up:

- The German current injection method (common mode signal on a/b lines) is used (VDE 0878 part 200).
  This method is described in Ref. [10].
- Dummy microphone: 82nF
- Dummy earpiece : 82nF
- Frequency range: 100kHz-200MHz.
- RF levels:
  Normal requirement: 3Vrms 100kHz-30MHz, 0.5Vrms 30-200Mhz.
- Amplitude modulation: 80%, 1kHz.
- Line current: 27mA.

Requirements:

- The allowed demodulation level towards the A/B lines or earpiece can be calculated from the required 40dB signal to interference ratio and the nominal line signal level. With a nominal signal of 100mV (send or receive, 1kHz) across the telephone line, this corresponds to a demodulation level of 1mV.
  With respect to 0dB/775mV this 1mV corresponds to -57.7dB in the graphs of the A/B lines results, to -55dB for the earpiece (measured receive gain is +2.7dB).

Test results:

- Are given as demodulated signal levels, with respect to 0dB/775mV, on the A/B line terminals of the PCB (Fig. 88), across the earpiece in the handset (Fig. 89).

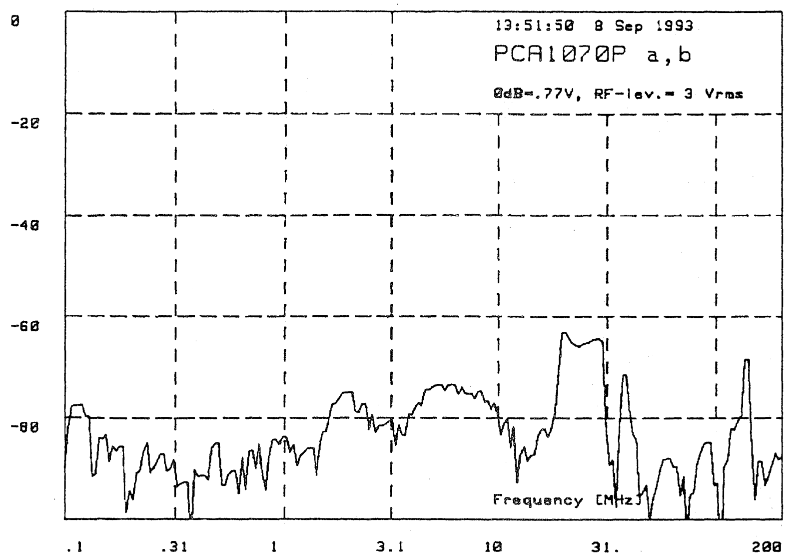


Figure 85 EMC characteristics: demodulated signal at the a/b lines (requirement: ≤-57.7dB[0.775V])

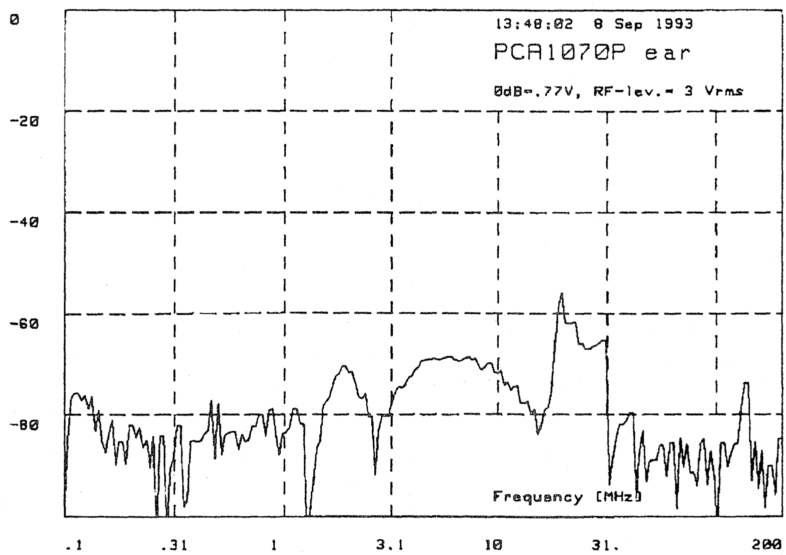


Figure 86 EMC characteristics: demodulated signal at the earpiece (requirement: ≤-55dB[0.775V])

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10 Acknowledgement

The authors thank the people from the PACT team (Rolf Becker, Jaap Bolt, Igor Ferencic, Stephan Haag, Gabor Hirsch, Eric De Mey, Jaap Mulder, Urs Voegeli, Alan Waller and Zhenhua Wang) for their work during the development of the PCA1070. Also many thanks to Mrs. Judit Munkacsy for her work on the layout of this circuit.

Appendices

APPENDIX A Overview PCD335x(A) µC family

Type	ROM	RAM	OTP	I/O	EEPROM	Features	Package
PCD3350A	8k	256	-	34	256	clock	QFP44
PCD3351A	2k	64	-	20	128	-	DIL/SO28
PCD3352A	4k	128	-	20	128	-	DIL/S028
PCD3353A	6k	128	-	20	128	-	DIL/SO28
PCD3354A	8k	256	-	36	256	-	QFP44
PCD3755A	-	128	8k	20	128	-	DIL/SO28

PCD335x(A) family of telecom and low-voltage µC
All have DTMF, Melody output, 8048 instruction set based, V_{DD} = 1.8V to 6.0V
(For DTMF and EEPROM programming V_{DD} = 2.5V to 6.0V).

APPENDIX B List of abbreviations

a/b	Line terminals (tip-ring)
A/D	Analog to digital
AGC	Automatic Gain Control (= automatic line loss compensation)
BRL	Balance Return Loss
BTL	Bridge Tied Load (symmetrical drive of load impedance)
CCITT	The International Telegraph and Telephone Consultative Committee
CLK	Clock signal input
CNET	Centre National d'Etudes des Télécommunications
C_{LSI}	Capacitor between LN and LSI
Cp	Internal Capacitor between LN and VSS
Cpt	Parallel Capacitor of Tax pulse filter
C_{REG}	Capacitor between LN and REG
Cs	Capacitor of sidetone balance impedance Zoss
Cst	Series Capacitor of Tax pulse filter
C_{VDD}	Capacitor between VDD and VSS
C_{VMC}	Capacitor between VMC and VSS
C_{VREF}	Capacitor between VREF and VSS
D	Germany
D1	Diode between VDD and VMC
dBA	Acoustical level, A-curve weighted
DC	Direct Current
DLT	Dynamic Limiter Threshold (bit)
DMO	Dial Mode Output
DOC	Dial Output Connection
DPI	Dial Pulse Input (bit)
DST	DC Start Time (bit)
DTMF	Dual Tone Multi Frequency
EEPROM	Electrical Erasable Programmable Read Only Memory
EMC	Electro Magnetic Compatibility
ESD	Electro Static Discharge
f	Frequency
fax	Facsimile
f_{CLK}	Clock Frequency
fp	Pole Frequency of set impedance Zs
f_{SCL}	Serial Clock Frequency
GB	Great Britain
G_{CTs}	Confidence Tone gain for Symmetrical drive (BTL)
G_{DTMF}	DTMF Gain
G_M	Gain Microphone channel
Gma	Gain send prog-amp
Gra	Gain receive prog-amp
G_{RA}	Gain Receive channel for Asymmetrical drive (SEL)
G_{RS}	Gain Receive channel for Symmetrical drive (BTL)
Gst	Sidetone Gain between microphone inputs and earpiece outputs
Gsupp	Sidetone suppression
HPL	Hearing Protection Level (bit)
I^2C	Inter IC connection
IC	Integrated Circuit

IEC	International Electrotechnical Commission
I/O	Input and/or Output
I_{DD}	Internal supply current
I_{LINE}	Line Current
I_{LN}	Current into pin LN
I_P	Current to external Peripheral circuitry
I_{SCR}	Current out of pin SCR
I_{SLPE}	Current into pin SLPE
I_{VMC}	Current into pin VMC
I_{VP}	Current out of pin VP
I_{PNP}	Current in PNP from collector to emitter
LC	Inductor and Capacitor
LI	Line Interface
LN	Positive Line terminal
LSI	Line Signal Input
Leq	Equivalent Inductor
Lpt	Parallel Inductor of Tax pulse filter
Lst	Series Inductor of Tax pulse filter
MCOR	Microcontroller On Reset
MCreset	Microcontroller reset
MIC-	Inverting input of Microphone preamp
MIC+	Non-inverting input of Microphone preamp
MNex	External N-MOST
NL	Netherlands
N-MOST	N-channel Metal Oxide field effect Transistor
NSA	Nummern Schalter Arbeitskontakt (also known as MUTE2 or DMO)
OMIC	Output Microphone preamplifier
OREC	Output Receive preamplifier
OTP	One Time Programmable
PABX	Private Automatic Branch eXchange
PCB	Printed Circuit Board
PD	Power Down
PDx	Power Down bits
P-MOST	P-channel Metal Oxide field effect Transistor
PNP_{TX}	PNP transistor at TX
POR	Power On Reset
PRES	Pact Reset (bit)
prog-amp	programmable amplifier
PSR	Power Supply Rejection
PTT	Telephone company
QR-	Inverting output of BTL receive output amplifier
QR+	Non-inverting output of BTL receive output amplifier
REG	Voltage Regulator decoupling
RFC	Resistive-not / Capacitive load select (bit)
RFI	Radio Frequency Interference
RF	Radio frequency
RG	internal logical signal PCA1070 indicating supply from an external source
RLR	Receive Loudness Rating
RMC	Reset output for Micro Controller
RM	Receive Mute

RRG	Reset RG (bit)
R/Wn	Read / Write-not
Ra	Resistor Ra of Zs
Rb	Resistor Rb of Zs
R_{DOC}	Resistor between DOC and LINE
Rexch	Feeding bridge Resistor in exchange
R_{LSI}	Internal resistor between LSI and VSS
Rm	Resistor between MIC+ and MIC-
Rp	Internal resistor between REG and VSS
Rsa	Resistor Rsa of Zoss
Rsb	Resistor Rsb of Zoss
R_{SCR}	Resistor between SCR and VSS
Rset-dc	dc-resistance of the telephone set
R_{SLPE}	Resistor between LN and SLPE
R_{SUP}	Resistor between SLPE and VDD
Rt	Resistor between QR+ and QR-
SCL	Serial Clock Line of I^2C-bus
SCR	Sending Current Resistor
SC	Switched Capacitor
SDA	Serial Data Line of I^2C-bus
SEL	Single Ended Load (asymmetrical drive of load impedance)
SLPE	Slope (DC resistance)
SLR	Send Loudness Rating
SM	Send Mute
S/N	signal to Noise ratio
STMR	Side-Tone Mask Rating
SW	Software
Tamb	Ambient Temperature
THD	Total Harmonic Distortion
TONE	Output pin of DTMF generator on μC
TST	Testpin
TX	Drive output of DC voltage stabilizer
Vab	Voltage at line terminals a/b
VBGAP	Bandgap voltage
VDD	Positive supply decoupling
VDD/VMC	VDD connected to VMC
VDE	Verband Deutscher Elektrotechniker
VMC	Input to sense supply voltage Micro Controller
V_{RING}	Voltage at point RING
VP	Supply for electret microphone
VREF	Voltage Reference decoupling
VSS	Negative line terminal
Vbe	Voltage between base and emitter of transistor
V_{CREG}	Voltage across C_{REG}
V_{CLK}	Voltage at pin CLK
$V_{DD}/2$	Voltage at VDD divided by 2
Vdiode	Voltage across a diode
Vexch	Supply Voltage in exchange
V_{LN}	Voltage at pin LN
$V_{LN}p$	AC peak voltage at pin LN

V_{LN}p-p	AC peak to peak voltage at pin LN
V_{LSI}	Voltage at pin LSI
V_{MIC}	AC Voltage between MIC+ and MIC-
Vp-p	Peak to peak voltage
Vrec	AC Receive Voltage
Vrec'	AC receive voltage in exchange
Vsend	AC Sending Voltage
V_{SLPE}	Voltage at pin SLPE
V_{SLPE}p-p	AC peak to peak voltage at pin SLPE
V_{SPEECH}	Voltage at point SPEECH
V_{SS}	Voltage at pin VSS
V_{VMC}	Voltage at pin VMC
Z_{LINE}	Telephone line impedance
Z_{LN}	Impedance between LN and VSS
Zoss	Balance impedance for Optimum Sidetone Suppression
Zo_{VP}	Output impedance of electret supply VP
Zref	Reference impedance of telephone line
Zs	Set impedance (programmable)
Zs'	Replica from Zs
µC	Micro Controller

APPLICATION NOTE Nr	ETT8707
TITLE	TEA1081: a supply IC for Peripheral Circuits in Electronic Telephone Sets
AUTHOR	F. van Dongen
DATE	October 1987

CONTENTS.

FIGURES.

1. INTRODUCTION.

The supply provisions of most of the transmission circuits are limited up to about 3 mA in combination with 3 V supply voltage, sufficient to supply a DTMF generator or a micro controller with its additional devices.

Feature phones with listening-in and often combined with extended dialling facilities require relatively high supply currents up to 20 mA at a minimum voltage of 3 V. When such sets have to be supplied from the telephone line the TEA1081 can be applied.

The TEA1081 performs the interface between telephone line and the peripheral circuits to be supplied. It can be connected directly to the line because of its high input impedance for audio frequencies.
However, the TEA1081 input contains no voltage regulator to determine the DC voltage over the line, so this IC has to be used in combination with a transmission circuit. The TEA1081 is provided with a power down facility to isolate this IC from the total application during pulse dialling.

The TEA1081 is a successor of the TEA1080 , the differences between them are described in this report.

Application examples of the TEA1081 in combination with the TEA1060 are given, in particular the effect of the TEA1081 application on the transmission characteristics of the set and the behaviour of the set during pulse dialling with and without use of power down facility of the TEA1081.

In general, the members of the TEA1060 family , are transmission circuits which perform the interface between microphone, receiver transducer, dialler IC (dtmf or pulse dial) and telephone line in electronic telephone sets. The ICs of TEA1060 family are fully described in ref. 1.

2. CIRCUIT DESCRIPTION OF THE TEA1081.

2.1. Block diagram and pinning.

The block diagram of the TEA1081 is shown in fig.1, while the pinning is given in fig.2.

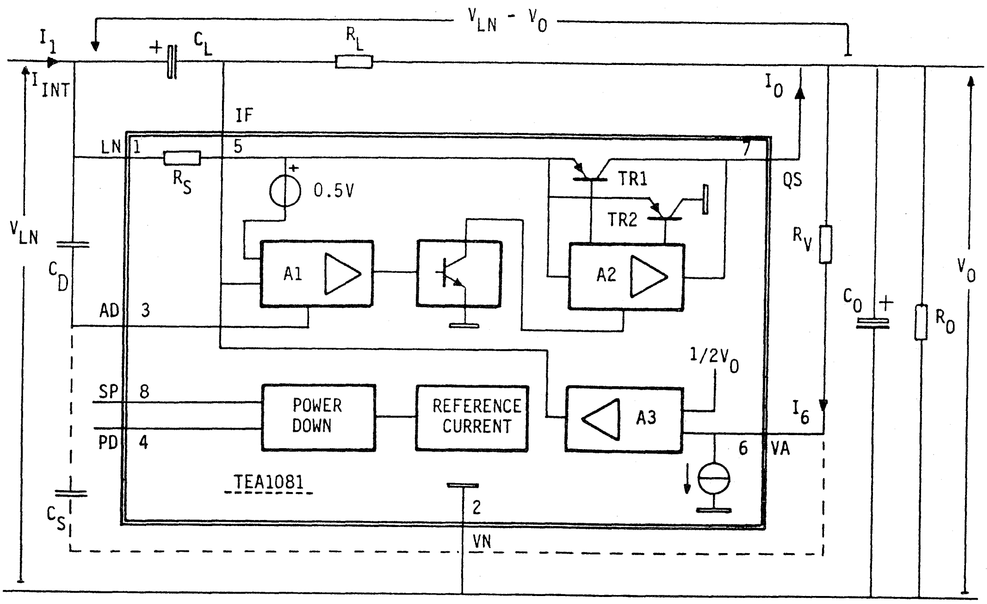


Fig.1 Block diagram of the TEA1081 including ext. components.

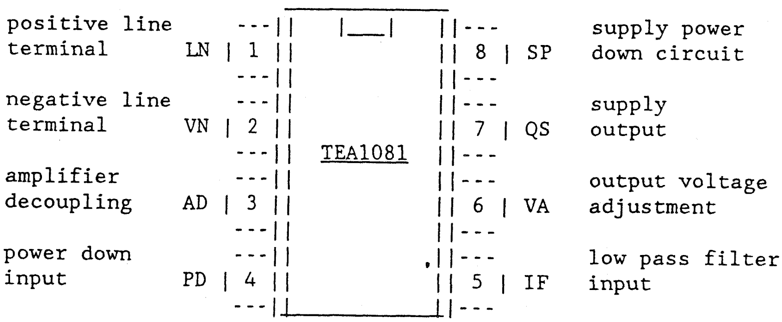


Fig. 2 Pinning diagram of the TEA1081

The internal functions according to fig.1 are:

- Amplifier A1 in combination with R_S and an external R and C to fulfil an inductor function.
- Amplifier A2 which controls both transistors TR1 and TR2 to provide low signal distortion up to high AC line levels. The supply current flows from input LN via R_S and TR1 to output QS. At large AC line levels a part of the input current will flow via TR2 to VN, as shown in fig.3.
- Amplifier A3 and current I_6 (from pin 6) together with an external R to provide a limited output voltage.
- Power down circuit to isolate the TEA1081 from the external circuitry when the PD input is activated.
- Reference current source to bias the internal circuitry.

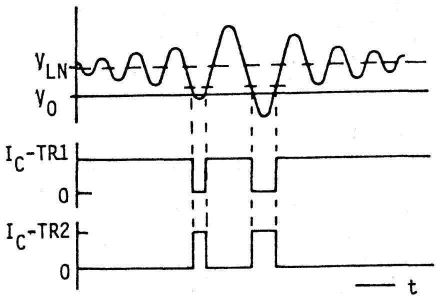


Fig.3 Collector currents of transistors TR1 and TR2.

2.2. Purpose of the external components.

A brief description of the purpose of the components according to fig.1 is given as follows:

C_L, R_L: The inductance value between input and output terminal is determined by the values of C_L, R_L and the internal resistance R_S (typ. 20 ohm). $L_I = C_L . R_L . R_S$ (H). Typical values for C_L and R_L are respectively 2.2 uF and 100 kohm. A high inductance value improves the BRL but will result also in a long start time.

R_V: Without R_V the output voltage V_O will follow the line voltage V_{LN} with a voltage drop between input LN and output QS.
With R_V in the application V_O will be limited at a max. DC level of $V_O = 2.I_6.R_V$ (V), with $I_6 = 20$ µA typ.

C_O: This capacitor serves as a storage capacitor. An optimum value for C_O has to be chosen for a minimum output ripple and short start time.

C_D, C_S: These capacitances are required to ensure stability of the TEA1081. Both are necessary for output currents $I_O < 1.5$ mA. C_S can be omitted at $I_O > 1.5$ mA. $C_D = 68$ pF and $C_S = 47$ pF.

Another possibility to ensure stability at $I_0 > 0$ mA is to put a series network of 68 pF and 10 kohm between LN and AD . The stability components have to be mounted close to the IC.

2.3. DC behaviour.

2.3.1. Input voltage V_{LN}.

The TEA1081 has to be applied in parallel with a transmission circuit, or other device, which stabilizes the DC input voltage. The max. input voltage is 12 V which includes the DC and AC-peak value.

2.3.2. Input current I_1.

The input current depends on the output current, however it will increase with the mean level of the input signal, in most cases a speech signal.
The available line current has to be sufficient to supply the TEA1081 supply part as well as the transmission application.

2.3.3. Output voltage V_0.

Fig.4 shows a simplified circuit diagram of the TEA1081 to explain the behaviour of the output voltage in relation with R_V. The part within the broken lines is only in use when R_V is applied.

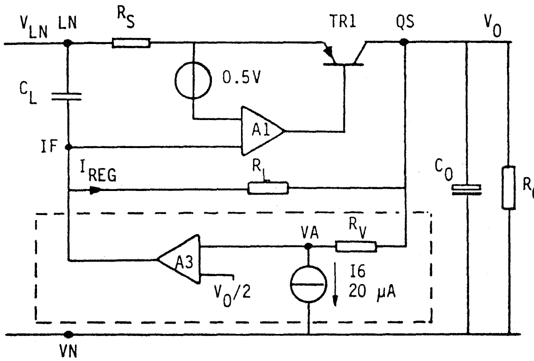


Fig.4 Simplified supply diagram.

Without R_V the voltage drop between LN and QS equals the 0.5 V bias voltage of amplifier A1 (at $I_0 = 0$ mA) due to the DC coupling between QS and IF via R_L. The total voltage drop between LN and QS is a result from this bias voltage and the voltage drop across R_S. The output voltage is:

$$V_0 = V_{LN} - 0.5 - I_1 . R_S \quad (V) \quad , \quad R_S = 20 \text{ ohm typ.}$$

With applied R_V (fig.1) a part of the output voltage ($V_0/2$) is compared with a reference voltage $V_0 - I_6.R_V$ by means of amplifier A3. A stable condition is reached if the reference voltage equals $V_0/2$. The output voltage will be limited at:

$$V_0 = 2.I_6.R_V \quad (V) \quad , \quad I_6 = 20\ \mu A \text{ typ.}$$

The output current I_{REG} (fig.4) causes a voltage drop across R_L and as a consequence an enlarged voltage drop between input LN and output QS.
The input DC level has to be sufficient for this function. The output voltage follows the DC input voltage at low levels, like without R_V. A limited output voltage is achieved (fig.A2) as soon as the input voltage:

$$V_{LN} > 2.I_6.R_V + I_1.R_S + 0.5 \quad (V) \quad \text{or}$$

$$V_{LN} - V_0 > I_1.R_S + 0.5 \quad (V)$$

At a specified DC input voltage and limited output voltage the max. input current is given by:

$$I_1(\text{max}) = (V_{LN} - V_0 - 0.5)/R_S \ (\text{mA}) \text{ with } V_0 = 2.I_6.R_V \ (V)$$

When I_1(max) is exceeded, the output voltage will be not kept at a constant level but will decrease due to the enlarged voltage drop across R_S.
Current I_{REG} has to flow into output load R_0 because the TEA1081 cannot sink current into the output stage. A minimum output current of about 0.2 mA is therefore advised.

2.3.4. Power Down (PD) circuit part.

The TEA1081 is provided with a PD input to reduce input and output current during pulse dialling.
When the PD is activated the reference current (fig.1) will be switched-off. The PD circuit has therefore its own reference circuit and has to be supplied via terminal SP. Terminal SP can be connected to QS in condition when $V_0 >$ V_{SP}min (= 2 V) during line interruptions. When $V_0 < V_{SP}$min, SP has to be connected to an external supply point (e.g. to V_{CC} of the TEA1060).
The logic levels of PD input have to be max. 0.3 V for the non-activated mode and min. 1.5 V for the activated mode of the power down function.
The inputs PD and SP can be left open when power down is not required.

2.3.5. Minimum current consumption.

At PD = LOW the supply current I_{INT}, consumed by the internal circuitry, is about 0.8 mA at V_{LN} = 4 V and will be 1.4 mA at V_{LN} = 12 V, both at I_O = 0 mA. The current I_{INT} will flow from input LN to ground VN.
At PD = HIGH the current drawn from the input pin will be reduced; $I_{INT} < 60\ \mu A$ at V_{LN} = 4 V.
At increased DC input levels the input current will increase mainly depending on the voltage difference between input and output. Fig.A1 shows the input current as well as the output current (drawn from C_O) as function of V_{LN} during power down. As shown in fig.A1 the output current can become positive (depending on V_{LN}) which means that the output capacitor will be charged depending on the load on QS. In general the input current at PD = HIGH can be approximated by:

$$I_{INT} = V_{LN}/150\ (mA) \quad \text{if} \quad V_{LN} - V_O <= 4V \quad \text{while}$$

$$I_{INT} = V_{LN}/150 + (V_{LN} - V_O - 4)/17.5 \quad (mA) \quad \text{if } V_{LN} - V_O > 4\ V$$

2.4. AC behaviour.

2.4.1. General.

The AC behaviour depends on the conditions under which the circuit is applied; these can be summarized as follows:

- circuit impedance, mainly determined by external components.
- input current as a result of output current, signal level and whether R_V is applied or not.
- signal distortion as a function of signal level, output current and DC input voltage.
- noise.

The TEA1081 is designed for minimal signal distortion when the instantaneous line voltage drops below the output voltage. During the time that $V_{LN} > V_O + 0.4$ V the input current will flow into the output load. During the rest of the period at which $V_{LN} < V_O + 0.4$ V the instantaneous input current I_1 will be rerouted to VN. The output capacitor C_O will be discharged by the output load during this time.
For small signal levels the input current measures about the same as the output current. However the mean value of I_1 and the voltage drop $V_{LN} - V_O$ will increase with the signal level due to the rerouted current, especially when R_V is not applied.
In practice the signal levels will vary with mean speech level while DC level V_{LN} is determined by the line current due to the DC slope of the voltage stabilizer of the transmission IC, and of course by the voltage adjustment of the voltage stabilizer.

2.4.2. Circuit impedance.

The circuit can be substituted for audio frequencies by an equivalent parallel network consisting of a conductor $L_I = C_L.R_L.R_S$ (H), a capacitor of about 10 nF (depending on C_D) and a resistor which depends on the output current. In fig.A2 is shown the input impedance (|Z| + phase) at 5 mA and 15 mA output current. The component values of the equivalent network are given too.

2.4.3. Influence of AC line level, ratio k.

The input current I_1 exceeds the output current I_O due to I_{INT}, the base currents of TR1 and TR2 and the internally rerouted current to VN which depends on the level of the line signal.
With or without R_V the ratio $k = I_1/I_O$ is about 1.2 for signal levels up to 200 mV_{rms}. At increased levels the current ratio can be up to a factor 2.
Without R_V factor k = 2 at about 1 V_{rms}, independent of V_{LN} (no signal clipping). The voltage difference between input and output terminal is then maximum:

$$V_{LN} - V_O = 0.5 + 2.I_O.20 \qquad \text{(V)}$$

With applied R_V the difference between input and limited output voltage has also effect on ratio k.
At $V_{LN} - V_O = 1$ V factor k = 2 at 1.2 V_{rms} and I_O = 20 mA.
At $V_{LN} - V_O = 2$ V factor k = 2 at 1.7 V_{rms} and I_O = 20 mA.

Note: The given examples at k = 2 are only useful for continuously signal levels. For speech signals the mean level of the line signal will determine factor k, which will vary with time very much.

2.4.4. Signal distortion, minimum instantaneous line voltage.

The total distortion is a result of the cross-over distortion, when the instantaneous AC signal drops below the DC output level, and the clipping of the AC signal due to the minimum instantaneous line voltage. The distortion (without signal clipping) is less than 2 % up to I_O = 30 mA while the minimum instantaneous line voltage is :

$$V_{LN} \text{ min-inst} = 1.2 + I_1.R_S \text{ (V)}, \quad \text{with } R_S = 20 \text{ ohm typ.}$$

To avoid clipping of the line signal, V_{LN} has to be:

$$V_{LN} > V_{LN} \text{ min-inst} + v_{LN} \text{ peak} \qquad \text{(V)}$$

2.4.5. Noise.

The noise of the TEA1081 psophometrically (P53-curve) measured at the input in parallel with 600 ohm is approximately - 83 dBmp. The noise voltage which is mainly generated by amplifier A1 (fig.1) is independent of C_L, R_L, R_V and I_O.

2.4.6. Start time.

For applications of the TEA1081 with a transmission circuit both circuits will have their contribution to the start time. The start time of the TEA1081-TEA1060 combination is given in chapters 4.10 and 5.3; see also figs. A10 and A14. The start time of the TEA1081, at which the output voltage is present and the IC is functioning, depends on the available line current, output load and the values of C_L and C_O.

2.5. Enlarged output current.

For applications where more than the maximum output current (> 30 mA) is required the possibility exist to put two TEA1081's in parallel according to fig.A3. Two series resistors of minimum 33 ohm are advised to keep the input currents of the two TEA1081's (at max. output current) equal within 20 %.
The input impedance of the total application is reduced with respect to one TEA1081; see chapter 2.4.2.

2.6. Internal protection diodes.

The terminals of the TEA1081 are as follows protected by internal diodes.

pin	to
1 - LN	-
2 - VN	-
3 - AD	LN and VN
4 - PD	SP and VN
5 - IF	LN and VN
6 - VA	QS and VN
7 - QS	VN
8 - SP	VN

3. TEA1081 COMPARED WITH TEA1080.

The differences for the electrical characteristics between TEA1081 and TEA1080 are as follows:

Characteristic	TEA1081	TEA1080	unit
Max. line voltage	12	<10	V
Noise voltage	-83	-72	dBmp
Min. instantaneous line voltage $I_0 = 5$ mA	1.4	1.8	V
Input resistance	positive	negative	
Distortion	improved	-	
Side tone distortion	improved	-	
Start up time	improved	-	

For application of the TEA1081 (a power down function is included) a comparison of the pinning of both ICs is given:

pin	TEA1081	pin name		TEA1080
1	positive line terminal	LN		same
2	negative line terminal	VN		same
3	amplifier decoupling	AD		same
4	PD input	PD	RD	not used
5	LP filter input	IF		same
6	outp. voltage adjustment	VA		same
7	supply output	QS	SO	same
8	PD supply	SP	ER	decoupling

4. TEA1081 APPLICATION CONNECTED BETWEEN LN AND V_{EE} OF THE TEA1060.

Fig.A5 shows the circuit diagram of this solution. The V_{EE} terminal is the common ground for the transmission part as well as for the supply part and the peripherals. The terminals of the TEA1060 (DTMF, QR+, QR-, MUTE and PD) can be directly wired to the peripheral circuits.

A part of the available line current is taken by the TEA1081 to supply the peripheral circuits. The rest of the line current will flow through the TEA1060 application with the consequence that the correct line loss compensation (if selected by R6; fig.A4) will be disturbed; see chapter 4.5.

4.1. Supply possibilities.

A demand is that the available line current has to be sufficient to supply the TEA1081 as well as the TEA1060 application. This means that the line current I_{LN} of the set must cover the maximum required input current I_1 of the TEA1081 and at least the minimum current I_L for the transmission circuit to guarantee the required transmission level.
The input current of the supply part will not be limited, so at excessively high supply currents for the TEA1081 application the current for the transmission part becomes too low which results in limitation of the transmission level (chapter 4.3) and a possible malfunctioning of the voltage stabilizer.

4.1.1. V_O and V_{CC} separated.

When V_O of the TEA1081 is not connected to V_{CC} of the TEA1060 two supply points are available, according fig.5.

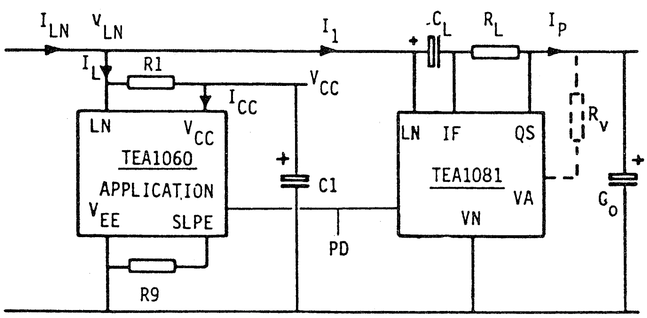


Fig.5 TEA1060 and TEA1081; V_{CC} and V_O separated.

The devices which consume relative high supply currents can be connected to V_O while V_{CC} can be applied for small supply currents as for a microphone pre-amplifier.
An advantage is that at too low line currents the TEA1081 application can be switched off by activating the PD function of the TEA1081. The voltage across R9 can be compared with an external reference voltage to control PD of the TEA1081. However a hysteresis function has to be built in because the current through R9 will increase with the value of the supply current I_1 after switching off the supply part.
For PD applications the PD inputs of both ICs can be interconnected.

4.1.2. V_O <u>and</u> V_{CC} <u>interconnected.</u>

Fig.6 shows the diagram of this supply possibility for which only one smoothing capacitor is used.

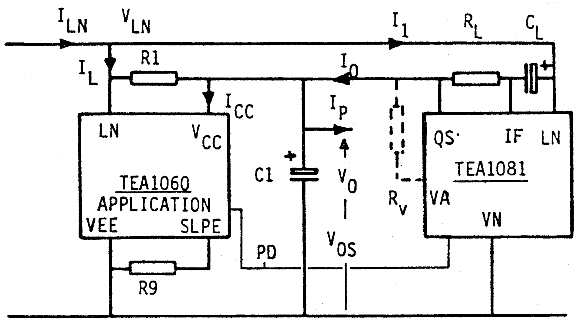


Fig.6. TEA1060 and TEA1081; V_{CC} and V_O interconnected.

Since the TEA1081 now also supplies the internal circuitry of the TEA1060 the maximum receiving level is improved; see chapter 4.4. Without R_V the supply voltage is about:

$$V_O = V_{LN} - (I_P + I_{CC}).k.20 - 0.5 \quad (V).$$

In which I_{CC} is about 1.2 mA and k is the ratio between input and output current of the TEA1081; see chapter 2.4.3.

With R_V included the supply voltage can be limited as explained in chapter 2.3.3. The TEA1081 is only able to limit the output voltage at the preset level V_{OS} when the output current of the TEA1081 is more than zero.

This means that the supply current $I_P > \frac{V_{LN} - V_{OS}}{R1} - I_{CC}$ (mA) at which I_{CC} = 1.2 mA, R1 = 620 ohm and $V_{OS} = R_V.40.10^{-6}$ (V). At lower I_P values the supply voltage will not be limited at the preset level V_{OS} and the output voltage:

$$V_O = V_{LN} - (I_P + I_{CC}).R1 \quad (V).$$

In this case all supply current is delivered via R1 and not by the TEA1081.

4.2. DC line voltage.

Without TEA1081 application all available line current I_{LN} flows in approximation through R9 (fig.A5). With included TEA1081 application a part of the line current will bypass R9 and the DC line voltage will drop from V_{LNO} to:

$$V_{LN} = V_{LNO} - I1.\ R9 \quad (V)$$

at which V_{LNO} is the line voltage of the TEA1060 at I_1 = 0 mA.

* V_{LNO} = 4.45 V typ. at I_{LN} = 15 mA.
* DC slope of V_{LN}/I_{LN} characteristic equals the value of R9.

4.3. Transmission behaviour.

Transmission gain at nominal conditions and 600 ohm line load is slightly affected by the impedance of the supply part, At 300 Hz the gain variation is about 0.2 dB for R_L = 100 kohm, C_L = 2.2 uF and 10 mA output current.

Maximum transmission level at low harmonic distortion (d< 2%) depends, within certain limits, on the available current through the transmit/regulator stage of the TEA1060 and on the minimum instantaneous line voltage of the TEA1081; see chapter 2.4.4. Without signal clipping it is given by:

$$v_{LN}\max = (I_{LN} - I_{CC} - I_1 - I_S) \cdot \frac{R1 \cdot Z_{LINE}}{R1 + Z_{LINE}} \cdot \frac{1}{1.41} \quad (V_{rms})$$

in which:
I_{LN} = available line current.
I_{CC} = supply current of TEA1060 (< 1.2 mA).
I_1 = input current of TEA1081 (function of v_{LN})
I_S = substrate current of TEA1060 (< 0.5 mA).
Z_{LINE} = line impedance.

Fig.A6 shows the maximum transmission level in dBm for d = 2 % at supply currents I_O is 5, 10 and 15 mA, without and with R_V = 75 kohm; Z_{LINE} = 600 ohm.

The transmission behaviour as a function of line current, including restrictions, are described in appendix 1.

4.4. Receiving behaviour.

The receiving gain of the TEA1060 is not affected by the application of the TEA1081.
The maximum receiving level can be improved when the TEA1081 supplies the internal circuitry of the TEA1060 (V_O and V_{CC} interconnected; chapter 4.1.2.)
Fig.A7 shows the receiving level of the TEA1060 application alone at d = 5 % (signal clipping) and the improvement when the TEA1081 is used to bias the TEA1060. No supply current is used for other purposes, so I_P shown in fig.6 equals zero.

4.5. Gain control.

Because not all line current flows through R9 the AGC function of the TEA1060 is disturbed.
Start and stop values of the send and receive gain versus line current will change depending on the part of the line current consumed by the TEA1081 application. Correction of the gain control curves will, in most cases, not be possible. Between the start and stop values of the line current the transmission and receiving gain depends on the current through the transmission part. In fact both gain factors are modulated by the output current of the supply IC and moreover by the mean level of the speech signal on the line.

The variation of the gain factors will be about 2 dB due to an AC line level variation from 0 to 6 dBm at R6 = 30 kohm, I_{LN}= 18 mA and I_O = 5 mA.

The gain control function for this application is not so useful and may be better switched off especially for low exchange voltages.

4.6. Send/Receive reference equivalents.

Due to the modified start and stop values of the gain control function (chapter 4.5) the send and receive reference equivalents are changed too. At short line lengths send and receive loudnesses and side tone level are increased. As explained in chapter 4.5 complete correction is not possible, the gain control function may have to be switched off.

Both gain factors have to be modified to acceptable loudness levels when R6 is not used (no gain control), because the gain is now constant and maximum over the whole line current range. Moreover the side tone circuit Z_{BAL} of the TEA1060 has to be optimised to get acceptable side tone levels for short as well as for long cable lengths (see ref.1 and next chapter).

4.7. Side tone distortion.

Without clipping of the transmission or receiving signal the side tone distortion will be caused by the harmonics of the line signal generated by the TEA1081 at which the anti side tone bridge of the transmission part is not balanced.
Side tone distortion is specified by some PTT administrations. They require the maximum level of the second and third harmonics with respect to the original side tone level.
Especially at optimum cable lenghts, at which side tone attenuation is maximum and side tone level minimum, side tone distortion can be a problem.
Fig.A8 shows the absolute level of the side tone and its second and third harmonics as a function of cable lengths.
The conditions are:
As shown in fig.A8: f = 500 Hz, line signal 0 dBm at 0 km.
Cable: 0.5 mm, 176 ohm/km, 38nF/km.
TEA1060: asymm. receiver output, no gain control, Z_{BAL} optimised, send and receive gain 3 dB decreased with respect to standard application fig.A4.
TEA1081: C_L = 2.2 uF, R_L = 100 kohm, no R_V, I_O= 5 , 15 mA.

4.8. Noise voltage.

As given in chapter 2.4.5. the noise voltage of the TEA1081 is low and will not affect the original noise voltage on line and receiver output of the TEA1060 at nominal transmission and receiving gain. Noise voltage of the TEA1060 depends on the gain settings, see ref.1.

4.9. Circuit impedance, BRL.

The impedance of the application is affected by the impedance of the supply part. The BRL of the TEA1060 alone and from the total application measured against 600 ohm is shown in fig.A9 .

4.10. Start time.

The time is measured between switching on the DC line current and the moment that the receiving signal is available and an output voltage of 3 V is reached, both as a function of line current ; see fig.A10. The conditions are:

* test circuit diagram according fig.A10.
* switching rate of S: 0.1 Hz.
* Z_M = 150 ohm symmetrical, Z_T = 150 ohm asymmetrical, (see fig.A4).

4.11. Stability.

Application of the TEA1081 with TEA1060 requires a series network of at least 68 pF and 10 kohm between LN and AD of the TEA1081 in order to ensure HF stability. Higher values of this capacitance will however reduce the BRL at maximum audio frequencies.

When R_V is used, LF oscillation of the line voltage (motor boat effect) can occur depending on signal level, DC voltage difference between line and supply output and the way in which the application is supplied (current or voltage source).

To prevent this oscillation one of the following solutions can be used:

* Lower value of C3 (4.7 uF) of the TEA1060; will decrease transmission gain and BRL at low audio frequencies and will influence the start up time.
* Increase of C_L or R_L of the TEA1081; increases the start up time of the application.
* Addition of a resistance (of about 30 ohm) between C_L (+ lead) and LN of the TEA1081; reduces the maximum input current at a specified DC input and output voltage; see chapter 2.3.3.

5. TEA1081 APPLICATION CONNECTED BETWEEN LN AND SLPE OF THE TEA1060.

The SLPE terminal of the TEA1060 is in this application the ground reference for the supply part as shown in fig.A11. All line current including the supply current is flowing through R9 of the transmission part which means that the gain control function is not disturbed.

Because of the DC shift between V_{EE} and peripheral ground SLPE, which depends on the total line current, extra circuitry has to be used for the interface of the signals to or from QR, DTMF, PD and MUTE of the TEA1060.
Fig.A12 gives an example of a listening-in circuit with the TEA1060, TEA1081 and the loudspeaker amplifier TDA7050. To couple the audio signal from QR of the TEA1060 (referred to VN) to the TDA7050, the reference inputs of this IC (terminals 3 and 4) are coupled to VN by a capacitor of 2.2 uF. To bias the reference inputs a resistor of 10 kohm is used.

To control the PD and MUTE inputs of the TEA1060 from the peripheral circuits, two examples are shown in fig.A13a and fig.A13b at which the signal is inverted or not inverted. The circuits have to be improved depending on the logic levels and output structure of the peripheral devices.

5.1. Supply behaviour.

In contrast with the application from chapter 4, the supply part is now powered by a constant voltage of about 4.2 V between LN and SLPE; see appendix 1. Without applied R_V the output voltage V_O (fig.A10) is given by:

$$V_O = V_{RL} - 0.5 - I_1.20 \quad (V) \quad \text{at} \quad V_{RL} = 4.2 +/- 0.2 \text{ V}$$

For this application both supply points V_O and V_{CC} cannot be interconnected because R9 will be short circuited for AC via capacitors C1 and C_O.
The DC line voltage will be the same as without TEA1081 application, all line current is flowing through R9.

5.2. AC behaviour.

Transmission gain is not affected by the supply part; see appendix 1. The maximum transmission level is about the same as given in chapter 4.3, see also appendix 1. The signal level across the supply part is about 0.6 dB more than for the application of chapter 4 due to the AC signal on terminal SLPE.

Receiving gain and maximum level are not affected.

Gain control function is the same as without supply part; all line current flows through R9.

Send/receive reference equivalents are not changed by the TEA1081 application according fig.A11.

Noise contribution of the TEA1081 is negligible for this application.

Circuit impedance and BRL are the same as without TEA1081 application. The values of C_L and R_L can be chosen freely. However a low value of those components results in a low impedance of the supply part at minimum audio frquencies and a relatively high AC input current of the TEA1081 which limits the maximum transmission level of the total application. Minimum values of 1 uF for C_L and 100 kohm for R_L are advised.

5.3. Start time.

The time is measured between line current switching on and the availability of the supply voltage (at V_O = 3V) and the AC voltage on the receiver outputs. The results are shown in fig.A14, while the conditions are:

* diagram of test circuit according fig.A14.
* switching rate of S: 0.1 Hz.
* Z_M = 150 ohm symmetrical, Z_T = 150 ohm asymmetrical, (see fig.A4).

6. APPLICATION OF THE TEA1081 IN A PULSE DIAL TELEPHONE SET.

An application of the TEA1060 with a pulse dialling circuit of the PCD3320 family is used to test the influence of the TEA1081 application on line current and supply voltage V_{CC} of the TEA1060 during pulse dialling. Fig.A15 shows the application diagram of the TEA1060, PCD3322 and TEA1081. For other types of the PCD3320 family see ref.2 and 3.

The supply point of the TEA1060 directly powers the PCD3322. Capacitor C20 and diode D20 are applied to back up this IC during hook on of the telephone set. Output DP of the PCD3322 is connected to the PD input of the TEA1060 and is used to interrupt the line current via a DMOS-FET (BST76A). The TEA1081 is connected between LN and V_{EE} of the TEA1060.

The PD input of the TEA1081 can be connected to the DP or M2 output of the PCD3322 depending on the application, while the power down part of the TEA1081 (SP terminal, see chapter 2.3.4) can be supplied by its output voltage V_O (terminal 7) or by V_{CC} of the TEA1060 when V_O drops below the minimum V_{SP} voltage (2 V) during pulse dialling.

The photo's on page 42 to 45 show the line current I_{LN}, the supply voltage V_{CC} and the output voltage V_O of the TEA1081 for several applications (see table 1; next page) during dialling. Moreover is shown the influence on I_{LN}, V_{CC} and V_O when the load R_O is switched off from the output capacitor of the TEA1081 during dialling by using DP or M2 of the PCD3322 (DP = dialling pulse ; M2 = HIGH during pulsing of each digit and LOW during an inter-digit pause).

Test conditions:

* TEA1060; Standard application fig.A4.
  R_M = 150 ohm, R_T = 150 ohm single ended driven.
* TEA1081; C_L = 2.2 uF, R_L = 100 kohm, R_V not used.
  C_O = 220 uF, I_O = 10 mA at PD = LOW.
* I_{LN} = 30 mA, supply bridge 600 ohm.

Photo 1 shows the measurement results without TEA1081, while for the other photo's the TEA1081 is applied. Key button "8" is pressed for photo's 1 up to 9 and key button "8" followed by "5" for photo 10 up to 13.

Photo nr.	measured I_{LN}	V_{CC}	V_O	PD-TEA1081 to:	SP-TEA1081 to:	R_0 switched off by:	see note
1	*	*					a
2	*	*		open	open		b
3			*	open	open		b
4	*	*		DP	V_{CC}		c
5			*	DP	V_{CC}		c
6	*	*		DP	V_O	DP	
7			*	DP	V_O	DP	
8	*	*		DP	V_O	M2	e
9			*	DP	V_O	M2	
10	*	*		M2	V_{CC}		c
11			*	M2	V_{CC}		c
12	*	*		M2	V_O	M2	
13			*	M2	V_O	M2	

Table 1. Test conditions of the pulse dial set.

notes:
a) TEA1081 not applied.
b) PD function of TEA1081 not used; R_0 not switched off.
c) R_0 not switched off.
e) M2 only available from PCD3322 and PCD3323.

7. CONCLUSIONS.

Two methods are given to combine the TEA1060 (TEA1060 family) with the supply circuit TEA1081. An overview of the differences between both applications and the influence of the applied TEA1081 are given in table 2. The main differences concern the disturbed gain control function for the first solution (LN - V_{EE}; chapter 4) and the required extra circuitry for the interface of the signals to or from QR, DTMF, PD and MUTE for the second solution (LN - SLPE; chapter 5;).

TEA1081 application connected between:	LN - V_{EE}	LN - SLPE
interface of logic signals	no problem	needs variable level shift
supply possibilities $V_O = V_{CC}$ use of R_V	 no problem no problem	 not possible not necessary
DC line voltage	decreased	not changed
maximum transmission level	function of I_{LN} and I_O (I_1)	function of I_{LN} and I_O (I_1)
max. receive level	increased when $V_O = V_{CC}$	not changed
gain control fu.	not useful	not changed
send/rec. equivalents	changed	not changed
noise voltage	not changed	not changed
BRL	decreased	not changed
start time	increased	increased

Table 2. Differences in behaviour of both solutions.

Note: "not changed" means no difference compared with typical TEA1060 application without TEA1081 part.

Application of the TEA1081 in a pulse dial telephone set and its influence on the set behaviour during pulse dialling is described in chapter 6.
Comparison of the photo's 4, 6, 8 and 12 with photo 2 shows that use of the power down facility of the TEA1081 is useful to prevent V_{CC} voltage values which are too low for the device(s) supplied from V_{CC} during pulse dialling. Controlling the power down of the TEA1081 as well as the switch for the output load, both by DP or M2 of the PCD3322, (photo's 6 and 7 or 12 and 13) gives the best results with respect to V_{CC} and V_0 behaviour.

7. REFERENCES.

Ref.1.: * Technical publication TP162 by P.J.M.Sijbers. Title: Versatile transmission ICs for electronic telephone sets.

* Laboratory report ETT8606 by P.J.M.Sijbers. Title: Application of the low voltage versatile transmission circuit TEA1067.

* Designers' Guide: TEA1060 Family, Versatile Speech /Transmission ICs for Electronic Telephone Sets. Author:P.J.M.Sijbers.
cataloque number: 9398 341 10011

Ref.2.: * Laboratory report TTE87125 by J.van Tiggelen. Title: Documentation for printed circuit board CAB3132. (TEA1060+PCD3321)

Ref.3.: * Elcoma data handbook IC03 "Integrated circuits for telephony".

APPENDIX 1. CIRCUIT CALCULATIONS.

As defined in chapter 5, the IC members of the TEA1060 family offer the possibility to connect a load (supply part) between the terminals LN and SLPE without much influence on the transmission characteristics of the total application. This chapter gives the results of the calculations done on the transmit/voltage regulator stage from the TEA1060.

In fig.AP1 Z_{LD} (within broken lines) represents the load between LN - SLPE which can be the TEA1081 application (chapter 5, fig.A11) or a device (R_{L2}) which is supplied via a RC filter (R_{L1} C_{RL}).

The electrical symbols used in the expressions hereafter are defined in fig.AP1 as well.

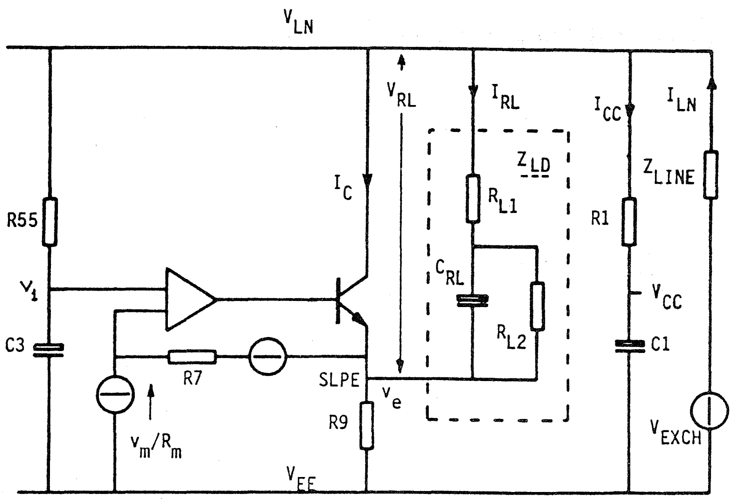


Fig.AP1. Equivalent circuit diagram of the transmit/regulator stage of the TEA1060 with a load connected between LN - SLPE.

A1.1. DC voltages V_{LN} and V_{RL}.

Line voltage $$V_{LN} = V_{RL} + (I_C + I_{RL}).\ R9 \quad (V) \tag{1}$$

In which V_{RL} the temperature compensated reference voltage is of 4.2 +/- 0.2 V.
The line voltage depends on the available line current which is about $I_C + I_{RL}$ and is thus independent of the load between LN and SLPE.

A1.2. Transmission gain.

The transmission gain between microphone amplifier input and line is derived as follows:

$$v_e = -\frac{v_m}{R_m}\ .\ R7 \tag{2}$$

where v_m is the microphone signal and R_m represents the transimpedance between mic. amplifier input and input of the transmit output stage.

$$i_{RL} + i_C = \frac{v_e}{R9} \quad (3), \qquad v_{LN} = (i_{RL} + i_C).\frac{R1\ .\ Z_{LINE}}{R1 + Z_{LINE}} \tag{4}$$

substitution of (2) and (3) into (4) gives a transmission gain of:

$$\frac{v_{LN}}{v_m} = -\frac{1}{R_m}\ .\ \frac{R7}{R9}\ .\ \frac{R1\ .\ Z_{LINE}}{R1 + Z_{LINE}} \tag{5}$$

As can be seen in (5) is the transmission gain independent of load Z_{LD}.

A1.3. Set impedance.

Calculation of the set impedance (LN - V_{EE}) is as follows:

$$Z_{SET} = \frac{v_{LN}}{i_{CC} + i_{RL} + i_C} \quad (6), \qquad i_{CC} = \frac{v_{LN}}{R1} \quad (7),$$

$$v_e = v1 = \frac{v_{LN}}{1 + p.C3.R55} \quad (8) \qquad \text{in which} \quad p = j.2.pi.f$$

Substitution of (3), (7) and (8) into (6) gives:

$$Z_{SET} = \frac{(1 + p.C3.R55).R1.R9}{R1 + (1 + p.C3.R55).R9} \tag{9}$$

The set impedance is not changed by the applied Z_{LD}. For audio frequencies Z_{SET} is about the value of R1 by assuming:

$\frac{1}{p.C3} \ll R55$, $\frac{1}{p.C1} \ll R1$ and $R55 \gg R1$.

A1.4. Maximum transmission level.

To calculate the maximum transmission level fig.AP1 is re-drawn for AC into fig.AP2, assuming:

$$\frac{1}{p.C_{RL}} \ll R_{L1} \quad \text{and} \quad \frac{1}{p.C1} \ll R1.$$

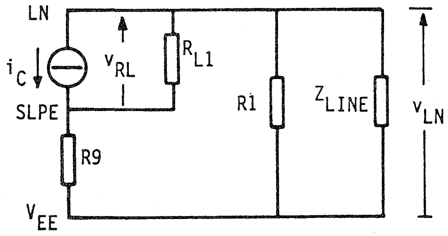


Fig.AP2 Equivalent circuit diagram for AC (derived from AP1)

The voltage between LN and SLPE is given by:

$$v_{RL} = - i_C.Z_{VL} \tag{10}$$

in which Z_{VL} is the total impedance between LN and SLPE.

$$Z_{VL} = \frac{(R9 + R1//Z_{LINE}).R_{L1}}{R9 + R1//Z_{LINE} + R_{L1}} \tag{11}$$

The line voltage $$v_{LN} = \frac{R1//Z_{LINE}}{R1//Z_{LINE} + R9} \cdot v_{RL} \tag{12}$$

substitution of (11) into (10) and into (12) results in:

$$v_{LN} = - i_C \cdot \frac{(R1//Z_{LINE}).R_{L1}}{R1//Z_{LINE} + R9 + R_{L1}} \tag{13}$$

The maximum AC collector peak current which can be generated equals the available DC current (fig.AP1), thus:

i_C peak = I_C (14) and $I_C = I_{LN} - I_{CC} - I_{RL} - I_S$ (15)

Where I_S (< 0.5 mA) the substrate current is flowing from LN to V_{EE}. Substitution of (15) into (14) and into (13) gives the peak level of the line signal as a result of the available line current.

$$v_{LN}\text{ peak} = (I_{LN} - I_{CC} - I_{RL} - I_S).\frac{(R1//Z_{LINE}).R_{L1}}{R1//Z_{LINE} + R9 + R_{L1}} \tag{16}$$

The maximum line signal (low distortion) of the TEA1060 is a function of the collector current I_C up to about 10 mA. At higher values the line signal will be limited at about 9 dBm due to voltage saturation of the output transistor shown in diagram fig.AP1.

The supply current $I_{RL} = \dfrac{V_{RL}}{R_{L1} + R_{L2}}$ with $V_{RL} = 4.2$ +/- 0.2 V

which is valid for resistive loads between LN and SLPE.
In case the TEA1081 application is connected between those points the current I_{RL} represents the input current I_1 of the TEA1081. Assuming that R9 << ($R1//Z_{LINE}$) and the input impedance of the TEA1081 $Z_I >> R1//Z_{LINE}$ equation (16) can be rewrited into:

$$v_{LN}\ \text{peak} = (I_{LN} - (I_{CC} + I_S + I_1)).\ \frac{R1.Z_{LINE}}{R1 + Z_{LINE}} \qquad (17)$$

in which ($I_{CC} + I_S$) is about 1.7 mA.

A1.5. Results.

For small supply currents (I_{RL} < 4 mA) network Z_{LD} as shown in fig.AP1 can be applied. With R_{L1} = 300 ohm, C_{RL} = 220 uF and I_{RL} = 4 mA the supply voltage across load R_{L2} measures about 3 V, while for 6 dBm line signal the line current has to be at least 20.7 mA.
With the TEA1081 applied between LN - SLPE the line current has to be 17 mA for 6 dBm line signal and 4 mA output current of the TEA1081. The output voltage in this case is 3.5 V.

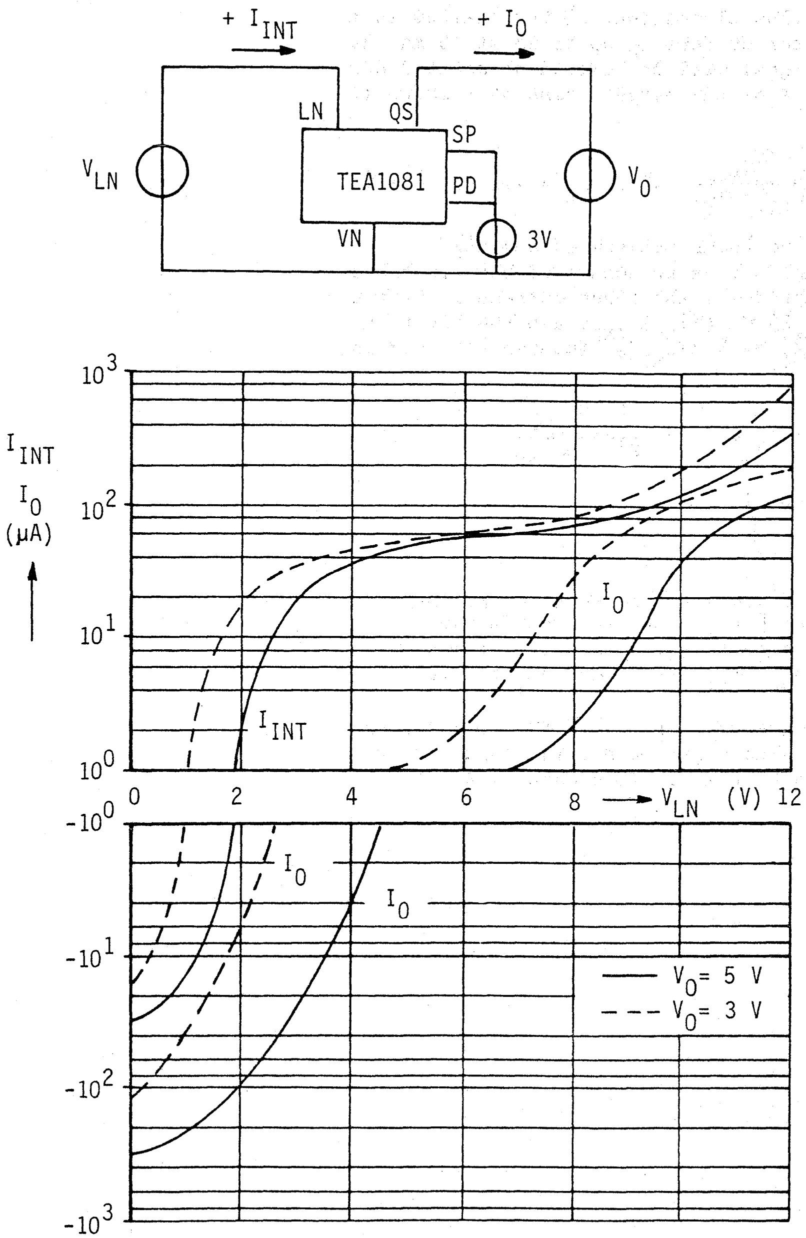
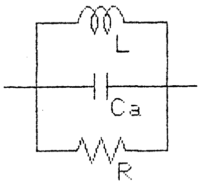


Fig. A1. Input (I_{INT}) and output (I_0) current versus line voltage at V_0= 3 and 5 V at power down.



	I_0 = 5 mA	I_0 = 15 mA	
L	3.90	4.07	H
C_a	9.5	11.0	nF
R	27.0	13.2	kohm

V_{LN} = 770 mV_{rms} C_L = 2.2 uF R_L = 100 kohm

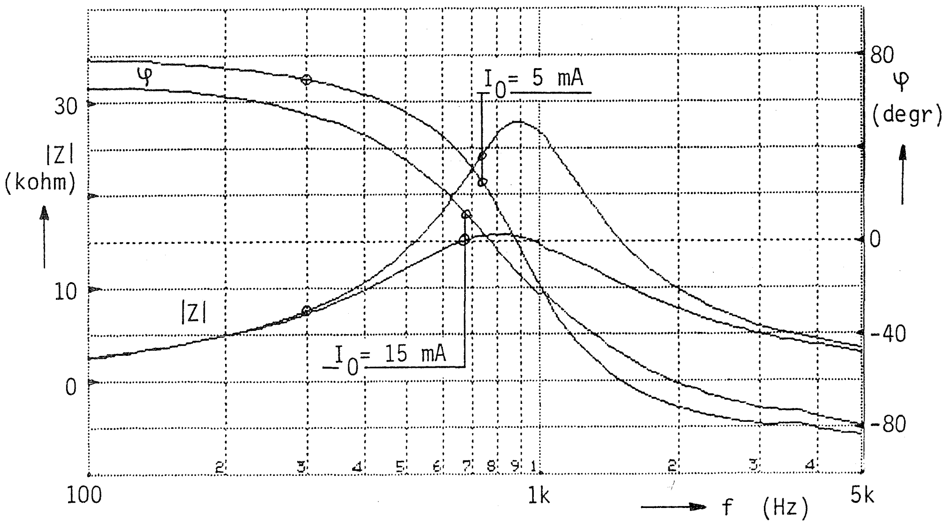


Fig. A2 Input impedance of the TEA1081.

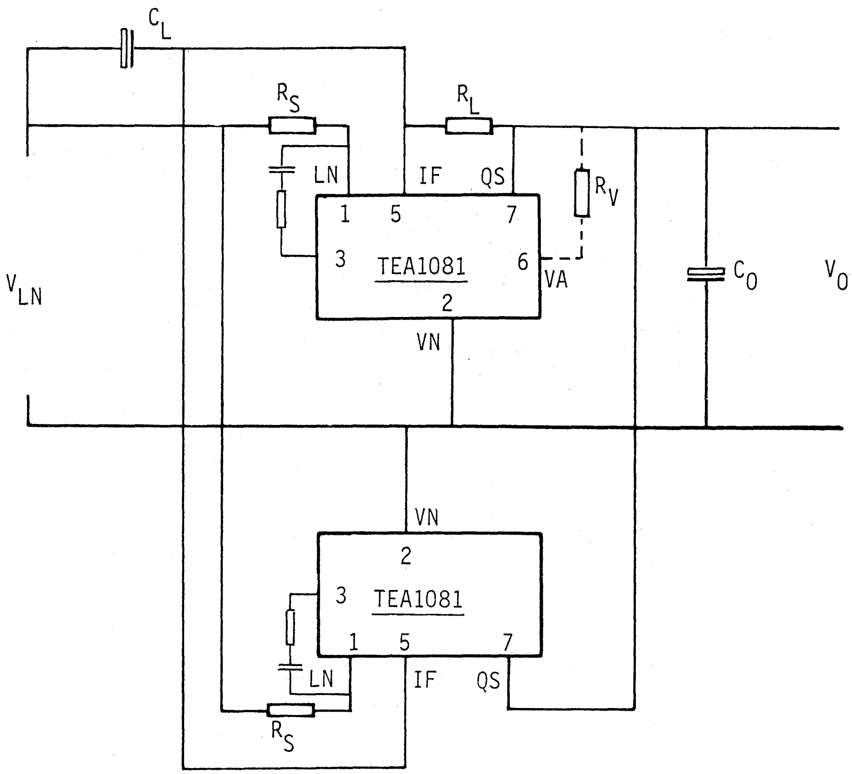


Fig. A3. Two TEA1081's in parallel for enlarged output current.

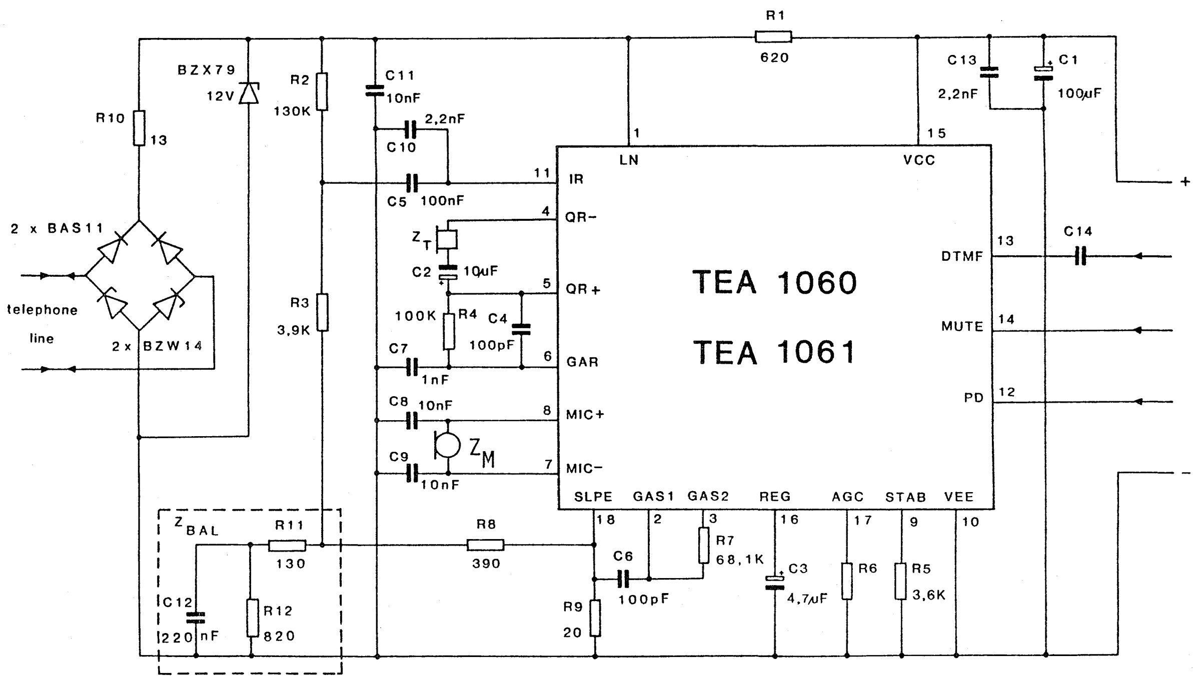


Fig. A4. Application diagram TEA1060/TEA1061, shown with a dynamic earpiece (Z_T 450) and DTMF dialling. Pulse dialling or register recall require a different protection arrangement.

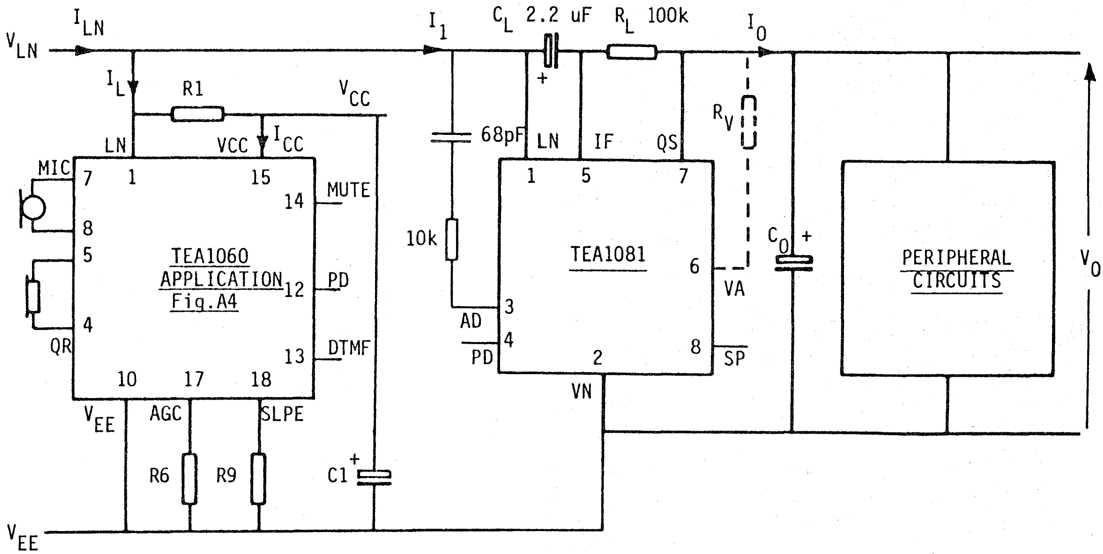


Fig. A5. Application diagram; TEA1081 application connected between LN - V_{EE} of the TEA1060.

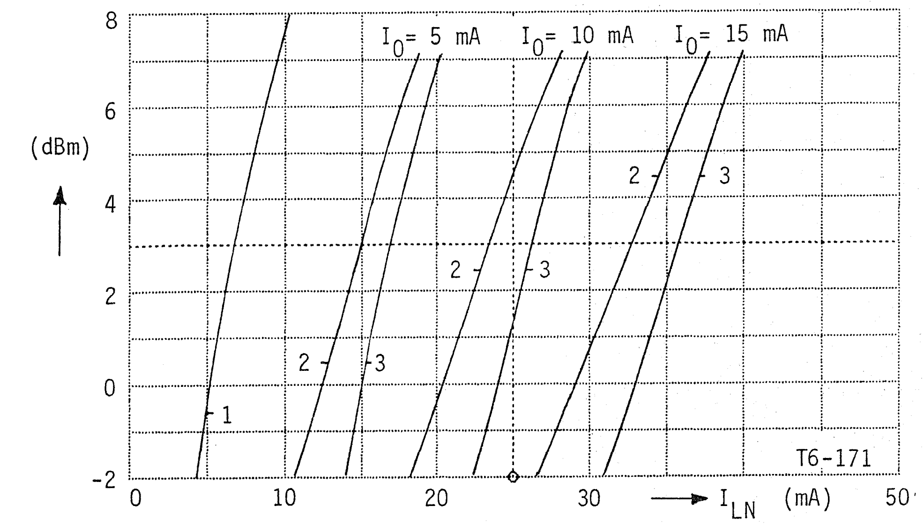


Fig. A6. Transmission level (d= 2 %) versus line current at f = 300 Hz.

Curve: 1, TEA1060 alone fig.A4
2, TEA1060 + TEA1081, R_V= 75 kohm, fig.A5
3, TEA1060 + TEA1081, without R_V , fig.A5

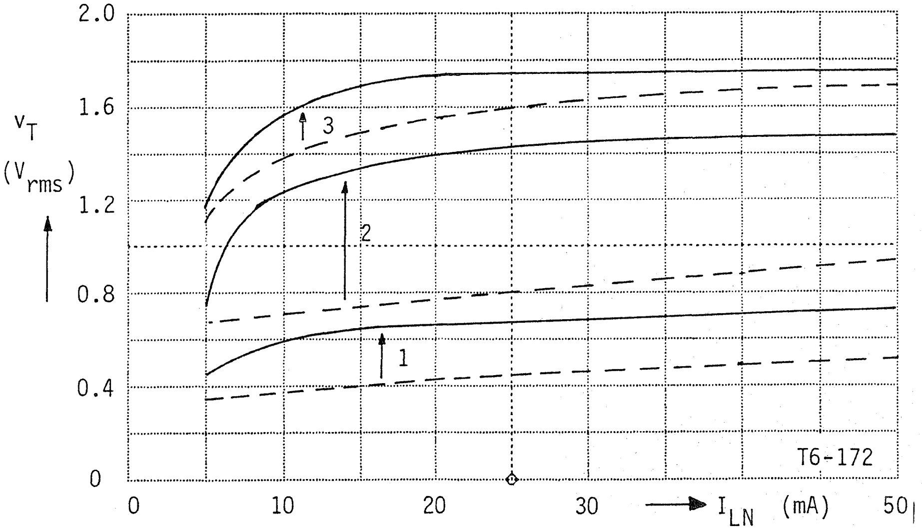
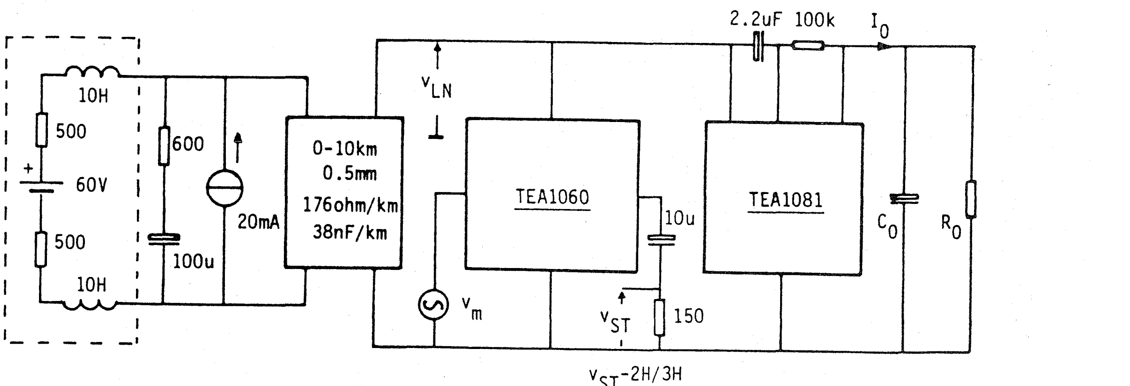


Fig. A7. Receiving level (d = 5 %) as a function of line current. f= 1 kHz, V_{LN}= 4.45 V at I_{LN}= 15 mA.

Curve: 1, Z_T= 150 ohm single ended
2, Z_T= 450 ohm differential
3, Z_T= 47 nF + 100 ohm diff.

——— TEA1060 + TEA1081 V_0 = V_{CC} , fig.6
- - - - TEA1060 alone



TEA1060: no AGC, Z_{BAL} optimised:

R8= 620, R11= 130, R12=1k, C12= 100nF

- v_m = 2.7 mV_{rms}, A_m = 49 dB, R7= 47k5
- A_T = -10 dB, R4= 70k

} Fig. A4

- v_{LN} = 775 mV_{rms} at 0 km
- f = 500 Hz

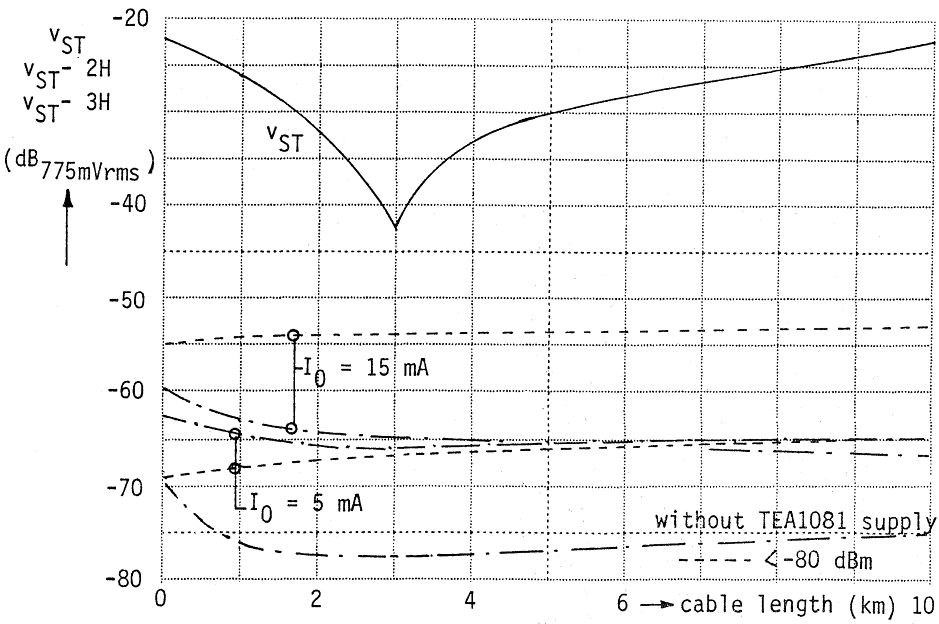


Fig. A8. Side tone distortion

——— side tone level v_{ST}

-—-— level of second harmonic v_{ST}-2H

------ level of third harmonic v_{ST}-3H

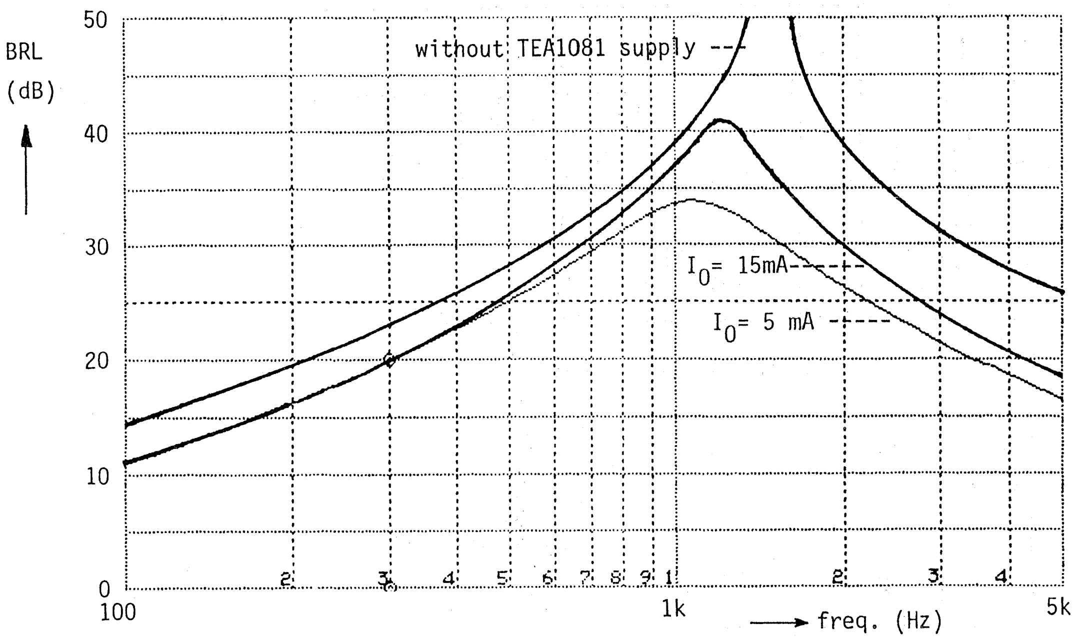


Fig. A9. BRL as a function of frequency (fig. A5).

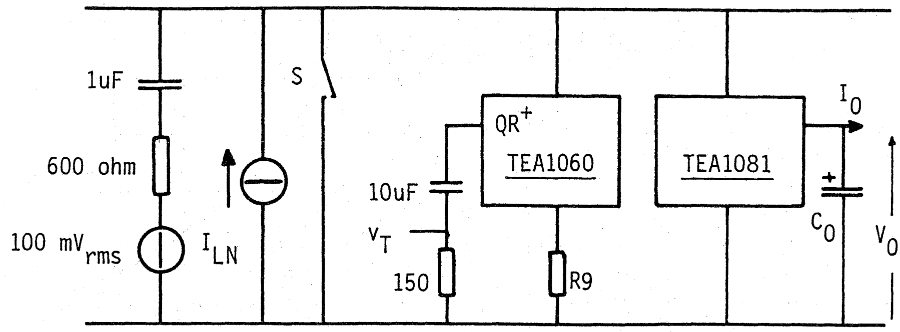
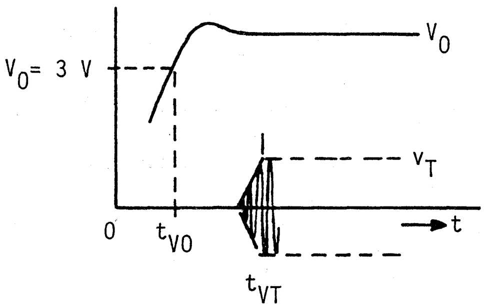


Diagram of test circuit.



Start time definitions.

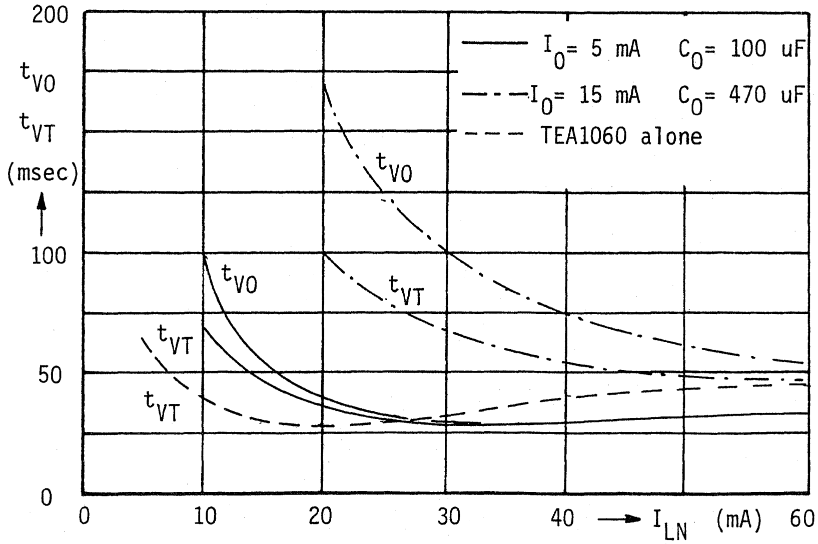


Fig. A10. Start time as a function of line current.

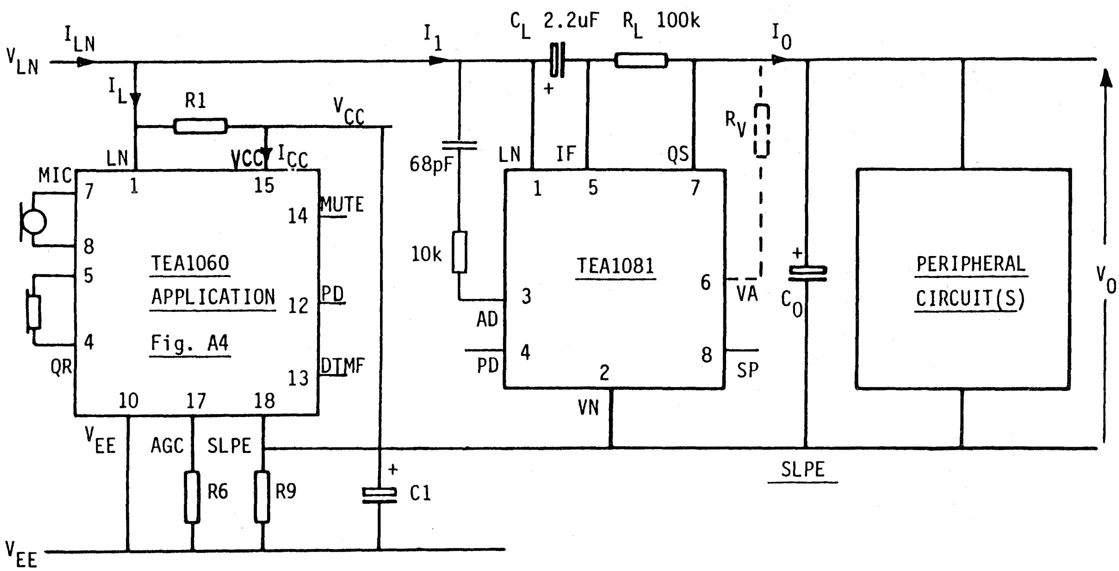


Fig. A11. Application diagram; TEA1081 application connected between LN - SLPE of the TEA1060.

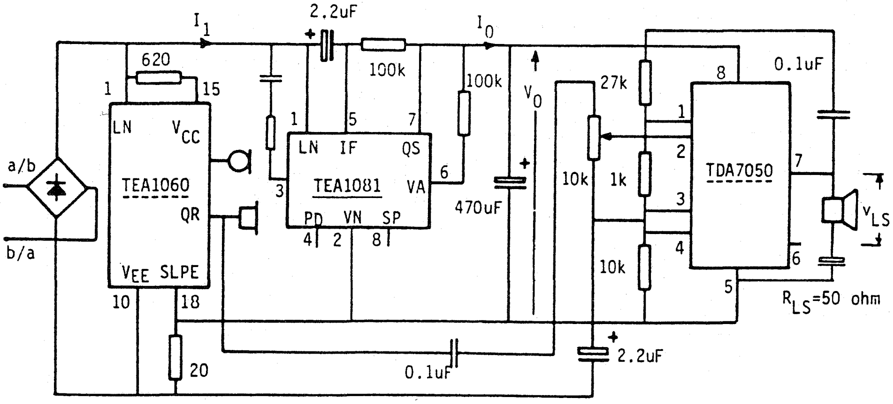


Fig. A12. Circuit diagram of a listening-in application.

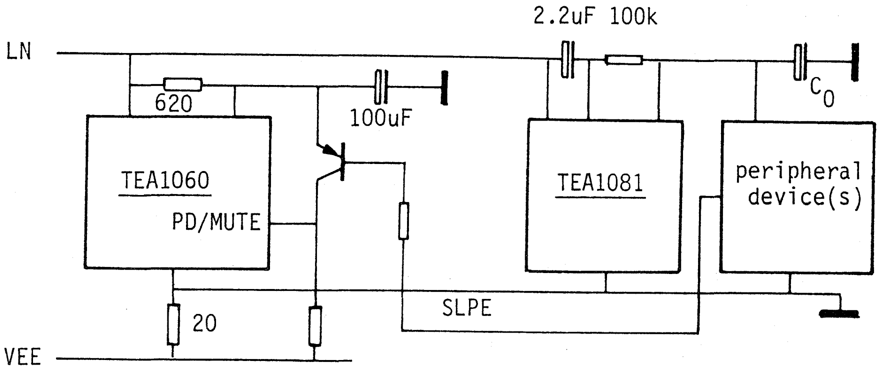


Fig. A13a. Interface circuit for PD/MUTE signal (inverting)

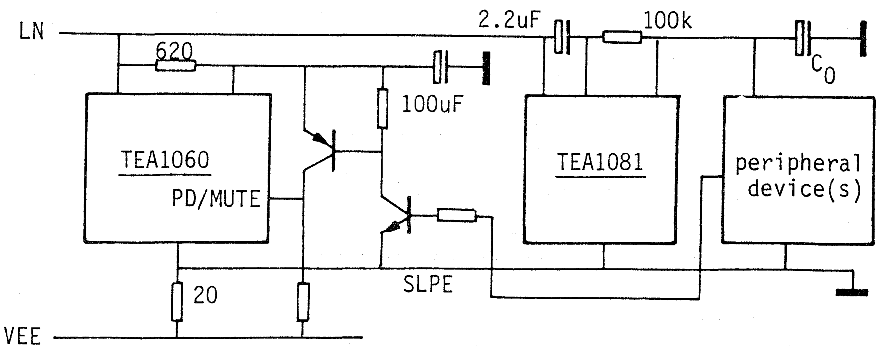
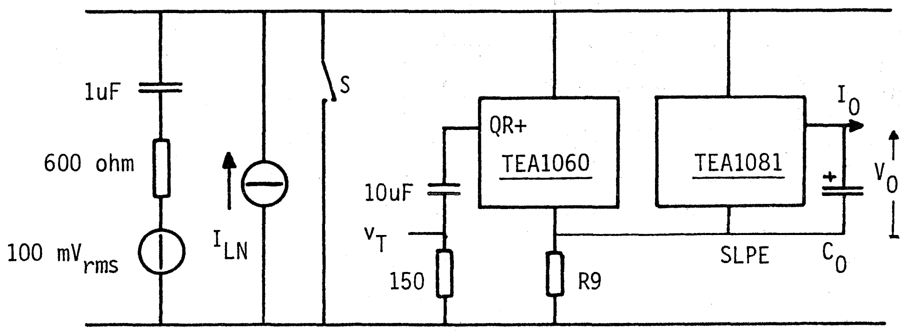
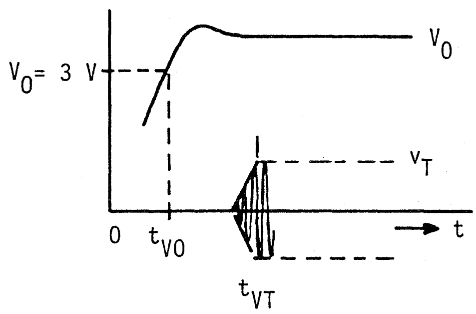


Fig. A13b. Interface circuit for PD/MUTE signal (non-inverting).



Diadram of test circuit.



Start time definitions.

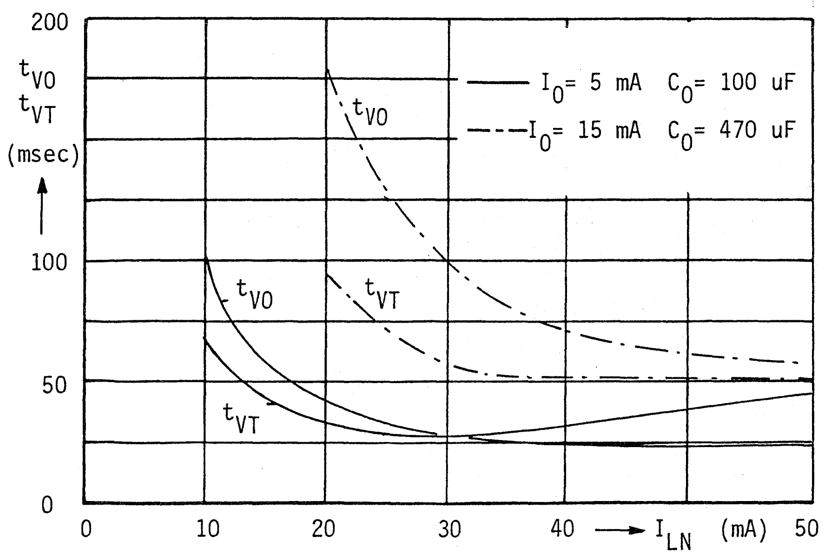


Fig. A14. Start time as a function of line current.

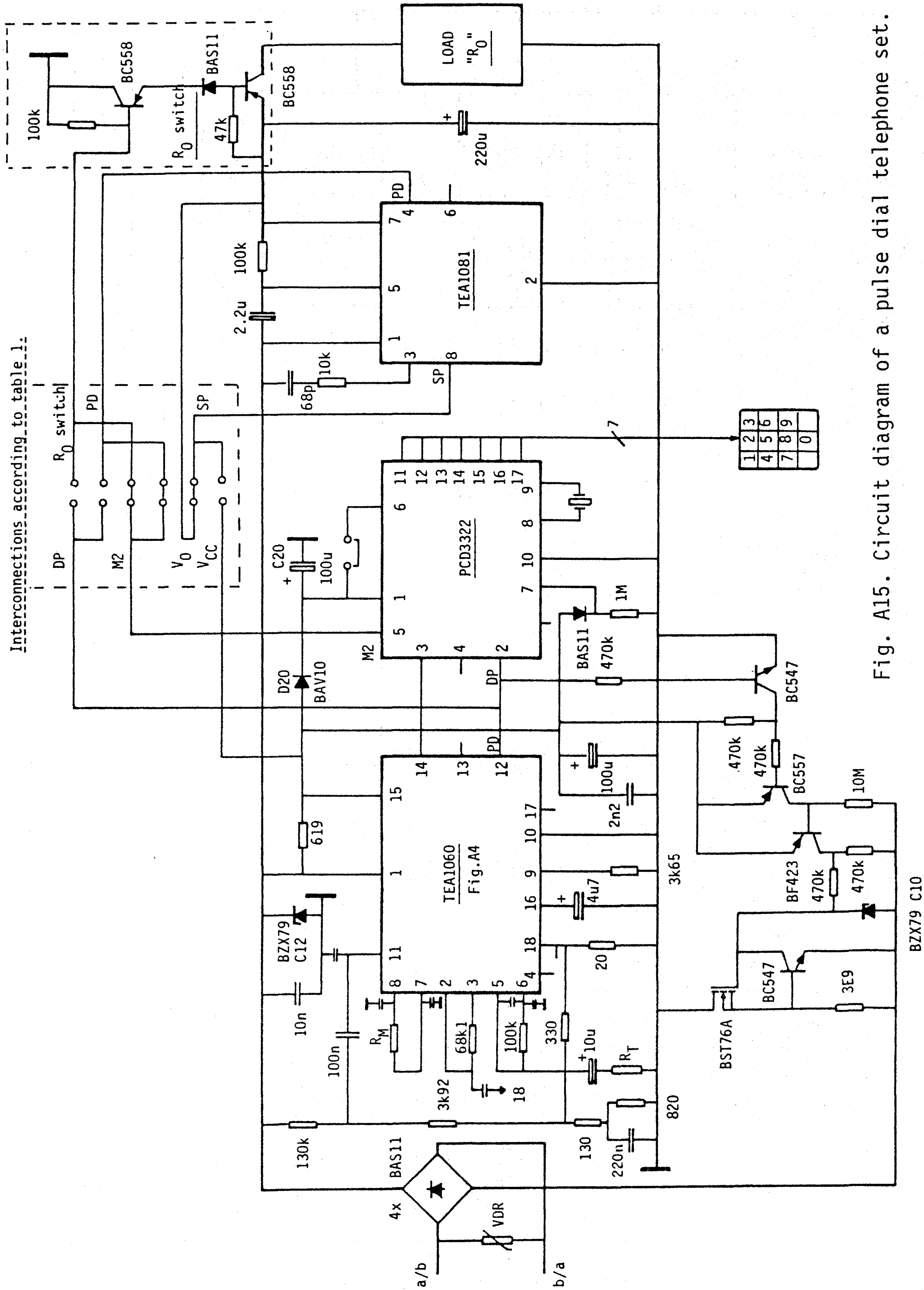
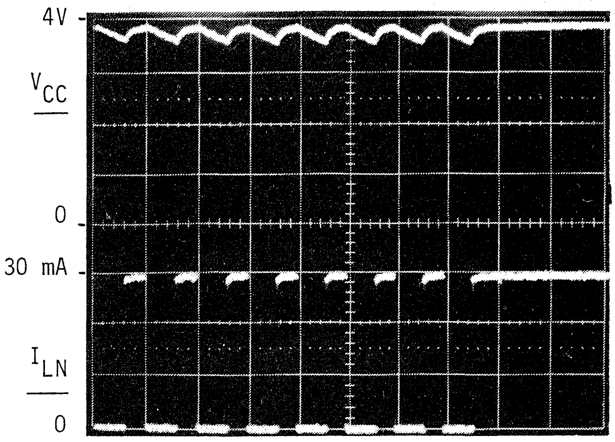


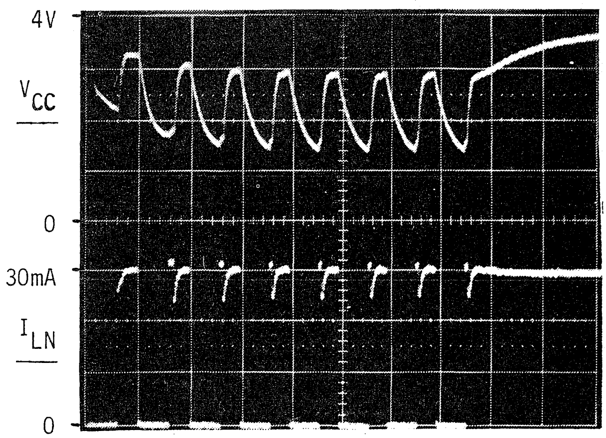
Fig. A15. Circuit diagram of a pulse dial telephone set.



---> 100 msec/div.

1

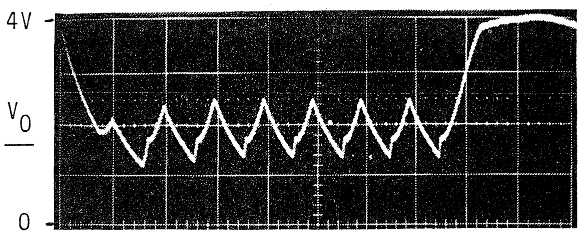
TEA1081 not used.



2

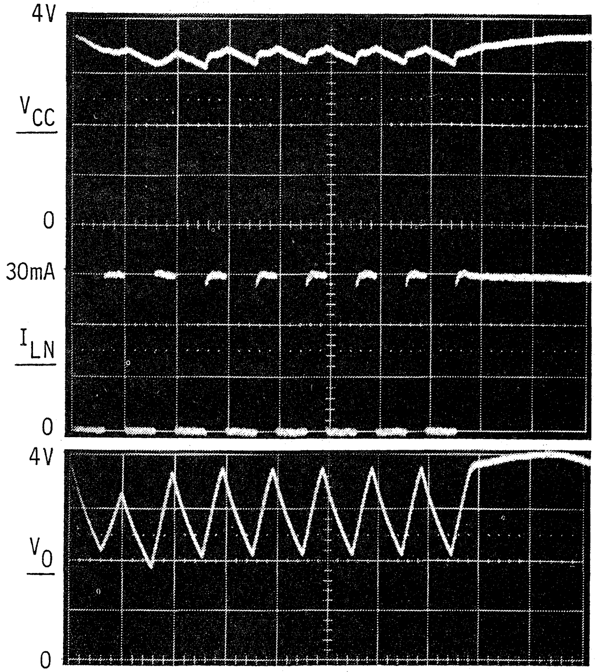
TEA1081 added, I_0 = 10 mA

PD function of TEA1081 not used



---> 100 msec/div.

3



---> 100 msec/div.

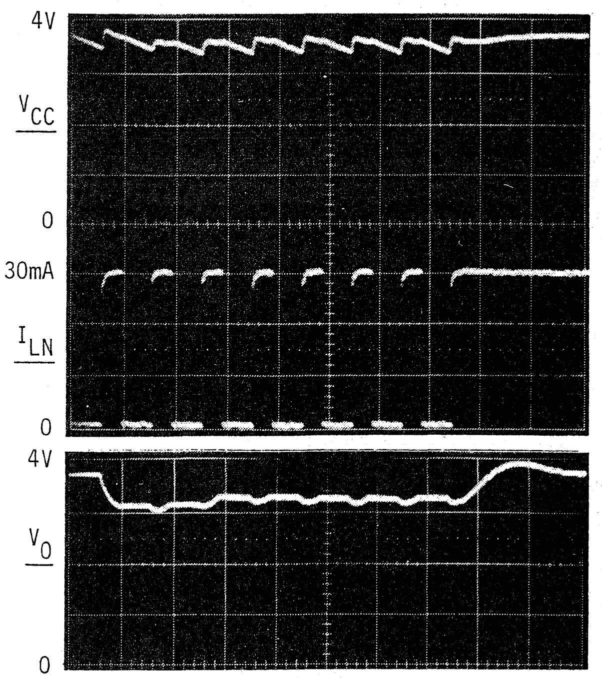
4

PD - TEA1081 connected to
DP - PCD3322.

SP-TEA1081 supplied by V_{CC}-TEA1060.

R_0 not switched.

5



---> 100 msec/div.

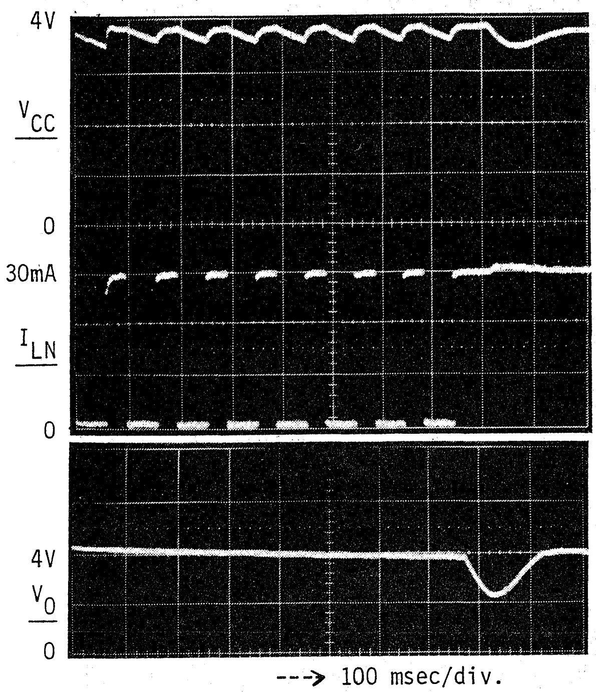
6

PD-TEA1081 connected to DP-
PCD3322.

SP-TEA1081 supplied by V_0.

R_0 switched by DP-PCD3322.

7



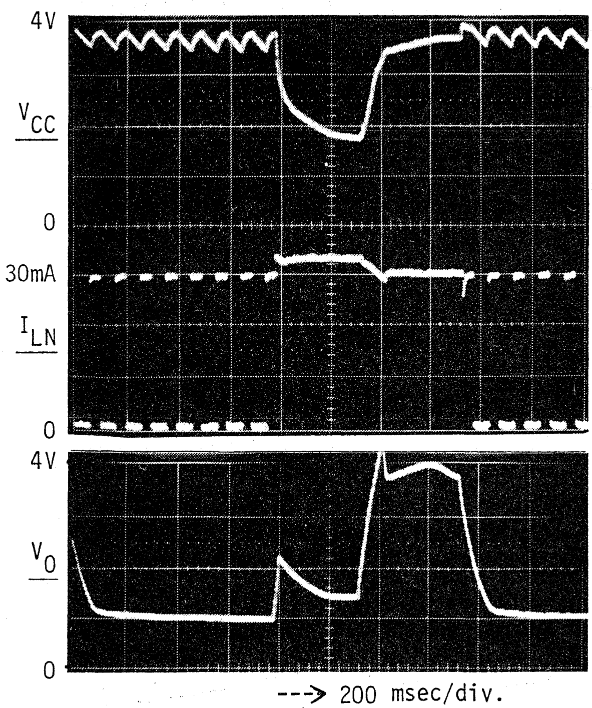
8

PD-TEA1081 connected to DP-PCD3322.

SP-TEA1081 supplied by V_0.

R_0 switched by M2-PCD3322.

9



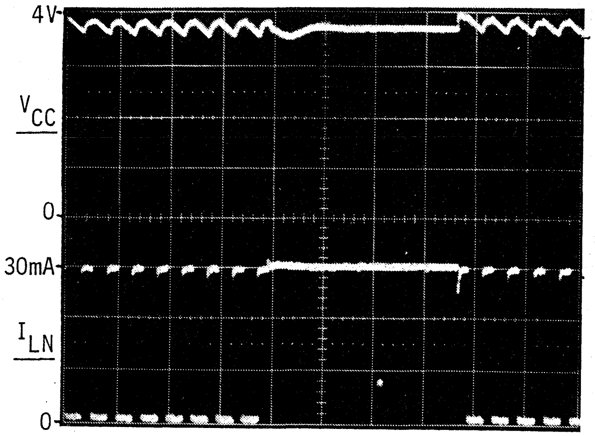
10

PD-TEA1081 connected to M2-PCD3322.

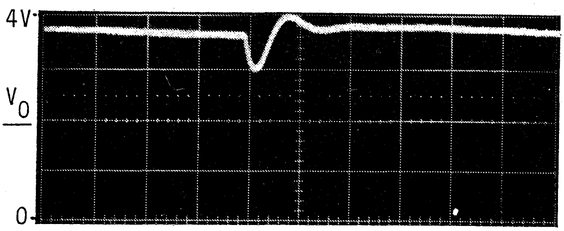
SP-TEA1081 supplied by V_{CC}-TEA1060.

R_0 not switched.

11



12



13

---> 200 msec/div.

PD - TEA1081 connected to M2 - PCD3322.

SP - TEA1081 supplied by V_0.

R_0 switched by M2 - PCD3322.

APPLICATION NOTE Nr	ETT89008
TITLE	Listening-in with the TEA1081, TDA7050 and TEA1064
AUTHOR	K. Wortel
DATE	May 1989

Table of Contents:

1. Introduction

This report describes an application for a listening-in function in electronic telephone sets with the TEA1064 transmission circuit, the TEA1081 supply IC and the TDA7050 loudspeaker amplifier.

Normally the TEA1081 is used in parallel with the speech circuit TEA1064 and acts as a large inductance so it will not influence the set impedance. It can be connected to the LN and VEE pins, or to the LN and SLPE pins of the TEA1064 speech/transmission circuit.
However, if the TEA1081 is connected between LN and SLPE it does not have to act as an inductance. Any kind of load can be connected between these pins, the TEA1064 output stage will eliminate the effect on balance return loss (see ref.1). Based on this fact this report describes a new patented application method of the TEA1081 in which it is used, in conjunction with the TEA1064, as a current splitter. As will be shown this application incorporates a lot of advantages.
The most important features:

- The TEA1064 output stage is biased at a very low current (1-2mA) while still a large sending level can be obtained.
- Current consumed by peripheral circuits can not cause the current through the TEA1064 to become too small for good operation.
- Maximum power is available for peripherals (loudspeaker amplifier TDA7050) under all sending and receiving conditions.

Section 2 of this report describes the principle of the TEA1081 used as a current splitter in conjunction with the TEA1064. Section 3 gives detailed information about interconnecting a TDA7050 amplifier to the TEA1081 and TEA1064. In section 4 an application example is given and test results are shown. Section 5 discusses the use of other TEA106x types. Finally section 6 gives some conclusions.

2. The TEA1081 and TEA1064 as a current splitter

This section describes the use of the TEA1081 as a current splitter when connected between LN and SLPE pins of the TEA1064. In paragraph 2.1 the principle is explained. The DC-behaviour will be discussed in paragraph 2.2 while paragraph 2.3 describes the AC-behaviour.

2.1 Principle of the current splitter

As mentioned in the introduction the TEA1081 does not have to act as an inductance when it is connected between LN and SLPE of the TEA1064. This means RL and CL of figure 1 can be left out and pin IF can be used for other purposes (Leq = RL*CL*Rs see also reference 2):

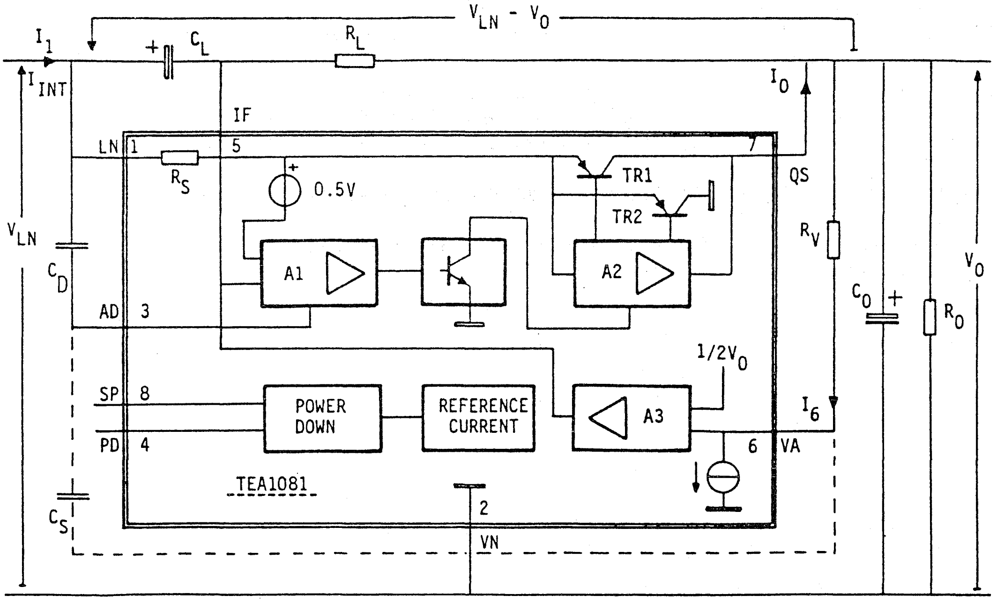


Figure 1: Block diagram of the TEA1081 including external components

Figure 2 on the next page shows the circuit diagram in which pin IF of the TEA1081 is connected to LN of the TEA1064. With the aid of an extra series resistor between LN of the TEA1081 and LN of the TEA1064 a current split effect occurs (next page).

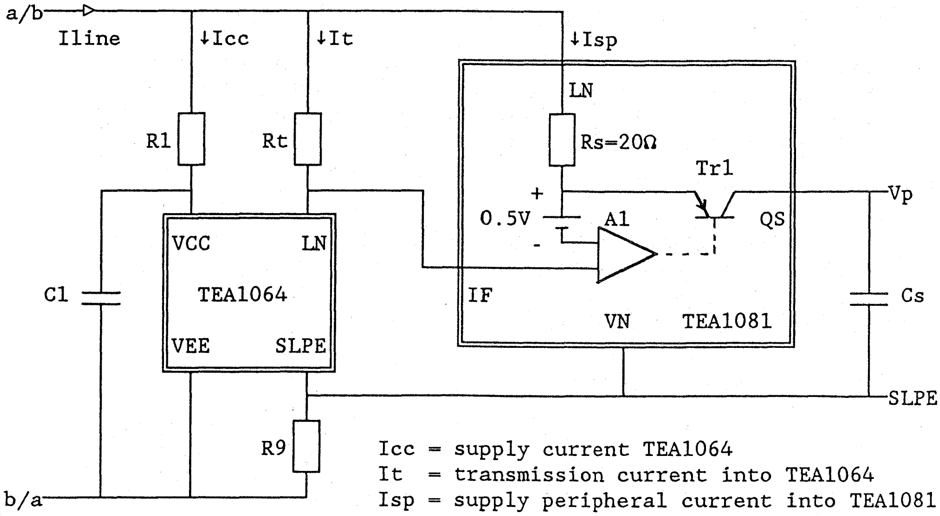


Figure 2: Application diagram TEA1081 and TEA1064 as a current splitter

Operational amplifier A1 of the TEA1081 (see fig 1 and 2) is used in a feedback loop keeping its input voltage 0V.
This means (fig. 2):

$$Isp*Rs + 0.5V = It*Rt \qquad \text{(Formula 1)}$$

$$Iline = Icc + It + Isp \qquad \text{(Formula 2)}$$

Combining formula 1 and 2:

$$Isp = (Iline-Icc) * \frac{Rt}{Rs + Rt} - \frac{0.5}{Rs + Rt} \qquad \text{(Formula 3)}$$

$$It = (Iline-Icc) * \frac{Rs}{Rs + Rt} + \frac{0.5}{Rs + Rt} \qquad \text{(Formula 4)}$$

With Icc = current consumption TEA1064 ≃ 1mA, and Rs=20Ω

2.2 DC-behaviour of the current splitter

The formulas of the previous paragraph show already the current split up. In practice Rt>>Rs so current Isp will be almost equal to Iline-Icc, and current It will be almost 0 (0.5/Rt neglected).
This means almost the total line current flows into the TEA1081 and thus is available for supply of peripherals.

If a practical value of 500Ω is chosen for resistor Rt the following figure can be drawn:

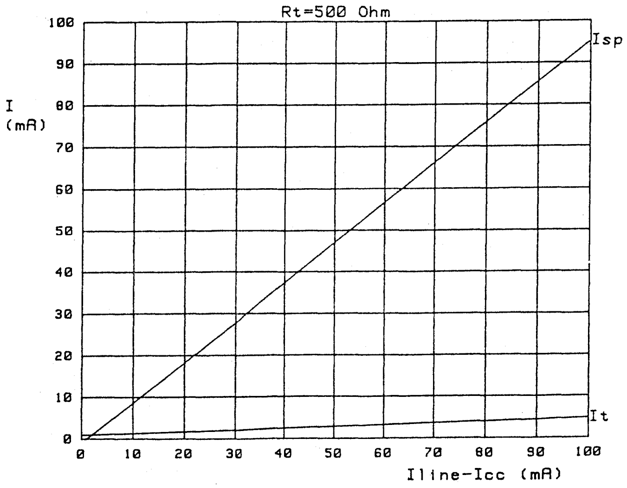


Figure 3: Isp and It versus line current with Rt=500Ω

- At Iline=21mA and Icc=1mA:

  It = 1.7mA
  Isp = 18.3mA

In this case 87% of the line current flows into the TEA1081. At higher line currents this current efficiency increases! At 41mA line current It= 2.5mA and Isp=37.5mA, so 91% of the total line current flows into the TEA1081. If this current is not needed to supply peripheral circuitry connected at the output of the TEA1081, it will be sunk to VN (conn. to SLPE of TEA1064).

However, the maximum output current of the TEA1081 is specified to be less than 30mA. Assuming 100% efficiency (input/output current) of the TEA1081 (low AC line levels) this means the input current of the TEA1081 has to be limited at 30mA.
This can be realized by applying an electronic zener in parallel with Rt to limit the maximum voltage drop between LN and IF of the TEA1081. The value of this zener can be calculated: Ispmax*Rs + 0.5 = Vzener (see fig.2) in which Isp maximum is 30mA and Rs=20Ω, so Vzener = 1.1V.
If the zener voltage is reached the residual part of the line current will flow through the TEA1064 output stage (see also application section 4).

Note:
If the input current of the TEA1081 is limited to 30mA this could mean less than 15mA (50%) is available at the output if a large AC signal is present at the line causing the voltage at LN of the TEA1081 to drop momentarily far below the output voltage! The residual 50% is then sunk to pin VN which is connected to pin SLPE of the TEA1064 (for more detailed information see ref. 2).

The DC current through the TEA1064 can be chosen by means of the value of resistor Rt (fig 2). For correct DC operation this current must be minimum 500uA, which is the current needed for the voltage stabilizer in the output stage and the current flowing to substrate (pin VEE) in the output transistor.

2.3 AC behaviour of the current splitter

As explained in the previous paragraph the DC line current is split up into a small transmission current and a large current going into the TEA1081. But, to obtain a sending signal the output stage of the TEA1064 has to be driven. To drive the output stage DC current is needed.
In an application without the TEA1081 as a current splitter, this current must be rather high: about 5-8mA depending on line level and set and line impedance; for instance to obtain a signal of 1.5Vrms with 600Ω set and line impedance:

$$I_{outputstage}\ (DC) = 1.5V*\sqrt{2}/300\Omega = 7.1mA$$

In the application as discussed here however the AC current for sending is also split up!

From figure 1 and 2 it can be calculated that for AC:

$$it*Rt = isp*Rs \text{ (0.5V short circuited for AC), which means:}$$

$$\frac{it}{isp} = \frac{Rs}{Rt} \text{ (constant ratio!), and}$$

$$it + isp - icc = iline$$

Since in practice Rt>>Rs only very little AC-current is modulated within the output stage of the TEA1064. Most of the AC-current flows through the TEA1081 (current isp). In fact this means that the current flowing into the TEA1081 is used twice:
- to supply peripherals
- to produce a sending signal.

The minimum allowed DC transmission current (It) is dependent on:
- the maximum AC sending level at the line
- the set and line impedance
- minimum available line current
- load at pin LN of the TEA1064

The load at pin LN of the TEA1064 exists almost completely out of the capacitor connected to LN which is added to improve Electro Magnetic Compatibility. The value of this capacitor normally is in the order of 1nF-10nF.

In practice current It will be 1-2mA at the lowest line current (e.g. 15mA at the longest line). This means resistor Rt has a practical value of about 200-600Ω.

3. Interconnections between TDA7050, TEA1081 and TEA1064

The output of the TEA1081 can be used to supply a TDA7050 loudspeaker amplifier to obtain listening in. The TDA7050 incorporates 2 amplifiers which can be used in a stereo (2 amplifiers single ended) or Bridge Tied Load mode.
Since in telephony applications the limiting factor to get power into a 50Ω loudspeaker is current and not voltage (especially at low line currents), it is preferable not to use BTL drive but only 1 amplifier of the TDA7050 in a single ended mode.

The TEA1081 has SLPE of the TEA1064 as ground reference (Ch.2), so the TDA7050 amplifier is biased at SLPE. Since the output Qr+ of the TEA1064 is referenced to VEE, signal SLPE has to be subtracted from Qr+ (signal SLPE is a part of the sending signal!). This can be done by applying signal SLPE as a common mode voltage to the TDA7050 inputs:

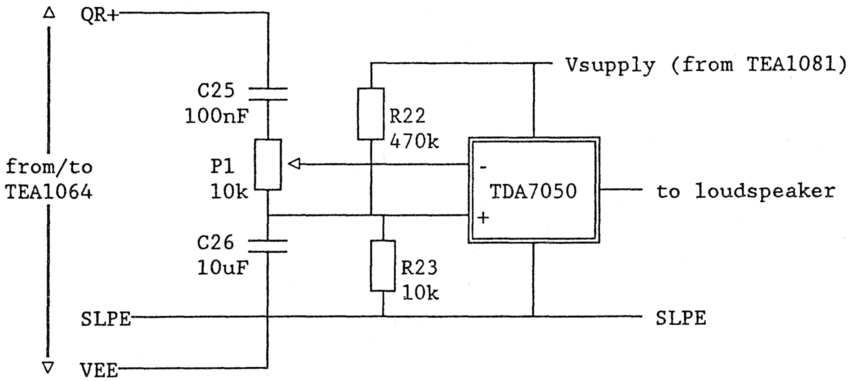


Figure 4: Interconnection between TDA7050 and TEA1064

If we take SLPE as a ground reference (in mind) a signal -SLPE is present at VEE and a signal -SLPE + QR+ is present at the receive output of the TEA1064.
Since the TDA7050 can be seen as an operational amplifier, with internal feed back, only signal Qr+ ought to be present at the output.

In practice it appears the common mode signal -SLPE is too big to be handled by the TDA7050. Distortion occurs at the output of the TDA7050 at the moment a large sending signal is present at the line (and thus at SLPE).
The reason for this distortion is the fact that the pnp input stage of the TDA7050 is momentarily driven below its linear operation area.
This problem is solved by applying a small positive DC bias at the inputs. In this way the Vce voltage of the pnp input transistors is increased. In figure 4 this is done by means of the resistors R22 and R23.

REMARK: If the TDA7050 is driven, current is drawn out of the TEA1081. As a result the supply voltage will drop only little until the TDA7050 consumes all available current (Isp). From that moment on the supply voltage will drop rapidly.
If the voltage at the output of the TEA1081 drops below a value of 1.6V

typical, its start-circuit will become active. In this case the TEA1081 does not act as a current splitter anymore but will use up the total line current to charge the smoothing capacitor at the output.
To prevent this from happening during normal operation an extra diode can be applied in series with the supply line to the TDA7050 (figure 5). With the TDA7050 working down to 1.5V (measurement) the voltage at QS of the TEA1081 is quaranteed to be above 1.5V + Vdiode.

In case of a series diode an extra smoothing capacitor is needed to get rid of ripple caused by the TDA7050 (see figure 5).

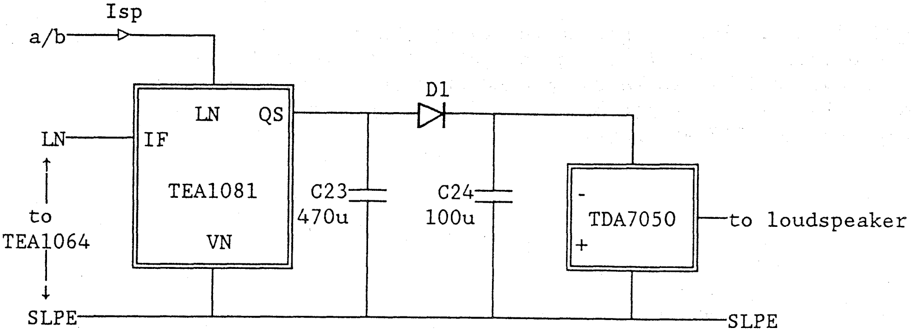


Figure 5: Supply connection of TDA7050

Especially at low line currents the diode does not have a great effect on the available output power of the TDA7050. The limiting factor of output power exists mainly out of the available current (Isp).

4. A listening-in application with TEA1064, TEA1081 and TDA7050

This section describes a complete listening-in application with the TEA1064, TEA1081 and TDA7050. The application was tested and measurement results are included.
The TEA1081 is used as a current splitter, as described in section 2, and supplies the TDA7050 loudspeaker amplifier (as well as other peripherals).

Paragraph 4.1 explains the circuit diagram while paragraph 4.2 gives all measurement results. In paragraph 4.3 some comments on using other TEA106x types without limiter are given.

4.1 Circuit diagram

The circuit diagram of the application as discussed here is shown in figure 6 as shown on the next page.

In principle the TEA1064 is used in its basic application as much as possible. Only some small modifications are necessary to enable the optimum listening-in application. Microphone gain, telephone gain, sidetone circuitry and impedance setting are not changed. Only the reference voltage is pulled up by means of resistor Rva (pins REG, SLPE) of 27.4kΩ to 4.05V typical (normally 3.3V). The application with regulated line voltage (VLN-SLPE is constant) is used.
In series with LN of the TEA1064 a resistor is connected, built up out of R20 and R21, to obtain the current split up facility in conjunction with the TEA1081 (see section 2).
An electronic zener of about 1.1V is formed by transistor T1 and resistors R20 and R21 to ensure the maximum specified output current of the TEA1081 is not exceeded. T1 is built up out of two transistors of the type BC547C. The reason for using 2 transistors is to have less influence of the emitter resistances (better zener effect).

Pin LN of the TEA1064 is connected to pin IF of the TEA1081 via a low pass filter formed by resistor R22 and capacitor C20. This filter is needed to ensures HF stability of the total circuit. Without this filter oscillations occur.
The TDA7050 is connected to the TEA1081 and the TEA1064 as described in section 3 of this report.

Extra peripherals (u-controller, dialer etc.) can be supplied via Vp and SLPE and can interface with the speechcircuit TEA1064 via the signals DTMF, MUTE and PD directly. As explained in the introduction of this section no extra level shifters are necessary.

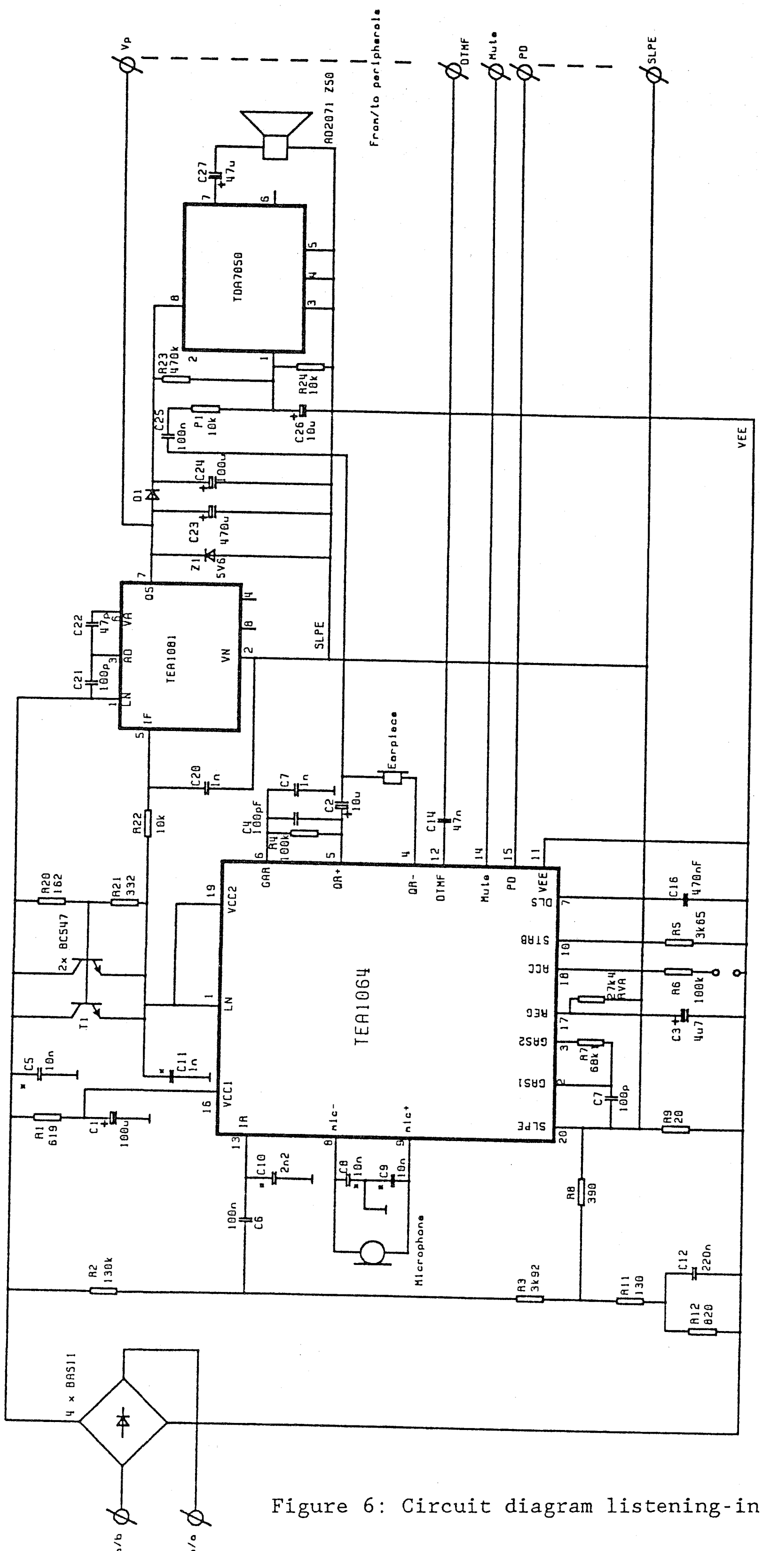


Figure 6: Circuit diagram listening-in application

4.2 Measurement results

This paragraph gives measurement results of the circuit as shown in figure 6 on the previous page. First the current split up is shown followed by the supply capabilities of the TEA1081 and the DC-characteristics. The amount of power into a loudspeaker and the most important AC parameters are measured as well.

Current split up

In order to get a good view on how the current split up is realised figure 6 and 7 show this split up without transistors T1 for different values of Rt (Rt=R20+R21 in figure 6). The formulas of paragraph 2.2 are still valid here:

$$Isp = (Iline-Icc) * \frac{Rt}{Rs + Rt} - \frac{0.5}{Rs + Rt} \quad \text{(Formula 3)}$$

$$It = (Iline-Icc) * \frac{Rs}{Rs + Rt} + \frac{0.5}{Rs + Rt} \quad \text{(Formula 4)}$$

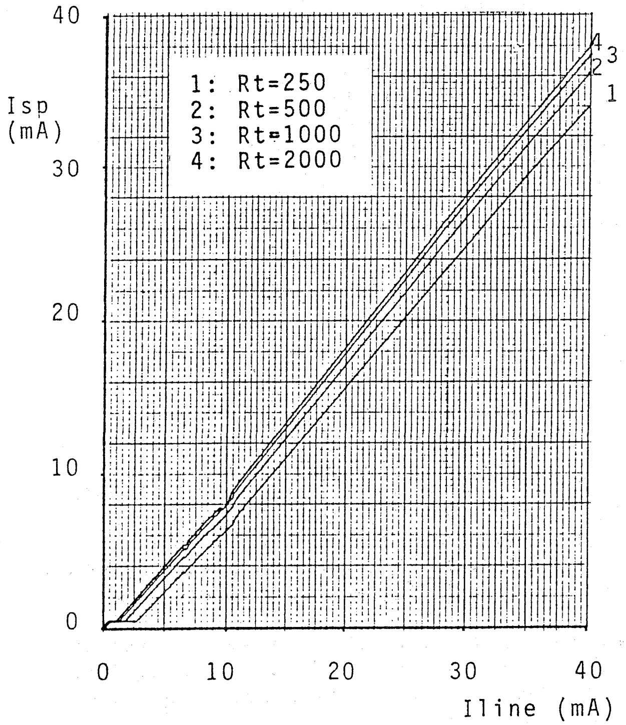


Figure 7: Isp versus Iline without limiting

It is clear the current into the TEA1081 is linear with the line current. At low line currents a start-up behaviour is noticed which is due to the fact that the TEA1064 is in parallel operation area and thus keeping the line voltage low. At DC line voltages above 2.5V (specifications) the TEA1081 gets active and starts splitting up the line current. This start-up effect can be noticed even better in figure 8 which shows the current through the TEA1064:

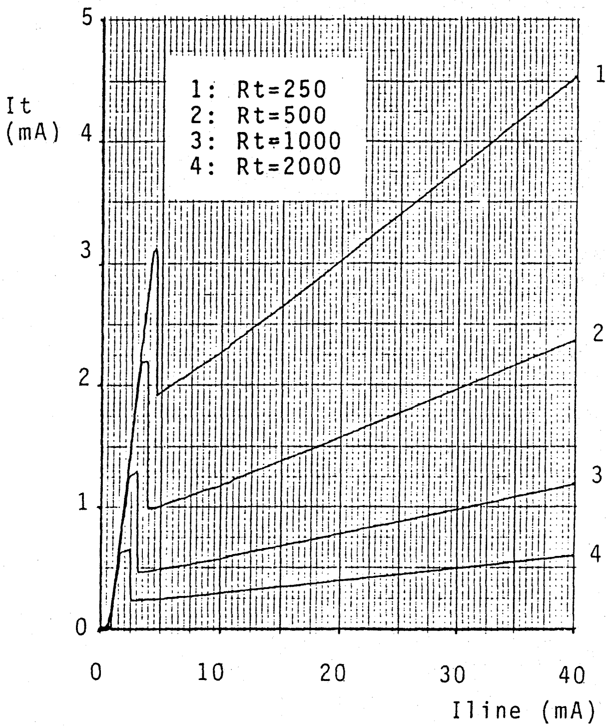


Figure 8: It versus Iline without limiting

Figure 8 shows that for Rt=500Ω the TEA1081 gets operative at Iline = 4mA. For line currents above 4mA It has a linear relation to Iline.

Since the output current of the TEA1081 has a maximum specified value of 30mA a current limitation has to be incorporated in the circuit diagram. Limiting the current at the output of the TEA1081 is not possible so the input current has to be limited. As described this can be realised by means of T1 which, in conjunction with R20 and R21, forms an electronic zener.

Without limitation the output current could reach 30mA. In such a case the input current will always be higher then 30mA due to the fact the TEA1081 consumes current (≈1mA) and some current (up to 50% dependent on AC line signal) is shunted to earth (SLPE of TEA1064).
So a limitation of input current of about 30mA guarantees the output current to stay below 30mA.
Figure 9 shows the effect of current limitation (zener effect). Curve 1 was measured without T1 (similar to fig.7), curve 2 shows the effect if 1 transistor of the type BC547C is used as shown in the circuit diagram. Curve 3 shows the same effect if T1 is built up with 2 BC547C in parallel.

The temperature dependance of this zener will cause a higher voltage drop at low temperatures. This will result in higher currents through the TEA1081. However, the output current of the TEA1081 will be maximum at large receiving signals (TDA7050 consumes current). At large receiving signals the efficiency of the TEA1081 will decrease (down to 50% max.) and thus the output current will be smaller than the input current. This means the zener voltage may deviate a few percent while still the output current will be protected to 30mA maximum (input current 120mA maximum).

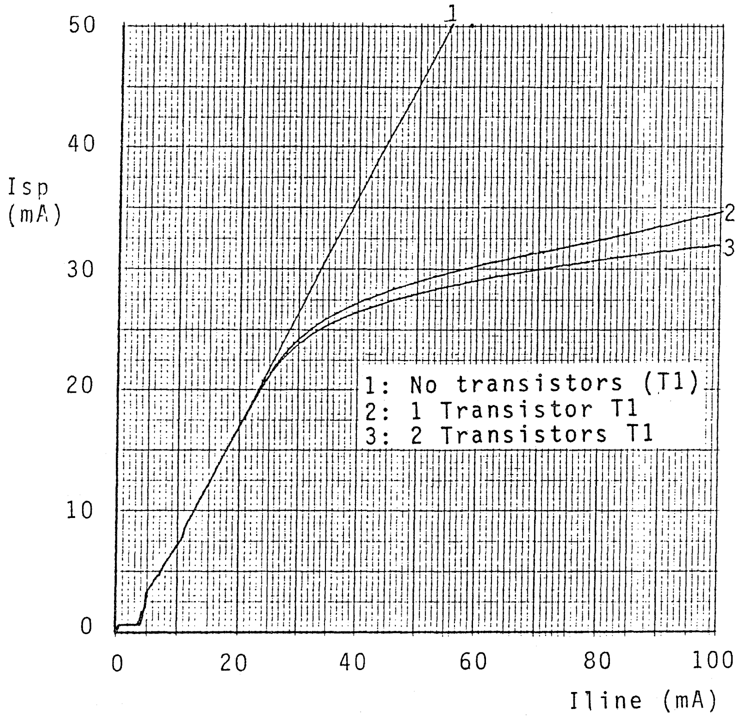


Figure 9: Isp versus Iline

Similar curves were measured for the current through the TEA1064 as shown in figure 10 (next page).

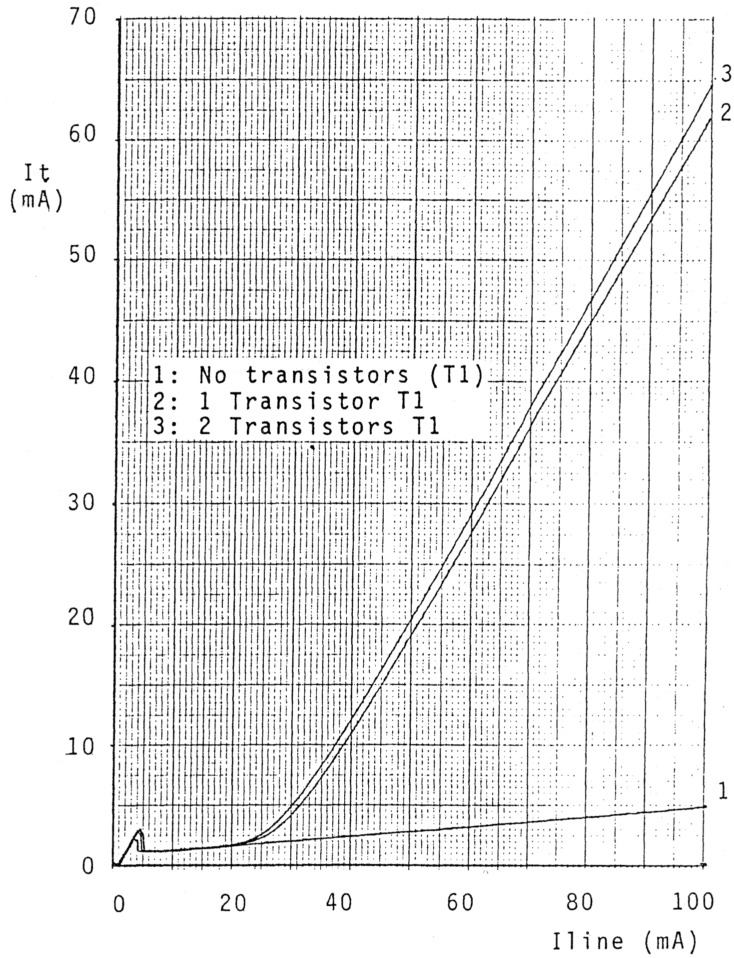


Figure 10: It versus Iline

Remark: The measurements which are described in the following pages all refer to a zener realized with 2 transistors as shown in figure 6. The transistors do not distort the sending signal!

Output current TEA1081

Figure 11 shows the output voltage of the TEA1081 (Vp) versus the output current (Ip) of the TEA1081 for several line currents.

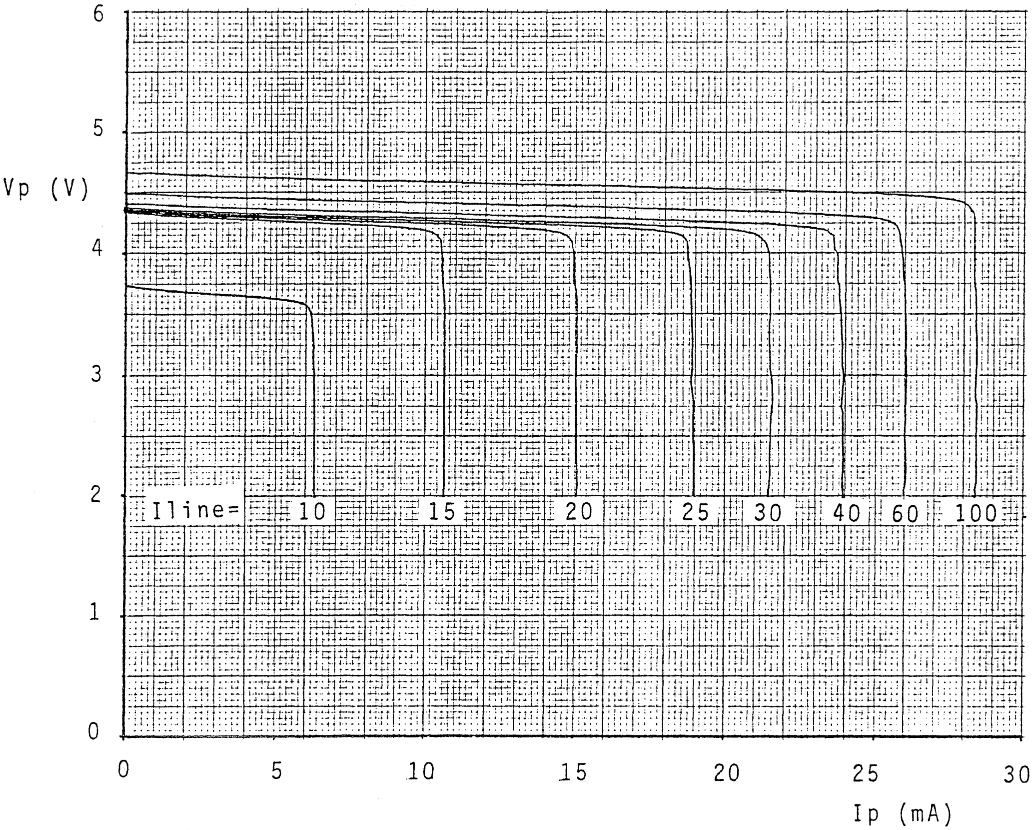


Figure 11: Vp versus Ip

The curves show clearly the rapidly decreasing voltage Vp as a result of using up all available current (Isp). The curves were measured down to Vp=2V. Lower values can be reached but voltages below 1.6V typ. must be avoided since the TEA1081 will stop splitting up current below this value (internal start circuit gets active). As already described in the previous sections this is the reason for applying a series diode to supply the TDA7050.

The curves of figure 11 change if a sending signal is present at the line. Figure 12 shows Vp versus Ip if a sending level of 0dBm is present at the telephone lines (next page).

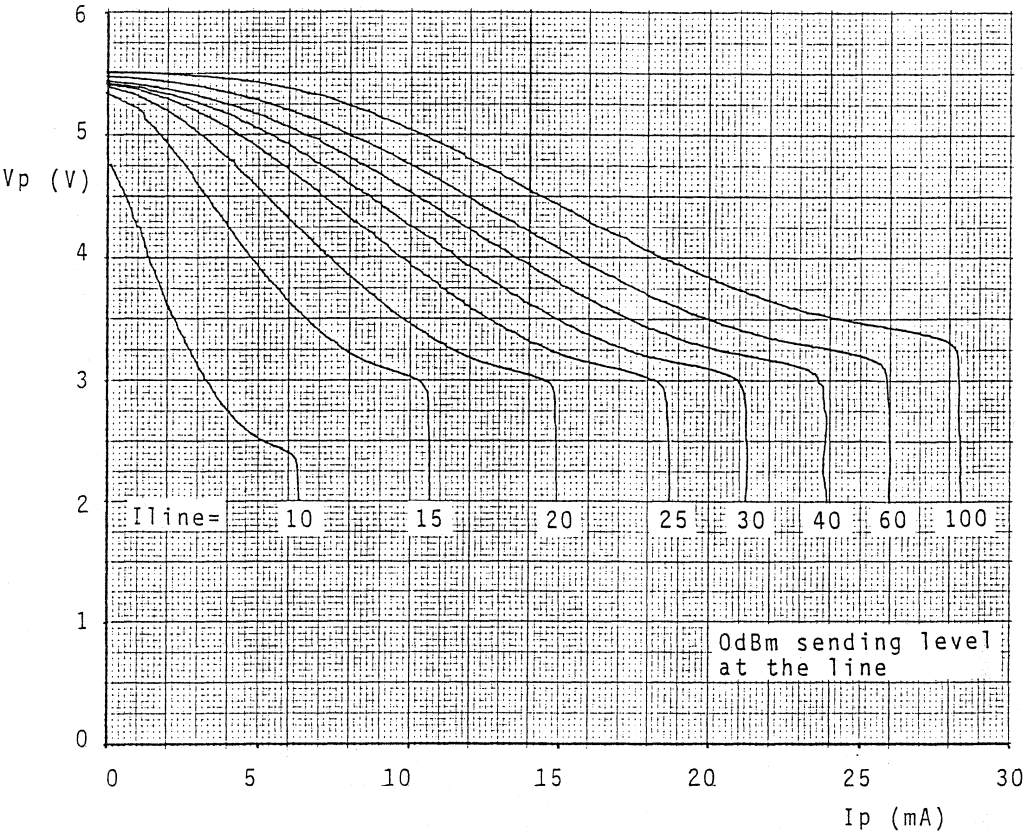


Figure 12: Vp versus Ip with 0dBm sending level at the line

Figure 12 shows that for low peripheral currents the output voltage of the TEA1081 increases compared to the situation were no sending level is present. This is due to the fact that the line voltage (voltage LN-SLPE) is momentarily higher. In this situation the DC current Isp flows completely to the output of the TEA1081, until the moment the line voltage drops below Vp+0.5V (0.5V = minimum voltage drop pin LN to QS TEA1081). In fact the TEA1081 acts as a sort of top-detector. However, the power efficiency of the TEA1081 decreases due to the fact that at line voltages lower than Vp+0.5V all current Isp flows to VN (=SLPE TEA1064).
To prevent the DC output voltage of the TEA1081 exceeding the limits of the maximum specified DC supply voltage for peripherals, a zener of 5V6 is applied (Z1 in circuit diagram fig.6).

At 15mA line current still 10mA/3V is available for peripherals. This is a very good figure since the TEA1064 and TEA1081 internally consume about 2mA and normally, without current splitting, the sending signal would need 3.6mA DC current through the output stage of the TEA1064 (0.775Vrms into 600Ω//600Ω).

For a 0dBm receiving signal at the line the situation gets even better because no AC signal is present at SLPE:

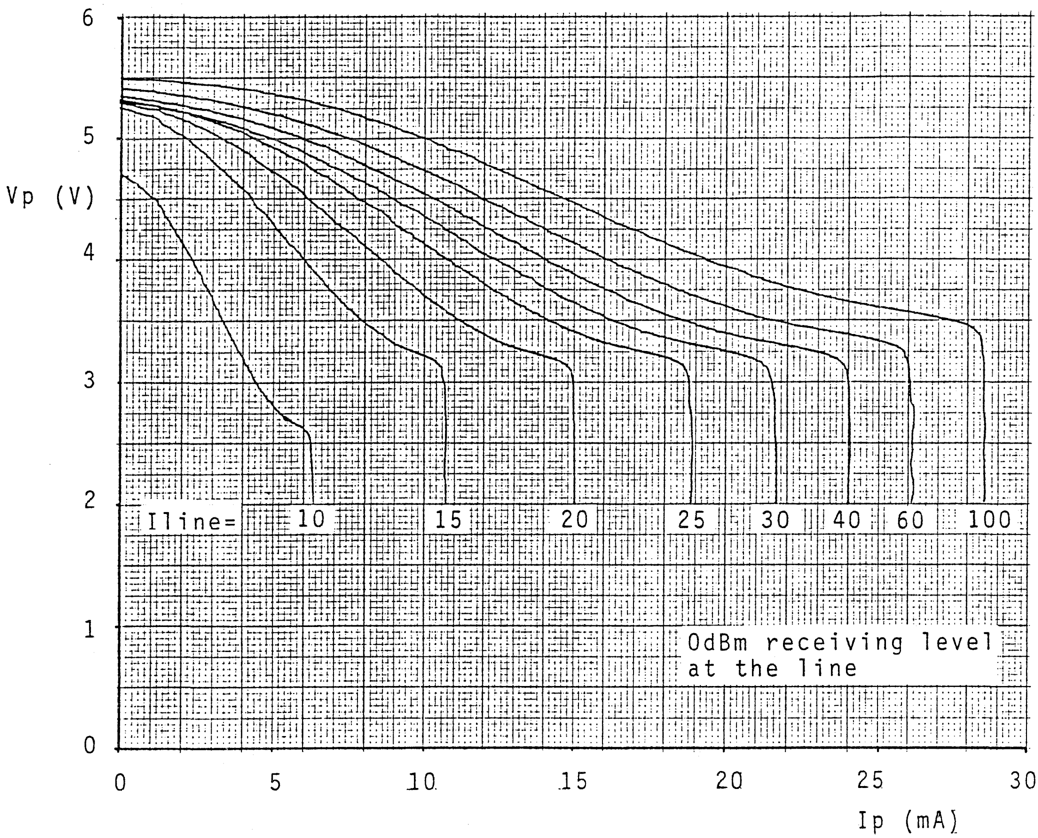


Figure 13: Vp versus Ip with 0dBm receiving level at the line

DC characteristics

Figure 14 shows some specific voltages of the application versus line current.

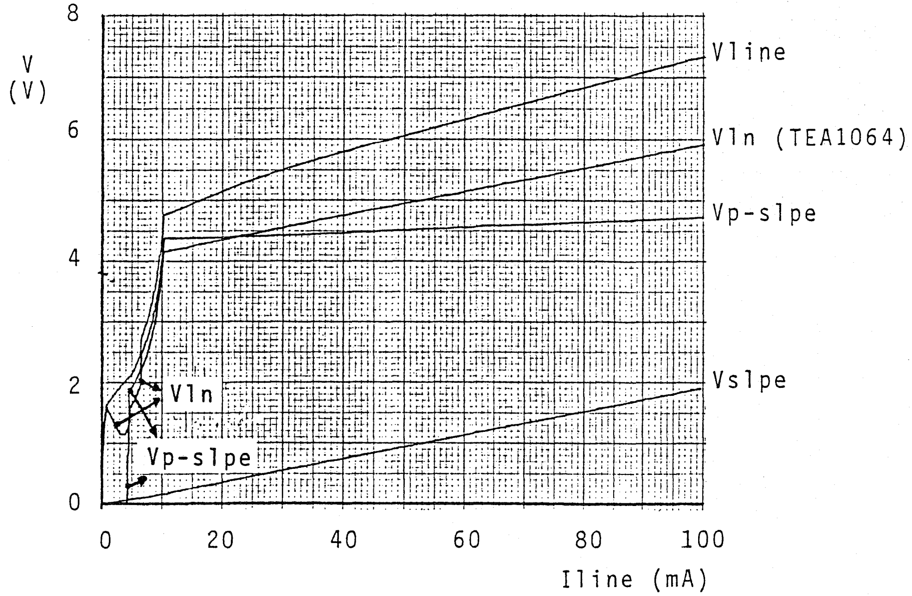


Figure 14: Most important voltages of the application versus line current

Vline is the voltage at pin LN of the TEA1081 and has a value of 5.2V at 20mA line current. Taking into account a voltage drop of 1.4V for the diode bridge this means 6.6V/20mA is present at the telephone line. This value can be considered to be an average of PTT requirements for line voltage in several countries.
The line voltage can be decreased by decreasing the reference voltage of the TEA1064 (Rva in fig.6). The available amount of power at the output of the TEA1081 however will also decrease in that case.
More output power for peripherals will be available if the DC line voltage is increased. This can be done by applying a series diode (or zener) in series with LN of the speechcircuit TEA1064.

Vp of figure 14 represents the voltage at the output of the TEA1081. This curve shows the TEA1081 gets operative at currents above 4mA (Vline > 2.1V).

Power into a 50Ω loudspeaker

The amount of power which can be put into loudspeaker connected to the TDA7050 is dependent on:
- line current
- AC line voltage
- Loudspeaker impedance

Curves 15 and 16 were measured at a constant AC receiving level at the line of 100mVrms and 500mVrms respectively.
Measurements of power were done by measuring the voltage across a 50Ω loudspeaker. Power is then defined as: $P = v^2/50\Omega$

The AC voltage at the output of the TDA7050 was adjusted by means of the potentiometer at its input (P1 circuit diagram fig.6) to a maximum at THD < 2%.

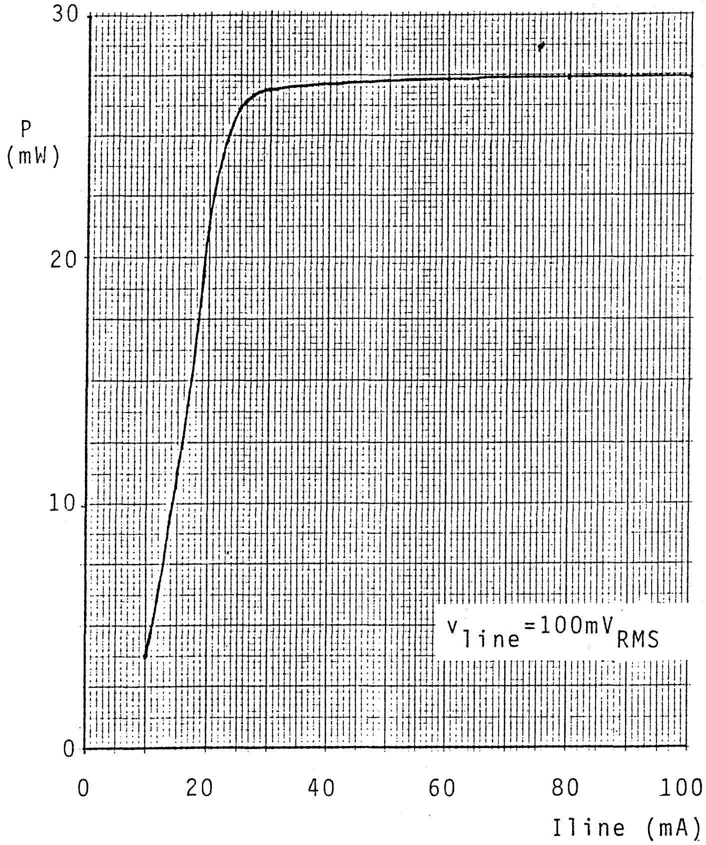


Figure 15: Power into 50Ω loudspeaker at 100mV line voltage

Figure 15 shows a curve which is very realistic since in practice the average value of receiving level at the line is about 100mV. Figure 16 shows that at large receiving levels the efficiency of the TEA1081 decreases and thus less power is available (next page).

At 20mA line current and 100mVrms line voltage more than 20mW power can be put into a loudspeaker of the type AD2071/Z50 (fig.15). This loudspeaker has a specified sensitivity of 90dBspl at 0.5m at 0.55W/1kHz. So at 20mW the sound pressure level is: 90dBspl - 10*log(550mW/20mW) = 75,6dBspl.
This is a very good figure since for instance the CNET requires a minimum level of 70dBspl at 0.5m with a line level of -20dBm (77.5mV).

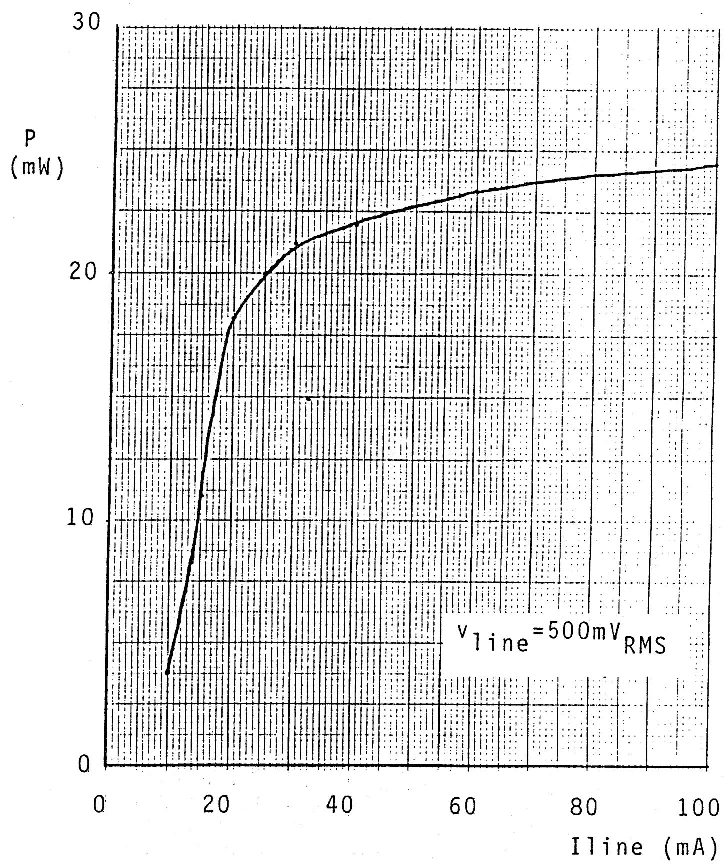


Figure 16: Power into 50Ω loudspeaker at 500mV line voltage

Under all receiving conditions sending is still possible!! The amount of available power at the output of the TEA1081 however will decrease (efficiency decreases at large line signals).

Maximum sending level:
Figure 17 (next page) shows a curve of maximum sending level (line load=600Ω, F=1kHz, THD<2%) versus line current. The curve was measured with the potentiometer P1 at maximum, to put maximum power (via sidetone) into the loudspeaker.
At frequencies above 1kHz the maximum transmit level decreases. This is caused by the RC filter between LN of the TEA1064 and IF of the TEA1081 disturbing the ideal current split-up. However, due to the dynamic limiter no distortion occurs! At the application as shown the following values of max. transmit level were measured at 15mA of line current:

300Hz : 1.85Vrms
3400Hz: 1.35Vrms
4000Hz: 1.15Vrms

By increasing the current through the TEA1064 this effect can be minimized.

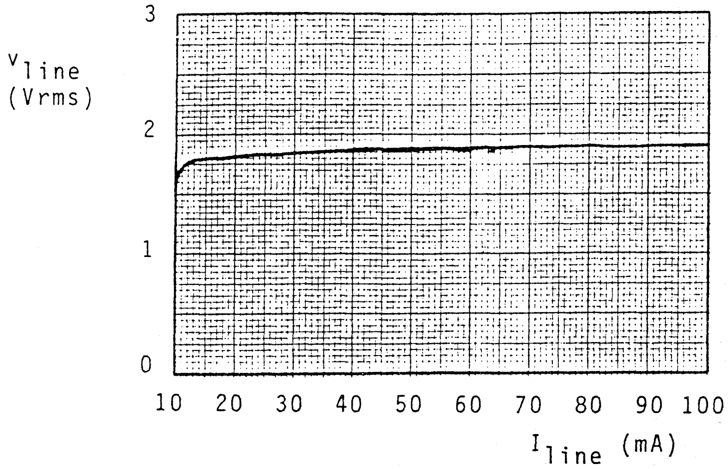


Figure 17: Max. sending level versus line current

Start up time
Figure 18 shows the start up behaviour of the application (all capacitors discharged) at a line current of 20mA. Curve "Vline" represents the line voltage while curve "Vp-Vee" shows the output voltage of the TEA1081. Both curves were measured with reference VEE of the TEA1064. For curve "Vp-Vee" this means the actual available output voltage is 0.4V (Iline * 20Ω) lower because the TEA1081 is referenced to SLPE of the TEA1064.

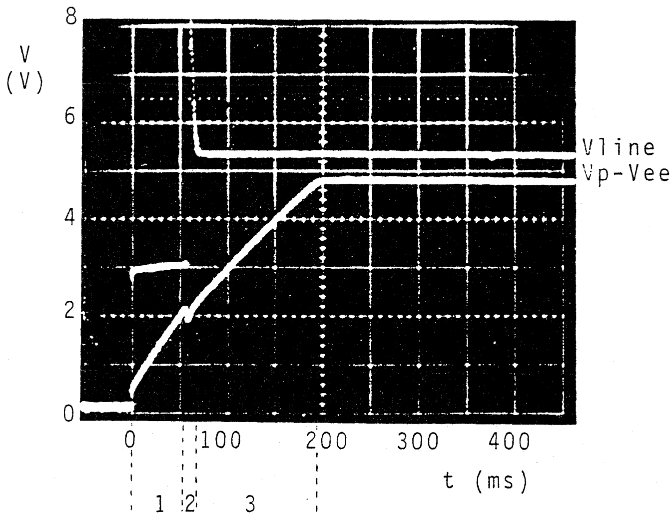


Figure 16: Start-up behaviour of line voltage and output voltage

As shown the DC start up time mainly exists out of 3 parts:

1. In the first part the TEA1081 is using up all line current to charge the

capacitor at its output (fig.6 C23 and C24). The line voltage is mainly controlled by the TEA1081 + voltage at pin SLPE (totally about 3V). During this time also the capacitor at pin VCC1 is charged as well as the capacitor at pin REG (see circuit diagram fig.5).
If the output voltage of the TEA1081 reaches 1.6V typ. (2V in fig.18) it gets active and starts splitting up the line current.

2. In the second part which follows directly after, the TEA1064 is switched in and the line voltage gets stable. This time is very short because the most important capacitors (at VCC1 and REG) are already charged for the most part.

3. In the third part all available peripheral current (Isp) is used to charge the output capacitor of the TEA1081 to its maximum value.

As shown in figure 18 the line voltage is stable after 70ms. The output voltage of the TEA1081 however reaches its maximum value after 200ms.
At higher line currents the total start up time will decrease!
The DC start up time can also be decreased by decreasing the value of capacitors C23 and C24, but this will cause more ripple at the supply voltage.

Balance Return Loss
Figure 19 shows BRL figures referenced to 600Ω as a function of frequency:

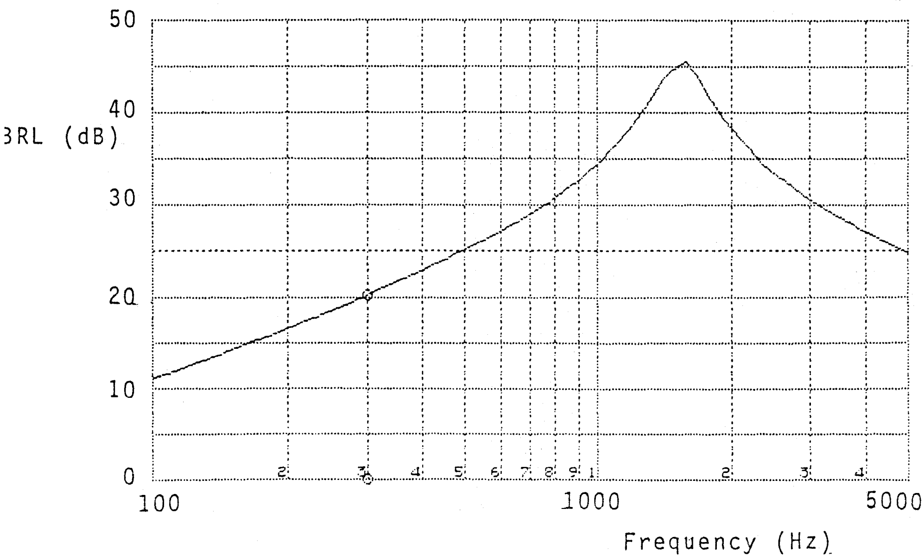


Figure 19: Balance return loss at 20mA line current

The curve shows only little deviation with the normal application of the TEA1064 which proves that the influence of the TEA1081 as a current splitter on BRL can be neglected. The curve was measured at 20mA line current. At higher line currents the same curves are valid (even a little better).

Noise

Measurements showed no deviation from the basic application of the TEA1064 concerning noise level at the line! This means at 15mA line current and the microphone gain set at 52dB (R7=68k) only -72dBmp is present at the line (psophometrically weighted, load at mic+, mic- =200Ω). This means the noise contribution of the TEA1081 can be neglected in this case (according to spec. only -83dBm in basic application).

5. Using other TEA106x family member:

In case another TEA106x (TEA1060/61/66/67/68) is used the absence of the dynamic limiter will cause the following problems:

- At very low line currents (below 15mA) clipping of the sending signal occurs at the top of the sinewave. Since in those cases the current through the application is momentarily zero, the TEA1081 is not supplied anymore (also current less due to split-up). As a result small oscillations can be expected.

- The sending signal at the line can also clip at the bottom of the sinewave. Normally (without TEA1081) this is caused by saturation of the TEA106x output transistor. However, with the TEA1081 current splitter application, the TEA1081 might clip before the TEA106x does. This can be noticed by a strange two level switching behaviour at the bottom of the clipped sinewave.
The effect will vanish if sufficient DC voltage drop is applied in series with pin LN of the TEA106x (e.g. 100mV is sufficient).

- A slewing effect occurs if the current in the TEA106x output stage is totally used for sending. This condition can in practice easily be reached due to the RC- stability filter between LN of the TEA106x and IF of the TEA1081 which affects the ideal current split-up. The effect of slewing can be noticed especially at large sending level at high frequency speech signals.
The effect can be diminished by increasing the DC current through the TEA106x but will never completely vanish.

With a dynamic limiter at the sending stage these problems do not occur. So, it is clear the use of a dynamic limiter is essential in the current split application.

6. Conclusions

The new application of the TEA1081 as a current splitter between LN and SLPE of the speech circuit TEA1064 incorporates some big advantages. The main advantage, compared to the normal use of the TEA1081 as an inductor, is the increased amount of peripheral current which is available under all sending and receiving conditions. Also the fact that the TEA1064 output stage can not become inactive due to lack of current is a big improvement.

The idea of using the peripheral current also for making a sending signal works very well. Especially for countries were PTT requirements require a rather low line voltage the application is very useful.

The overview below gives the three possible applications for the TEA1081 with comments. A minus (-) means a disadvantage, a plus (+) means an advantage compared to the other applications:

TEA1081 as an inductor + TEA1064 with relative (compared to Iline) high current into TEA1081 (to peripherals)		TEA1081 as a current splitter + TEA1064 with 1-2mA Itransmission	
LN-VEE	LN-SLPE	LN-VEE	LN-SLPE
-Current 'It' is not known, might reach 0	-Current 'It' is not known might reach 0	-not possible	+fixed current 'It'
-no/less send-level (low It)	-no/less send-level (low It)		+normal send-level
-AGC not operative	+AGC operative		+AGC operative
-wrong parallel operation	+parallel operation		+parallel operation
-Interface to dialer	+interface to dialer		+interface to dialer

It is clear the application with the TEA1081 as a current splitter between LN and SLPE incorporates a lot of advantages.

As discussed in chapter 5 a dynamic limiter is essential for correct behaviour of the application.

References:

1. Philips components report: "TEA1060 family, Versatile Speech/Transmission ICs for Electronic Telephone Sets, Designers Guide" ; by P.J.M. Sijbers; 12nc. 9398 341 10011

2. Philips components report: "TEA1081: a Supply IC for Peripheral Circuits in Electronic Telephone Sets" ; by F. van Dongen; 12nc. 9398 058 30011

APPLICATION NOTE Nr	ETT/AN91016
TITLE	TEA1085A/TEA1085 - A listening-in facility for electronic telephone sets
AUTHOR	F. van Dongen
DATE	September 1991

Table of Contents:

<u>List of tables and figures</u>

1. INTRODUCTION

Appendix F gives a list of all abbreviations in this report.

The TEA1085A and TEA1085 are bipolar ICs offering a listening-in facility of the received line signal via a loudspeaker. They have to be applied in line powered telephone sets in combination with a speech/transmission circuit of the TEA1060 family (named the TEA106X).

The TEA106X provides the interface functions between telephone line and dialling IC, microphone, earpiece and between line and loudspeaker via the TEA1085 or TEA1085A.
Most of the line current is available to power the loudspeaker due to the application structure of the TEA106X with the TEA1085A or TEA1085.

The TEA1085A and TEA1085 are members of the TEA108X "line monitoring IC-family". They are in this report named as the 'TEA1085/A', unless there is a difference between both ICs.
An overview of the TEA108X ICs is given in table 1.

Table 1. Overview of TEA108X line monitoring ICs and there facilities.

Product:	TEA1082	TEA1083	TEA1083A	**TEA1085/A**
Application area 1)	call progress monitoring			listening-in
PD facility	x		x	x
LSE or MUTE facility		x	x	x 2)
Dynamic limiter				x
Howling limiter				x
VBB setting				x
SEL	x	x	x	x
BTL				x
Number of pins 3)	8	8	16	24

Notes:

1) Call progress monitoring, to enable audible monitoring of the progress of the call attempt by a loudspeaker, is recommended for telephone sets with automatic on-hook dialling facilities by the 'ETSI'. Only a minimum loudspeaker level (50dBA) has to be guaranteed. Facilities are not required.
Documentation of the call progress monitoring ICs is listed in ref.3 of ch.7.

Listening-in sets, reproducing the receive speech signals during conversation mode, have to offer the user more facilities.
Howling limiting, to reduce annoying loudspeaker and line signals, and dynamic limiting of the loudspeaker signal with respect to supply conditions are required. Acoustic output levels for listening-in are in the order of 70-75dBA, corresponding with a loudspeaker level of $\approx 1V_{rms}$ ($P_O \approx 20mW$) across a 50Ω loudspeaker, Philips type AD2071/Z50.
The TEA1085 and the TEA1085A cover both application areas.

2) The MUTE of the TEA1085A has a logic input, while the MUTE of the TEA1085 is provided with a toggle input.

3) Consult product specification concerning package outlines.

This report describes the TEA1085/A and an application example of the TEA1085/A in a listening-in set.
An application of the TEA1085A with the transmission circuit TEA1064A is realized on a demonstration board. This DEMO board is described in ref.2; it is available under identification number PR4516X.
An adjustment procedure and some application hints for use with the TEA1085/A are listed in appendix E.

The data sheet of the listening-in IC can be found in ref.1 (ch.7).

2. BLOCK DIAGRAM AND PINNING OF THE TEA1085A and TEA1085

The block diagram is shown in fig.1, while the pinning is given in fig.2.

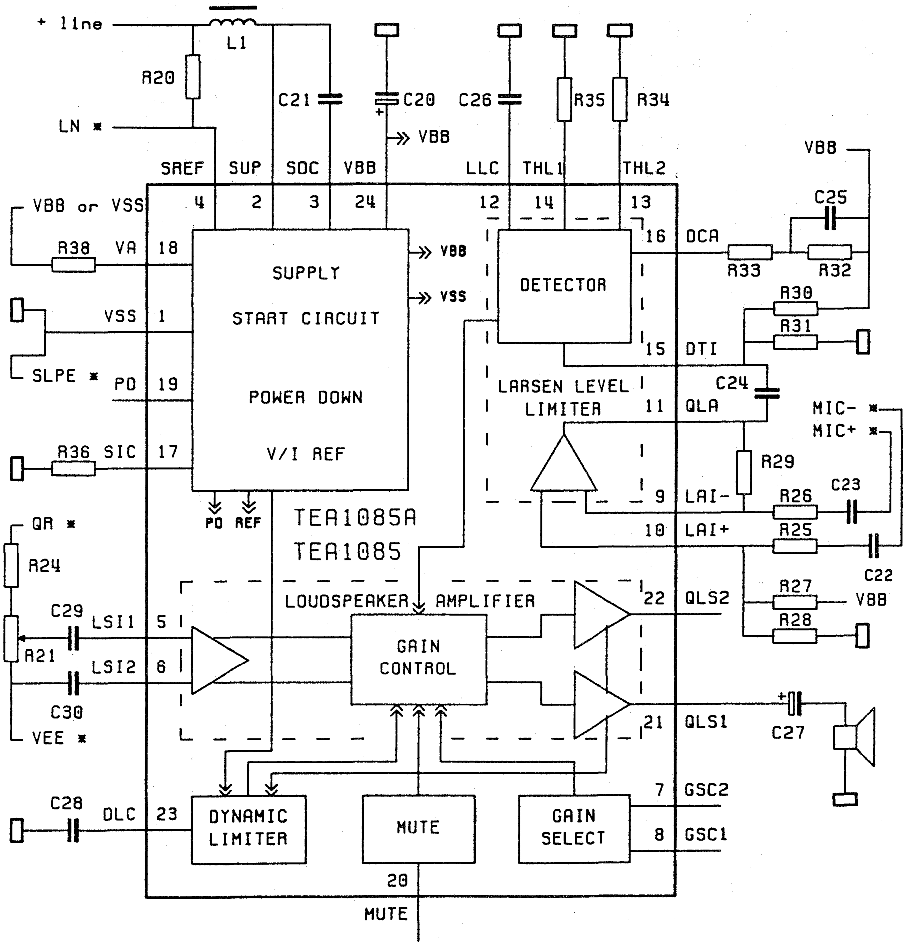


Fig.1 Block diagram of the TEA1085/A with external components.
('*' means a connection with a pin of the TEA106X)

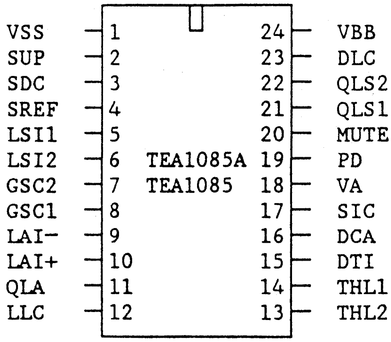


Fig.2. Pinning diagram

PINNING

1	VSS	Negative supply
2	SUP	Positive supply
3	SDC	Supply amplifier decoupling
4	SREF	Supply reference input
5	LSI1	Loudspeaker amplifier input 1
6	LSI2	Loudspeaker amplifier input 2
7	GSC2	Logic input 2 for gain select
8	GSC1	Logic input 1 for gain select
9	LAI−	Larsen limiter preamplifier inverting input
10	LAI+	Larsen limiter preamplifier non-inverting input
11	QLA	Larsen limiter preamplifier output
12	LLC	Larsen limiter capacitor
13	THL2	Larsen limiter residual threshold level
14	THL1	Larsen limiter attack delay threshold level
15	DTI	Larsen limiter detector input
16	DCA	Larsen limiter detector current adjustment
17	SIC	Larsen limiter current stabilizer
18	VA	VBB voltage adjustment
19	PD	Power down input
20	MUTE	Mute input
21	QLS1	Loudspeaker amplifier output 1
22	QLS2	Loudspeaker amplifier output 2
23	DLC	Dynamic limiter capacitor
24	VBB	Stabilized supply voltage

2.1. Overview of internal functions and external components

Referred to the blockdiagram of fig.1 an overview of the internal functions and the purpose of the external components is given:

- **Supply**: forms the interface between the line terminal and internal circuitry; it controls the split up of line current between bias current of the transmission IC and supply current of the TEA1085/A. This block includes furthermore a voltage reference and a stabilized supply voltage VBB for internal and external use.

  R20 - Determines the bias current of the transmission IC.
  C21, L1 - Stability components.
  C20 - Buffer capacitor of the supply voltage VBB.
  R38 - Resistor to adjust the VBB voltage.
  R36 - Determines an internal reference current.

- **Start circuit**: provides, after hook-off, a correct start up of the internal circuitry and charges quickly the buffer and timing capacitors .

- **Power down circuitry**: reduces the internal current consumption from input SUP and supply voltage VBB during pulse dialling or register recall.

- **Loudspeaker amplifier**: transfers, via the pre-amplifier, the received speech signal from the transmission IC into a current for the power amplifiers which drive the loudspeaker connected as a SEL or BTL.

  R21 - Potentiometer to control the level of the loudspeaker signal.
  C29, C30 - Coupling capacitors between receiver output of the transmission IC and loudspeaker amplifier inputs.
  C27 - Coupling capacitor of the loudspeaker.

- **Gain select**: reduces the gain of the loudspeaker amplifier according the logic state of the two GSC inputs.

- **Dynamic limiter**: limits the distortion of the loudspeaker signal at overdrive, overload or limited supply conditions.

  C28 - Determines the attack and release times.

- **Mute function**: offers the possibility to switch over between standby and LI condition. The MUTE input is provided with a logic function in the TEA1085A and with a toggle function in the TEA1085.

- **Larsen Level Limiter**: reduces the receive gain of the loudspeaker amplifier when a howling signal, due to acoustic coupling between loudspeaker and handset microphone, is detected. For normal speech signals, the LLL has no effect on the conversation behaviour.

  C22, R25, C23, R26 - First section of high pass filter and coupling between microphone and pre-amplifier.
  R26, R29 - Gain setting of pre-amplifier.
  R27, R28 - Bias resistors of pre-amplifier input.
  C24 - Coupling capacitor between pre-amplifier output and detector input.

C26 - Determines the attack and release times.
R30, R31 - Bias resistors of the detector input. In combination with C24 second section of high pass filter.
R32, R33 - Detector current setting.
R33, C25 - Third section of high pass filter.
R35, R34 - Setting of attack and residual thresholds.

3. INTERCONNECTION WITH THE TRANSMISSION IC

The application of the TEA1085/A has to be connected between the positive line terminal and terminal SLPE of the TEA106X, as shown in fig.3.

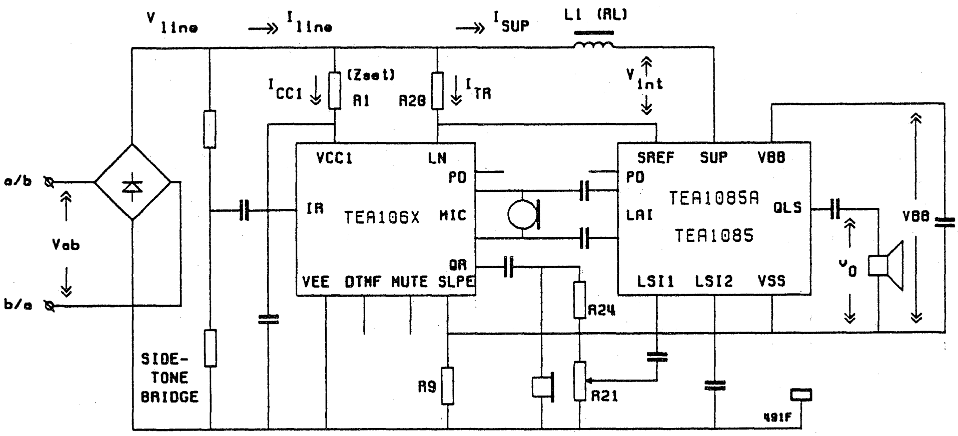


Fig.3 Interconnections of the TEA1085/A with the TEA106X.

Chapter 5 describes an application of the TEA1085/A with the TEA1064A. The TEA1064A is chosen because of the common reference of the dialler interface signals of the TEA1064A as well as the TEA1085/A (MUTE, PD, DTMF).
Instead of the TEA1064A, other members of the TEA1060 family (TEA106X) may be applied. However, interface circuitry between TEA1085/A, TEA106X and control peripherals may be necessary, depending on the supply reference of the peripherals.

3.1 Line current split up

The TEA1085/A is supplied by the line current, which can be in the range of 10 mA up to 140 mA. Most of the line current flows through the TEA1085/A (from SUP to VSS), due to the LINE-SLPE application structure. Only a small constant part of the line current is used by the TEA106X.

As shown in fig.3, the line current is divided into the currents I_{CC1}, I_{TR} and I_{SUP}. I_{CC1} supplies the internal circuitry of the TEA106X, while I_{TR} biases the output stage of the TEA106X. I_{SUP} is consumed by the listening-in IC. These currents can be calculated as follows, whereby the DC resistance RL of coil L1 is taken into account and the relative small DC current in the side tone bridge is neglected:

$$I_{TR} = (I_{line} - I_{CC1}) \cdot \frac{RL}{R20 + RL} + \frac{V_{int}}{R20 + RL} \qquad (A)$$

$$I_{SUP} = (I_{line} - I_{CC1}) \cdot \frac{R20}{R20 + RL} - \frac{V_{int}}{R20 + RL} \qquad (A)$$

where:

V_{int} is an internal bias voltage between SUP and SREF. V_{int} = 315mV typical, R20 = 150Ω.

In case RL << R20: $I_{TR} \approx 2mA$ and $I_{SUP} \approx I_{line} - I_{CC1} - 2mA$.

The supply current available to power the loudspeaker is less than I_{SUP}. As will be explained in ch.4 in more detail, this is due to the following factors:

- Efficiency of the supply of the TEA1085/A depending on DC settings and level of the line signal.
- Standby current I_{SUP0} of the TEA1085/A.
- Amplitude of the transmit current to modulate the line current.
- Current consumption of the peripherals, if any, connected to VBB.

3.2. Line signal transfer

The line signal, which has to be monitored by the loudspeaker, is available at the receiver output (QR) of the TEA106X. This received line signal is coupled into the LSI inputs of the loudspeaker amplifier.

To reduce the so called howling effect between loudspeaker and microphone the microphone signal is applied to the LAI inputs for further processing in the LLL.

3.3. Logic interfaces

The PD circuits of the TEA106X and TEA1085/A reduce the current consumption from VCC1 (and VBB) during pulse dialling. They have the same input structure and can be controlled in parallel, in principal, from the same source.

On the contrary the MUTE functions of the applied ICs are different. The MUTE of the transmission IC disables speech during dialling, while the MUTE of TEA1085/A can be used to switch the loudspeaker amplifier on (LI mode) or off (standby mode).

4. DESCRIPTION OF THE TEA1085/A

4.1. Start up behaviour

During start up of the telephone set the major part of the line current is available to initiate the transmission functions of the applied transmission IC.
The current taken by the TEA1085/A is between 4mA and 6mA depending on the line current and the voltage difference between SUP and VSS during start up.

The start circuit of the TEA1085/A provides:

- Quick charging of the timing capacitors C28 and C26 of the dynamic limiter respectively Larsen Level Limiter.
- Quick charging of the VBB storage capacitor C20.
- Start up of the MUTE circuitry of the TEA1085 in standby mode. See par.4.4.
- Suspension of the operation of the voltage stabilizer and loudspeaker amplifier, to prevent disturbance of the start procedure.

Fig.4 illustrates the increase of the VBB voltage whereby capacitor C20 is charged. The timing of the start procedure of the TEA1085/A depends mainly on the value of C20.
At VBB < 0.65V capacitor C20 is charged with ≈ 4mA, while C26 and C28 are charged with 30μA respectively 60μA. The charge currents of C26 and C28 are doubled and I_{C20} is enlarged to ≈ 6mA when VBB exceeds the 0.65V limit.

Up to VBB ≈ 2.1V the major part of the line current is available to start the TEA106X.
The supply part of the TEA1085/A starts to operate at VBB ≈ 2.1V; the currents I_{TR} and I_{SUP} then will reach their nominal values.
A stop circuit terminates the start period at VBB ≈ 2.7V; the start currents of C26 and C28 are switched off. The charge current of C20 depends now on I_{SUP} reduced by the internal current consumption of the TEA1085/A.

The total start time of the LI application, whereby VBB reaches its nominal value of 3.6V, depends on I_{SUP} and the value of C20. At I_{SUP} = 15mA the start time measures less than 300ms. Higher current levels reduce the start time.

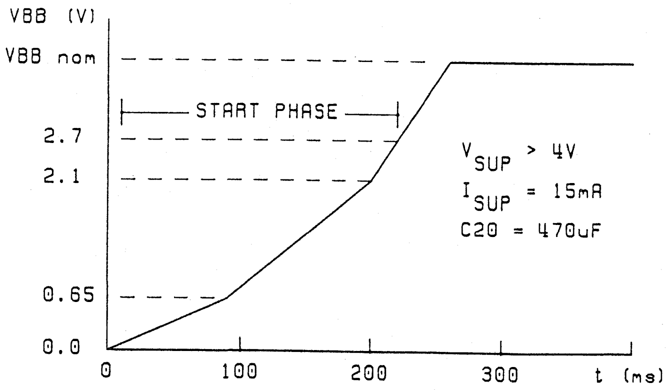


Fig.4 Behaviour of the VBB voltage during start up.

The start circuit will be activated again, after termination of the start up period, when VBB drops below a threshold voltage of 2.1 V. The charge currents are switched on and the MUTE (toggle) circuitry of the TEA1085 is set into the standby mode; see also par.4.4.

The levels at which the start circuit is switched on (VBB < 2.1 V) or switched off (VBB > 2.7 V) are realised with multiple diodes. Both levels are independent of VBB settings, but vary with the junction temperature with about -6mV/K respectively -8mV/K.

The line voltage can show LF relaxations at low line currents between 6mA and 8mA when the TEA1085/A is applied in combination with transmission ICs which are not equipped with a low voltage function, such as the TEA1060, TEA1061 and the TEA1068.
Consult, if necessary, the application hints in appendix E.

4.2. Supply

Fig.5 shows the internal configuration of the supply part of the TEA1085/A. Input current I_{SUP} supplies the internal circuitry and the external devices, if any, connected between VBB and VSS.
The internal current consumption I_{SUP0} is 4.2 mA typical at I_{SUP} = 15 mA, and without loudspeaker signal. I_{SUP0} consists of the currents $I_{BIAS} \approx 0.4$ mA and $I_{BB0} \approx 3.8$ mA, which represent the standby current of the internal circuits connected to SUP respectively to VBB.

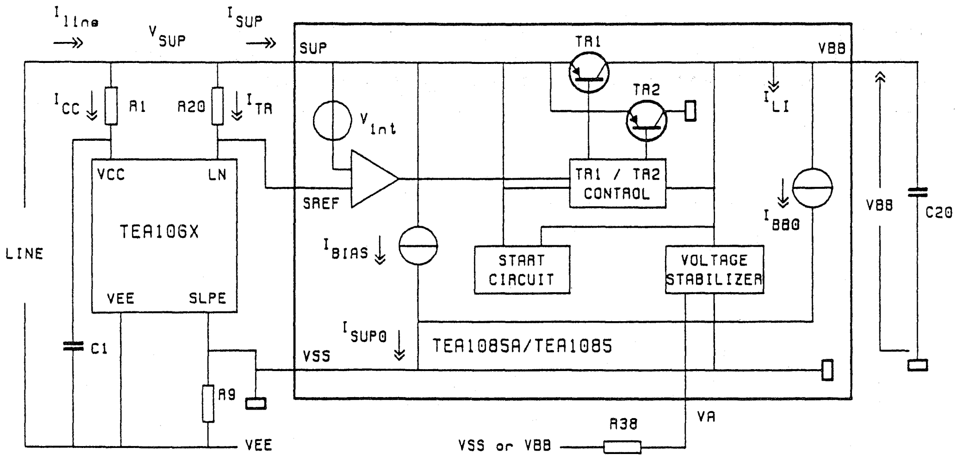


Fig.5 Internal configuration of the supply part.

I_{SUPO} is rather independent of V_{SUP}, but increases with I_{SUP} ($I_{SUPO} \approx 4.7$ mA at $I_{SUP} = 50$ mA) due to the increased bias current of TR1 and TR2.
I_{BBO} depends on the VBB voltage as described in par.4.2.3.

Current I_{LI} is, in principal, available to power the loudspeaker and the external devices connected to VBB.
VBB is stabilized at 3.6 V (typical) by means of a shunt regulator; it drains all surplus of current to VSS which is not consumed by the internal circuitry, loudspeaker and external devices.

4.2.1. Efficiency of the supply circuitry

The supply block incorporates a current switch, consisting of TR1 and TR2, controlled by the instantaneous voltage difference between the speech signal at SUP and the supply voltage VBB.

Transistor TR1 conducts I_{SUP} to the VBB supply rail. TR2 diverts momentarily I_{SUP} to VSS whenever the instantaneous voltage at SUP drops below VBB. This switch minimizes the distortion of the line signal, but reduces also the current efficiency of the supply part. This efficiency, defined as the ratio between output and input current (I_{LI}/I_{SUP}), depends on the DC voltage between SUP and VBB and the magnitude of the signal on SUP or the line. Maximum efficiency is reached for:

$$v_{SUP\text{-}peak} = v_{line\text{-}peak} \leq V_{SUP\text{-}VBB} - 250 \text{ mV},$$

Transistor TR2 starts conducting (reducing the efficiency), when the voltage difference between SUP and VBB becomes less than 250mV.

Fig.6 shows the supply current I_{LI}, available to power the loudspeaker, versus line signal level for VBB voltages between 3V and 6V. Input current I_{SUP} is varied between 15mA and 50mA, while voltage space $V_{SUP-VBB}$ is set between 0.5V and 1.5V.

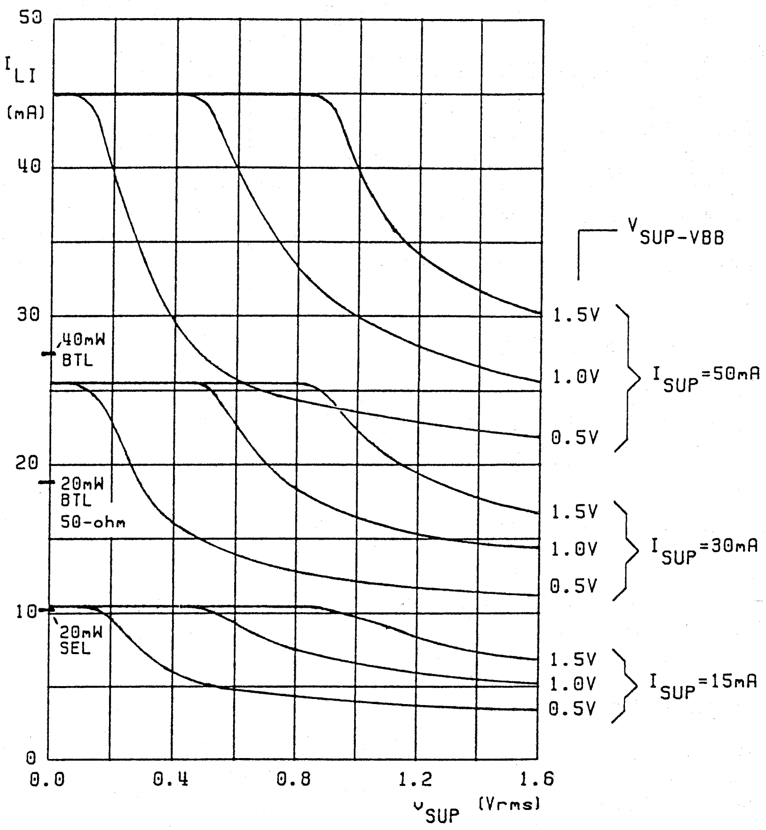


Fig. 6 Supply current I_{LI} to power the loudspeaker versus line signal level.

This figure also indicates the required supply currents (I_{LI}) to generate an output power of 20 mW in SEL mode (10.5mA), and 20 mW (18.6mA) respectively 40 mW (27.6mA) in BTL mode into a 50 Ω loudspeaker.

The major part of the line current flows via the TEA1085/A to terminal SLPE of the TEA106X. Bias current I_{TR} is kept constant, thus I_{SUP} is modulated by the microphone signal resulting in a lower supply current.
I_{LI} is reduced, during sending, with the average value of the modulation current. Depending on the magnitude of the line signal and set and line impedance, the supply current is:

$$I_{LI\text{-}send} = I_{LI} - \frac{v_{line} \cdot \sqrt{2}}{\pi . Z_{set}//Z_{line}}$$

This relation is valid only for line signals with a constant amplitude. The influence on the supply current for speech signal can be neglected, due to the non-continuous wave forms.

4.2.2. Maximum line signal, distortion of the line signal

The maximum voltage swing on the line is determined by the transmission IC (with or without dynamic limiter) or by the LI circuit. The TEA1085/A has a minimum instantaneous working voltage between SUP and VSS of 1.4V. This level is independent of VBB settings, but vary with about -4mV/K.
The maximum signal level which can be generated between SUP and VSS depends on the DC voltage between SUP and VSS; it is expressed by:

$$v_{SUP\text{-}peak} = V_{SUP\text{-}VSS} - 1.4 \quad \text{(V)}$$

Signal distortion can be caused by signal clipping and/or cross over distortion from the current switches TR1 and TR2.
The contribution of the TEA1085/A to the harmonic distortion of the line signal can be neglected for audio frequencies, as long as the line signal is not clipped.

4.2.3. Stabilized supply voltage

A stabilized supply voltage of 3.6 V is available at VBB (pin 24). VBB can be increased by means of external resistance R38 connected between VA and VSS or decreased when R38 is connected between VA and VBB. The value of R38 to adjust VBB is for:

$$\text{VBB} > 3.6\text{ V:} \quad R38_{VA\text{-}VSS} = \frac{38}{VBB - 3.6} \quad (\text{k}\Omega)$$

$$\text{and for VBB} < 3.6\text{:} \quad R38_{VA\text{-}VBB} = \frac{(VBB - 1.25)\ .\ 30}{3.6 - VBB} \quad (\text{k}\Omega).$$

Fig.7 shows the adjusted VBB voltage as a function R38.
The upper part of fig.7 concerns the increase of VBB with R38 connected between VA and VSS, while the lower part of fig.7 represents the decrease of VBB by R38 connected between VA and VBB.

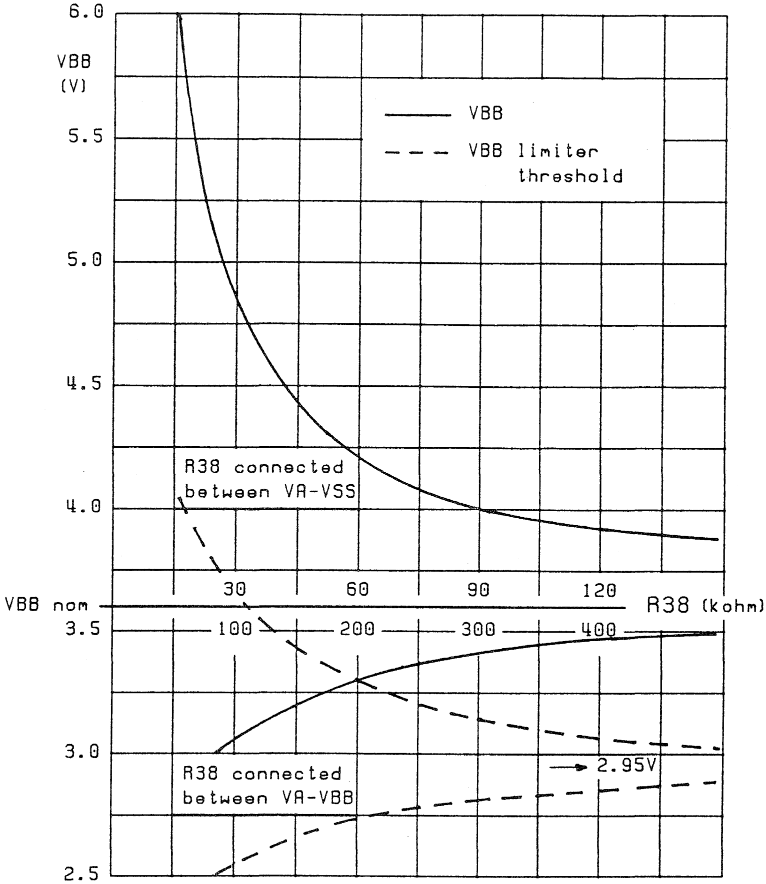


Fig.7 Adjusted VBB voltage as a function of R38.

A modification of the VBB voltage, with respect to 3.6 V typical, has the following restrictions and consequences:

- The minimum and maximum level, between which VBB may be adjusted, is limited to 3.0V respectively 6.0V with respect to the specified I_{SUP} and ambient temperature range.
- An increased VBB level means a reduced voltage difference between SUP and VBB which can have influences on the supply efficiency.
- Supply current I_{BBO} varies with VBB according 1mA/V.
- R38 connected between VA and VBB increases the current consumption from VBB, during PD, by (VBB − 1.25)/R38.
- The VBB threshold of the dynamic limiter (2.95V nominal) depends on the VBB setting; fig.7 shows the modified voltage threshold in correlation with the value of R38.

- The bias resistor R32 of the Larsen Level Limiter has to be modified as described in par.4.8.3.

Remark: The VBB voltage can also be adjusted by means of a current from or into terminal VA, instead of using resistor R38. To modify the VBB voltage with +1V or -1V, with respect to 3.6V nominal, the current taken from VA or delivered into VA has to be 31μA typical.

VBB as supply point for external devices

VBB may be used to supply external devices. Fig.6 shows in this case the maximum supply current (I_{LI}), which can be delivered by VBB, as a function of the line signal at different DC conditions.

Supply current I_{LI} has to be sufficient to power the loudspeaker amplifier as well as the external device(s). The LI performance will suffer if too much current is consumed by the external devices.
The VBB voltage decreases in case of overload.

4.2.4. Current consumption during Power Down

The PD function reduces the internal current consumption from SUP and VBB during pulse dialling or flash instructions. It prevents a quick discharge of the buffer capacitor C1 of the transmission IC and buffer capacitor C20 of the TEA1085/A. The storage of both capacitors is required to bridge the supply gap during line breaks.

When PD is activated, the internal current consumption from pin SUP is reduced to 55μA, while the current taken from VBB is reduced to less than 400μA; both at nominal DC conditions as described in the data sheet (ref.1).

I_{SUP} depends on the voltage between SUP and VSS with < 20μA/V, but depends also on the voltage difference between SUP and VBB. The current consumption from SUP increases if the voltage difference $V_{SUP-VBB}$ > 3V, because of internal protection diodes. Fig.8 shows the current consumption as a function of the voltage on SUP for different values of VBB.

Buffer capacitor C20 (470μF) is discharged with approximately 1V/s when PD is activated.
The decrease of the VBB voltage has to be limited because of the minimum working voltage of the PD circuitry (2.1V) and the voltage threshold which triggers the start circuit.

If VBB drops below 2.1V, the start circuit will be activated, as described in par.4.1.
The dependency of I_{BBO} with VBB is < 50 μA/V.

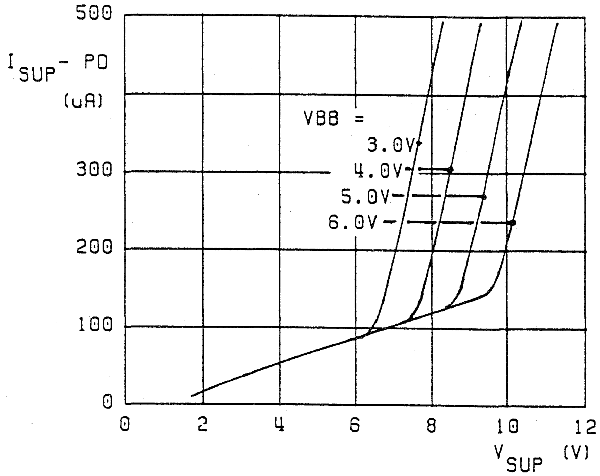


Fig.8 Current consumption from SUP, during PD, as a function of V_{SUP} at VBB = 3V, 4V, 5V and 6V.

4.2.5. Internal protection, voltage and current limitations

All IC terminals are protected against ESD, with respect to VBB and VSS by means of internal diodes, with exception of the VBB, SUP and SREF terminals.

VBB and SUP are protected to VSS only, while SREF is protected to SUP and VSS.

The maximum voltage allowed at the logic inputs PD, MUTE, GSC1 and GSC2 is VBB+0.4 V. When the logic inputs have to be controlled by levels exceeding the allowed maximum, a resistor in series with the logic input is necessary to limit the input current to a maximum of 100μA.

The specified minimum is VSS−0.5V, however, a minimum level of VSS−0.2V is advised to guarantee a normal operation of the TEA1085/A in combination with the TEA106X.
Control levels of less than VSS−0.2V disturb the DC operation of the applied transmission IC due to parasitic currents flowing from SLPE to VEE.
In such cases, where control levels drop below the minimum level, interface components or circuits are required.

4.3. Mute function of the TEA1085A

The MUTE of the TEA1085A is provided with a logic input to operate with a microcontroller for instance. The loudspeaker amplifier is disabled (standby mode) whenever the MUTE input is low (input open or connected to VSS).
A HIGH level at the input enables the amplifier in the listening-in mode.

During standby, the gain reduction of the loudspeaker amplifier is more than 60dB with respect to the nominal gain of 35dB.

4.4. Mute function of the TEA1085

The MUTE of the TEA1085 has a toggle input. The 'toggle' MUTE is activated by means of a positive going edge of the input signal, which switches the TEA1085 over between standby and LI.

The loudspeaker amplifier is disabled during standby; the gain reduction from the LSI inputs to QLS1 or QLS2 output is the same as of the TEA1085A and measures more than 60dB with respect to the nominal gain of 35dB.

The MUTE can be controlled by a simple push button switch. The mute input has a pull up structure with a maximum current drain of 28 µA to VSS.

A timing diagram of the mute input signal is shown in fig.9. The voltage levels are referred to VSS.

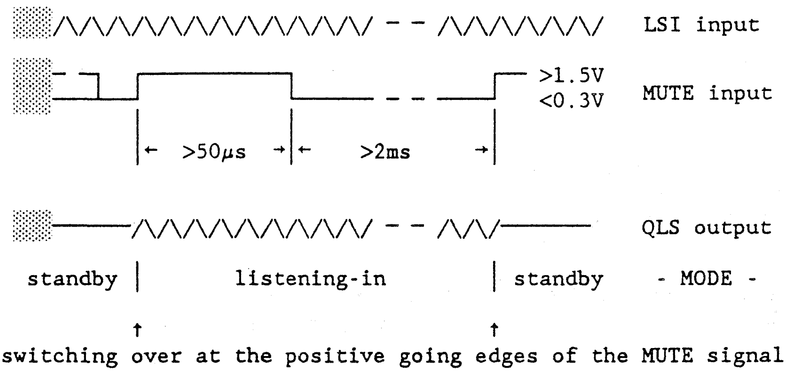


Fig.9 Timing diagram of the mute input signal (TEA1085).

Fig.10 and 11 show some alternative possibilities to control the MUTE by a simple push button switch.
The TEA1085 is set into the standby mode during start up, whenever the MUTE input is open or connected to VSS via a switch or resistor, as shown in fig.10.

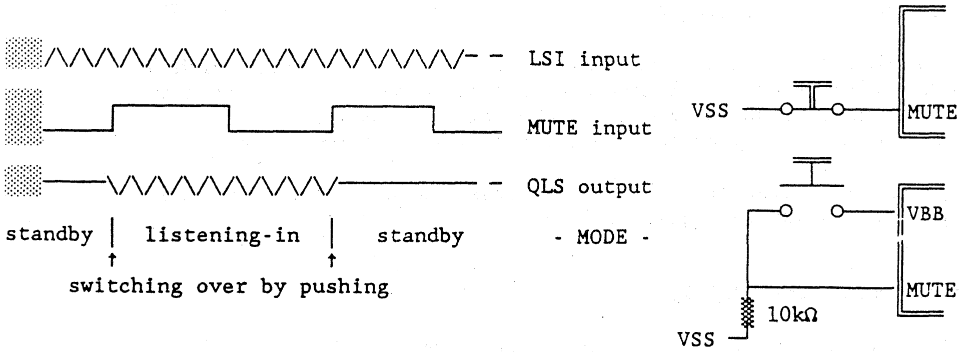


Fig.10 Control of the MUTE (TEA1085) by a switch with a 'break' or 'make' contact. Start up in standby mode.

Alternation from standby into LI condition, within the start up period, can be obtained when the MUTE pin is connected to VBB via a resistor of 10 kΩ according to fig.11.
The LI function is than operational directly after hook-off.

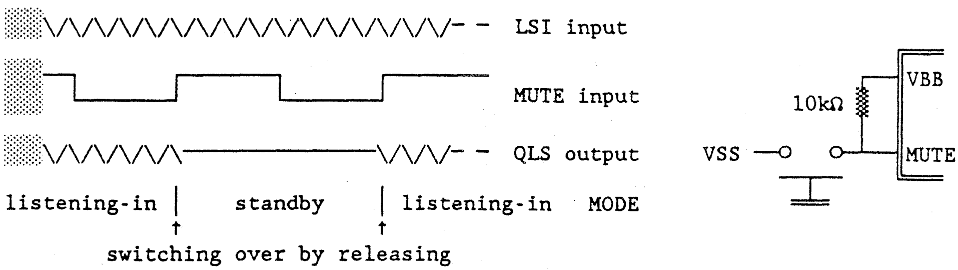


Fig.11 Control of the MUTE (TEA1085) by a switch with a 'make' contact. Start up in listening-in mode.

Debouncing can be realized by means of a capacitor of 10nF connected between MUTE of the TEA1085 and VSS.

The MUTE will be set into the standby mode when the VBB voltage drops below a threshold voltage of 2.1V.

4.5. Dynamic limiter

The dynamic limiter consists of a peak limiter, current limiter and VBB limiter. It limits the harmonic distortion of the loudspeaker signal in case of overdrive of the input signal, overload of the amplifier or at too low supply conditions.

Capacitor C28 (330nF), connected to pin 23 (DLC), determines the attack time (t_{att}) and release time (t_{rel}). Gain reduction is effected or returns to its nominal level by means of discharging respectively charging of C28. The attack and release times are specified in the data sheet, ref.1.
The dynamic range of the voltage at pin DLC is 1.3V +/- 250mV; the gain is reduced with 6dB at V_{DLC} = 1.3V at room temperature.

Pin DLC can be applied to control the gain externally, for example as an extra mute function. The release time at which the gain returns to its nominal level is relatively long in case C28 is fully discharged to VSS level. This release time depends on the value of C28, the charge current (1.1µA) and the amount of discharge of C28.

This paragraph describes the three limiter functions and their dynamic behaviour, while the static effect on the loudspeaker signal will be shown in par.4.6.3 "Output capabilities".

Peak limiter

The maximum level of the loudspeaker signal will be limited by the peak limiter (called also: voltage limiter) at sufficient supply currents.

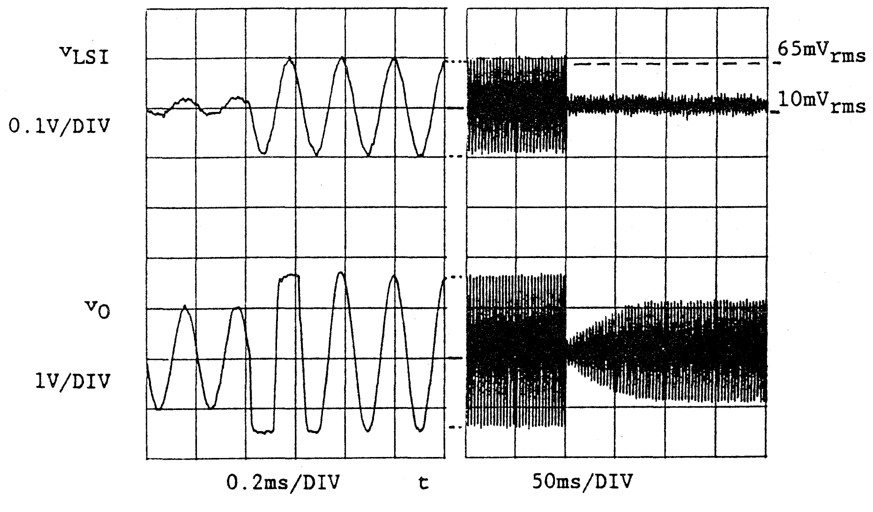


Fig.12 Attack and release of the peak limiter at I_{SUP} = 20mA and SEL with 50Ω.

The gain of the loudspeaker amplifier is reduced rapidly when the drive currents of the power amplifiers exceed internal current thresholds, due to saturation effects or at overload of the amplifier.
The maximum swing of the loudspeaker signal depends on the nominal VBB voltage in case of saturation. At overload, it is limited by the product of maximum output current and loudspeaker impedance. See also fig.17.

When the peak limiter is activated the loudspeaker signal is clipped during a relative short time, see fig.12, after which the gain and signal distortion are reduced. The gain returns to its nominal level in 75 ms, when the input level is reduced below the overdrive level.

Current limiter

The maximum loudspeaker signal level is limited by the current limiter at low supply currents. It limits the signal level by controlling the gain of the amplifiers whereby the current consumption of the loudspeaker amplifier corresponds with the available supply current.
This limiter is activated when the surplus of supply current, normally drained to VSS by the internal voltage stabilizer, becomes zero.

The attack time of this "slow" limiter is relatively long (typical 500 ms) to avoid system instabilities. This slow action can result in VBB drops at low line currents and high level speech signals, activating the VBB limiter. The behaviour of the current limiter is shown in fig.13. The VBB voltage (not plotted in fig.13) decreases when the limiter is activated and returns to the nominal voltage after 330ms. The amplifier reaches its nominal gain factor when the input level is reduced.

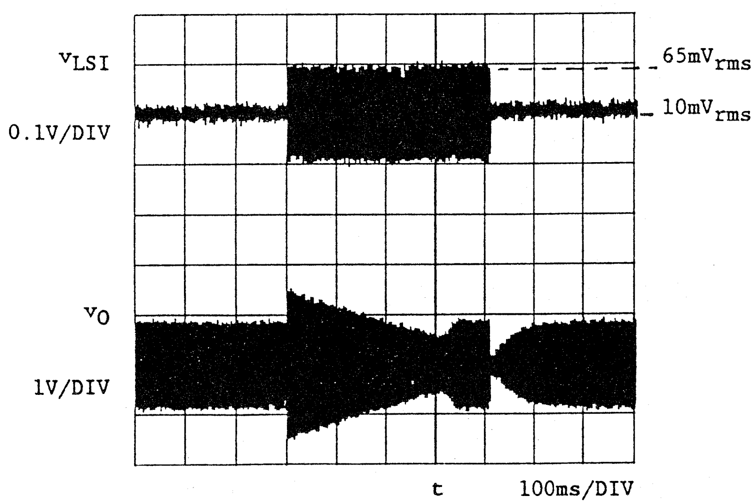


Fig.13 Behaviour of the current limiter at I_{SUP} = 12mA and SEL with 50Ω.

VBB limiter

The VBB limiter is an addition to the current limiter to prevent too low VBB levels. At low supply (or line) currents and signals which amplitudes vary constantly, such as speech, the momentarily current consumption of the power amplifiers can exceed the supply current. The VBB voltage then drops due to the discharge of the VBB capacitor C20.

The consumption of the power amplifiers has to be diminished instantaneously preventing too low VBB levels whereby the amplifiers work not correctly and signal distortion occurs.

Fig.14 shows the behaviour of the VBB limiter. When v_{LSI} increases, the VBB drops due to the low supply current. The current limiter starts to reduce the loudspeaker level but is too slow to prevent VBB drops.
If VBB drops below a threshold of 2.95V, the VBB limiter reduces the gain and the amplifier current consumption instantaneously, so that VBB can restore.
When C20 is charged and VBB exceeds the 2.95V (at nominal VBB setting), timing capacitor C28 will be recharged with 1.1μA. The limiter remains in the gain reduced condition up to about 200ms.

The threshold level of the VBB limiter follows the adjusted VBB voltage as shown in fig.7 of par.4.3.

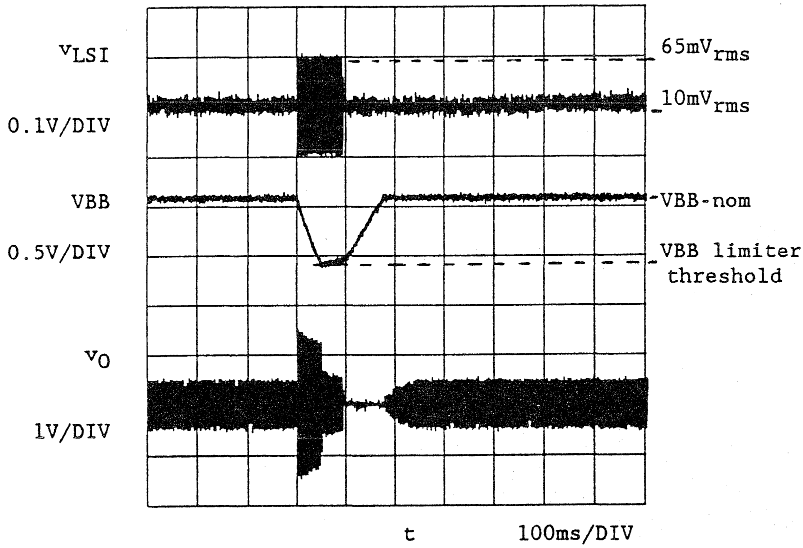


Fig.14 Reaction of the VBB limiter at low supply currents, I_{SUP} = 9mA, SEL with 50Ω.

4.6. Loudspeaker amplifier

The loudspeaker amplifier has symmetrical inputs LSI1 and LSI2 (pin 5 and 6). One of them has to be connected with the receiver output -QR- via a potentiometer, while the other input has to be connected with VEE via coupling capacitor C30. VEE is the reference of output QR as shown in fig.3.

A loudspeaker can be connected, via coupling capacitor C27, as a SEL between VSS and QLS1 (pin 21) or QLS2 (pin 22), or as a BTL between QLS1 and QLS2.

4.6.1. Input impedance, maximum input signal

The input impedance, symmetrically measured between LSI1 and LSI2, is typically 19kΩ (2 * 9.5kΩ) with a tolerance of maximum ± 20%. Both inputs are biased to an internal voltage reference of about 1.3V.

The signal transfer of the input stage is linear (THD ≤ 2%) for input levels up to $200mV_{rms}$; higher levels result in signal distortion caused by this stage.

4.6.2. Volume control, gain control function

The gain of the loudspeaker amplifier between LSI inputs and QLS output(s) is typically 35dB for SEL and 41dB in case of BTL. The gain can be reduced by means of the gain select function via the GSC1 and GSC2 inputs (pin 8 and pin 7) in steps of 6dB.

Volume control of the loudspeaker signal can be obtained by attenuation of the input signal via a potentiometer or/and by means of the gain control function. Fig.3 shows the input attenuator consisting of R21 and R24.

The value of R21 and R24 has to be in accordance with the total receive gain of the application of the transmission IC depending on the sensitivity of the ear capsule.
Both values have to be chosen in such a way that for a mean speech level on the line, the maximum input signal at LSI measures $18mV_{rms}$. This corresponds with a loudspeaker signal of $1V_{rms}$ at SEL and $2V_{rms}$ in case of BTL if the gain select function is not activated.

The gain control function can be used to reduce the amplifier gain, in 3 steps of 6dB each. This function will be activated if a voltage level of more than 1.5V is applied at the GSC input(s), as shown in table 2.

The inputs can be controlled by a microcontroller but can also be wired to VBB for a permanent reduced gain. Both inputs have a pull down structure with a maximum input current of 8μA referred to VSS. When gain control is not required, the inputs may be left open or connected to VSS.

GSC1 input level	GSC2 input level	Gain at SEL (dB)	Gain at BTL (dB)	Gain - reduction (dB)
L	L	35	41	0
L	H	29	35	6
H	L	23	29	12
H	H	17	23	18

Table 2.

Gain select of the loud-speaker amplifier via the GSC1/GSC2 inputs.

The gain values, as listed in table 2, can only be obtained when the dynamic limiter is not functioning. Refer to fig.18 and fig.19, if necessary.

4.6.3. Output capabilities

The output stages of the two power amplifiers are optimised for use with a 50Ω loudspeaker, for example Philips type AD2071/Z50. Loudspeakers with another impedance can be applied too, as shown in this paragraph.

The output impedance of the amplifiers is less than 1Ω, measured with 50Ω load, for SEL as well as BTL.

Fig.15 shows the voltage gain from LSI inputs to QLS output(s) as a function of frequency. The output load is 50Ω load connected as a SEL or BTL, while the output levels are 500mV$_{rms}$ respectively 1V$_{rms}$ at 1kHz.

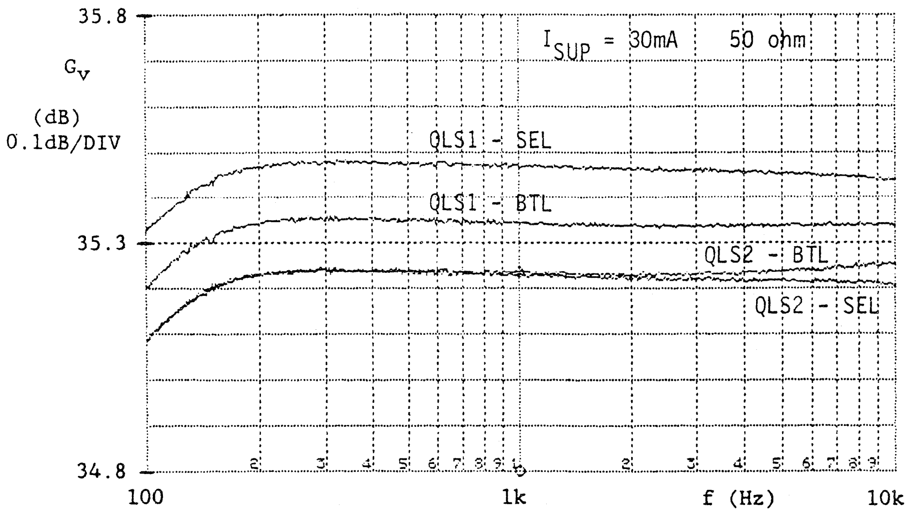


Fig.15 Frequency characteristics of the loudspeaker amplifier. Output signal of QLS1 is in anti-phase with QLS2.

Fig.16 shows $P_{O\text{-}max}$ and $v_{O\text{-}max}$ both as a function of I_{SUP}. The output load is 50Ω, THD ≤ 2%, while VBB is varied between 3.0V and 4.5V.

Referred to fig.16, the output power is limited by the current limiter in the region where P_O increases with I_{SUP}; the supply current is here not sufficient to deliver the desired output swing.
The output swing is limited by the peak limiter in the region where P_O is constant with I_{SUP}.

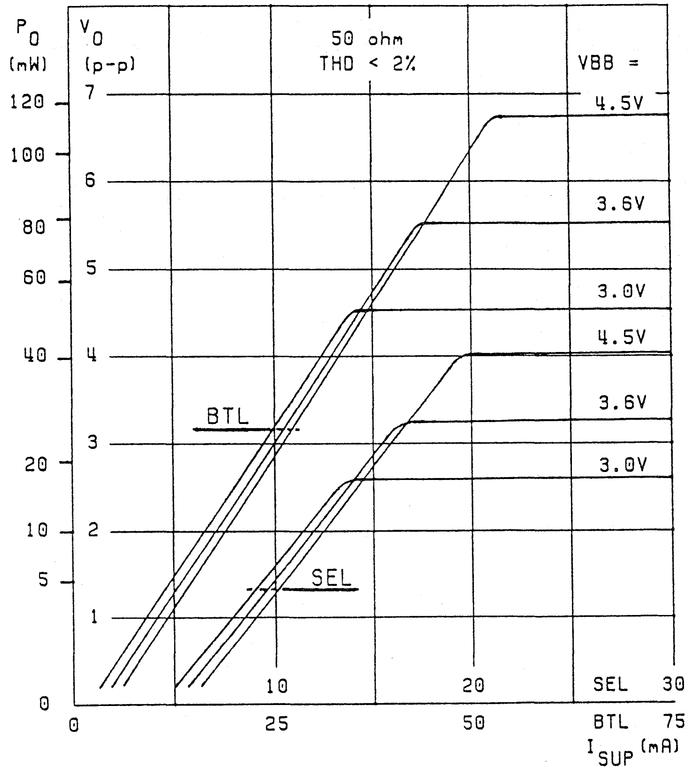


Fig.16 Maximum output power and output voltage versus I_{SUP} at VBB = 3V, 3.6V and 4.5V.

Fig.17 shows $P_{O\text{-}max}$, and the corresponding required supply current I_{SUP}, both as a function of loudspeaker impedance R_{LS} at nominal VBB level.
The output power is limited by the peak limiter with R_{LS} > 20Ω connected as a SEL and with R_{LS} > 40Ω in case of BTL.
P_O is restricted by the maximum output current of the power amplifier, which is approximately 70mA-peak, with lower loudspeaker impedances than 20Ω respectively 40Ω. Both curves are measured at an input level of 30 mV_{rms}.

Appendix A can be consulted to calculate the supply current depending on demanded output voltage v_O and loudspeaker impedance R_{LS}. This appendix also describes the usefulness of coupling capacitor C27, in BTL applications, with respect to maximum output power at low line currents.

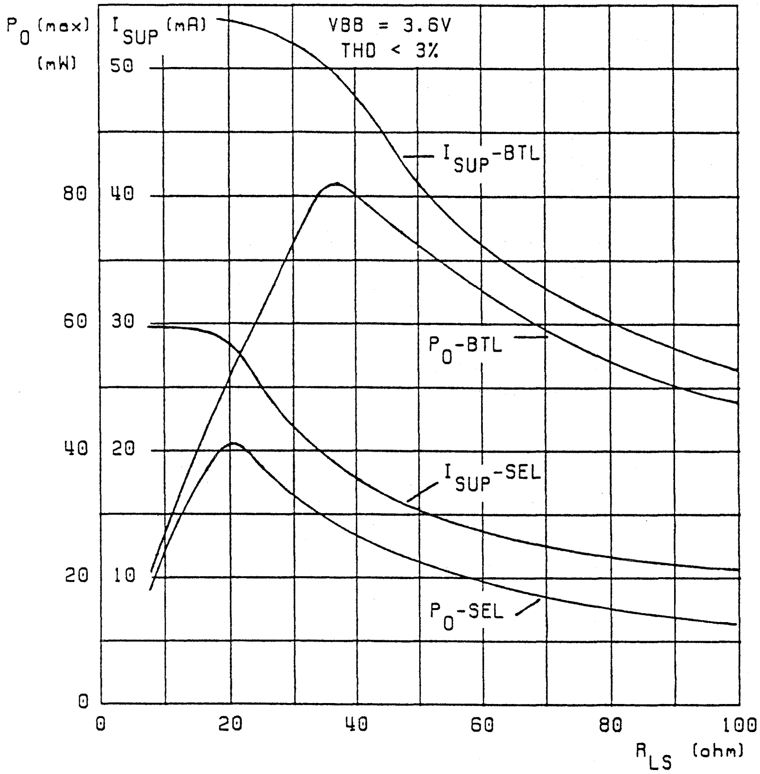


Fig.17 Maximum output power and required I_{SUP} versus loudspeaker impedance at VBB = 3.6 V

Fig.18 and fig.19 show the effect of the dynamic limiter on the output voltage v_O when the input signal v_{LSI} is enlarged with 15dB with respect to $18mV_{rms}$.
Both graphs show the effect of the gain control function on the output voltage, reducing the gain with respectively 0dB, 6dB, 12dB and 18dB, according to table 2.

Remark: Due to the LLL, the gain of the loudspeaker amplifier can be reduced and remains reduced during tests of the LI facility with microphone signals with a constant amplitude. The influence of the LLL can be cancelled by means of a short circuited resistor R32.

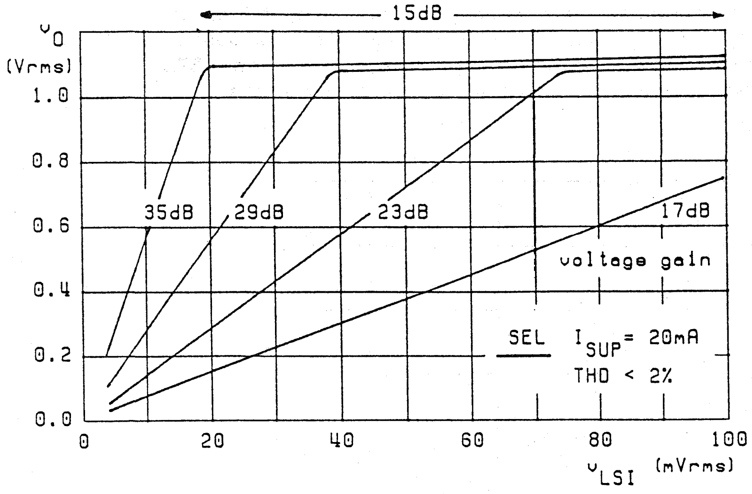


Fig.18 Static behaviour of the peak limiter at overdrive on the LSI inputs at SEL mode.

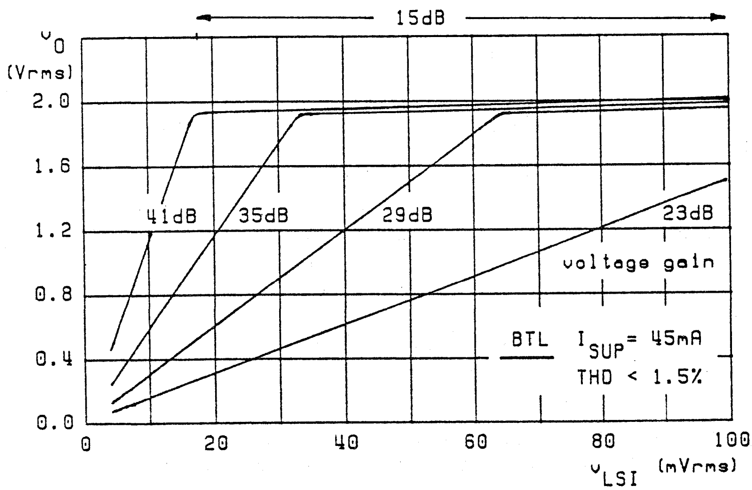


Fig.19 Static behaviour of the peak limiter at overdrive on the LSI inputs at BTL mode.

4.7. Power down function

The PD input (pin 19) is available for use in pulse dialling and register recall applications, where the line current is interrupted. During these line current breaks the TEA1085/A is without continuous power and has to be supplied by the charge of capacitor C20. The drop of the VBB voltage due to the discharge of C20 will be less in case the PD function is applied, which results in less ripple on VBB during pulse dialling. See par.5.5.

This function will be activated when a high level of > 1.5V is applied to the PD pin, as long as VBB > 2.1V. The PD input has a pull down structure with a maximum input current of 8μA referred to VSS.

The current consumption during PD is described in par.4.2.4.

When the power down facility is not required, the PD pin can be left open or connected to VSS.

4.8. Larsen Level Limiter

The Larsen Level Limiter (LLL) reduces the howling or Larsen signals, by means of limiting the gain of the loudspeaker amplifier.

Howling is an audible oscillation resulting in annoying loudspeaker and line signals. It will occur when the total loop gain, from microphone to loudspeaker via the electrical domain and from loudspeaker back to the microphone via the acoustical domain, is more than one.

The TEA1085/A has to perform a LI function in the telephone set. This means that at normal using conditions (speech signals), the LI function has to operate without howling effects and thus without interruptions from the LLL.

The LLL has to intervene in the LI function when howling occurs only during movement of the handset (containing microphone and earcapsule) back to the base of the telephone set containing the loudspeaker. Consult ref.11.
The LLL system makes use of the signal from the microphone in the handset to detect howling. It must have a low sensitivity for speech signals.

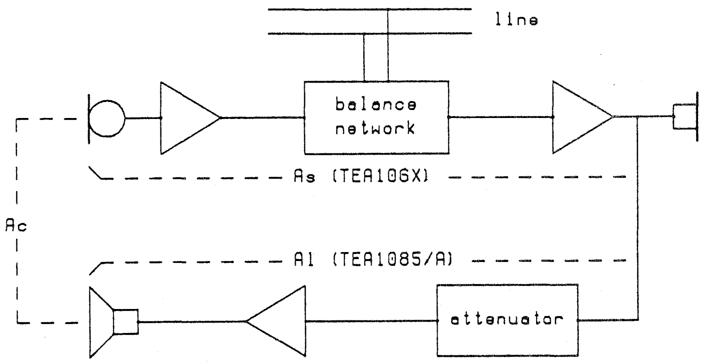


Fig.20 Signal path responsible for howling.

Fig.20 shows the total loop (via As, Al and Ac) which can cause howling. The total loop gain can be divided in:

- Transfer As: side tone transfer in the transmission part from microphone input to receiver output, which depends on its turn on sensitivity and frequency response of the microphone, transmit gain, receive gain, frequency, anti side tone network and line length.
- Transfer Al: LI transfer from receiver output, via the attenuator, to loudspeaker output. Al depends on the loudspeaker amplifier gain, sensitivity, frequency response and enclosure of the loudspeaker.
- Transfer Ac: Depends on the distance between loudspeaker and microphone.

Some transfer parameters could be modified to prevent howling; like side tone reduction by means of improved balance network(s) or a lower gain of the loudspeaker amplifier.

But the parameters, such as gain settings, are more or less fixed and have to be according to capsule sensitivities and the loudness ratings of the PTT organizations. Furthermore, the output power of the LI part also has to fulfill the requirements as described in appendix C.

On the other hand, an enlarged LI gain has a positive effect on the LI behaviour but can introduce audible instabilities in the form of howling during normal use of the telephone set.

The LLL system of the TEA1085/A offers a good alternative to reduce howling to an acceptable level.

The LLL makes use of the typical differences between speech and howling signals (via the microphone in the handset). Howling signals have a constant amplitude. The frequency is in general more than 1.5kHz, depending on the distance between loudspeaker and microphone.
Basic speech, however, is a more noiselike signal with varying amplitude and frequencies up to ≃ 4kHz, which has its maximum acoustic energy at f < 1kHz.

Howling reduction is specified by the CNET, see appendix D and ref.11.

4.8.1. Dynamic behaviour

The LLL contains the following circuitry as shown in fig.21:

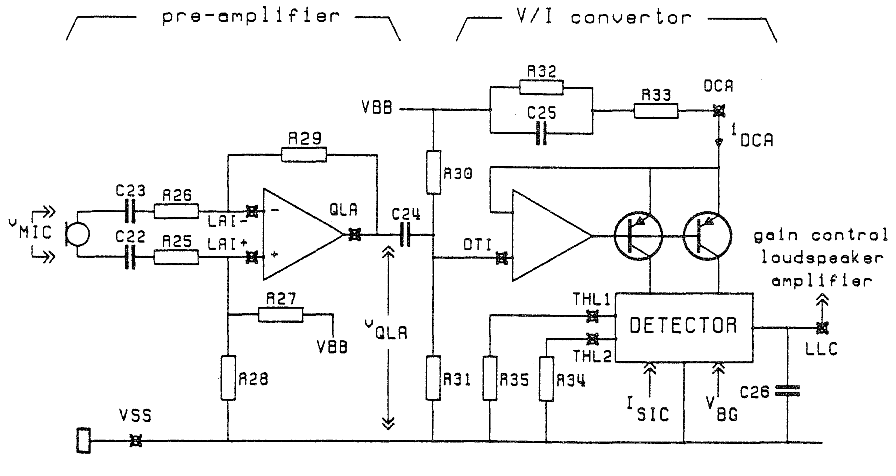


Fig.21 Configuration of the LLL with pre-amplifier, V/I convertor and detector.

- Pre-amplifier with inputs LAI+ and LAI− and output QLA. It amplifies the microphone signal; the gain is determined by the external components R26 and R29, while the input is biased at VBB/2 by means of R27 and R28.

- V/I convertor with signal input DTI and detector current input DCA. It transforms the output voltage from the pre-amplifier into AC currents. The convertor is biased by resistors R30 and R31 which determine the DC current flowing into terminal DCA.

- Three cascaded first order high pass filters by which the AC current into pin DCA is filtered with respect to the microphone signal The high pass filters are formed by the elements C23.R26 (or C22.R25), C24.R30//R31 and C25.R33. The three separated cut off frequencies are 500Hz, 1kHz and 3kHz which can be modified as given in the next paragraph.

- A detector which processes the output currents from the V/I convertor. It compares the input voltage at DTI with two thresholds, specified as V_{DTI1} (attack level) and V_{DTI2} (residual level) as being voltage levels at the DTI input. The threshold levels are determined by R35 and R34, connected to THL1 respectively THL2.

- The detector uses an integrator to measure the time at which the microphone signal is present at the inputs. The attack, attack delay and release time of the detector are determined by C26 and R36. The voltage across C26 is used to control the gain of the loudspeaker amplifier.

The LLL has to separate speech from howling by using a HP filter and an integrator.

A howling signal is received by the microphone, amplified, filtered, compared with V_{DTI1} and checked for its time duration.
As soon as the signal at DTI exceeds V_{DTI1}, the detector starts discharging C26 as shown in fig.22 by V_{C26}. For normal speech the total discharge will be small, whereby the gain of the loudspeaker amplifier remains constant.

The second threshold V_{DTI2} is disabled during the LI mode.

The system switches over from LI to Larsen mode for signal levels of ≥ 100 mV_{rms} remaining at the DTI input for more than the attack delay time t_{ad}.
The gain of the loudspeaker amplifier is than reduced within 20ms (t_{LAa}).
The time t_{ad} ensures little sensitivity of the system for own speech, because t_{ad} is more than the 'burst time' of speech signals.

The residual loudspeaker signal in this Larsen mode will be determined by the second threshold V_{DTI2}, which is enabled in this mode. The circuit acts as a dynamic limiter with peak detector which controls the gain in correlation with this threshold.
When howling stops, by moving the handset away from the base, the gain of the loudspeaker amplifier returns to the nominal value within release time t_{LAr}.

The settings of the thresholds V_{DTI1} and V_{DTI2} by means of R35 respectively R34 are given par.4.8.4.

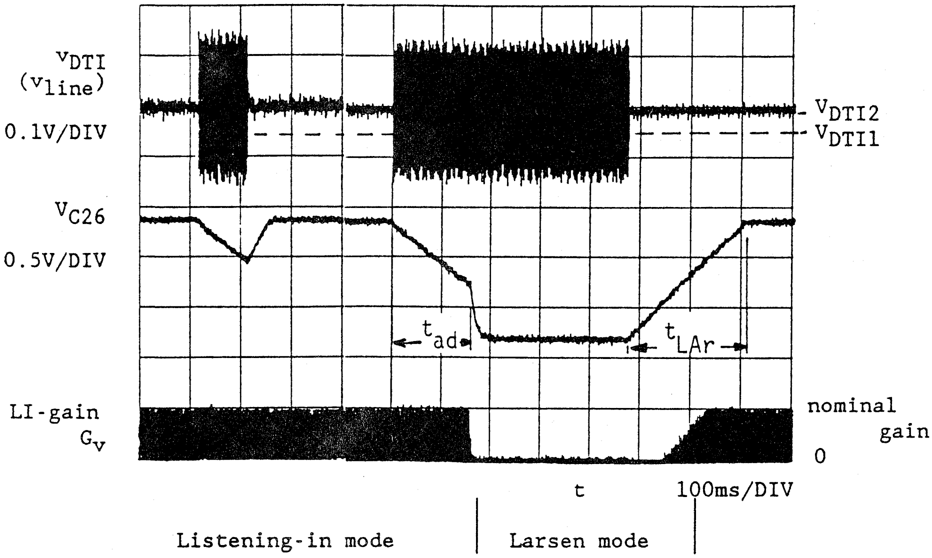


Fig.22 Dynamic behaviour of the LLL.

4.8.2. Pre-amplifier and HP filter

Pre-amplifier gain

The gain (A_{pre}) between the microphone terminals and amplifier output QLA has to be adjusted to the same value as the microphone gain of the applied transmission IC. The attenuation of the three cascaded high pass filters must not be taken into account. The signal level on QLA equals the level of the line signal.

With a fixed value of R25 = R26 = 10kΩ, the gain can be adjusted according:

$$A_{pre} = \frac{v_{QLA}}{v_{mic}} = \frac{R29}{R26}$$

With R29 = 1.65MΩ, the gain is set to 44dB, which corresponds to the microphone gain of the applied transmission circuit; see chap.5.
Input LAI+ is biased at VBB/2 by means of R27 and R28.
The recommended value of R27//R28 equals R29.

High pass filter

Besides of the integrator in the detector, a HP filter is required to reduce the sensitivity of the LLL system for own speech.
The three separated cut off frequencies are experimentally determined at f1 = 500Hz, f2 = 1kHz and f3 = 3kHz by means of the components values:
C24 = 4.7nF, R30 = 100kΩ, R31 = 220kΩ, C25 = 330nF, R33 = 510Ω and C23 = C22 = 4.7nF.

The cut off frequencies may be adjusted using the equations:

$$f1 = 1/(2.\pi.C24.R30//R31),$$
$$f2 = 1/(2.\pi.C25.R33) \text{ and}$$
$$f3 = 1/(2.\pi.C23.R26) = 1/(2.\pi.C22.R25)$$

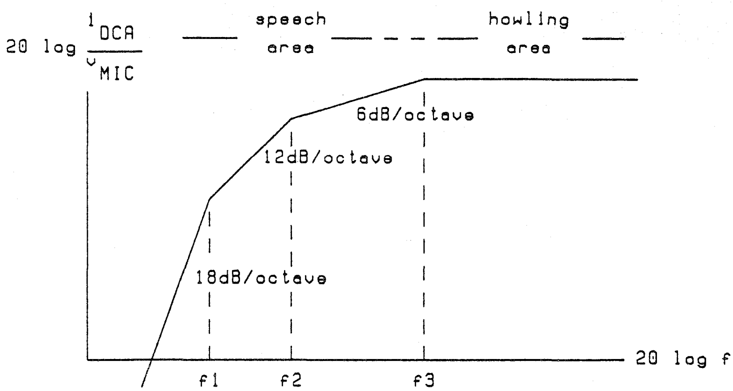


Fig.23 Third order HP filter.

4.8.3. Basic settings

The timing of the detector of the LLL (t_{ad} and t_{LAr}) and the threshold levels (V_{DTI1} and V_{DTI2}) are based on the value of C26 and R36.
R36 determines an internal reference current. Advised is to keep C26 = 1μF and R36 = 120kΩ. Refer to appendix D if necessary.

The detector current into pin DCA consist of a DC current of I_{DCA} = 11.2μA, at VBB = 3.6V typical, and an AC component i_{DCA} which depends on v_{DTI}.
Advised is to keep I_{DCA} constant when VBB has to be adjusted (par.4.2.3) because I_{DCA} determines the threshold levels as given in appendix D.
In order to keep I_{DCA} = 11.2μA, R32 (100kΩ at VBB = 3.6V) has to be modified depending on the required VBB level, according:

$$R32 = \frac{R30}{R30 + R31} . VBB . 90.10^3 - R33$$

4.8.4. Threshold settings

To change the threshold voltage V_{DTI1} and V_{DTI2} by means of R35 and R34 the requirements have to be known with respect to attack delay time (t_{ad}) and residual voltage level (V_{DTI2}).
As explained in appendix D, t_{ad} has to be chosen between a maximum delay time for small howling signals, and a minimum delay time for large howling signals.
The maximum delay time, within which howling has to be reduced, is specified by some PTT's. The minimum delay time determines the sensitivity of the LLL system for speech signals.

The first threshold V_{DTI1} and value of R35 can be calculated by:

$$V_{DTI1} = v_{LA1} . \sqrt{2} \ . \ \sin \left[0.5 - \left(\frac{R36 \ . \ C26}{6 \ . \ tad} + \frac{1}{3} \right) \right] . \ \pi$$

$$R35 = \frac{R33 \ . \ 2.52}{V_{DTI1} + (R33 \ . \ 11.2.10^{-6})} \qquad \text{if } R33 = 500\Omega: \quad R35 = \frac{1260}{V_{DTI1} + 5.6.10^{-3}}$$

Fig.24 shows the value of R35 as a function of t_{ad}. The conditions of this graph are: Larsen signal at the line v_{LA1} = 100mV_{rms}, R36.C26 = 120ms with R36 = 120kΩ and C26 = 1μF, R33 = 500Ω.

For a required t_{ad} = 160ms at v_{LA1} = 100mV_{rms}, the value of R35 = 52kΩ.
The calculated level V_{DTI1} = 18.5mV.

After the attack delay time the gain of the loudspeaker amplifier is reduced to a level at which the residual Larsen signal will be determined by a second threshold. This threshold V_{DTI2} = 6.9mV with R34 = 100kΩ. The relation of R34 with V_{DTI2} is given by:

$$R34 = \frac{R33 \ . \ 2.52}{V_{DTI2} + (R33 \ . \ 11.2.10^{-6})} \qquad \text{if } R33 = 500\Omega: \quad R34 = \frac{1260}{V_{DTI2} + 5.6.10^{-3}}$$

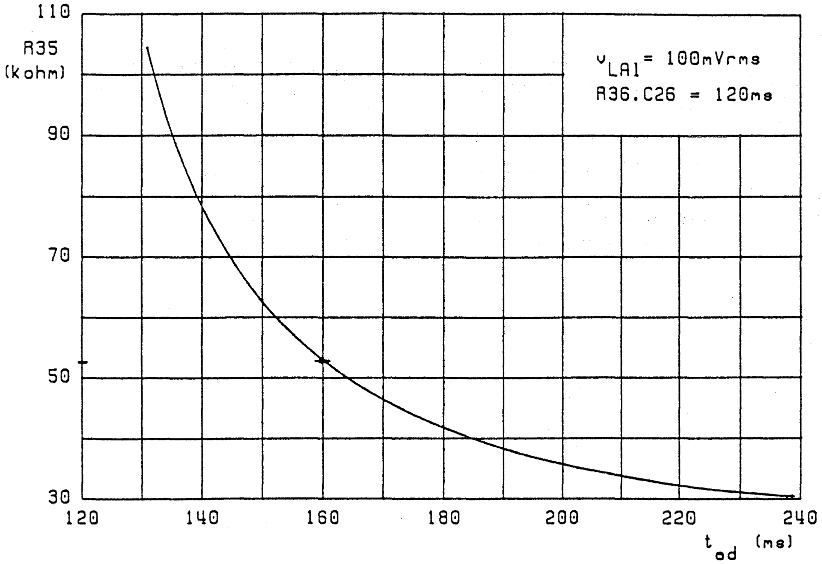


Fig.24 Value of R35 versus attack delay time t_{ad}.

5. APPLICATION EXAMPLE OF THE TEA1085/A

5.1. Description of the application

Fig.25 shows an application example of the TEA1085/A in combination with the TEA1064A speech/transmission IC and the PCD3310 as pulse/dtmf dialler. The ringer is not included. A software controlled ringer can be easily implemented when using the PCD33XXA-family of telephony μC instead of the PCD3310 (ref.9).

The TEA1064A, including its external components, provides the transmission interface functions between telephone line and microphone, ear capsule, dialler and between line and loudspeaker via the TEA1085/A.
It incorporates a line voltage stabilizer, microphone/DTMF amplifier, dynamic limiter, receiving amplifier and has facilities for line-loss compensation, mute control and side tone suppression. A smoothed voltage VCC1 supplies the internal circuitry which can also be used to supply an electret microphone.

Fig.25 shows a dynamic microphone and dynamic ear capsule. Consult ref.4 for application details or when an electret microphone or asymmetrically connected earpiece has to be applied.
The set impedance is determined by a 600Ω termination. Complex set impedance will be discussed in par.5.3.2.

The received line signal is wired from the QR output to the LSI inputs of the TEA1085/A via R24 and potentiometer R21 which is available for manual control of the loudspeaker signal. A 50Ω loudspeaker is connected as a SEL.

The PCD3310 is a dual-standard dialling IC for either DTMF or pulse dialling. It is a member of the PCD3310-family specified in the data sheets. Any standard matrix keypad can be connected.
The dialler IC is supplied by the VBB voltage of the TEA1085/A via a backup diode and smoothing capacitor. Terminal SLPE of the TEA1064A is the ground reference of the DTMF and logic signals between dialler, transmission and LI circuit.

The interface between dialler (DP/FLO-pin) and line interrupter (BSN254) is realized with discrete components.

The MUTE of the TEA1085, which has a toggle function, could be controlled by a push button switch for instance. The TEA1085A has a logic MUTE which could be controlled by a μC.
The gain select inputs GSC1 and GSC2 can be used to reduce the loudspeaker amplifier gain with 6dB steps. They can be wired to the VBB voltage for a permanent gain reduction or controlled by a μC via the keypad.

The PCD3310 has to be replaced by a μC, for instance a member of the PCD33XXA-family like the PCD3344A, PCD3351A etc., when the MUTE and GSC functions have to be controlled via keypad entries.

A general purpose demonstration board (PR45163) with an application of the TEA1085A and TEA1064A is available and described in ref.2

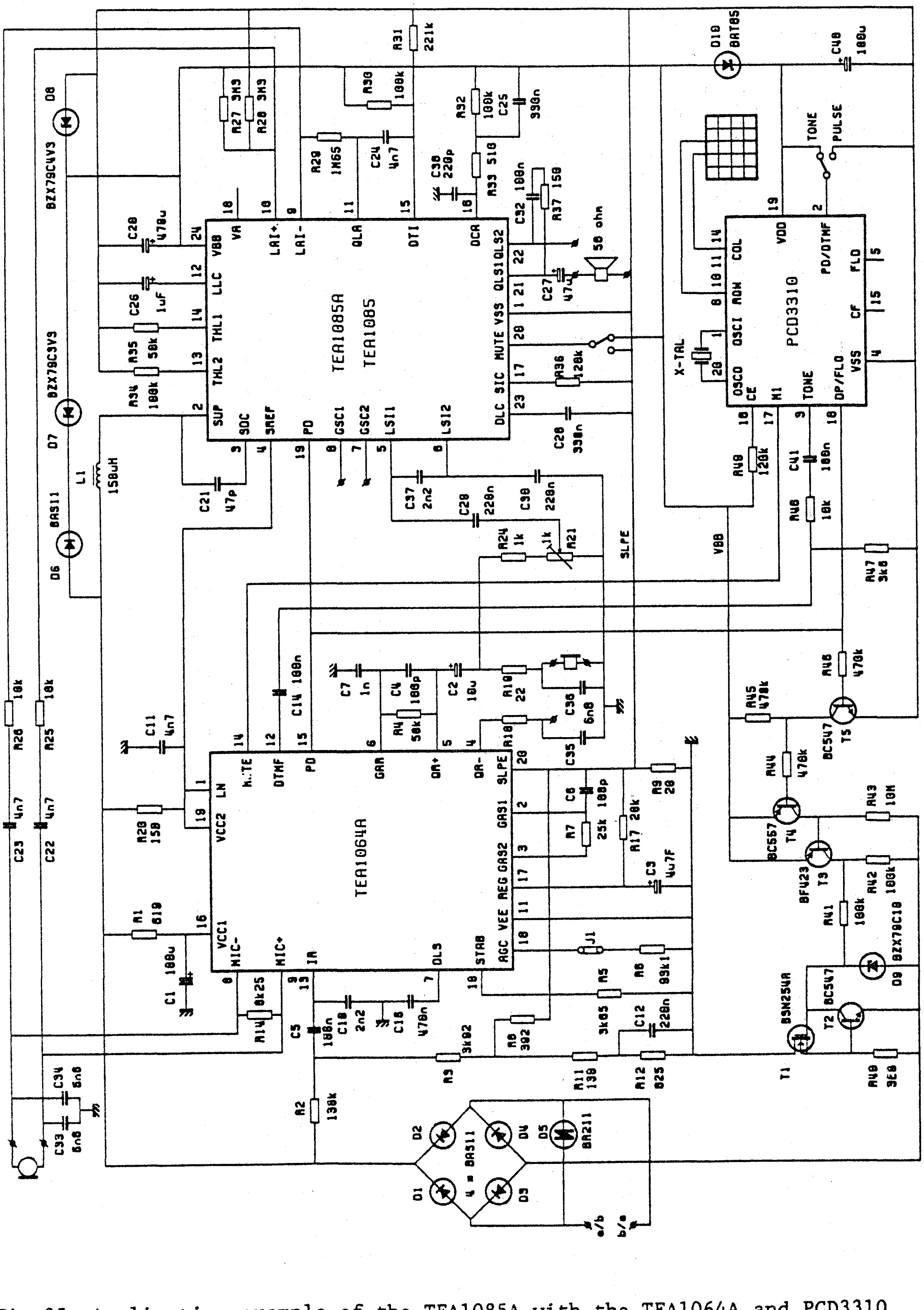


Fig.25 Application example of the TEA1085A with the TEA1064A and PCD3310.

5.2. DC behaviour

5.2.1. DC settings

The reference voltage of the TEA1064A ($V_{LN\text{-}SLPE}$) is increased from 3.3V typ. to 4.2V by means of R17 = 20kΩ; in this case $V_{SUP\text{-}VBB}$ is 900mV.
$V_{SUP\text{-}VBB}$ is too low (300mV) without R17 resulting in a low efficiency of the supply.
A mean signal level of 400mV_{rms} is permitted on the line without reduction of the supply efficiency; see par.4.2.1.
The VBB voltage is not adjusted; the nominal voltage is 3.6V, while the supply voltage for the dialler (VDD) measures 3.4V.

Fig.26 shows Vab, VBB, VDD, I_{SUP}, I_{TR} and the available supply current I_{LI}, to power the loudspeaker and peripherals, as a function of line current. It shows also the low voltage function of the TEA1064A for parallel operation.

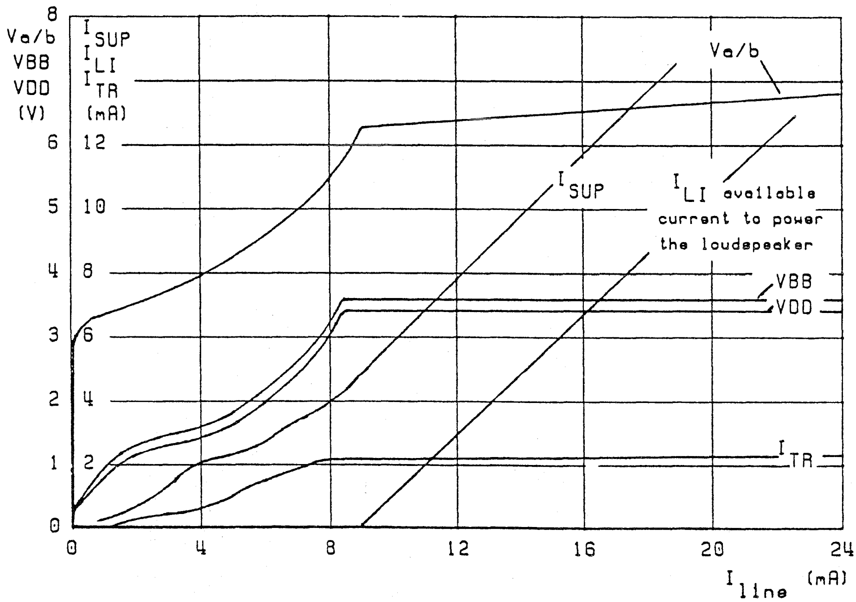


Fig.26 DC voltages and currents versus line current.

The supply current for LI is adapted to the available line current. At lower line currents the LI possibilities are limited giving a higher priority to the normal transmission functions (send and receive). LI is not possible at I_{line} < 10mA, as indicated in fig.26, while the normal transmission functions are operational down to about 4mA.

The voltage at the a/b-b/a terminals measures 6.7V at I_{line} = 20mA, which includes the voltage drop across the line interrupter and diode bridge.
I_{SUP} ≈ 16mA at I_{line} = 20mA, which is sufficient to generate a continuous

output power of 20mW across 50Ω connected as a SEL.

The output stage of the transmission IC is biased by means of DC current I_{TR} which measures about 2mA.
The amplitude of the AC voltage on LN equals the line voltage, while the AC current in the output stage depends on the load of terminal LN determined by the frequency, the value of C11 and the internal impedance of the transmission IC. The value of I_{TR} must be sufficient at maximum audio frequencies.
For details is referred to appendix B.

5.2.2. Start up

The start current of the LI circuit is limited to 4mA up to 6mA in order to speed up the initiation of the transmission functions.
Sending and receiving is possible within 100ms after going off hook.

Without extra measures the charge of the VBB and VDD smoothing capacitors C20 (470µF) and C40 (100µF) takes more than 300ms at I_{line} = 20mA (par.2.3).

Extra current is supplied into these capacitors via zener diode D7 (3.3V) to speed up the charge process. The excess of current, during start up, which normally flows to ground via the protection zener diode is applied to charge C20 and C40.

The normally applied protection diode of 8.2V (connected between LN and SLPE) is divided in D6, D7 (3.3V zener) and D8 (4.3V zener). Diode D6 blocks the current through D7 at negative swings of the line signal.

A VDD level of 2.5V (VDD-minimum) is reached within 160ms at 20mA line current. Lower start times can be realized by a reduction of the capacitor values.

5.2.3. Supply possibilities

VDD supplies the PCD3310 via a backup diode connected to VBB, as shown in application diagram fig.25. The VBB voltage is kept constant by the internal stabilizer and smoothed by C20. The VBB ripple caused by the loudspeaker signal can be ignored at Iline > 17mA for SEL configuration and at I_{line} > 35mA in case of BTL.
The VBB limiter reduces the ripple to ≤ 650mV (see par.5.4); the mean level depends on the line current.
Fig.28 and 29 show the minimum and mean levels of VBB and VDD as a function of I_{line}.

5.3. Transmission performances

5.3.1. Influence of the TEA1085/A

The influence of the TEA1085/A on the transmission behaviour of the set is given in appendix B. The major difference of this application with a single TEA1064A is that the modulation current to generate the line signal flows via the TEA1085/A and not through the transmit stage of the transmission IC. Set impedance, noise, transmit gain and frequency responses are not affected.

The combination of the two active circuits, TEA1085/A and TEA1064A, can cause HF instability on the line which results in a worsening of the transmission behaviour. This instability is cancelled by means of the combination of coil L1 in the SUP wire and C11 between LN and VEE. Coil L1 (150μH) has no influence on the transmission behaviour.
L1 may not be saturated at maximum line current.

Capacitor C11 appears in parallel with the set impedance; see appendix B 'Set impedance'.

5.3.2. Set impedance / complex set impedance

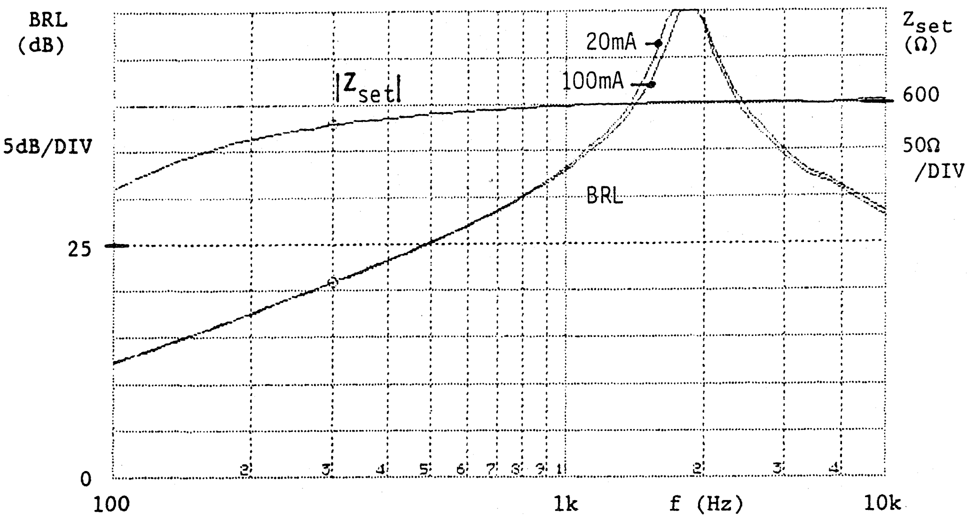


Fig.27 BRL and set impedance versus frequency at I_{line} = 20mA and 100mA.

Set impedance Z_{set} consists, for audio frequencies, of the coil function of the TEA1064A in parallel with R1 and the impedance of C11, as proved in appendix B. The 'real' Z_{set} of about 600Ω of this application, is determined by R1. The BRL is more than 20dB for the audio frequency range. Fig.27 shows the BRL and set impedance at I_{line} = 20mA and 100mA.

Complex set impedance can be realized by a complex network instead of R1. A consequence, however, is a reduction of the BRL and a frequency dependent send and receive gain at complex Z_{ref} respectively complex set termination. Compensation methods are available as indicated in ref.6.

5.3.3. Send and receive gain

The send gain between the MIC inputs and the line measures 44dB at Z_{line} = 600Ω. The gain factor may be adapted by resistor R7.

The receive gain is adjusted to -7dB by R4, measured from line to one of the QR outputs. A modification of the receive gain will also change the LI gain from line to loudspeaker amplifier output. Compensation can be achieved by adjustment of resistor R24.

5.4. Listening-in performance

5.4.1. Listening-in possibilities versus line current

Chapter 4, with the description of the TEA1085/A, shows the output power of the loudspeaker amplifier versus supply current measured with input or line signals with a constant amplitude and frequency.

This paragraph shows the LI behaviour at lower line currents tested with basic speech instead of continuously signals to give a better impression of the LI possibilities.

The tests are done with a relative high line level and maximum gain between line and amplifier output. Measured is the mean level of the loudspeaker signal, at SEL and BTL configuration, versus line current.
The conditions are:

- Voltage gain from telephone line to loudspeaker, at v_{line} < -20dBm, is 22dB at SEL and 28dB at BTL. Potentiometer R21 in maximum position.
- Line signal modulated with speech: vline-mean = -15dBm to -10dBm and vline-peak = 1.5V
- Iline < 35mA

Fig.28 (SEL) and fig.29 (BTL) show the test results. The loudspeaker signal is reduced at lower line currents due to the activated dynamic limiter.

The speech signal from the loudspeaker, connected as a SEL, is still intelligible down to I_{line} = 10mA.

BTL requires the double supply current as SEL for the same signal level. In this case, LI is possible at line currents of 20mA or more.
The combination of low supply currents and loudspeaker signals with a large dynamic range cause many VBB drops down to 2.95V. At I_{line} < 20mA the speech signal is, or becomes, unintelligible because of the many and relative long signal interruptions.
Refer to par.4.5 where the VBB limiter is explained.

Fig.28 and fig.29 indicate, furthermore, the mean and the minimum level of the supply voltages VBB and VDD as a function of line current. The supply voltages decrease due to the generated loudspeaker signals at the relative low line or supply currents.

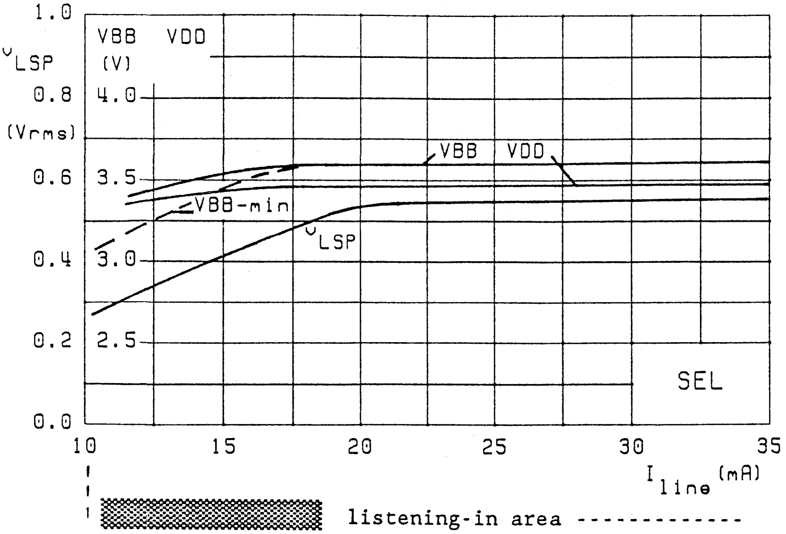


Fig.28. Influence of I_{line} on the listening-in performance; mean and minimum supply voltages. SEL configuration.

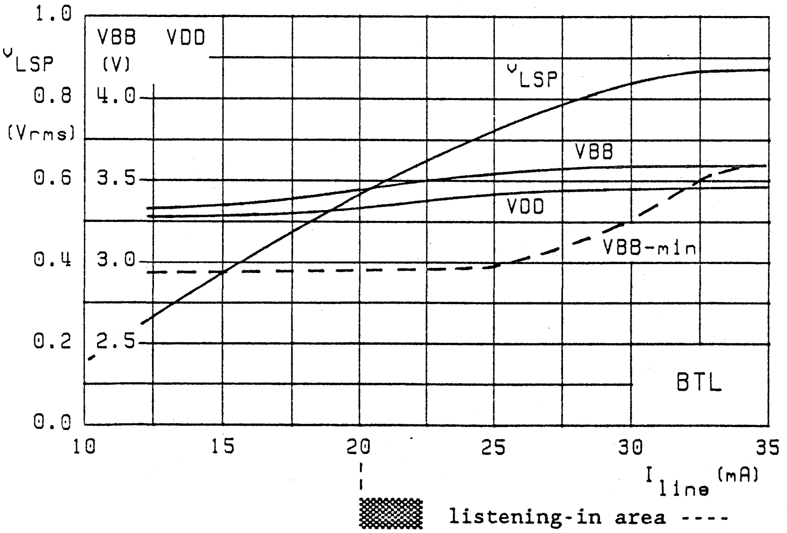


Fig.29. Influence of I_{line} on the listening-in performance; mean and minimum supply voltages. BTL configuration.

The transfer from line to loudspeaker and to earpiece versus frequency is shown in fig.30.

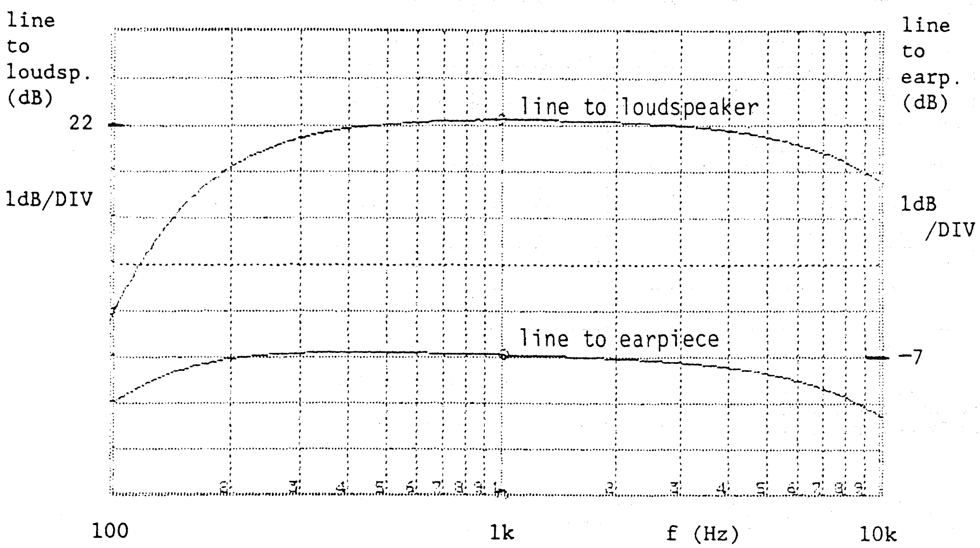


Fig.30 Frequency characteristic of the receive channel from line to loudspeaker and to earpiece in SEL mode.

5.4.2. Output noise level

The output noise voltage $v_{no} \approx 1mV_{rms}$ in case of SEL and $2mV_{rms}$ in BTL mode. Both values are measured at maximum gain from line to amplifier output. The noise voltage here is a result of the generated noise at the line (-78dBmp at 600Ω) and the total LI gain of 22dB or 28dB; the contribution of the loudspeaker amplifier can be ignored.

5.4.3. Howling performance / behaviour

The application of fig.25 is provided with a handset, containing a microphone and earpiece, and a loudspeaker housed in a cabinet.
By moving the handset to the loudspeaker, a howling signal is generated by the loudspeaker and on the line.

A howling signal at the line, its rise and reduction by the LLL, is shown in fig.31. V_{C26} represents the gain control signal of the loudspeaker amplifier in this graph. The LLL is in the Larsen mode when the signal is reduced whereby the residual level is determined by threshold level V_{DTI2}.
When the handset is moved away from the loudspeaker, the gain of the amplifier returns to its nominal level.

Fig.32 shows the sensitivity of the LLL for own speech. The speech signal from the microphone starts to reduce the level of V_{C26}, but the amplifier gain remains nominal. The Larsen mode will not be attacked in case of speech signals because of the HP filter and the integrator of the LLL.

The conditions of fig.31 and fig.32 are:

- Microphone sensitivity: 2mV/Pa. Earpiece sensitivity: 16Pa/V.
- 50Ω loudspeaker, type AD2071/Z50 (see appendix C), connected as a SEL.
- Short line, terminated with 600Ω. I_{line} = 30mA.

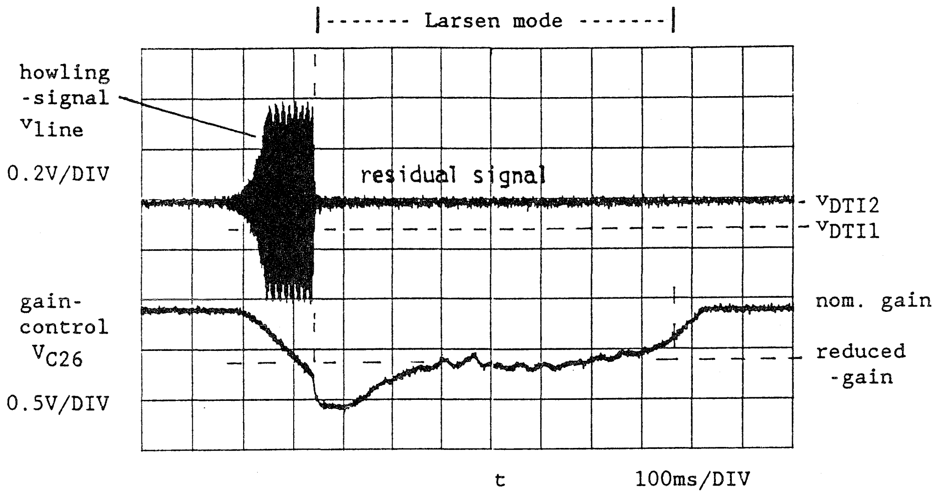


Fig.31 Howling reduction.

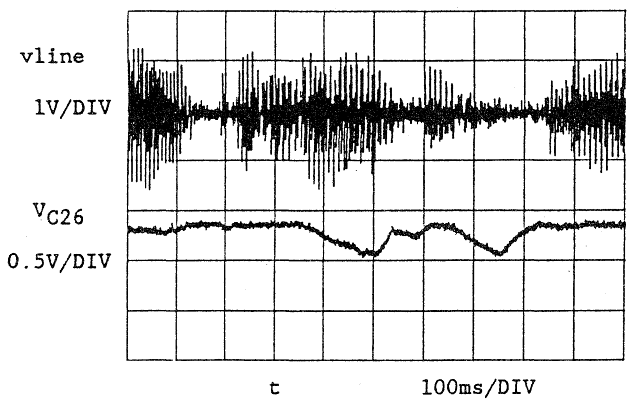


Fig.32. Sensitivity of the LLL for own speech.

5.5. DTMF / pulse dialling

The DTMF and dial pulse signals from the PCD3310 are input to the TEA1064A and TEA1085/A without interface circuitry because of the common reference between the ICs.
The measured DTMF line levels are -10dBm and -8dBm, respectively for the low-frequency and high-frequency group.

The TEA1085/A can be applied in on-hook dialling applications where the line signal has to be monitored by the LI circuit during pulse dialling.
The LI circuit is set in the LI mode before dialling starts and must remain in LI mode during the on-hook dialling procedure.
When the TEA1085 (toggle MUTE) has to be used in such an application, attention has to be paid to the MUTE input level.
The toggle MUTE of the TEA1085 will be reset during dialling, from LI to standby, when the input level is normally low (VSS level) between the MUTE pulses.

The reason is that the PD inputs of the TEA1085 and TEA1064A are activated before the line current is interrupted, because of the time delay between DP signal of the PCD3310 and control signal of the interrupter (BSN254A).
When DP goes HIGH, the TEA1085 and TEA1064A are switched off while the line current is not yet interrupted by the MOS device.
This causes a voltage transient at pin SUP of the TEA1085 which resets the MUTE from LI into the standby mode via an internal capacitive coupling in the start circuit.

This reset problem can be avoided when the MUTE input of the TEA1085 is not normally low but normally open or connected to a high voltage level. The MUTE has to be controlled by a (push button) switch with a 'make' contact as shown in fig.11 or by a μC output which returns to VDD between the MUTE pulses.

The TEA1085A, which has a logic MUTE input, could be applied without any reset problem.

Fig.33 shows the wave forms of the line current and line voltage during pulse dialling (digit "8"). It indicates furthermore, the influence on the supply voltages VBB and VDD when the TEA1085/A and TEA1064A are powered down (via PD input) by the dialling circuit during line interruptions.

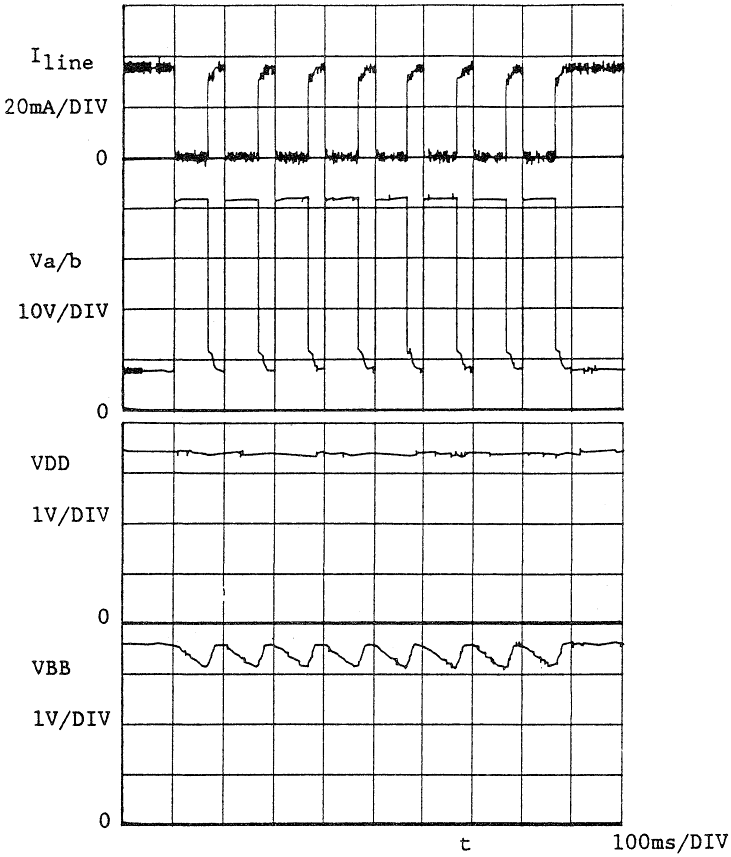


Fig.33 Line current, line voltage and supply voltages during pulse dialling.

6. IMMUNITY TO RF SIGNALS

High-intensity electromagnetic fields induce common mode AM-RF signals in the a/b lines.
These common mode signals can become differential mode (AM) RF signals as a results of asymmetrical parasitic impedances to ground (through the hand holding the handset for example), which can be detected in the applied ICs.

Measures have to be taken in the application to reduce the levels of the unwanted (detected) earpiece and line signals by a reduction of the AM detection sensitivity.

The TEA1085/A has to be applied in combination with a transmission IC, dialler and ringer. The immunity to RF signals of such a combination is different from a single transmission IC. Some basic measures for the TEA1064 and transmission ICs of the TEA1060 family are given in ref.4 respectively ref.7.

The EMC behaviour of demonstration board PR45163, with the TEA1085 / TEA1064A combination, is described and elucidated with measurement results in ref.2.
This DEMO board is carried out with a double sided wiring and two ground reference layers both connected to the VEE terminal of the TEA1064A.
The mounted EMC components, as shown in fig.25, are C33 and C34 to protect the microphone inputs. C10, C35 and C36 are connected to the input respectively the outputs of the receiver, while C37 protects the LSI inputs. C38 reduces the demodulated level at the detector input DCA.

The signal-interference ratio is more than 40dB measured across the a/b terminals, across the earpiece and the loudspeaker. The demodulated level at terminal DCA of the LLL is less than -60dBm.

7. REFERENCES

Ref.1 Data sheet TEA1085, TEA1085A.

Ref.2 PCALE Application Note ETT/AN91012. 'User's Manual -PR4516X- DEMO board with the TEA1085A-TEA1064A listening-in application' by F.v.Dongen.

Ref.3 Documentation of the TEA1082, TEA1083 and TEA1083A:
- Preliminary data sheet TEA1082, PCALE Application Report ETT/AN90022.
- Preliminary data sheet TEA1083/TEA1083A, PCALE Application Report ETT/AN91008.
- Documentation of the DEMO board PR4534X: Call progress monitoring with the TEA1082 - TEA1064A application, PCALE Application Report ETT/AN91009.
- User's Manual PR4535X -DEMO board with TEA1083A-TEA1064A call progress monitoring application-, PCALE Application Report ETT/AN91010.

Ref.4 PCALE Laboratory Report ETT89009. 'Application of the versatile speech /transmission circuit TEA1064 in full electronic telephone sets' by F.v.Dongen and P.J.M.Sijbers.

Ref.5 PCALE Laboratory Report, 12nc 9398 341 10011. 'TEA1060 family; versatile speech/transmission ICs for electronic telephone sets; Designers guide' by P.J.M.Sijbers.

Ref.6 PCALE Application Note ETT/AN90005. 'The TEA1064A with complex set impedance and complex line terminations' by K.Wortel.

Ref.7 PCALE Laboratory Report ETT89016. 'Measures to meet EMC requirements for the TEA1060 family speech/transmission circuits' by M.Coenen and K.Wortel.

Ref.8 PCALE Application Note ETT/AN90017. 'Software controlled ringer for German market' by CH. Voorwinden and K.Wortel.

Ref.9 PCALE Application Note ETT/AN90019. 'Two Chip Telephone Set with a Software Controlled Ringer using the PCD3347 Micro-controller' by C.A.Keogh .

Ref.10 DATA HANDBOOK 'ICs for telecom - IC03a/b.

Ref.11. CNET Specification Technique ST/LAA/ELR/302 (V-2).

8. ACKNOWLEDGMENT

The author thanks the people of the design team of the TEA1085: P.Jouen, J.P.Coulmance, P.J.M.Sijbers and M.Tyndall for the development and design of the circuit parts. Furthermore thanks to A.Charlot and A.Gauthier for their layout work and J.Breard for the realization of the test possibilities.

APPENDIX A: SUPPLY CURRENT VERSUS LOUDSPEAKER IMPEDANCE

The loudspeaker amplifier consists of two push pull class B output stages (outputs QLS1 and QLS2) with common emitter configurations.
The electrical output power generated in the loudspeaker is given by:

$$P_O = v_O^2 / R_{LS} \quad (W)$$

The maximum voltage swing across the loudspeaker as SEL approximates the VBB voltage, while in BTL mode the maximum voltage swing is twice the maximum voltage swing in SEL mode as long as the maximum output current of the power amplifiers (≈ 70mA-peak) is not reached.

The average supply current to drive the loudspeaker is pulled from VBB. The loudspeaker connected as BTL needs twice the average supply current as the loudspeaker connected as SEL for the same output level. This average supply current is in:

$$\text{SEL mode:} \quad I_{av} = \frac{v_O . \surd 2}{\pi . R_{LS}} \quad \text{and in BTL mode:} \quad I_{av} = \frac{2 . v_O . \surd 2}{\pi . R_{LS}} \quad (A).$$

The supply current of the loudspeaker amplifier has to be delivered by the available line current. Taking into account the extra current to drive the power amplifiers, the minimum current I_{LI} to generate the loudspeaker signal v_O can be approximated as follows:

a) In SEL mode:

$$I_{LI\text{-}min} = \frac{k . v_O . \surd 2}{\pi . R_{LS}} \quad (A) \qquad [a1]$$

The factor k represents the loss in the power output stage; k = 1.15 for SEL.

b) In BTL mode, with the coupling capacitor C27 in the application:

$$I_{LI\text{-}min} = \frac{2 . k . v_O . \surd 2}{\pi . R_{LS}} \quad (A) \qquad [a2]$$

c) In BTL mode, <u>without</u> capacitor C27:

$$I_{LI\text{-}min} = \sqrt{\left[\left(\frac{2 . k . v_O . \surd 2}{\pi . R_{LS}}\right)^2 + \left(\frac{V_{offs}}{R_{LS}}\right)^2\right]} \quad (A) \qquad [a3]$$

V_{offs} (150mV max) is the offset voltage between QLS1 and QLS2, k = 1.1 in case of BTL configuration.

The DC current through the loudspeaker can reach a value of 3mA with R_{LS} = 50Ω, caused by the offset voltage. This current loss can not be used to power the loudspeaker; it limits the LI possibilities at low line currents.

The required current into pin SUP, without taking into account the efficiency of the supply (see fig.6), has to be more than $I_{LI\text{-}min} + I_{SUPO}$. The surplus of current is drained to VSS by the voltage stabilizer.

Notes:

- The given approximations to calculate I_{LI} are only valid for loudspeaker signals which are not limited by clipping.
- The voltage drop across C27 can be ignored if $2.\pi.f.C27.R_{LS} >> 1$.
- The SEL mode is preferable above BTL mode when limitation in I_{line} and I_{SUP} occurs.
- When limitation in line voltage occurs, thus when VBB has to be reduced, BTL is preferable above the SEL mode.

APPENDIX B: CIRCUIT CALCULATIONS TRANSMISSION BEHAVIOUR

Fig. B1 shows the equivalent circuit diagram of the output stage of the TEA1064A combined with the TEA1085/A supply. The impedance of L1 is ignored for audio frequencies.
This appendix is included to show the influence of the TEA1085/A on the transmission behaviour of this combination in comparison with a single TEA1064A. Signals are referred with respect to VEE.

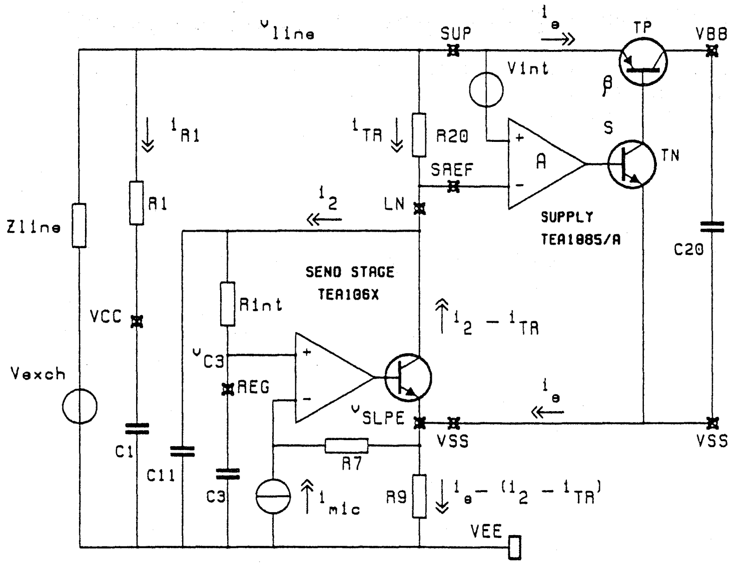


Fig. B1 Equivalent circuit diagram of the TEA1085/A supply part combined with the output stage of the TEA106X.

Transmit mode

The influence of the TEA1085/A on the send gain, between line and SLPE, is derived as follows:

$$v_{SLPE} = - i_{mic} . R7 \qquad [b1]$$

i_{mic} is generated by the microphone signal via the microphone amplifier.

$$i_e - (i_2 - i_{TR}) = \frac{v_{SLPE}}{R9} \qquad [b2] \qquad v_{line} = - (i_e + i_{TR}) . Z_{line}//R1 \qquad [b3]$$

Equation [b3] can be rewritten into, using [b2]:

$$v_{line} = -\left(\frac{v_{SLPE}}{R9} + i_2\right) \cdot Z_{line}//R1 \qquad [b4]$$

where $$i_2 = \frac{v_{LN}}{Zx} \quad \text{and} \quad Zx = \left(Rint + \frac{1}{p.C3}\right) // \frac{1}{p.C11} \qquad [b5]$$

The voltage on terminal LN approximates the line voltage, thus $v_{LN} = v_{line}$, because $i_{TR} \ll i_e$ which is proved by:

$i_e = i_{TR}.R20.A.S.\beta = A_{tot}.i_{TR}$ [b6], with $A_{tot} = R20.A.S.\beta > 10.000$.

The gain between line and SLPE can be derived by substitution of equation [b5] into [b4] giving:

$$\frac{v_{line}}{v_{SLPE}} = -\frac{Z_{line} // Zx // R1}{R9} \qquad [b7]$$

Conclusions:
- The transmit or send gain from SLPE to the line is the same as without TEA1085/A according to equation [b7].
- The microphone signal at SLPE results in a modulation current which flows via the TEA1085/A and not, as usual, through the transmit stage of the transmission IC (as long as RL << R20), because $i_e \gg i_{TR}$ according to equation [b6] with $A_{tot} > 10.000$.

Set impedance

$$Z_{set} = \frac{v_{line}}{i_e + i_{TR} + i_{R1}} \quad [b8], \quad i_e + i_{TR} = \frac{v_{SLPE}}{R9} + i_2 \quad [b9], \quad i_{R1} = \frac{v_{line}}{R1} \quad [b10]$$

i_{TR} can be ignored with respect to i_e, according equation [b6], so:

$$v_{SLPE} = v_{C3} = \frac{v_{line}}{1 + p.C3.Rint} \qquad [b11]$$

$$\text{and } i_2 = \frac{v_{line}}{(Rint + 1/(p.C3)) // (1/p.C11)} \qquad [b12]$$

Substitution of the equations [b12] and [b11] into [b9], and substitution of the result of [b9] and [b10] into [b8] results, with a little bit of fantasy into:

$$Zset = \frac{R9.(1 + p.C3.Rint)}{(1 + p.C3.R9)} // R1 // \frac{1}{p.C11} \qquad [b12]$$

Conclusion:

- The set impedance of this application is the same as the impedance of the TEA1064A application without TEA1085A or TEA1085. It consists, for audio frequencies, of the coil function "p.C3.Rint.R9" in parallel with Rint, R1 and the impedance of C11.
- C11, connected between LN and VEE, appears thus in parallel with R1 (or a complex network) which determines the set impedance for audio frequencies.

APPENDIX C: GAIN OF THE LOUDSPEAKER AMPLIFIER

The loudspeaker amplifier of the TEA1085/A is coupled between the receiver output of the transmission IC and the loudspeaker.
Its gain (v_O/v_{LSI}) depends in general on the:
- mean signal level of the earcapsule, which depends on the required sound pressure p_{TE} and transducer sensitivity.
- required sound pressure p_{TL} from the loudspeaker.

The voltage gain of the TEA1085/A is fixed and is chosen for a sensitive earcapsule delivering a specified sound pressure by means of a relatively low voltage level. It measures 35dB in SEL mode and 41dB in BTL; these values are based on the following data:
- CCITT (P34) recommendations / CNET requirements: $p_{TL} \approx$ 75dBSPL (at 55dBA room noise) measured at 50cm distance. Amplifier noise $P_O \leq$ 30dBA.
- Loudspeaker AD2071/Z50 (Philips type); 50Ω, p_{TL} = 90dBSPL at P_O = 550mW, 1kHz and 50cm distance.
- Earcapsule with a sensitivity of 30dBPa/V (≈32Pa/V).
- p_{TE} = 0.58Pa at a mean line signal level of $100mV_{rms}$.

The required voltage across the loudspeaker (producing 75dBSPL) is calculated, by using the expressions:

$$p_{TL2} = p_{TL1} + 10 \log \frac{P_{O2}}{P_{O1}} \qquad \text{and} \qquad v_O = \sqrt{(P_{O2} \cdot R_{LS})}$$

where: p_{TL1} = 90dBSPL, p_{TL2} = 75dBSPL, P_{O1} = 550mW and R_{LS} = 50Ω,

It results in a voltage level of $v_O = 0.93V_{rms}$ (P_{O2} = 17.4mW//50Ω)

With an earcapsule sensitivity of 32Pa/V and an output pressure of 0.58Pa, the voltage across the earcapsule (and amplifier input) is $v_T = 18.1mV_{rms}$.

So, the voltage gain of the loudspeaker amplifier has to be more than 20 log (v_O/v_T) = 34dB. The gain is fixed at 35dB for the SEL configuration. The amplifier delivers 6dB more in BTL mode.

An attenuation network (R21, R24 in fig. 25) can be applied between receiver output and amplifier input in case of less sensitive earcapsules. An earcapsule with a sensitivity of 24dBPa/V (16Pa/V) requires a $v_T = 36.2mV_{rms}$ for the same sound pressure of 0.58Pa.
The input signal of the amplifier has to be attenuated by 6dB, via R21 and R24, to deliver the same sound pressure of 75dBSPL.

A noise level of 30dBA from the loudspeaker corresponds with a psophometric output noise voltage of $v_{no} = 3.9mV_{rms}$ (approximately), and an input noise voltage of $70\mu V_{rms}$. This maximum output level is the allowed contribution of receiving channel of the transmission IC and loudspeaker amplifier.
The specified output noise of the TEA1085/A is $170\mu V_{rms}$ ($3\mu V_{rms}$ at the input), so the noise voltage of the transmission IC must be less than $70\mu V_{rms}$.

APPENDIX D: TIMING OF THE LLL DETECTOR

The LLL of the TEA1085/A has to detect howling and to reduce the loudspeaker amplifier gain in order to avoid annoying loudspeaker and line signals.
Furthermore, the LLL must not be activated by speech signals from the microphone.
The LLL makes use of a HP filter and an integrator to separate howling signals from speech signals.
This appendix concerns the detector. It calculates the threshold settings with respect to PTT requirements and immunity for own speech; see ref.11 and paragraph 4.8.

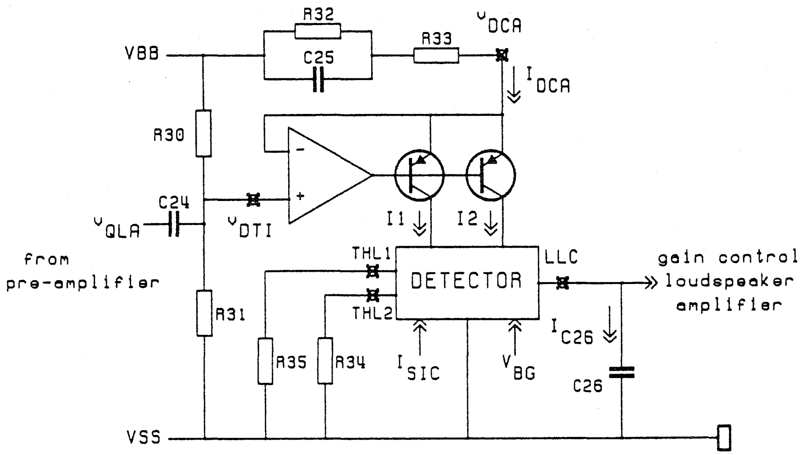


Fig.D1 Detector circuitry of the LLL.

Howling reduction is specified by the CNET, see ref.11:
- The CNET defines howling (or Larsen) as a signal on the line, due to acoustic instabilities, of more than $100mV_{rms}$ during more than 100ms.
- No howling may occur at normal use of the telephone set; distance between loudspeaker and microphone d > d0; d0 = 20cm.
- At d ≤ d0, the howling signal must be reduced to an acceptable signal level within 200ms.

Requirements are clear, but interpretation and hardware realization is another case.

The LLL is designed according to the following data:
- A howling signal (v_{LA1}) on the line of ≥ $100mV_{rms}$ has to be suppressed within 200ms.
- This howling signal, with frequencies of > 1.5kHz, is present at DTI and DCA. $v_{DCA} = v_{DTI} = m1.v_{LA1}$, m1 = 1 in this story.
- To separate howling from speech, an attack delay time (t_{ad}) is required of > 120ms, because of a mean 'burst time' of speech signals of about 100ms. The attack delay time has to be > 120ms, even for howling levels on the line of >> $100mV_{rms}$.
- After the attack delay time, the howling level is reduced to an acceptable

signal level defined as V_{DTI2}.

The attack delay time has to be: 120ms < t_{ad} < 200ms. Chosen is t_{ad} = 160ms at v_{AL1} = 100mV$_{rms}$. Thus with 100mV$_{rms}$ at DTI, DCA and the line, the gain of the loudspeaker amplifier has to be reduced at t_{ad} = 160ms.

Calculation of threshold V_{DTI1} versus t_{ad}

C26 is charged by I_{SIC}, during LI (or speech) mode, up to $V_{C26\text{-}max}$:

$$I_{SIC} = VBG / R36 \quad [d1], \qquad V_{C26\text{-}max} = V_{BG} + V_{BG}/2 \quad [d2]$$

As soon as howling occurs whereby $v_{DTI} > V_{DTI1}$, C26 is momentarily discharged by a current of $3.I_{SIC}$.

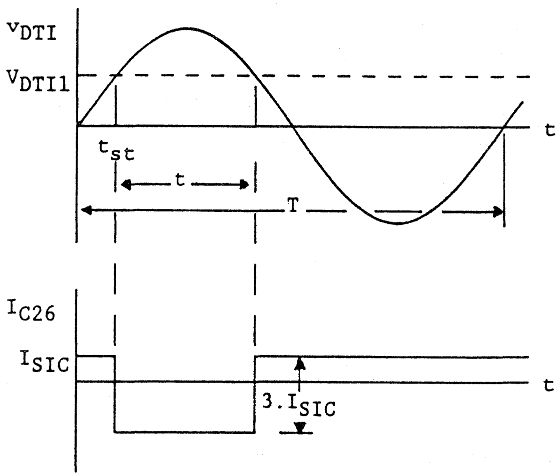


Fig.D2 Timing diagram of the current into C26 versus v_{DTI}.

The total discharge current of C26 is: $$I_{C26} = I_{SIC}.(3.\frac{t}{T} - 1) \qquad [d3]$$

The Larsen mode, where gain reduction starts, is reached if $V_{C26} \le V_{BG}$.
C26 has to be discharged from $V_{BG} + V_{BG}/2$ to V_{BG} during t_{ad} as given by:

$$t_{ad} = \frac{C26.(V_{BG} + V_{BG}/2 - V_{BG})}{I_{C26}} \qquad [d4]$$

Substitution of equations [d1] and [d3] in [d4] gives:

$$\frac{t}{T} = \frac{C26.R36}{6.tad} + \frac{1}{3} \qquad [d5]$$

Ratio t/T is expressed by the RC-time of R36.C26 and the attack delay time t_{ad}; it will be used to calculated the first threshold V_{DTI1}.
Referred to fig.D2, V_{DTI1} is expressed by:

$$V_{DTI1} = v_{DTI}.\sqrt{2}.\sin(2.\pi.\frac{t_{st}}{T}) \quad [d6] \quad \text{with} \quad t_{st} = 0.5.(\frac{T}{2} - t)$$

$$V_{DTI1} = v_{DTI}.\sqrt{2}.\sin(0.5 - t/T).\pi \qquad [d6]$$

Substitution of equation [d5] in [d6], where $v_{DTI} = m1.\ v_{LA1}$, results in:

$$V_{DTI1} = m1.v_{LA1}.\sqrt{2}.\sin[0.5 - (\frac{C26.R36}{6.tad} + \frac{1}{3})].\pi \qquad [d7]$$

Using $C26 = 1\mu F$, $R36 = 120k\Omega$, $v_{LA1} = 100mV_{rms}$, m1=1 and a required t_{ad} = 160ms gives a threshold value $V_{DTI1} = 18.5mV$.

When $v_{LA1} >> 100mV_{rms}$ and V_{DTI1} remains 18.5mV, the attack delay time t_{ad} approaches the RC time of R36.C26 = 120ms. The requirements are thus fulfilled.

Calculation of R35 versus V_{DTI1}

Threshold V_{DTI1} is realized by means of the external setting resistor R35. Current I1, consisting of a DC component and an AC component depending on v_{DTI}, is compared with threshold current $V_{BG}/R35$.

The calculation of R35 is given with respect to fig.D1.

I1 is a function of I_{DCA} and v_{DCA}, $\quad I1 = 0.5.(I_{DCA} + v_{DCA}/R33) \quad$ [d8]

where $v_{DCA} = v_{DTI}$, $f > 1/(2.\pi.R33.C25)$ and $f >> 1/(2.\pi.R32.C25)$..

$$\frac{V_{BG}}{R35} = 0.5.(I_{DCA} + \frac{v_{DTI}}{R33}) \quad [d9], \quad I_{DCA} = VBB.\frac{R30}{R30 + R31}.\frac{1}{R32 + R33} \qquad [d10]$$

The detector starts when $v_{DTI} > V_{DTI1}$.
R35 can be expressed as a function of threshold V_{DTI1}

$$R35 = 2.\frac{V_{BG}}{V_{DTI1}/R33 + I_{DCA}} \qquad [d11]$$

With V_{DTI1} = 18.5mV, V_{BG} = 1.26V, VBB = 3.6V and the components values: R30 = 100kΩ, R31 = 220kΩ, R32 = 100kΩ, R33 = 500Ω and C32 = 330nF results in a value R35 = 52kΩ.
The current I_{DCA} = 11.2μA.

Calculation of R34 versus V_{DTI2}

Current I2 (fig.D1) consists of a DC component and an AC component which depends on v_{DTI}. The level of I2 is compared with the threshold current V_{BG}/R34.
The calculation of R34 is done in a similar way as the calculation of R35.

$$R34 = 2.\frac{V_{BG}}{V_{DTI2}/R33 + I_{DCA}} \qquad [d12]$$

Calculation of the release time t_{LAr}

C26 is discharged, in the Larsen mode, to a minimum level of about $V_{BG}/2$. Returning from the Larsen to the LI mode, C26 is recharged to $V_{BG} + V_{BG}/2$ by means of a current $I_{C26} = I_{SIC}/2$. The maximum release time is given by:

$$t_{LAr} = \frac{C26.(V_{BG} + V_{BG}/2 - V_{BG}/2)}{I_{C26}} \qquad [d13]$$

Substitution of $I_{C26} = \dfrac{V_{BG}}{2.R36}$ in equation [d13] results in t_{LAr} = 2.R36.C26

With the components values of the application, R36 = 120kΩ and C26 = 1μF, the maximum release time is 240ms.

t_{LAr} can be adapted by means of R36 or C26. A consequence is however, that t_{ad} also is modified.

APPENDIX E: ADJUSTMENT PROCEDURE AND APPLICATION HINTS

Adjustment procedure

	consult:
- Adapt line voltage to requirements	data sheet transm. IC
- Adjust VBB (if necessary) by R38	par.4.2.3, fig.7
Take into account a minimum $V_{SUP-VBB}$	par.4.2.1
- Select loudspeaker drive: SEL or BTL	
depending on: loudspeaker impedance	par.4.6.3, fig.17
required output power	appendix A and C
minimum line current	
VBB voltage	
- Adjust send gain	data sheet transm. IC
- Adjust pre-amplifier of LLL by means of R29.	par.4.8.2
Adapt R27 and R28, if necessary	
- Modify R32, depending on VBB setting	par.4.8.3
- Adjust receive gain	data sheet transm. IC
- Modify attenuation network of R21 and R24	par.4.6.2, appendix C
Select permanent gain reduction by GSC inputs, if necessary	par.4.6.2
- Modify settings of LLL, if necessary	par.4.8, appendix D

Application hints

- High line currents; $I_{SUP} > I_{SUP-max} = 120mA$:
  Connect a resistance R in the SUP wire and a diode (BAS11) across R20 (cathode to LN); $R \simeq 5\Omega$.

- Low line currents:
  The dynamic limiter function can be improved, at low line currents, by means of an RC network ($R = 10k\Omega$, $C = 1\mu F$) connected between DLC and VSS. This RC network replaces the 330nF as shown in application diagram fig.25.

- Manual MUTE control of the TEA1085 (toggle MUTE) by a push button switch:
  Connect a debounce capacitor (10nF) between MUTE and VSS.

- On-hook dialling applications with the TEA1085 (toggle MUTE):
  Consult paragraph 4.4 (fig.11) when start up is required in the LI mode.

- On-hook pulse dialling applications with the TEA1085 (toggle MUTE):
  Keep the MUTE input referred to VBB voltage potential to prevent MUTE reset during dialling pulses. Consult paragraphs 5.5 and 4.4 (fig.11).

- Test of LI circuit part:
  Short circuit resistor R30 when the LI function has to be tested with microphone signals with a constant amplitude. This to prevent gain reduction of the LI loudspeaker amplifier by the LLL.

- Transmission ICs (TEA1060/61/65/66/68) without option for parallel operation :
  The line voltage can show LF relaxations at very low line currents around 7mA (paragraph 4.1). A resistor of 1kΩ has to be connected between SUP and VBB, when stable operation is required at that low line current range.

- Transmission ICs with VEE as common reference (TEA1060/61/62/65/66/67/68):
  Interface circuitry could be necessary between μC and logic inputs of the transmission IC and TEA1085/A. Consult the data sheet of the TEA1085/A concerning the minimum input voltages with respect to VSS; see also paragraph 4.2.5.

- Transmission IC with stabilized supply option (TEA1064A/63):
  The line voltage, and also the voltage space between SUP and VBB, vary with the supply current from VCC2. Do not apply the stabilized supply option when $V_{SUP-VBB}$ can become critical with respect to the efficiency of the supply circuit; see chapter 4.2.1.

- Transmission ICs without dynamic limiter (TEA1060/61/62/66/67/68):
  The line signal can be heavily distorted at large voltage swings (>7dBm) when the minimum instantaneous working voltage of the TEA1085/A (1.4V) is reached; see paragraph 4.2.2. A resistor of 100Ω between SREF and LN of the TEA106X can be useful.

APPENDIX F: LIST OF ABBREVIATIONS AND SYMBOLS

A	: voltage gain at low frequencies
AM	: Amplitude Modulation
β	: current gain (IC/IB)
BRL	: Balance Return Loss, with respect to Z_{ref}
BTL	: Bridge Tied Load; loudspeaker connected between the two outputs
CCITT	: International Telegraph and Telephone Consultative Committee
ch.	: chapter
CNET	: Centre National d'Etudes des Telecommunications
dBA	: sound pressure level, weighted according to A-curve, 0dBA = $2.10^{-5}N/m^2$ at 1kHz
dBm	: power level relative to 0dBm = 1mW into 600Ω
dBmp	: power noise level, psophometrically weighted (P53-curve)
dBSPL	: sound pressure level 0dBSPL = $2.10^{-5}N/m^2$
DP	: Dial Pulse
DTMF	: Dual Tone Multi Frequency
EMC	: Electro Magnetic Compatibility
ETSI	: European Telecommunications Standards Institute
f	: frequency
fig.	: figure
H	: logic High level; H ≥ 1.5V
HP	: High Pass
I_{BB}	: current consumption from VBB
I_{BIAS}	: bias current of supply stage
IC	: integrated circuit
I_{C20}	: charge current of C20
I_{CC1}	: supply current of the TEA106X
I_{LI}	: effective supply current to power the loudspeaker
$I_{LI\text{-}transmit}$	: supply current to power the loudspeaker during transmitting
I_{line}	: line current
I_{SUP}	: current into terminal SUP
I_{TR}	: bias current of the output stage of the TEA106X
L	: logic Low level; L ≤ 0.3V
LI	: listening-in
LLL	: Larsen Level Limiter to reduce howling
LSE	: Loudspeaker enable
mV/Pa	: sensitivity of microphone (also used: dB rel. 1V/Pa)
p	: $p = 2.j.\pi.f \quad j = \sqrt{-1}$
Pa	: (Pascal) sound pressure level, 1Pa = $1N/m^2$
Pa/V	: sensitivity of earcapsule (also used: dB rel. 1Pa/V)
par.	: paragraph
PD	: power down
P_O	: generated power into the loudspeaker
p_{TE}, p_{TL}	: sound pressure from earcapsule (p_{TE}) in Pa respectively from loudspeaker (p_{TL}) in dBSPL

QR	:	receiver output of the TEA106X
$R38_{VA-VBB}$	:	R38 connected between VA and VBB
$R38_{VA-VSS}$	:	R38 connected between VA and VSS
ref.	:	reference, listed in chapter 7
RF	:	Radio Frequencies
rms	:	root mean square
R_{LS}	:	loudspeaker impedance
S	:	transconductance (mA/V)
SEL	:	Single Ended Load; loudspeaker connected between one of the outputs and VSS
SLPE	:	terminal of the TEA106X
t_{ad}	:	attack delay time of the LLL
t_{att}	:	attack time of the dynamic limiter
TEA106X	:	transmission IC of the TEA1060 family; TEA1060/61, TEA1062, TEA1063, TEA1064A, TEA1065, TEA1066, TEA1067 or TEA1068
TEA108X	:	IC of the line monitoring family: TEA1082, TEA1083, TEA1083A, TEA1085 or TEA1085A
TEA1085/A	:	LI circuits TEA1085A and TEA1085
THD	:	Total Harmonic Distortion
t_{LAa}	:	Larsen attack time
t_{LAr}	:	Larsen release time
t_{rel}	:	release time of the dynamic limiter
μC	:	microcontroller
V_{BG}	:	bandgap voltage
VEE	:	ground reference terminal of the TEA106X
v_{LAl}	:	Larsen signal on the line
$v_{line-peak}$	:	peak level of the line voltage
v_{no}	:	output noise voltage, psophometrically weighted (P53-curve)
v_O	:	output voltage of the loudspeaker amplifier
V_{offs}	:	offset voltage
$V_{SUP-VBB}$	:	DC voltage between SUP and VBB
Z_{line}	:	impedance of the telephone line
Z_{ref}	:	reference impedance
Z_{set}	:	impedance of the telephone set or application

APPLICATION NOTE Nr	ETT94001
TITLE	Application of the TEA1096 transmission and listening-in circuit
AUTHOR	F. van Dongen, L. C. van Leeuwen
DATE	January 1994

Table of Contents:

1 Introduction

The TEA1096 as well as the TEA1096A are bipolar telephony IC's for use in line powered telephone sets. They offer a transmission function and a group listening-in (or monitoring) facility of the received line signal via a loudspeaker.
The TEA1096 and TEA1096A incorporate a line interface with active output stage, a stabilized supply, send and receive amplifiers, a double anti side tone circuit, line loss compensation, a loudspeaker amplifier, and dynamic limiters for transmit and loudspeaker signal.

This report gives a detailed description of the TEA1096 and TEA1096A and an application example. The description is given by means of block diagrams and discussion of the details of the sub-blocks.

For product details is referred to the Device Specification ref.1.

1.1 Package

The TEA1096 is available in the packages: DIL28 (SOT117N) and SO28 (SOT136A).

1.2 TEA1096 structure

The TEA1096(A) has a single ground reference structure. Transmission, listening-in and applied microcontroller make use of the same ground reference; no interface components or interface circuits are required. Furthermore the TEA1096(A) has an active output stage which offers a flat frequency response of the microphone signal in case of complex set and line impedance.

1.3 Differences TEA1096 - TEA1096A

There are two versions of the TEA1096 available:

a. TEA1096
b. TEA1096A

The difference concerns DC settings and the volume control of the loudspeaker signal:

TEA 1096	TEA 1096 A
Loudspeaker volume is controlled by an external potmeter. (see paragraph 3.5)	Loudspeaker volume is controlled by an external voltage at VCI (pin 2) (see paragraph 3.5), or/and an external potmeter.
The VBB supply voltage is adjustable via V_{BA} (see page 11, V_{BB} voltage stabilizer).	The VBB voltage is fixed (see page 11, V_{BB} voltage stabilizer).

2 Block diagram

The block diagram of the TEA1096(A) is shown in Figure 47. The pinning diagrams are given by means of Figure 1. Also a short description of the block diagrams is given including the functions of the external components.

2.1 Pinning of the TEA1096(A)

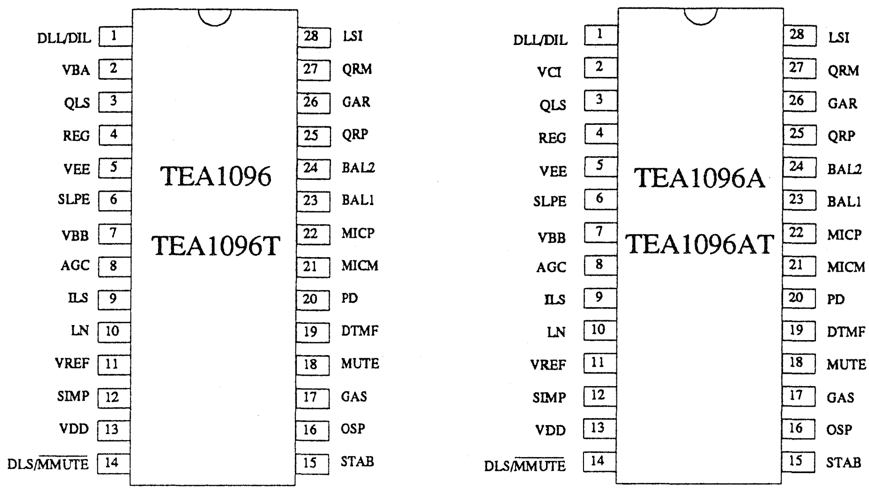


Figure 1 Pin configuration of the TEA1096 and TEA1096A.

Pin description

SYMBOL	PINNING TEA1096	PINNING TEA1096A	DESCRIPTION
DLL/DIL	1	1	Dynamic limiter and disable input for loudspeaker amplifier
VBA	2		VBB voltage adjustment
VCI		2	Volume control input for loudspeaker amplifier
QLS	3	3	Loudspeaker amplifier output
REG	4	4	Decoupling line voltage stabilizer
VEE	5	5	Negative line terminal (ground reference)
SLPE	6	6	Stabilized voltage, connection for slope resistor
VBB	7	7	Stabilized supply voltage for listening-in circuitry
AGC	8	8	Automatic gain control
ILS	9	9	Input line signal
LN	10	10	Positive line terminal
VREF	11	11	Reference voltage output
SIMP	12	12	Set impedance input
VDD	13	13	Supply voltage for speech circuitry/peripherals
DLS/$\overline{\text{MMUTE}}$	14	14	Dynamic limiter for sending and microphone mute
STAB	15	15	Reference current adjustment
OSP	16	16	Sending preamplifier output
GAS	17	17	Sending gain adjustment
MUTE	18	18	Mute input to select speech or DTMF dialling
DTMF	19	19	Dual tone multi frequency input
PD	20	20	Power down input
MICM	21	21	Inverting microphone amplifier input
MICP	22	22	Non inverting microphone amplifier input
BAL1	23	23	Connection for balance network 1
BAL2	24	24	Connection for balance network 2
QRP	25	25	Non inverting receiving amplifier output
GAR	26	26	Receiving gain adjustment
QRM	27	27	Inverting receiving amplifier output
LSI	28	28	Loudspeaker amplifier input

2.2 Blocks

As can be seen in Figure 47 the IC's comprise five parts: the DC line interface and supply, the AC line interface, the send channel, the receive channel and the listening-in channel. These blocks are briefly described.

DC line interface and supply

This block consists of the line voltage stabilizer, the current switch with TR1 and TR2, the V_{BB} supply voltage stabilizer, the low line current supervisor and a voltage reference.
The IC is supplied from the line via the diode bridge (not drawn in block diagram) and slope resistor R_{slpe}. The DC voltage between SLPE and VEE is stabilized. A coil function between SLPE and VEE is created by C_{reg}, R_{slpe} and an internal resistance. The transmission part is supplied by V_{DD} which is connected to SLPE via the low pass filter R_{dd}-C_{vdd}.

A current switch isolates the line interface from the stabilized supply voltage V_{BB} available for the listening-in part. C_{vbb} is the buffer capacitor of V_{BB}. A low line current supervisor reduces the SLPE voltage and the current into LN at low line currents. A PD function can be activated to put the IC in low current consumption mode. Pin REG can be used to change the voltage on SLPE. Pin VBA of the TEA1096 can be used to change the V_{BB} voltage.

AC line interface

An active output stage modulates the line via output LN by means of a bias current taken from the line. The set impedance is active generated and determined by external resistor (or network) R_{simp} connected to SIMP. Pin ILS is the feedback input of the active output stage and serves also as input of the receiver stage. A dynamic limiter reduces the distortion of the microphone signal. Timing capacitor C_{dls} is connected to DLS.

Send channel

The send channel includes a microphone pre amplifier with symmetrical inputs MICP and MICM and a DTMF pre amplifier with input DTMF. The send gain is determined by R_{gas} connected to GAS. The mode, speech or dialling, can be selected by the MUTE input.
To compensate the line loss an AGC function controls the gain of the microphone and receive channel as a function of the line current. It delivers also a control signal to balance between receiver inputs BAL1 and BAL2. R_{agc} connected to pin AGC determines the start and stop current. An internal reference current is adjusted by R_{stab} connected to STAB.
Pin OSP is available for side tone suppression.

Receive channel

The line signal available at pin ILS is coupled to the receive channel via an internal attenuator. The receive signal is amplified by the output amplifiers and available at the earpiece outputs QRP and QPM. The gain is adjusted by R_{gar} connected between QRP and GAR. To suppress side tone the receive signal is compared with the microphone signal from

OSP via the two side tone networks $Z_{set'} + Z_{bal'-short}$ and $Z_{set'} + Z_{bal'-long}$. A confidence tone is available at the earpiece outputs during DTMF dialling.

Listening-in channel

This block consist of a preamplifier, loudspeaker amplifier, dynamic limiter and in case of the TEA1096A a volume control interface. The receive signal from QRP is coupled to input LSI via an attenuator with potentiometer (TEA1096) and couple capacitor C_{lsi}. The loudspeaker amplifier gain of the TEA1096 is fixed while the loudspeaker amplifier gain of the TEA1096A can be varied with a DC level at pin VCI. The timing capacitor C_{dll} of the dynamic limiter is connected to DLL.

3 Description of the TEA1096(A)

All properties, values, data, figures etc. in this report are valid for the TEA1096 as well as the TEA1096A, unless stated otherwise.

3.1 Supply

The power distribution is schematically represented in Figure 2.

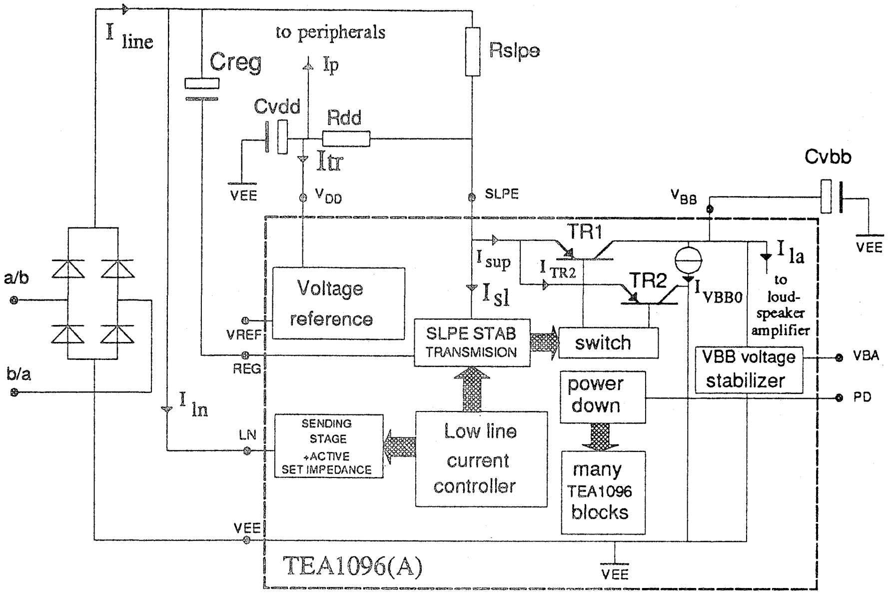


Figure 2 Block diagram of the power distribution of the TEA1096.

The power consumption is build up of the following parts:

1. A power supply, called V_{DD}, for the external peripherals to be connected to the TEA1096, for instance a microcontroller. The concerned current is represented by I_p. R_{dd} and C_{vdd} form a low pass to provide a supply decoupling. The block 'Voltage reference' does not stabilize V_{DD}. The voltage V_{DD} depends on the value of R_{dd}, I_p, I_{tr} and V_{SLPE}.

2. A supply current for the transmission functions of the IC: I_{tr}, consumed from V_{DD}.

3. A power supply for the loudspeaker amplifier, indicated by V_{BB}. I_{la} is the available current for the loudspeaker amplifier. This is the most power consuming part.

4. A current for the sending stage: I_{ln}.

5. A supply current for the SLPE stabilization: I_{sl}.

6. A supply current for the listening in part: I_{VBB0}.

Under nominal conditions the currents have the following magnitudes:

I_{ln}	= 5 mA
I_{tr}	= 2.4 mA
I_{sl}	= 0.3 mA
I_{VBB0}	= 2.5 mA
Total	10.2 mA

I_{sup} depends on the available line current (I_{line}). In case of normal speech signals on the line I_{TR2} = 0 mA and I_{sup} flows completely through TR1. If $I_{la}+I_{VBB0} < I_{sup}$ then the remaining part of I_{sup} is consumed by the block 'V_{BB} voltage stabilizer', as result of which the 'V_{BB} voltage stabilizer' is able to stabilize V_{BB}.

From this it follows that for good performance the following requirement has to be satisfied:

$$I_{line} > 10.2\text{ mA} + I_p + I_{la}$$

where I_p is determined by the design in which the TEA1096(A) is applied. The remaining part of I_{line} can be used for I_{la}, the supply current of the loudspeaker amplifier.

If the above mentioned requirement is not met, the listening-in feature does not work, performance is reduced and the voltages behave as indicated in Figure 7.

V_{slpe} is stabilized nominal at 4.4V, and V_{BB} at 3.6V.

Voltage reference

The block 'voltage reference' provides V_{REF} with the relation:

$$V_{REF} = ½\ V_{DD}$$

Current consumption out of the pin V_{REF} is limited to +/- 200μA.

SLPE stabilizer

The block 'SLPE STAB' stabilizes the voltage on SLPE. The speech signal remains present on SLPE. V_{SLPE} can be adjusted by a resistor between SLPE and REG, called R_{REG} (see Figure 48).

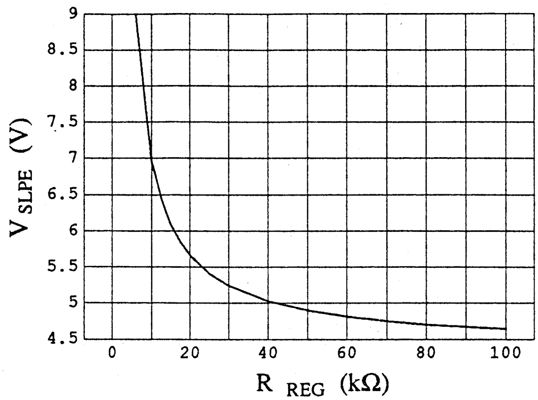


Figure 3 V_{SLPE} **can be adjusted by** R_{REG}.

The dependency of V_{SLPE} from R_{REG} is shown in Figure 3. R_{REG} does not affect V_{BB}. V_{SLPE} can also approximated by the equation:

$$V_{SLPE} \approx V_{SLPE\text{-}nom} + \frac{25k\Omega}{R_{REG}}$$

Where: $V_{SLPE\text{-}nom} = 4.4V \pm 0.25V$

A resistor connected between pin REG and VEE can be used to decrease the SLPE voltage while maintaining V_{BB} to its nominal value. When adjusting the SLPE voltage to a lower value, care should be taken that the difference to V_{BB} is at least 0.8V. A resistor connected between REG and VEE affects the active set impedance (see Figure 8). A resistor connected between REG and SLPE does not affect the active set impedance.

This block controls also the block 'switch', that is necessary to avoid distortion of large transmission signals. For more details about the block 'switch' see Figure 10.

V_{BB} voltage stabilizer

The block 'V_{BB} voltage stabilizer' stabilizes the voltage V_{BB} which is nominal 3.6V. On V_{BB} no speech signal is present. V_{BB} is decoupled by C_{vbb}. In case of line currents larger than 10.2mA, the 'V_{BB} voltage stabilizer' or the listening-in function (or both) consume the remaining line current.

In case of TEA1096 a resistor connected between pin VBA and VEE, called R_{vbh} in Figure 48, can be used to increase the V_{BB} voltage. When adjusting the V_{BB} voltage to a higher value, care should be taken that the difference to V_{SLPE} is at least 0.8V:

$$V_{SLPE} - V_{BB} > 0.8V$$

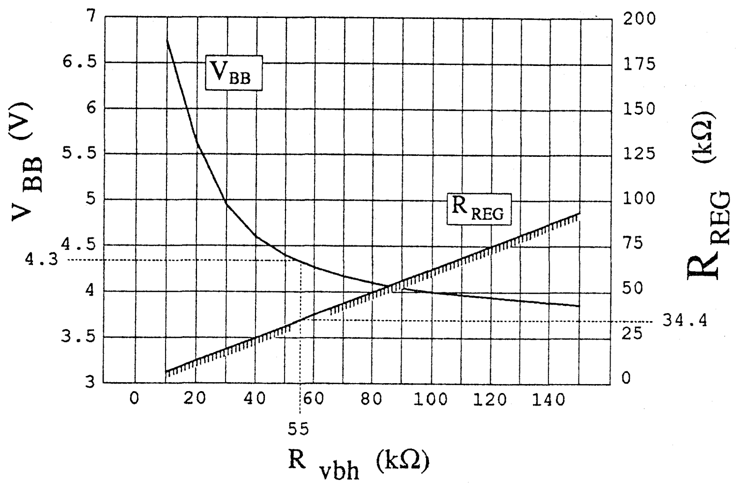


Figure 4 **V_{BB} versus R_{vbh} with the maximum allowable value of R_{REG} for keeping a voltage difference of 0.8V between V_{SLPE} and V_{BB}.**

In Figure 4 V_{BB} is shown as function of R_{vbh}. V_{BB} can be approximated by:

$$V_{BB} \approx V_{BB\text{-}nom} + \frac{40k\Omega}{R_{vbh}}$$

Where: $V_{BB\text{-}nom}$=3.6V±0.2V

As result of the requirement that the typical voltage between SLPE and VBB has to be larger than 0.8V, there is a maximum value for R_{REG}.

Example: Suppose a V_{BB} is required of 4.3V (see Figure 4). The appropriate R_{vbh} is 55kΩ, and R_{REG} has to be less than 39.3kΩ.

A resistor connected between pin VBA and pin V_{BB}, called R_{vbl} in Figure 48, will decrease the V_{BB} voltage see Figure 5.

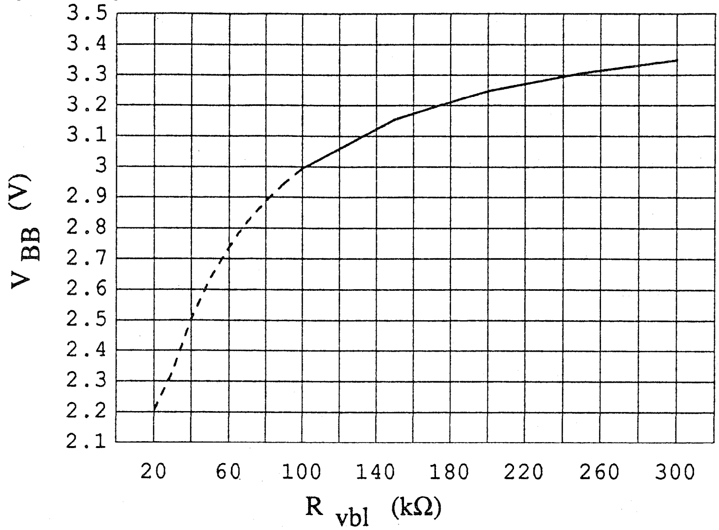


Figure 5 V_{BB} **can be adjusted to a lower value by** R_{vbl}.

The V_{BB} limiter threshold detector level is 2.8V±10%. If V_{BB} is below the threshold, the listening-in circuit is disabled. R_{vbl} is in parallel with internal resistors that have a tolerance of ±20%. A sufficient safety margin for the adjusted V_{BB} is assured if V_{BB} is nominal above 3V. This implicates a minimum value for R_{vbl} of 100kΩ.

Sending stage

The block 'sending stage' draws the current I_{ln} from the line in order to be able to modulate the line current by a speech signal. A positive part of a speech signal agrees with drawing less current from the line, a negative part agrees with drawing more current.

Low line current controller

The block 'low line current controller' affects the 'sending stage' and the 'SLPE stabilizer'. The line current can be very small in case of placing two (or more) telephones in parallel. To let the TEA1096 work properly, though with less performance, the 'low line current controller' decreases I_{ln} and the voltage at SLPE.

Power down

To reduce the current consumption during pulse dialling, the TEA1096 is provided with a power down PD input. The PD input has a pull down structure. When the voltage on PD is high, the current consumption from V_{DD} capacitor C_{vdd} is <150 µA and from the V_{BB} supply point <500 µA. The capacitors C_{vdd}=100 µF and C_{vbb}=470µF are sufficient to power the TEA1096(A) during pulse dialling.

3.2 Line voltage regulation

The line voltage V_{line}, that is the voltage between LN and VEE in Figure 2, is regulated as illustrated in Figure 6. There are two area's known: a low current area (below the knee), and a high current area (above the knee).

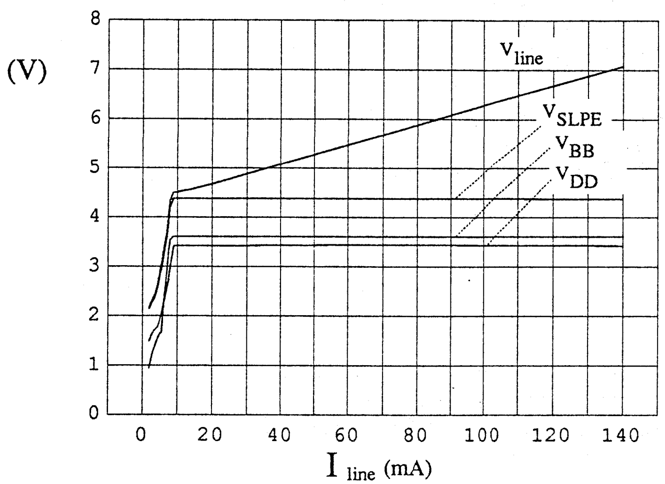


Figure 6 V_{line}, V_{BB}, V_{DD} and V_{SLPE} versus line current.

The low current area is related to parallel operation with another telephone set, the high current area is related to normal operation.

The low current area is controlled by the 'low line current controller'. V_{BB} and V_{DD} decrease with decreasing line current. Also I_{ln} decreases with decreasing line current. The size of I_{ln} determines the amplitude of the sent speech signal. If I_{line} < 10 mA there is no current available for the listening-in part.

In Figure 7 the low current area is enlarged, so that also the behaviour of V_{DD} and V_{BB} can be seen.

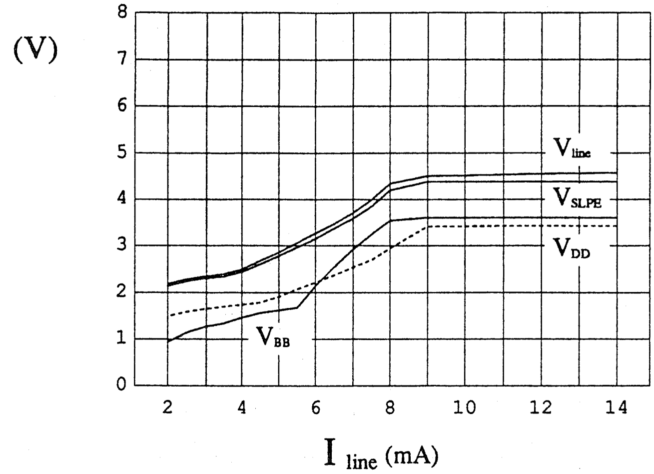


Figure 7 low current area enlarged.

The difference between V_{SLPE} and $V_{LN\text{-}VEE}$ is determined by R_{SLPE}. R_{SLPE} is advised to be 20Ω. Changing R_{SLPE} causes another value of the knee voltage in Figure 7, it affects the set impedance (L_{eq} in Figure 8), and the AGC curves (paragraph 3.8 and Figure 36).

In case of a practical application (for instance Figure 49) where a protection circuit, consisting of a zener diode and a protection resistor (R_{prot}), and a line current interrupter for pulse dialling, is added, the line voltage V_{line} behaves different from Figure 6, because R_{prot} and the interrupter are put in series with R_{SLPE}.

The impedance between the pins LN and VEE from Figure 2 can be substituted by the following circuit:

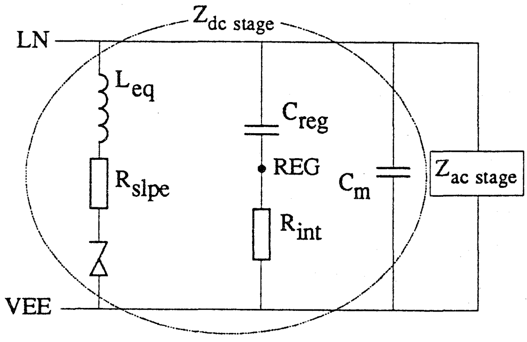


Figure 8 **Equivalent impedance between LN and VEE.**

Where:

R_{int} is typically 40 kΩ.
C_{reg} is 3.3 µF (externally to be connected).
R_{slpe} = 20 Ω (externally to be connected).

$$L_{eq} = R_{int} R_{slpe} C_{reg} = 40k\Omega \ x \ 20\Omega \ x \ 3.3\mu F = 2.64H$$

$$Z_{set} = Z_{dc\ stage} \,//\, Z_{ac\ stage}$$

$Z_{dc\ stage}$ has for audio frequencies negligible influence therefore: $Z_{set} \approx Z_{ac\ stage}$

C_m is an internal capacitor, determined by stability requirements. Its capacitance appears multiplied between LN and VEE. The cross over frequency, caused by C_m is situated above 10 kHz, thus the influence of C_m can be neglected in relation with the telephony frequency band.

Figure 9 shows the impedance between LN and VEE as function of frequency, for Z_{simp} = 6 kΩ. The target impedance Z_{set} is 600Ω. The influence of C_m can be seen above 10 kHz, and of L_{eq} below 50 Hz.

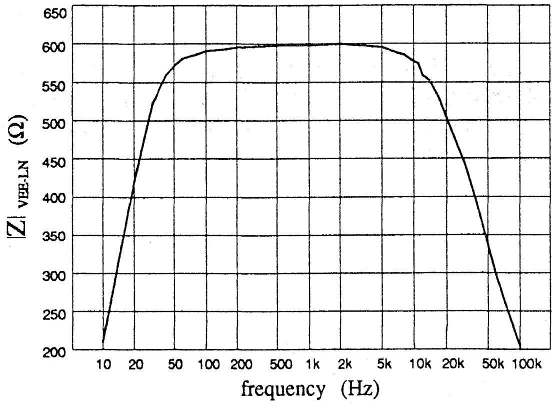


Figure 9 **Impedance between LN and VEE, with Z_{simp}=6kΩ, I_{line}=20mA, C_{reg}=3.3μF.**

Active set impedance

The set impedance Z_{set}, has to be adjusted externally, by placing a circuit equivalent of the required set impedance, multiplied by a factor 10, between the pins 'SIMP' and 'VREF'(see Figure 47 and Figure 48):

$$Z_{set} = 0.1 \; x \; Z_{simp}$$

Where: Z_{set} is the (ac) set impedance.
Z_{simp} is a circuit externally to be connected.

Z_{set} may be real or complex, without affecting transmission characteristics, if line terminations are equal. This is a result of the fact that there is not made use of the principle of the Wheatstone bridge for side tone suppression, which should cause frequency roll-off, but the attenuated microphone signal is subtracted from the attenuated line signal. The line signal is attenuated internally, the microphone signal is attenuated externally. The amplitude of the line signal depends on the line impedance. In case of a complex line impedance, the microphone signal can be attenuated also by a complex circuit, in such a way subtracting delivers a frequency independent side tone suppression. For more details is referred to paragraph 3.9.

Start up

As mentioned above, there are two supply points from which current can be drawn for driving peripherals: V_{DD} and V_{BB}. The difference between them is the way of stabilizing and the time they need to stabilize after the telephone set has gone off-hook. V_{DD} is the fastest one and therefore the best supply point for peripherals, such as a microcontroller, that has to be activated immediately.

During starting up, the transmission functions are initiated first, next the listening-in function.

The start up sequence is as follows:

- charge of C_{reg} via R_{int} and charge of C_{vdd} via R_{dd} and R_{slpe}
- quick charge of C_{dls} (Figure 47) when $V_{DD} > 2.4$ V
- charge of C_{vbb} when the line voltage stabilizer starts to function; $V_{DD} > 2.4$V.
- quick charge of C_{dll} (Figure 47) when $V_{BB} > 2.5$V.

The transmission and listening-in functions are available when their dynamic limiters are ready to operate; thus when their capacitors, C_{dls} respectively C_{dll} are charged to operating level.

Supply efficiency

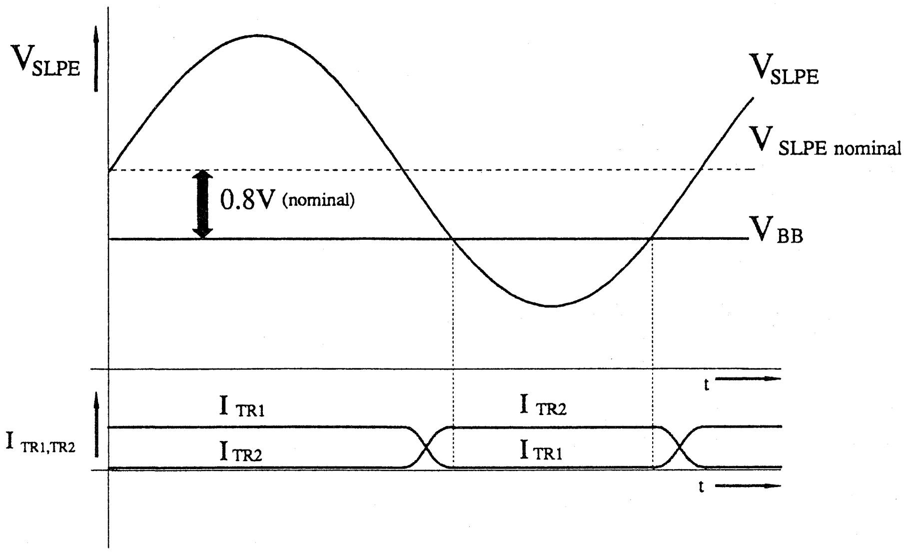


Figure 10 **TR2 prevents distortion by taking over the current through TR1, in case of large transmission signals.**

Figure 10 shows the situation of a large transmission signal on the line. The speech signal causes V_{slpe} to drop below V_{BB}. Referring to Figure 2 it will be obvious that, during this moment, no current can flow through TR1. For not disturbing the input impedance, TR2 takes over the current, so that it flows to VEE. The sum of I_{TR1} and I_{TR2} is constant. Because current flows to VEE there is less current available for the listening-in part.

3.3 Send block

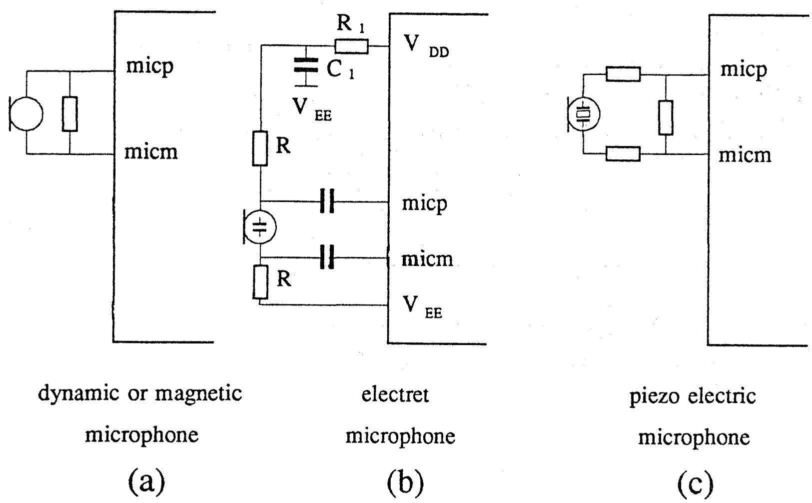


Figure 11 **Microphone arrangements for several microphone types.**

Refer to Figure 47. The TEA1096(A) has symmetrical microphone inputs MICP, MICM with a high impedance. The input resistance at MICP and at MICM is nominal 32kΩ with maximum tolerances of ±20%. With this high input impedance it is possible to determine the matching of several microphone types very accurately by means of external components. The circuit is suitable for dynamic, magnetic or piezo electric microphones with symmetrical drive; electret microphones with built-in source follower or preamplifier can be used in asymmetrical mode. The arrangements with several microphone types are shown in Figure 11.

In case of electret microphones V_{DD} has to be decoupled by C_1 and R_1, because of the high microphone amplifier gain.

Gain adjustment microphone- and DTMF amplifier.

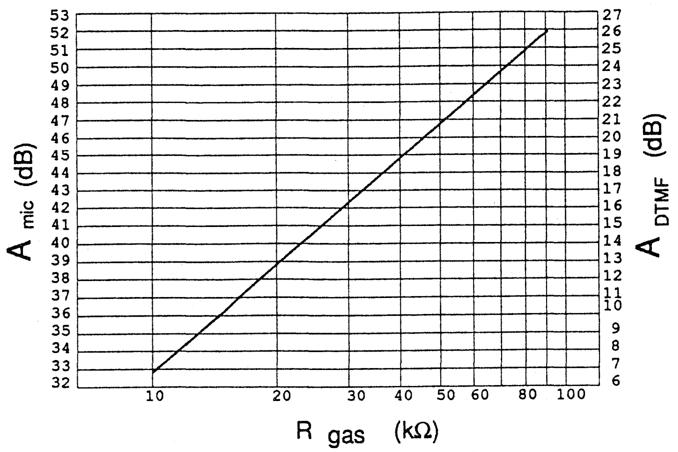


Figure 12 **Microphone gain and DTMF gain as a function of the value of R_{gas}.**
f=1kHz, I_{line}=20mA, Z_{set}=Z_{line}=600Ω, C_{gas} is not present.

In speech mode (input MUTE is low, or not connected because it is internally pulled down), the overall gain from MICP-MICM to LN can be adjusted from 33 dB up to 52 dB to suit specific requirements with respect to various microphone sensitivities. The gain is proportional to the value of R_{gas} and equals 52 dB with R_{gas}=90.9kΩ and 33 dB with R_{gas}=10kΩ, while I_{line}=20 mA.
The microphone gain is shown as a function of R_{gas} in Figure 12.

The microphone gain is linear with R_{gas} with the relation:

$$A_{mic} = 8.75 \ x \ 10^{-3} \ x \ R_{gas} \ x \ \frac{|Z_{line}|}{|Z_{set}+Z_{line}|}$$

The effect of the Automatic Gain Control (AGC) gain can not be seen in Figure 12, because AGC is not active. For more details concerning AGC is referred to paragraph 3.8.

The maximum input voltage between MICM and MICP up to 2% total harmonic distortion is 17 mV (rms), with I_{line}=20mA, f=1kHz, if the gain is smaller than 37 dB.

The microphone amplifier can be disabled by shorting pin DLS to VEE (secret function) and can be muted into DTMF mode by applying a high level on pin MUTE.

The TEA1096(A) has an asymmetrical DTMF input with an input resistance of 20 kΩ. In DTMF mode, the overall gain from DTMF to LN is proportional to R_{gas} and thus also to

A_{mic}. The relation is:

$$A_{mic} = A_{DTMF} + 26dB$$

Switch-over from mic-mode to DTMF-mode is click-free.

The maximum input voltage between DTMF and VEE up to 2% harmonic distortion is 176 mV (rms), with I_{line}=20mA, f=1kHz.

Microphone frequency roll-off

A capacitor C_{gas} (Figure 48) in parallel with R_{gas} can be used to provide a first order low pass filter, to support the requirements for the telephony frequency range. The 3dB-point is completely determined by R_{gas} and C_{gas} and equals:

$$f_{3dB\,microphone} = \frac{1}{2\pi R_{gas} C_{gas}}$$

Noise

To obtain optimum noise performance the microphone inputs have to be loaded. Figure 13 shows the noise voltage at the output LN in the cases that the microphone inputs are connected to microphones with impedances of respectively 200Ω and 8.2kΩ. It is measured psophometrically (P53 curve).

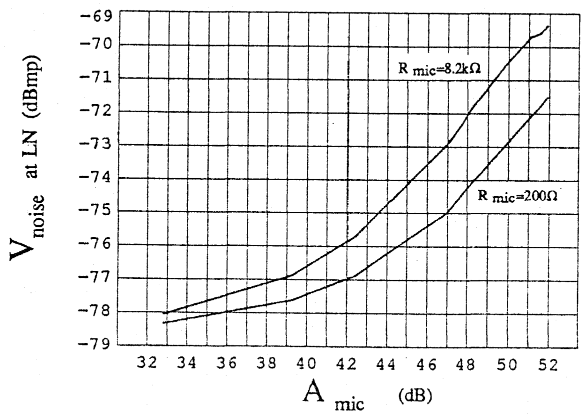


Figure 13 Psophometrically (P53 curve) weighted noise on the transmitter output as a function of microphone amplifier gain, at I_{line}=20mA. Z_{line}=Z_{set}=600Ω.

Dynamic limiter

The maximum level of the send signal at LN is limited at 3.65 V_{pp} at nominal DC settings, or it is limited by the bias current (I_{ln} in Figure 2) flowing into LN at I_{line} < 14 mA. To prevent distortion (as a result of clipping) of the transmitted signal a dynamic limiter is incorporated. The use of such a limiter improves side tone performance considerably, because side tone suppression is obtained by subtracting a signal, similar to the transmitted signal, from the ear piece signal.

When peaks of the transmitted signal on the line start to saturate the output stage, the gain of the sending amplifier is reduced rapidly. The time in which gain reduction is effected (the attack time) is in the order of a few milliseconds. The microphone channel stays in the gain-reduced mode until the peaks of the transmitted signal no longer cause saturation. The gain of the microphone channel then returns to its normal value within the release time.

Both attack and release time are proportional to the value of the capacitor Cdls. The total harmonic distortion of the microphone amplifier in reduced gain mode stays below 2% up to 10 dB of input voltage overdrive (provided that V_{MICP}, V_{MICM} is below 10 mV_{rms}, and gain is 52 dB).

In Figure 14 and Figure 15 the attack time and release time are schematically indicated.

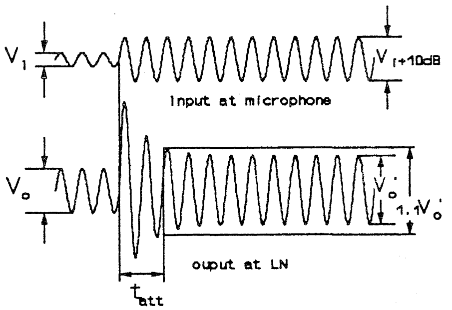


Figure 14 Result of the dynamic behaviour of the dynamic limiter on the microphone input signal. Definition of attack time (t_{att}).

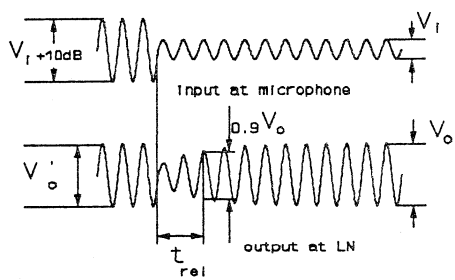


Figure 15 Result of the dynamic behaviour of the dynamic limiter on the microphone input signal. Definition of the release time (t_{rel}).

Microphone mute function

The dynamic limiter of the TEA1096(A) also provides a microphone mute (secret function) when pin DLS is shorted to VEE. The microphone gain is then 80 dB lower. The release time after a microphone mute is in the order of a few 10 ms.

Parallel operation

In case of parallel operation, the operating voltage of the TEA1096 is adjusted at a lower level. This affects the performance of the microphone amplifier. In Figure 16 the maximum output voltage at LN is shown as a function of line current flowing into the TEA1096, with a 600Ω telephone set connected in parallel.

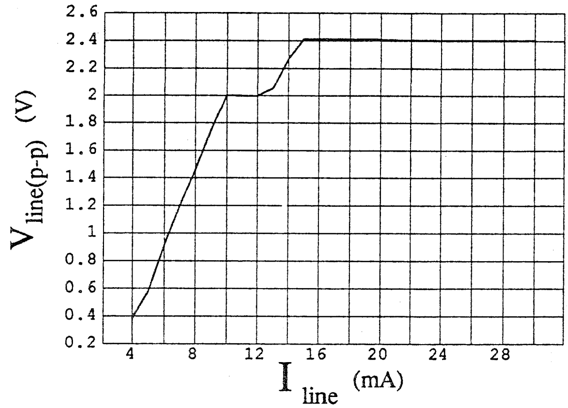


Figure 16 **Maximum output voltage peak-peak versus line current, with one telephone set of 600Ω connected in parallel. Distortion is kept constant at 3% by adjusting the microphone input amplitude.**

Transmit gain is set at 52 dB in case of a normal 600Ω load, but with a telephone set of 600Ω connected in parallel the gain decreases with 6 dB. The maximum output swing is not determined by the DC-voltage at pin LN but by the available current in the output stage of the TEA1096 (I_{ln} in Figure 2). The dynamic limiter prevents distortion also during parallel operation.

3.4 Receive block

The receive gain is defined between the line connection LN and the earpiece complementary outputs QRP (non-inverting) and QRM (inverting). With a R_{gar} resistor, between GAR and QRP, equal to 90.9 kΩ the gain from LN to QRP is -2.5 dB. The outputs may be used to connect a dynamic, magnetic or piezoelectric earpiece. When the earpiece impedance exceeds 450Ω, differential drive (BTL connection) can be used to obtain double the voltage. As both outputs are in opposite phase, the gain from LN to QRP, QRM is 3.5 dB.

Input / output structure

The receive channel consists of two preamplifiers, attenuators, and two output amplifiers. The line signal at pin ILS is attenuated and offered to the internal inputs of the preamplifiers. The attenuation is flat over the audio frequency range.

The inputs BAL1 and BAL2 are externally available and have to be connected to the side tone network(s) (for more details about BAL1 and BAL2 see paragraph 3.9 'anti side tone circuit').

The complementary (class B) outputs, the non-inverting output QRP and the inverting output QRM, can be applied either for single ended load or for bridge tied load, depending on impedance, sensitivity and type of the earpiece.

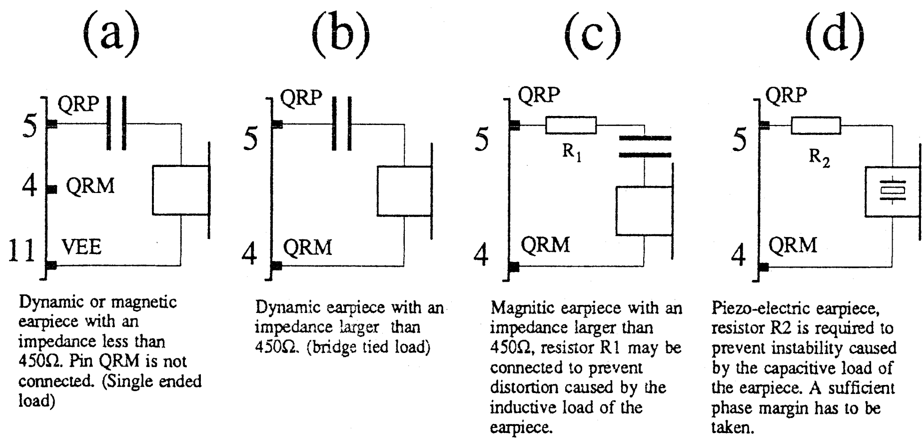


Figure 17 **Connection of several types of earpieces.**

It can drive either dynamic, magnetic or piezoelectric earpieces as shown in Figure 17. Earpieces with an impedance up to 450Ω must be driven in single ended mode (low impedance dynamic or magnetic capsules). This is shown in Figure 17a.

For impedances above 450Ω, with a high impedance dynamic, magnetic or piezoelectric capsule, differential drive is possible, as shown in Figure 17b, c, d.

The additional series resistor R_1 shown in Figure 17c can be used to prevent distortion of the output signal. If the maximum output current of the output stage is exceeded, in case of inductive loads, the output voltage and output current are not suited any more. This causes distortion in the earpiece which can be prevented by R_1.

A piezoelectric earpiece represents a capacitive load. Capacitive loading of the receiving output stage is permitted up to a maximum of 100nF between QRP and QRM. A resistor of 300Ω has to be connected in series to prevent instabilities.

Receive gain adjustments

By means of the R_{gar} resistor, the gain of the receiving amplifier can be adjusted to suit the sensitivity of the transducer which is used. The allowed range is comprised between -14 dB an +6 dB for single ended load, and between -8dB and +12 dB for bridge tied load.
In Figure 18 $A_{rx\text{-}BTL}$ represents the gain between LN and QRM-QRP in case of Bridge Tied Load (BTL). $A_{rx\text{-}BTL}$ depends on R_{gar}.

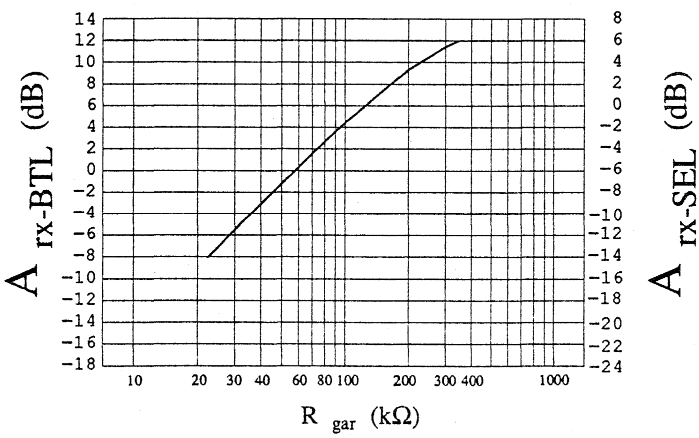


Figure 18 **Gain between LN and QRP-QRM ($A_{rx\text{-}BTL}$), when the automatic gain control (AGC) is disabled. The load between QRP and QRM (bridge tied load) is 450Ω. f=1kHz, line current is 20mA.**

The difference between BTL and SEL (Single Ended Load) is 6 dB see Figure 18:

$$A_{rx-BTL} = A_{rx-SEL} + 6dB$$

Up to R_{gar} = 100kΩ A_{rx-SEL} can be calculated by the formula:

$$A_{rx-SEL} = 8.74x10^{-6}\ x\ R_{gar} \frac{Z_{earpiece}}{Z_{out}+Z_{earpiece}}$$

Where: $Z_{earpiece}$ = impedance of the earpiece, including series resistors
Z_{out} = output impedance of the receiving amplifier (typical 5Ω)

If R_{gar} becomes larger than 100kΩ the open loop gain affects $A_{rx-BTL/SEL}$ so deviation between several samples of the TEA1096, increases above 1 dB.

Maximum input signal

The received speech signal on the line is internally attenuated before it enters the inputs of the preamplifiers. At low receive gain the distortion is mainly determined by the preamplifiers, because the speech signal is there relatively large. In case of minimum receive gain of A_{rx-BTL}= - 8 dB, the output signal at QRP has a total harmonic distortion of 2% at v_{line} = 830 mV_{rms} and 10% at v_{line} = 1.9 V_{rms}.

Maximum output signal

The maximum voltage swing of the output stages depends on the impedance of the earpiece and the supply voltage VDD which is derived from VSLPE via R_{dd}.

The maximum output swing versus V_{SLPE} is shown in Figure 19 at R_{dd}=392Ω, I_{line}=20mA, total harmonic distortion is constant 2%. V_{SLPE} is changed by adjusting R_{reg}. The maximum output swing will be higher under speech conditions because in that case the ratio of peak to RMS value is higher than with a continuous signal.

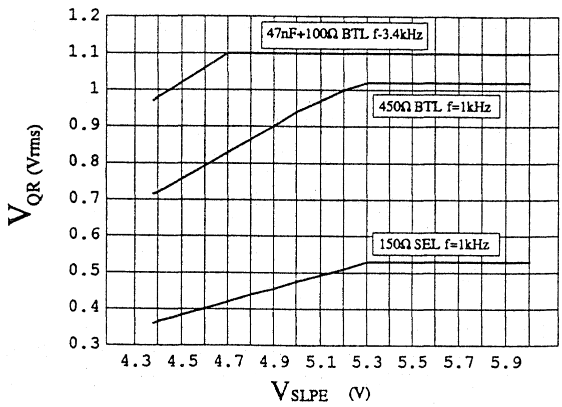


Figure 19 **Maximum output voltage for several loads in BTL and SEL. Total harmonic distortion is constant 2%, line current is 20 mA, SEL-gain=-2dB, BTL-gain=+4dB.**

Stability and frequency roll-off

Two external capacitors, C_{gar1} = 100 pF and C_{gar2} = 1 nF, ensure stability. Capacitor C_{gar1} is connected between QRP and GAR and capacitor C_{gar2} is connected between GAR and VEE.

The C_{gar1} capacitor is also used to obtain a first order low-pass filter in combination with R_{gar}. The cut-off frequency equals:

$$f_{cut\text{-}off} = \frac{1}{2\pi R_{gar} C_{gar1}}$$

The cut-off frequency can be adjusted by the C_{gar1} capacitor, but the relationship:

$$C_{gar2} = 10 \ x \ C_{gar1}$$

must be maintained for stability requirements.

Noise from microphone amplifier into earpiece.

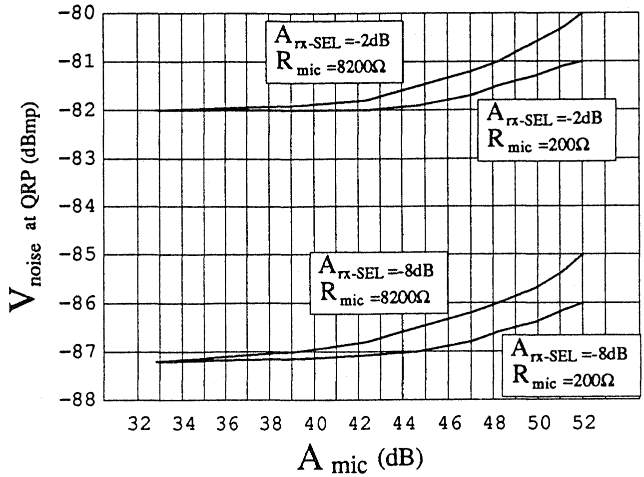


Figure 20 **Psophometrically weighted noise at the QRP earpiece output versus microphone amplifier gain.**

Figure 20 shows the noise generated by the microphone amplifier measured at the earpiece amplifier output QRP, as a function of microphone amplifier gain and for two microphone resistances.

Confidence tone

During DTMF dialling, the dialling tones can be heard at a low level in the earpiece. The sound level at the earpiece outputs QRP/QRM depends on the gain that has been set for the receiving amplifier and on the tone level applied to the DTMF input.

The gain between DTMF input and the receiving output QRP is given by:

$$20\log A_{CT} = 20\log A_{rx\text{-}SEL} - 20dB$$

Where:	A_{CT}	=	gain between DTMF input and QRP output
	A_{rx_SEL}	=	gain between LN and QRP-VEE

3.5 Loudspeaker amplifier

A block diagram of the loudspeaker amplifier is represented in Figure 47 on page 58. As can be seen the loudspeaker part consist of three parts in case of the TEA1096 and of four parts in case of the TEA1096A. The parts are:

1. Preamplifier
2. Output amplifier
3. Dynamic limiter
4. Volume control interface (only for TEA1096A)

Input / output structure

As can be seen in the figure on the page 58 the loudspeaker amplifier input is asymmetrical. The input impedance is typical 10kΩ. The DC level of the input is referenced to 1.25V. If the input voltage is smaller than 80 mV_{rms}, the total harmonic distortion will be below 3%, for all line currents.

The loudspeaker amplifier input has to be coupled to the earpiece output QRP. In case of the TEA1096 a potentiometer circuit has to be added for volume control. In case of the TEA1096A volume is controlled via the 'volume control interface' which acts as an attenuator. The voltage on pin VCI determines the attenuation. The gain of the earpiece amplifier, determined by R_{gar}, influences also the loudspeaker output signal.

The output QLS, in series with a capacitor, drives the loudspeaker.

Listening-in gain and volume control

The total gain from line to loudspeaker depends on receiving gain (line → QRP), attenuation from receiver output (QRP) to loudspeaker amplifier input (LSI) and gain of the loudspeaker amplifier. The last is fixed for TEA1096 and variable for TEA1096A.

The value of R_{LSI} and R_{pot} have to be in accordance with the listening-in gain of the application, which depends on the sensitivity of the ear capsule. Choose $R_{pot} << Z_{LSI}$. Z_{LSI} is the loudspeaker amplifier input resistance and is typical 10 kΩ.

R_{LSI} and R_{pot} have to be chosen in such a way that for a mean speech level on the line, the maximum input signal at LSI measures 20 mV_{rms}. This corresponds with the maximum loudspeaker amplifier output signal for $R_{loudspeaker}$=50Ω, V_{BB}=3.6V and I_{line} > 20 mA.

TEA 1096

The loudspeaker amplifier gain from LSI to QLS is fixed at 35.5 dB.

Volume has to be controlled by an external potentiometer as illustrated in the block diagram on page 58. Note that three high pass filters arise in the path from earpiece amplifier (=loudspeaker preamplifier) output to the loudspeaker. These are:

1. C_{lsi2} in combination with R_{lsi} and R_{pot}
2. C_{lsi1} in combination with the input resistance of input LSI (10 kΩ)
3. C_{qls} in combination with the loudspeaker impedance.

Choose the cut-off frequencies of the high passes at about 2 or 3 octaves below 300Hz.

TEA 1096 A

The total gain is 35.5 dB minus the attenuation caused by Volume Control Interface (see the block diagram in Figure 47).

The attenuation is controlled by an external voltage V_{VCI}. The attenuation as a function of V_{VCI} is:

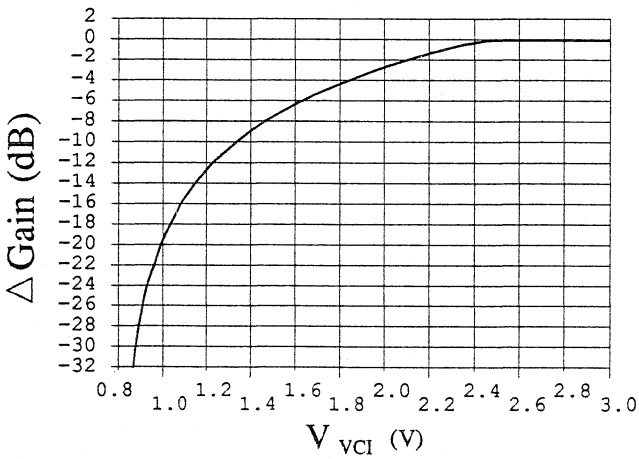


Figure 21 **Influence of V_{VCI} on loudspeaker amplifier gain.**

The voltage at VCI can be regulated by a microcontroller, modulating the duty cycle of an output port. The configuration is as follows:

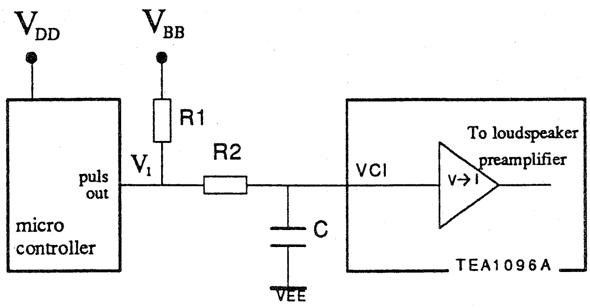


Figure 22 **Volume control by a micro controller.**

In this circuit V_{VCI} yields:

$$V_{VCI} = V_{BB}\frac{t_1}{t_2\left(1+\frac{R_1}{R_2}\right)+t_1}$$

Assumed that t_1 and t_2 are defined as:

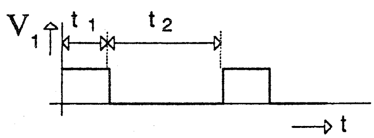


Figure 23 **definition of t_1 and t_2**

And: - V_1 is the output voltage of an ideal open drain output of the microcontroller.
- V_{VCI} is the voltage at input VCI
- R_2C is about 100 times the period time t_1+t_2

Note that two high pass filters arise in the path from earpiece amplifier (=loudspeaker preamplifier) output up to the loudspeaker. These are:

1. C_{lsi1} in combination with the input resistance of input LSI (10 kΩ)
2. C_{qls} in combination with the loudspeaker impedance.

Choose the cut-off frequencies of the high passes at about 2 or 3 octaves below 300Hz.

Output capabilities

The output stage of the power amplifier is optimised for use with a 50 Ω loudspeaker, for instance Philips type AD2071/Z50. Loudspeakers with another impedance can be applied too, as shown in Figure 24

The output impedance of the amplifier is about 3 Ω, measured with 50 Ω load.

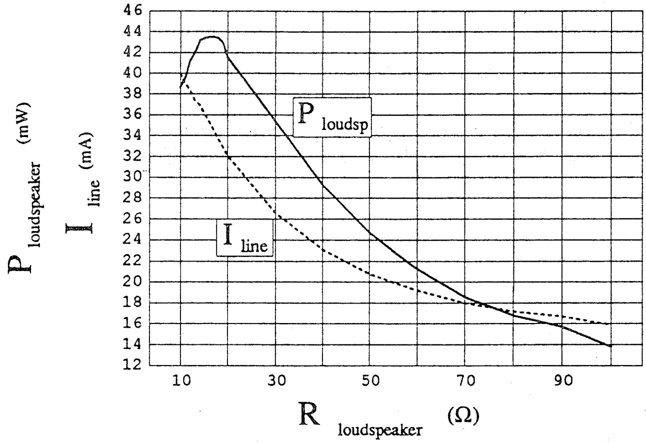


Figure 24 **Maximal loudspeaker output power with minimal required line current, versus loudspeaker resistance. Total harmonic distortion < 3%, frequency: 1 kHz.**

Figure 24 shows the maximum loudspeaker amplifier output power with its corresponding minimal required line current as a function of the loudspeaker resistance , at nominal V_{BB} and at a distortion level of ≤ 3%. The dynamic limiter prevents distortion of the loudspeaker signal.

Figure 25 shows the loudspeaker output power as a function of the line current, for several values of V_{BB} in case of TEA1096. As can be seen, the maximal output power depends on V_{BB}. V_{SLPE} must be adjusted in relation with V_{BB}, with the relation: $V_{SLPE} > V_{BB} + 0.8V$.

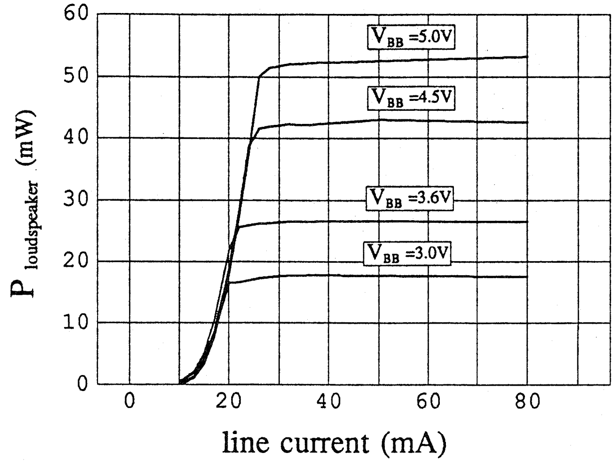


Figure 25 Loudspeaker amplifier output power with V_{BB} versus line current. Total harmonic distortion < 3%, frequency=1 kHz, loudspeaker resistance=50Ω.

V_{BB} is adjusted by R_{vbh} or R_{vbl}, see Figure 48. The curves from Figure 25 are measured with the following values for R_{vbh}, R_{vbl} and R_{reg}:

V_{BB}	R_{vbl}	R_{vbh}	R_{reg}	V_{SLPE}
3.0 V	93.1 kΩ	not present	not present	4.4 V
3.6 V	not present	not present	not present	4.4 V
4.5 V	not present	47.4 kΩ	21.0 kΩ	5.7 V
5.0 V	not present	30.1 kΩ	21.0 kΩ	5.7 V

The effect of the dynamic limiter, up to an overdrive of 20 dB, is represented in Figure 26. If the input voltage at LSI becomes larger than 20 mV, the output voltage at QLS is limited, and the distortion increases.

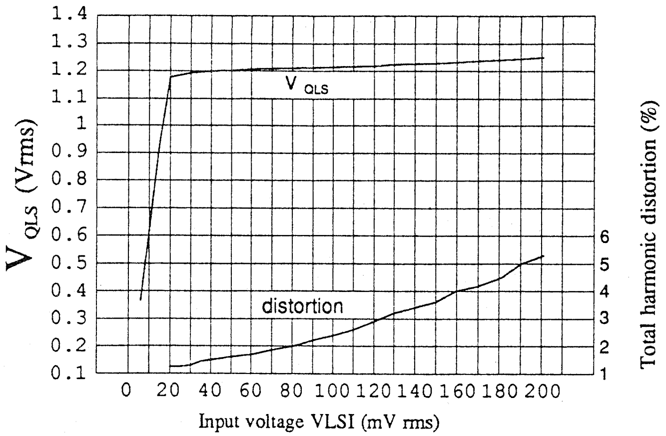


Figure 26 **Loudspeaker amplifier output voltage (V_{QLS}) in 50 Ω and total harmonic distortion versus V_{LSI}. The frequency is 1 kHz, line current is 30 mA**

Dynamic limiter

The dynamic limiter minimizes the distortion of the output signal by reducing the gain of the loudspeaker preamplifier when the output signal starts clipping or when too low supply conditions are detected. The amount of gain reduction is determined by the voltage on pin DLL. This voltage is varied by charging and discharging the capacitor C_{DLL}. The combination of charge and discharge currents and the capacitance value of C_{DLL} sets the timing of the dynamic limiter.

The dynamic limiter consists of a peak limiter, current limiter and V_{BB} limiter.

The peak limiter is activated in case the loudspeaker amplifier output voltage meets V_{BB} while the line current is large enough for not activating the current limiter.

The current limiter is activated in case the loudspeaker amplifier supply current becomes too high in relation with the line current.

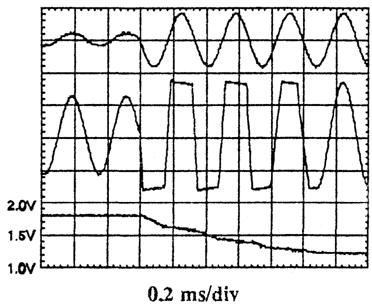
The V_{BB} limiter is activated in case V_{BB} becomes too low. It is an addition to the above mentioned current limiter. The difference is that it acts fast, while the current limiter acts slow, in order to prevent the V_{BB} voltage from dropping. If V_{BB} were to drop below 2.8 V the TEA1096 would not be able to work properly.

The peak limiter, current limiter and V_{BB} limiter affect the voltage over C_{DLL} see Figure 47 and Figure 48. This voltage controls the loudspeaker amplifier gain.

Peak limiter

Clipping of the loudspeaker output signal occurs when this signal reaches V_{BB} as a result of which the output transistors are driven into deep saturation. To prevent hard clipping, deep saturation has to be prevented. The saturation voltage of the output transistors strongly depends on the output current. In the peak limiter of the TEA1096 these effects are taken into account to have maximum output swing under any condition. When the peak limiter detects the beginning of saturation, the capacitor C_{DLL} is discharged with a current of approximately 1 mA. The gain is reduced until the tops of the sines are not flattened any more, as a result of which distortion is decreased, see Figure 27. The loudspeaker amplifier stays in its reduced gain mode until the peaks of the loudspeaker signal no longer start to cause saturation.

The gain then returns to its normal value by charging C_{DLL} see Figure 28. In this way it is ensured that always the maximum reachable output power can be obtained at low distortion.



V_{LSI}
0.1 V/div

V_{QLS}
1 V/div

V_{DLL}
0.5 V/div

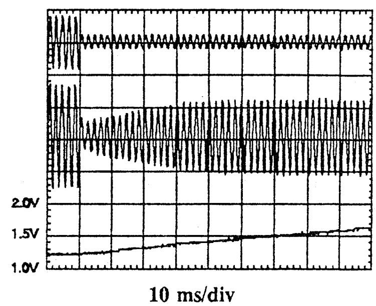


Figure 27 **Attack time of the peak limiter at a line current of 30 mA, and a loudspeaker load of 50Ω. f=3kHz, Cdll=470nF.**

Figure 28 **Release time of the peak limiter at a line current of 30 mA, and a loudspeaker load of 50Ω. f=500Hz, Cdll=470nF.**

Current limiter

Under low supply conditions, the gain of the loudspeaker amplifier is reduced to prevent unsettling the TEA1096 power supply. In case of low line currents the gain of the loudspeaker amplifier is reduced. In this way it is reached that always the maximum output power, corresponding with the available line current, is achieved.

The current limiter is relatively slow, to avoid system instabilities. This slow action can cause V_{BB} drops in case the V_{BB} limiter becomes activated too (see the paragraph 'V_{BB} limiter'). In Figure 29 the effect of the current limiter is shown. The V_{BB} voltage decreases when the limiter is activated and returns to the nominal voltage, in this case after 250 ms. The amplifier reaches its nominal gain factor when the input level is reduced.

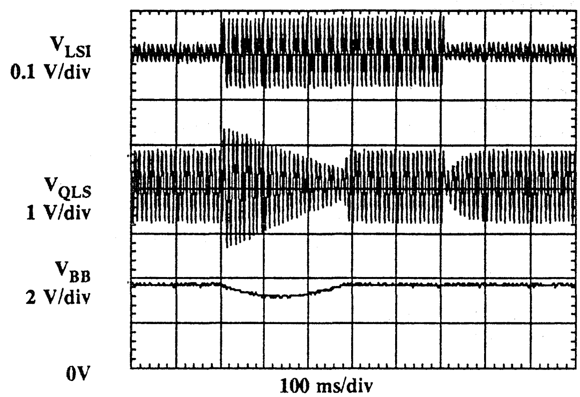


Figure 29 **Operation of the current limiter, correcting VBB at a line current of 16 mA. Cdll=470nF, loudspeaker load is 50Ω.**

V_{BB} limiter

The V_{BB} limiter is an addition to the current limiter to prevent too low V_{BB} levels. A too low V_{BB} level would prevent proper working of the circuit parts supplied by V_{BB}. At low line currents in combination with speech signals containing large amplitudes, the momentary current consumption of the power amplifiers can exceed the supply current, as a result of which V_{BB} would drop due to the discharge of C_{VBB} see Figure 47 and Figure 48. In this case the power consumption of the loudspeaker amplifier has to be diminished instantaneously.

In Figure 30 can be seen the effect of the V_{BB} limiter. When V_{BB} drops below its threshold, V_{QLS} decreases abrupt to a lower value. Directly after reaching the threshold C_{DLL} can be recharged again (by 1 μA), keeping V_{QLS} low. After V_{BB} is increased up to its nominal value (t_2) the current limiter keeps the amplitude of V_{QLS} at an acceptable level: the stable situation between t_2 and t_3. In Figure 30 the input amplitude V_{LSI} is constant between t_1 and t_3.

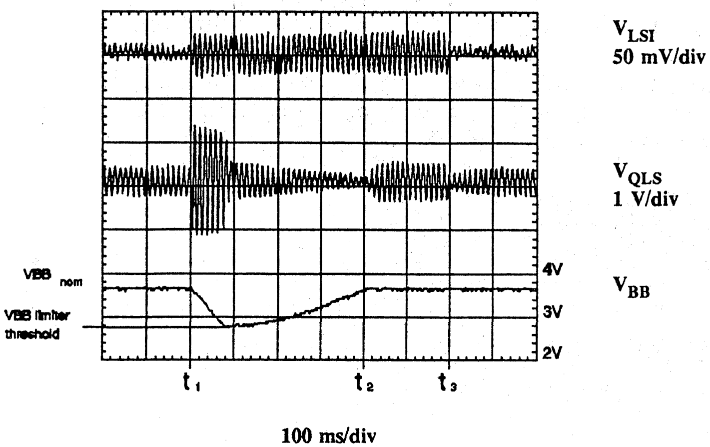


Figure 30 **Effect of the V_{BB} limiter at low supply currents. C_{DLL}=470 nF, line current is 13 mA, loudspeaker load is 50Ω.**

Disable function

The DLL input can also be used to disable the loudspeaker function. If DLL is connected to VEE, for instance by a switch, FET or microcontroller output, the ac loudspeaker output voltage V_{QLS} becomes zero. If an active device with open drain is used for pulling down the voltage at DLL, note that the charge current of C_{DLL} is about 2 μA at nominal V_{DLL} level (1.8V). The leakage current of the driver has to be negligible with respect to the charge current. If DLL is pulled down to VEE the charge current is increased up to 75 μA for fast recovering the normal operating voltage required by the dynamic limiter, see Figure 31 for the exact behaviour.

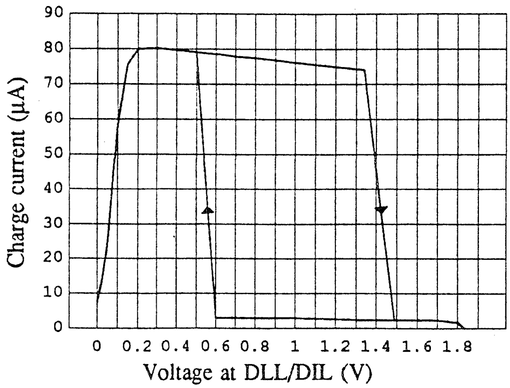


Figure 31 **Hysteresis of the charge current of C_{DLL} versus voltage at DLL/DIL (pin 1). V_{BB}=3.6V.**

3.6 Mute

Electronic switching between dialling and speech can be obtained by controlling the MUTE input. If a high level (>1.5V) is applied to the MUTE input, then both the microphone and receiving amplifier inputs are inhibited and the DTMF input is simultaneously enabled. After dialling the MUTE input can be made low (<0.3V) or left unconnected to disable the DTMF amplifier and regain the connection of the receiving amplifier and the microphone amplifier. The internal switching takes place with negligible clicking at the earpiece outputs and on the line.

In case of low line currents the MUTE function is operational down to a line current of 3.2 mA.

3.7 Power down

The power down input (PD at pin 20) can be used to decrease current consumption of the TEA1096(A). This can be useful in case of pulse dialling, where the telephone line is momentarily interrupted. During these interruptions the power supply is also interrupted and the TEA1096(A) is supplied by the charge available in C_{vbb} and C_{vdd} (Figure 2). The discharge time of these capacitors will be longer if the power-down function is used.

When input PD is made "HIGH" (>1.5V; input current<20μA) the current consumption from C_{vdd} is reduced up to a maximum of 150 μA, and from C_{vbb} up to a maximum of 500 μA. Furthermore, C_{reg} is internally disconnected to prevent discharge of C_{reg} during line interrupts. Result of this is that after each line interrupt, the voltage regulator is able to start without delay at the same DC-line voltage as before the interrupt. This minimizes the contribution of the IC to the shape of the current pulses during pulse dialling. Of course with a highly inductive exchange feeding bridge, the induction coils mainly determine current waveform. Under these conditions the voltage regulator may have some switch-on delay and cause a voltage overshoot at the line connection (LN) of the IC.

3.8 Automatic Gain Control (AGC)

The gain figures of the microphone amplifier and the receiving amplifier which were derived in the preceding chapters are applicable only when the AGC is inoperative, that means with pin AGC open. When the resistor R_{agc} is connected between AGC and VEE, the line current dependent gain control of both microphone amplifier and the receiving amplifier become operative. The DTMF amplifier is not affected.

Below a specific value of line current, $I_{line\text{-}start}$, the gain is equal to the values calculated with the formulas given before (paragraph 3.3 and 3.4). If the current $I_{line\text{-}start}$ is exceeded, the gain of both the controlled amplifiers decreases with increasing DC line current. Gain control stops when another value of line current ($I_{line\text{-}stop}$) is exceeded. The gain control range of both amplifiers is typically 6 dB. This corresponds with a line length of 5km of 0.5mm diameter copper twisted-pair cable with a DC resistance of 176 Ω/km and an average AC attenuation of 1.2 dB/km at 1kHz. The slope of the gain control characteristic has been chosen for an optimum tracking between the line attenuation and the required amplifier gain for a system with a 2x300Ω feeding bridge. In case lines with other parameters are used, a small additional tracking error will be introduced.

R_{agc} has to be optimised in accordance with the cable characteristics, the feeding bridge resistance and the exchange supply voltage. Normally R_{agc} can be read out from Figure 32 up to Figure 35. These figures are valid under the following circumstances:

rectifier:	One diode voltage of the rectifier is 0.7V (see Figure 36)	
cable:	DC resistance:	176Ω/km
	AC attenuation:	1.2 dB/km for 1 kHz.
	capacitance:	38nF/km
exchange:	feeding bridge:	400, 600, 800 or 1000Ω
	supply voltage:	36, 48 or 60V

In case of differences with these parameters see the procedure on page 43.

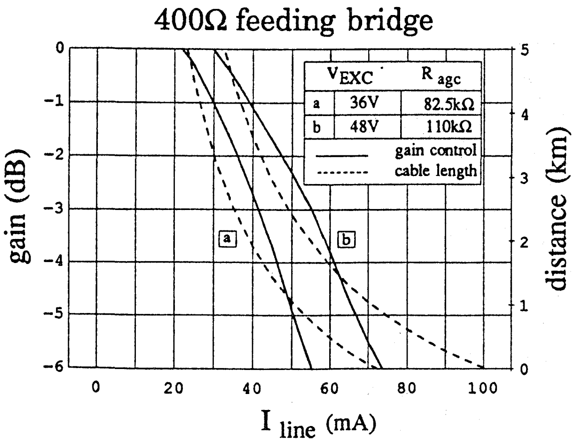


Figure 32 **Gain control characteristics. The resistance of the feeding bridge is 400Ω.**

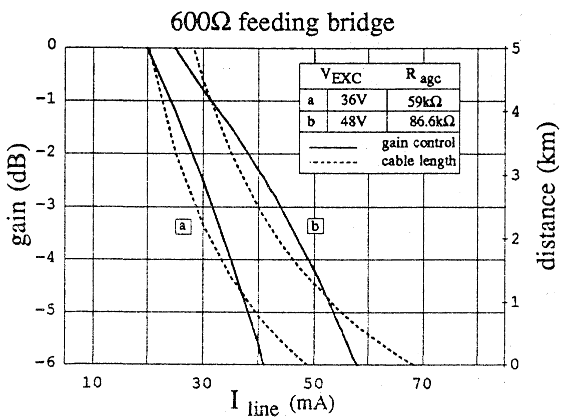


Figure 33 Gain control characteristics. The resistance of the feeding bridge is 600Ω.

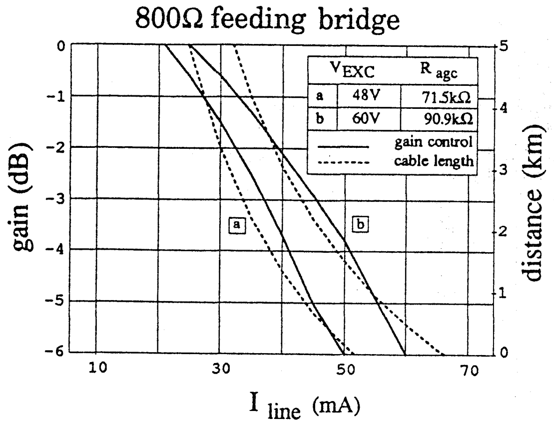


Figure 34 Gain control characteristics. The resistance of the feeding bridge is 800Ω.

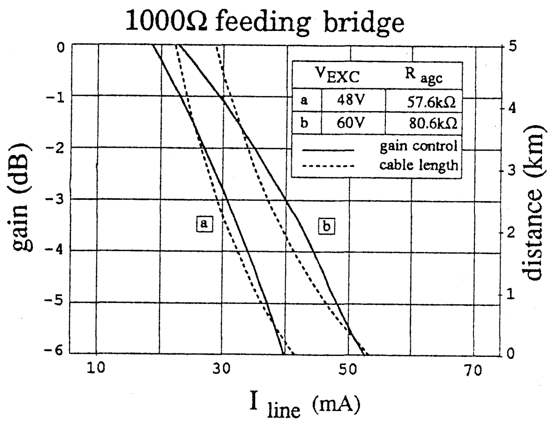


Figure 35 Gain control characteristics. The resistance of the feeding bridge is 1000Ω.

Figure 32 up to Figure 35 can be resumed by the following table:

Table I **R_{agc} for several values of the exchange supply voltage and feeding bridge resistance.**

R_{agc}				
	R_{exc}			
V_{exc}	400 Ω	600 Ω	800 Ω	1000 Ω
36 V	82.5 kΩ	59 kΩ		
48 V	110 kΩ	86.6 kΩ	71.5 kΩ	57.6 kΩ
60 V			90.0 kΩ	80.6 kΩ

In case of differences with the nominal line- and exchange values mentioned on page 39, R_{agc} can be determined in the following way:

First I_{line} has to be determined, see Figure 36.

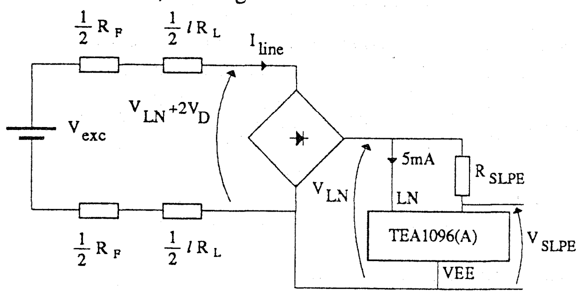


Figure 36 **Determining the line current to find the right R_{agc}.**

$$V_{LN} = V_{SLPE} + (I_{line} - 5mA)R_{SLPE}$$

$$I_{line} = \frac{V_{exc} - V_{LN} - 2V_D}{R_F + l\,R_L}$$

Substituting V_{LN}:

$$I_{line} = \frac{V_{exc} - V_{SLPE} - 2V_D + 5mA\;R_{SLPE}}{R_F + l\,R_L + R_{SLPE}}$$

Where: V_D =one diode voltage of the rectifier (V)
V_{exc} =exchange supply voltage (V)
R_F =feeding bridge resistance (Ω)
R_L =DC line resistance (Ω/km)
l =line length (km)
R_{SLPE} =20Ω

Similar with this is the equation:

$$l = \frac{V_{exc} - V_{SLPE} - 2V_D + 5mAR_{SLPE}}{R_L I_{line}} - \frac{R_F + R_{SLPE}}{R_L}$$

l can be drawn as a function of I_{line}. The attenuation of the AGC can also be drawn as a function of I_{line} in the same figure, for several values of R_{agc}. Choose the R_{agc} which has the curve that approximates at most the line attenuation see Figure 32 up to Figure 35 for the nominal values.

If AGC is not wanted, pin AGC of the TEA1096(A) can be left open to disable the AGC function. The amplifiers give their maximum gain and furthermore the double side tone principle is not longer active. Only one network is used. Pins BAL1 and BAL2 must then be shorted together.

3.9 Anti side tone circuit

Before the line signal reaches the earpiece, the attenuated microphone signal is subtracted from it, to suppress side tone. Attenuation is chosen in such a way that the microphone signal is the smallest possible part of the earpiece signal. Optimum side tone suppression depends on the line impedance. Because line impedance varies with the distance between subscriber and exchange the side tone suppression circuit of the TEA1096(A) consists of two parts: a part for long lines and a part for short lines. The TEA1096(A) chooses, depending on the available line current, the side tone suppression circuit that suppresses the side tone at most, with a smooth turnover between them. Balancing between the two inputs is controlled continuously by means of the AGC function. The take-over region is at 3 dB attenuation of the AGC curves.

Connection of the side tone suppression networks

The earpiece amplifier input is ILS.

The microphone signal is available at pin OSP. At this point the two side tone suppression networks are to be connected. The appropriate attenuated microphone signals have to be offered to the inputs BAL1 and BAL2. BAL1 is intended for the short line network, and BAL2 for the long line network.

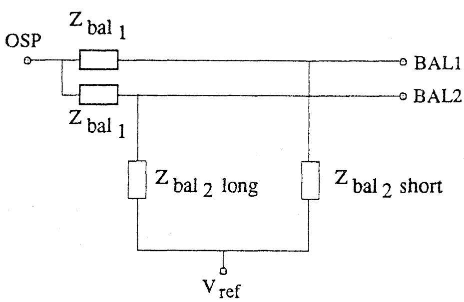


Figure 37 Configuration of the side tone suppression network.

Figure 37 shows the anti side tone networks that have to be connected to BAL1 and BAL2. The inputs BAL1/BAL2 are inputs of differential amplifiers. To the opposite inputs of the differential amplifiers the line voltage, divided by a certain factor, is internally offered. Optimum side tone suppression is achieved if both voltages are equal. According the architecture of the TEA1096(A) the following requirement must be met:

$$\frac{Z_{BAL_2}}{Z_{BAL_1}+Z_{BAL_2}}=\frac{Z_{line}}{2(Z_{set}+Z_{line})}$$

Where: $Z_{BAL2}=Z_{BAL2\ long}$ or $Z_{BAL2\ short}$
$Z_{set}=0.1 \times Z_{simp}$ = (complex) set impedance (for audio frequencies)
Z_{line} is the (complex) line impedance.

To determine Z_{simp} see page 16, paragraph: 'active set impedance'.

With: $Z_{BAL2} = \frac{1}{2}Z_{line}$ it follows that: $Z_{BAL1} = Z_{set} + \frac{1}{2}Z_{line}$

The circuit of Figure 37 becomes:

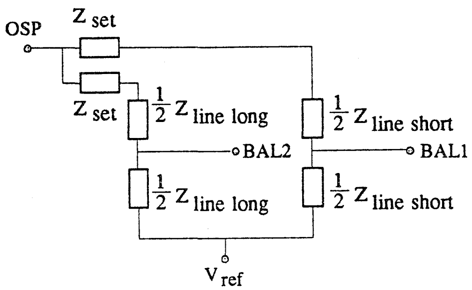


Figure 38 **Optimum side tone suppression.**

The circuit of Figure 39 is equivalent with the circuit of Figure 38. All impedances are multiplied by a factor k to have $kZ_{line\ long}$ or $kZ_{line\ short}$ in the order of 10 kΩ.

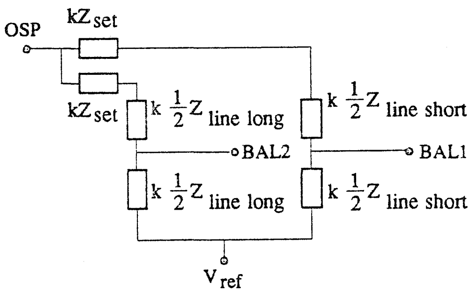


Figure 39 **The side tone suppression circuits multiplied by a constant factor k.**

kZ_{set} represents a scaled value of the set impedance.

$kZ_{line\ long/short}$ has to be calculated from the line impedance corresponding with a specified line length.

Z_{line} varies strongly with line length and line type. A value for $Z_{line\ long/short}$ have to be chosen that corresponds with an average line length giving satisfactory side tone suppression for long and short lines. The components of $Z_{line\ long/short}$ can be calculated as follows:

Suppose Z_{line} is represented by:

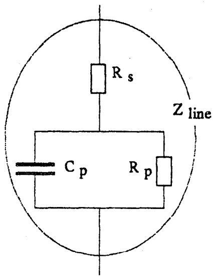


Figure 40 Circuit equivalent of a complex line impedance.

Figure 42 is an equivalent configuration for Figure 40, but is more easy to use when you want to realize the voltage divider of Figure 41 (see also Figure 38).

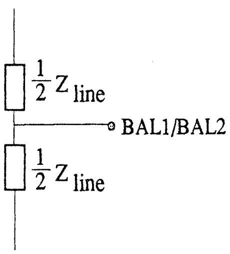


Figure 41 Voltage divider for optimum side tone suppression.

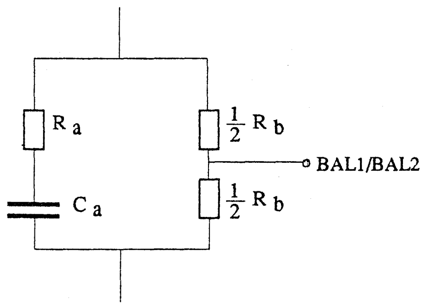


Figure 42 Circuit equivalent of complex line impedance combined with voltage divider to meet side tone suppression requirements.

Where:

$$R_b = (R_s + R_p) \qquad R_a = R_s + \frac{R_s^2}{R_p}$$

$$C_a = \frac{R_p \times C_p}{2R_s + R_p + \frac{R_s^2}{R_p}}$$

Choosing short-long line impedances ($Z_{line\ short}/Z_{line\ long}$)

Example:

The side tone networks have to be optimised for Z_{set} = 220Ω+820Ω//115nF and for a cable with the following characteristics:

core diameter of the line:	0.5 mm
resistance:	176 Ω/km
capacitance:	38 nF/km
average attenuation:	1.2 dB/km (1kHz)
length:	between 0 km and 10 km

The exchange supply voltage is 48 V, and the feeding bridge is 600 Ω. According table I a R_{agc} of 86.6 kΩ is chosen. The take over current (reached if AGC attenuates 3 dB) is about 45 mA (see Figure 33), and agrees with 2.5 km cable length.

To compensate the long lines, a Z_{line} is chosen corresponding with a line length of 7 km, while for the short lines a Z_{line} is taken corresponding with a line length of 1 km.

The tables below show the equivalent component values of $Z_{line\ long}$ and $Z_{line\ short}$, the calculated values according Figure 42 and the applied component values of the side tone networks using a scaling factor k = 10.

7 km line equivalent		
$Z_{line\ long}$	calculated values	applied values k = 10
(Figure 40)	(Figure 42)	
R_s = 297 Ω	½R_b = 928 Ω	k · ½R_b = 9.09 kΩ
R_p = 1560 Ω	R_a = 353 Ω	k · R_a = 3.65 kΩ
C_p = 174 nF	C_a = 123 nF	C_a/k = 12 nF

1 km line equivalent		
$Z_{line\ short}$	calculated values	applied values k = 10
(Figure 40)	(Figure 42)	
R_s = 55 Ω	½R_b = 392 Ω	k · ½R_b = 3.92 kΩ
R_p = 730 Ω	R_a = 59 Ω	k · R_a = 591 Ω
C_p = 31 nF	C_a = 31 nF	C_a/k = 3.3 nF

The components of $k{\cdot}Z_{set}$ (Figure 39) are given by 2.2 kΩ + 8.25 kΩ // 12 nF.

In case of PABX systems with relative short line lengths, the side tone network can consist of only one network.

When AGC is not required or not allowed (R_{agc} removed), the balance network can be connected to BAL1 or BAL2, while BAL1 and BAL2 are short-circuited. The balance network can also be connected to BAL2 while BAL1 is connected to VREF. If R_{agc} is removed there is no more switching.

4 Application example of the TEA1096(A).

Figure 49 shows an application example of the TEA1096(A) in combination with the PCD3330-1 as multi-standard repertory dialler/ringer with EEPROM. For more details of the PCD3330-1 is referred to ref.3. The circuit has the following features:

1. Speech/transmission
2. Listening-in/call progress monitoring
3. Ringer generation and detection with volume control
4. dialling features including:
   -pulse, DTMF and mixed mode dialling
   -Last number redial
   -Repertory dialling
   -On-hook dialling

Three important parts of the application can be distinguished:

1. Transmission part with TEA1096
2. Microcontroller with EEPROM and key board
3. Ringer part.

4.1 Transmission part with TEA1096(A)

DC settings

In Figure 43 the dc settings can be seen as function of the line current (see also Figure 2). V_{SLPE} and V_{BB} are not adjusted: the nominal value is shown. I_{LI} represents the current which is available for powering the loudspeaker amplifier.

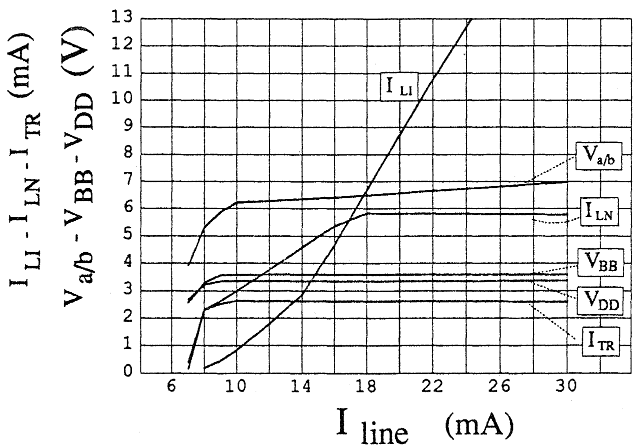


Figure 43 I_{LI}, I_{LN}, I_{TR}, $V_{a/b}$, V_{BB}, V_{DD} versus line current.

I_{LI} depends on the line current. At low line currents the highest priority is given to the normal transmission functions (send and receive). This means that below 8 mA line current listening-in is disabled, while the normal transmission functions are operational down to about 7 mA.

The value of $V_{a/b}$ includes the voltage over the bridge and line interrupter. Therefore the minimal $V_{a/b}$ voltage is 3.9V because the minimal micro controller supply voltage is 1.8V.

Start up

The smoothing capacitors charge-up time, after going off-hook, for C_{VDD} (100µF) is 200 ms and for C_{VBB} (470µF) 450 ms, at a line current of 20 mA.

The transmission functions sending and receiving are available at 70 ms after going off-hook, at a line current of 20 mA.

Set impedance

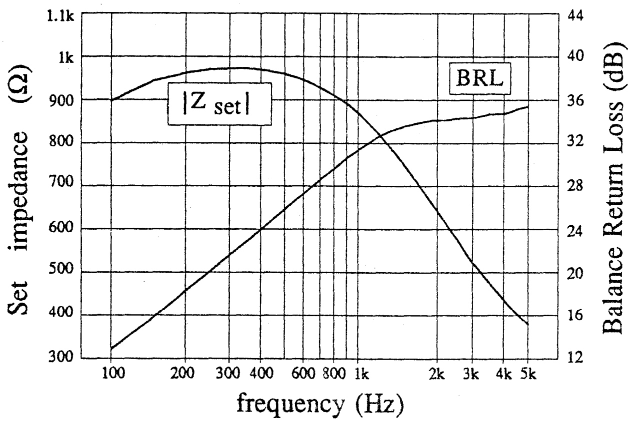


Figure 44 **Balance return loss for a complex impedance (220Ω + 825Ω//115nF) at a line current of 20 mA.**

In Figure 44 the balance return loss and the absolute value of Z_{set} are shown for a complex set impedance. The BRL is more than 20 dB for the audio frequency range. Real as well as complex set impedances can be realized by means of the network between SIMP and VREF

DTMF/pulse dialling

The DTMF/pulse dialling signals are generated by the PCD3330-1, and connected to the appropriate inputs of the TEA1096. The DTMF signal, measured at the line, is adjusted by R37 at a level of -10dBm for the low frequency part and at -8 dBm for the high frequency part. During dialling (pulse or DTMF), in this application, the MUTE input is activated (made HIGH) so that the receiving amplifier inputs and the microphone inputs are inhibited, while DTMF is simultaneously enabled. After dialling the MUTE input is made low, disabling the DTMF amplifier and enabling the receiving- and microphone amplifiers. In Figure 45 the case of pulse dialling is shown. At the moments that the line is interrupted, PD input is activated as result of which power consumption of the TEA1096(A) is reduced at this moments. Therefore the influence of the power breaks on V_{BB} and V_{DD} is small.

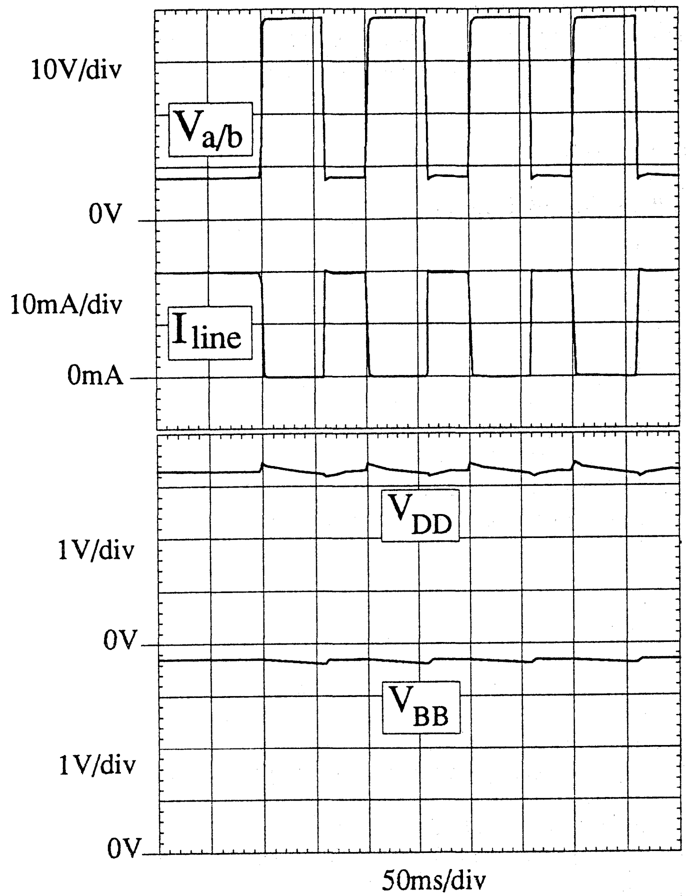


Figure 45 **Line current, line voltage and supply voltages during pulse dialling.**

Because the line interrupter is slower than the PD circuit, and the line interrupter is activated simultaneously with the PD input, small increases are present on V_{DD} just before the line is interrupted.

On-hook dialling

The TEA1096(A) can be applied in on-hook dialling applications where the line signal is monitored, via the loudspeaker, during pulse dialling. The LI circuit is set in the line monitoring mode before dialling starts and must remain in this mode during the on-hook dialling procedure.

4.2 Microcontroller with EEPROM and key board

The microcontroller configuration provides the following features:

1. Pulse dialling.
2. DTMF dialling.
3. Mixed mode dialling.
4. Flash or register recall.
5. Connect a/b line to earth function.
6. Mute function.
7. Disconnect function.
8. Standard 4x4 keyboard for:0 to 9 and *,#,A,B,C and D
9. Function keys for: Flash, Hook, Mute, Tone, Disconnect, LNRedial, Memory recall, Store, Access Pause, 1 Key repertory dialling and Program Ringer.
10. On-hook dialling control.
11. Redial Cursor method (maximum 24 digits) stored in internal EEPROM.
12. Storage for 13 repertory dial numbers (16 digits each) or 10 repertory dial numbers (20 digits each) in internal EEPROM.
13. Access pause generation and termination: manually or by Atlanta procedure.
14. Ringer input frequency detection.
15. Three-tone ringer with 4 different ringer sequences.
16. Ringer melody generation with four signal speeds and four output volume steps, keypad controlled.
17. Country specifications which can be stored in EEPROM are:
    - Will * and/or # be transmitted when switching over to DTMF dial mode.
    - Mark to space ratio (3:2 or 2:1).
    - 6 Tone time selections (60/90,70/70, 80/80, 100/100, 100/140 or 140/140 ms).
    - 4 Flash time selections (100, 115, 270 or 600 ms).
    - Mute output type selection (M1, $\overline{M1}$, M2 or $\overline{M2}$).
    - Dial Pulse output selection (DP or $\overline{DP}$).
    - DTMF keys or Function keys selection.
    - Access Pause time selection (1.5/1.0, 2.5/1.5, 3.0/3.5 or 6.0/6.0 s).
    - 10 Number repertory dialler selection (1 or 2 keys).

- Two repertory number programming procedures (General or Germany).
- Repertory length 16 or 20 digits.
- Ringer input frequency detection selection.
- Ringer output selection (via DTMF or special -RT- output).
- 4 Possible ringer melodies.
- 4 Possible ringer repetition rates.
- 4 Possible ringer volumes.

Pulse dialling or DTMF is selected by jumper J_2.

4.3 Ringer part

The ringer part can be divided into three parts:

Supply:
When a ringer signal is present on the a/b lines, this ac signal supplies the ringer hardware via capacitors C_4 and C_{28}, zener diodes D_{23} and D_{24}, resistor R_6 and diode bridge D_8, D_{11}, D_{12} and D_{25}. Capacitors C_4 and C_{28} block the dc current flow via the ringer hardware, two capacitors are necessary because in this on-hook dialling application both connections can give dc current flow. Zener diodes D_{23} and D_{24} are necessary because otherwise the ringer hardware short-circuits the ac signals (DTMF, speech) of the TEA1096(A). Resistor R_6 is present to make the total impedance ok. Finally there is a diode bridge D_8, D_{11}, D_{12} and D_{25} to rectify the ac ringer signal.
The PCD3330-1 is supplied via the voltage divider network D_{10} and D_{17}. Zener diode D_{10} together with zener diode D_{17} keep the maximum voltage across the ringer output stage below 28V and across the PCD3330-1 below 6V.

Ringer frequency measurement:
To activate the ringer frequency measurement, the ringer input frequency has to be supplied to the CE/RF input of the PCD3330-1. This is done by connecting the CE/RF input of the PCD3330-1 to one of the ac connection points of the ringer rectifier bridge (here at diodes D_{11} and D_{12}) via resistor R_{14}. When the ringer input frequency is between the ringer detection LOW and HIGH, the PCD3330-1 will go to the ringer melody generation mode.

Ringer output stage:
The ringer melody is amplified via T_5 and T_4 and supplied to the buzzer. The output volume control is achieved with T_6 and T_7 which are controlled from PCD3330-1 outputs RVOL1 and RVOL2.

4.4 Application hints

Howling limiter

Howling can occur when the handset microphone is close to the loudspeaker while the listening-in circuit is active.

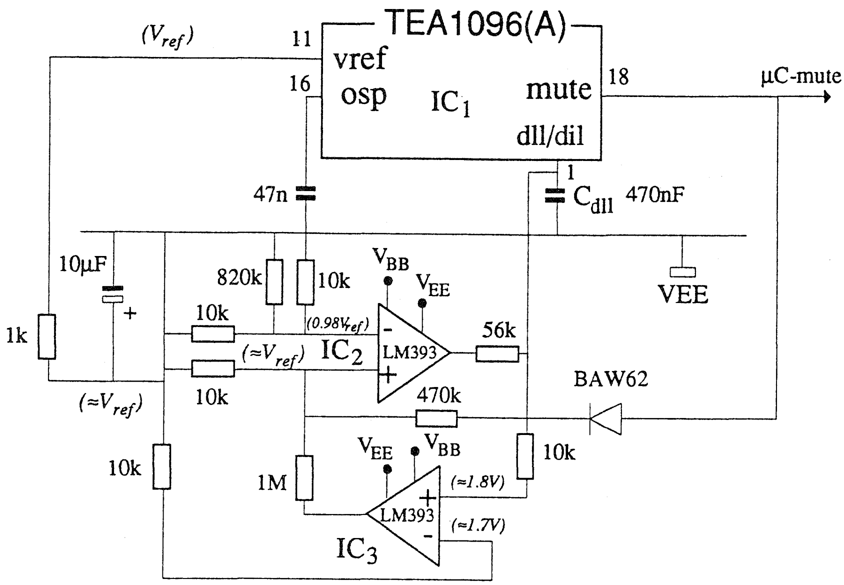


Figure 46 A proposal for a howling limiter.

A proposal of how to implement a howling limiting circuit with the TEA1096(A) is given in Figure 46.

The dc settings of several points, if the circuit is not active, are given between brackets. The circuit becomes activated only if a large amplitude of the microphone signal is detected. The amplified microphone signal is present at pin OSP. MUTE is an input that is driven by a not shown logic circuit, for instance a microcontroller. MUTE is LOW during speech. If there is no howling, the situation shown in Figure 46, IC_3 and the MUTE-line do not affect the output of IC_2. The input voltage at IC_2 equals:

$$[1-0.98]\ x\ V_{ref} = [1-0.98]\ x\ \frac{1}{2}V_{DD} \approx 0.02\ x\ 1.7V \approx 34mV$$

If the signal on output OSP causes a signal at the minus input of IC_2 that exceeds this 34 mV, the O.C. output of IC_2 becomes LOW and discharges C_{dll} via 56kΩ. The voltage over C_{dll} decreases and the peaklimiter decreases the loudspeaker output voltage. When the voltage over C_{dll} becomes smaller than V_{ref} (≈1.7V) the output of IC_3 becomes LOW, causing hysteresis for IC_2.

The MUTE line is HIGH during DTMF dialling. This causes a higher voltage on the plus input of IC_2, which inhibits the output of IC_2 to become LOW during DTMF dialling. Otherwise the circuit should recognize the DTMF signal as howling, as result of which the DTMF signal should be attenuated.

5 References

[Ref.1] Specification TEA1096/TEA1096A

[Ref.2] User's manual of demo board PR46051 transmission/listening-in with TEA1096(A)
Author: F. van Dongen
report number: ETT/AN93005

[Ref.3] Objective Specification of PCD3330-1 of January '93
A multi-standard repertory dialler/ringer with EEPROM

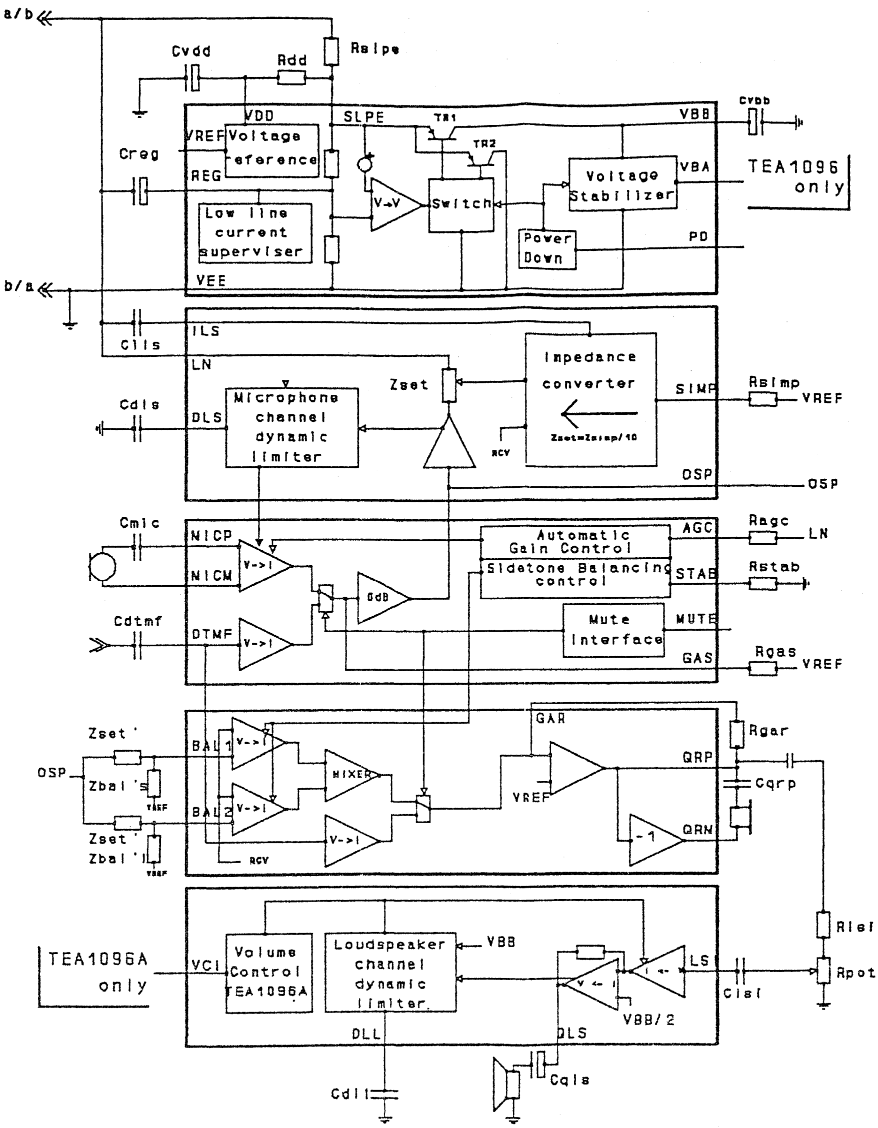


Figure 47 block diagram TEA1096 and TEA1096A

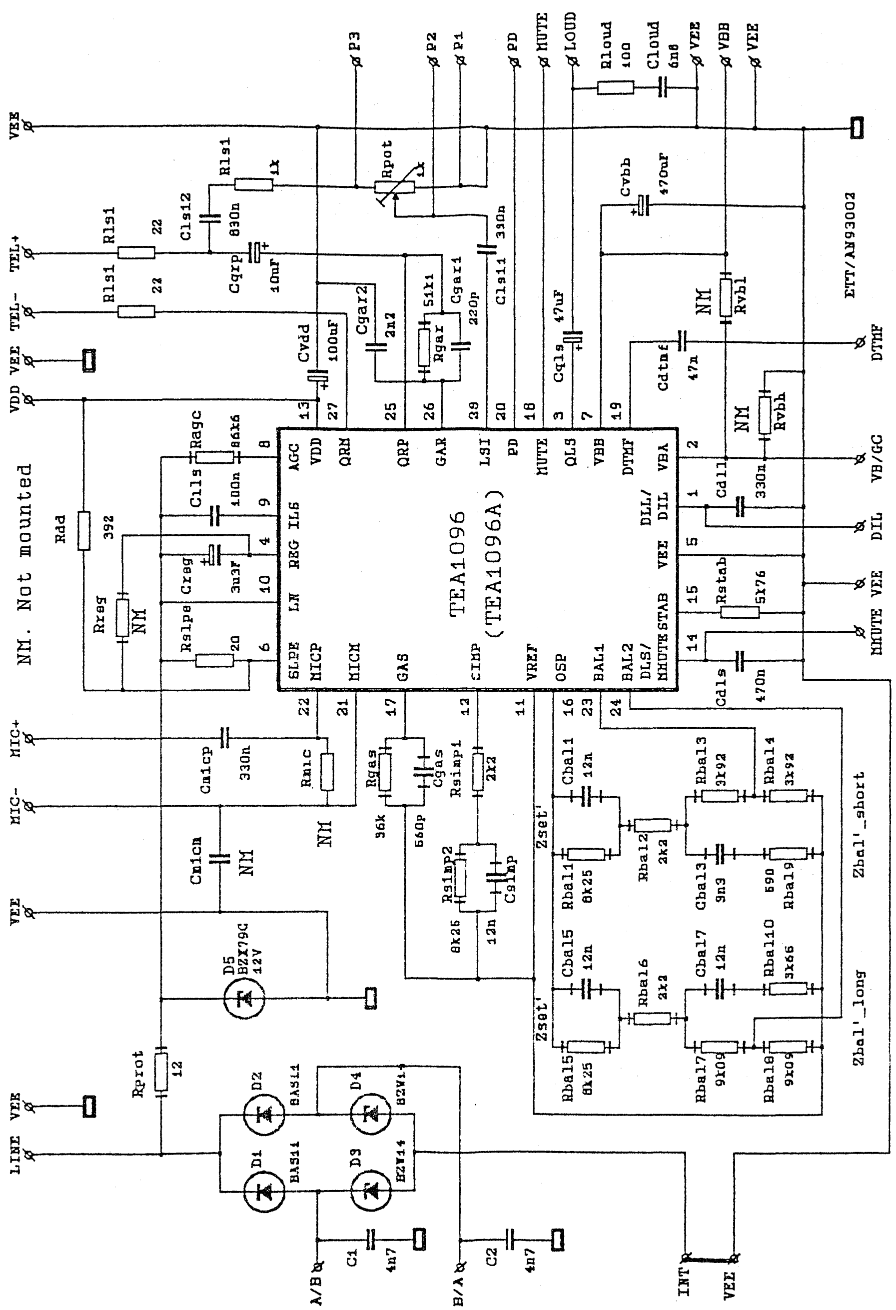


Figure 48 Circuit diagram of printed circuit board PR45951

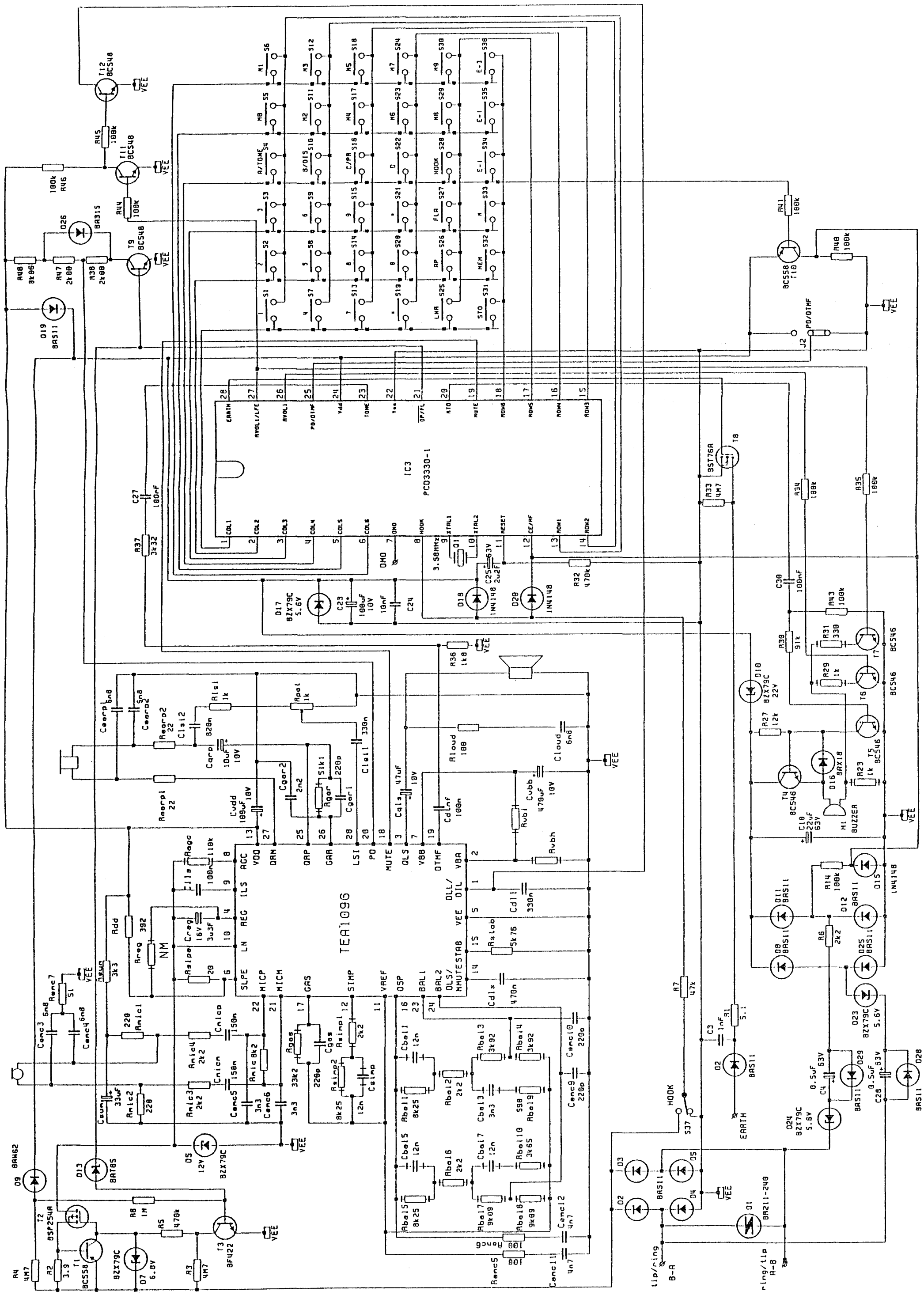


Figure 49 Application of the TEA1096 with the microcontroller PCD3330-1.

APPLICATION NOTE Nr	ETT/AN93015 (Supersedes report of November 1993 with same Nr)
TITLE	Application of the TEA1093 handsfree circuit
AUTHOR	R. van Leeuwen
DATE	November 1995

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1. INTRODUCTION

The TEA1093 is a circuit which, in combination with a member of the TEA106x/TEA111x or PCA1070 transmission circuits, offers a handsfree function. For a description of the combination of the TEA1093 and PCA1070 is referred to [Ref.8]. It incorporates a supply, a microphone amplifier, a loudspeaker amplifier and a duplex controller with signal and noise monitors on the transmit and the receive channel. The function of the handsfree circuit will be illustrated with the help of figure 1.

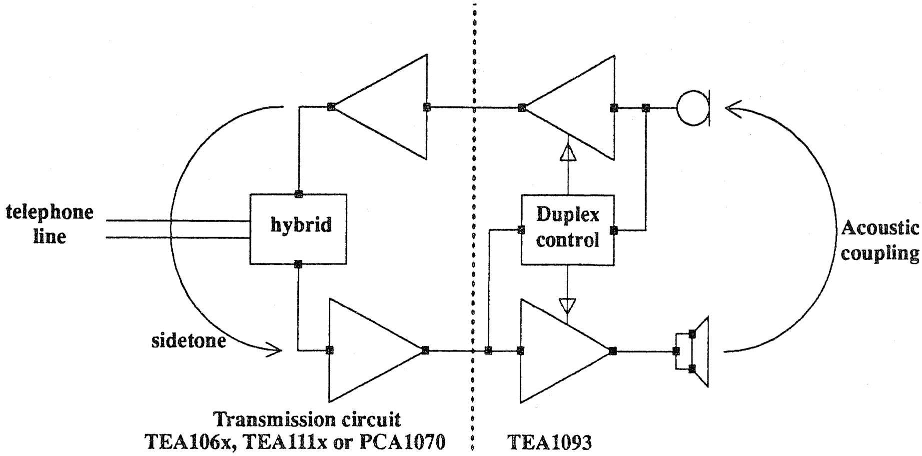


Fig.1 Handsfree telephone set principle.

The left side of figure 1 shows a principle diagram of a part of the TEA106x/TEA111x/PCA1070 circuit by means of a receiving amplifier for the earpiece, a transmit amplifier for the microphone and a hybrid. The right side of figure 1 shows a principle diagram of a part of the TEA1093 handsfree circuit by means of a microphone amplifier, a loudspeaker amplifier and a duplex controller.

As can be seen from figure 1, a closed loop is formed via the amplifiers, the anti side-tone network and the acoustic coupling between loudspeaker and microphone of the handsfree circuit. When the loop gain is higher than one, the set starts howling. In a full-duplex application, this would be the case. To avoid howling, the loop-gain has to be much lower than one and therefore has to be decreased. This is done by the duplex controller.

The duplex controller of the TEA1093 monitors the signal and noise on both the transmit and the receive channel in order to detect which channel contains the 'largest' signal. As a result the duplex controller reduces the gain of the channel which contains the 'smallest' signal. This is done such that the sum of the transmit and receive gain remains constant.

As a result, the circuit can be in three stable modes to be referred to throughout this report:

1. Transmit mode (Tx-mode): the gain of the microphone amplifier is at its maximum and the gain of the loudspeaker amplifier is reduced.
2. Receive mode (Rx-mode): the gain of the loudspeaker amplifier is at its maximum and the gain of the microphone amplifier is reduced.
3. Idle mode (Ix-mode): the gain of the amplifiers are halfway their maximum and reduced value.

The difference between the maximum gain and the reduced gain is called the switching range.

This report gives a detailed description of the TEA1093 and its application with a transmission circuit. This transmission circuit can be a member of the TEA106x- and TEA111x-family or the PCA1070 for high end telephone sets. The description is given by means of the block diagram of the TEA1093 (ch. 2 on page 9) and by discussing every detail of the sub-blocks (ch.3 on page 12). The application is discussed by giving a guideline for application (the application cookbook ch.4 on page 36) and by giving a number of worked-out applications including listening-in (ch.5 on page 50). For application of the TEA1093 combined with the PCA1070 is referred to [Ref.8]. Also EMC aspects are discussed (ch.6 on page 59). Finally a chapter with a quick reference data of the TEA106x family (ch.7 on page 64) and a list of abbreviations (ch.8 on page 64) are provided.

2. Block diagram

In this chapter the block diagram of the TEA1093 is shown by means of figure 2. The pinning of the TEA1093 is given by means of figure 3. Also a short description of the block diagram is given including the function of the external components.

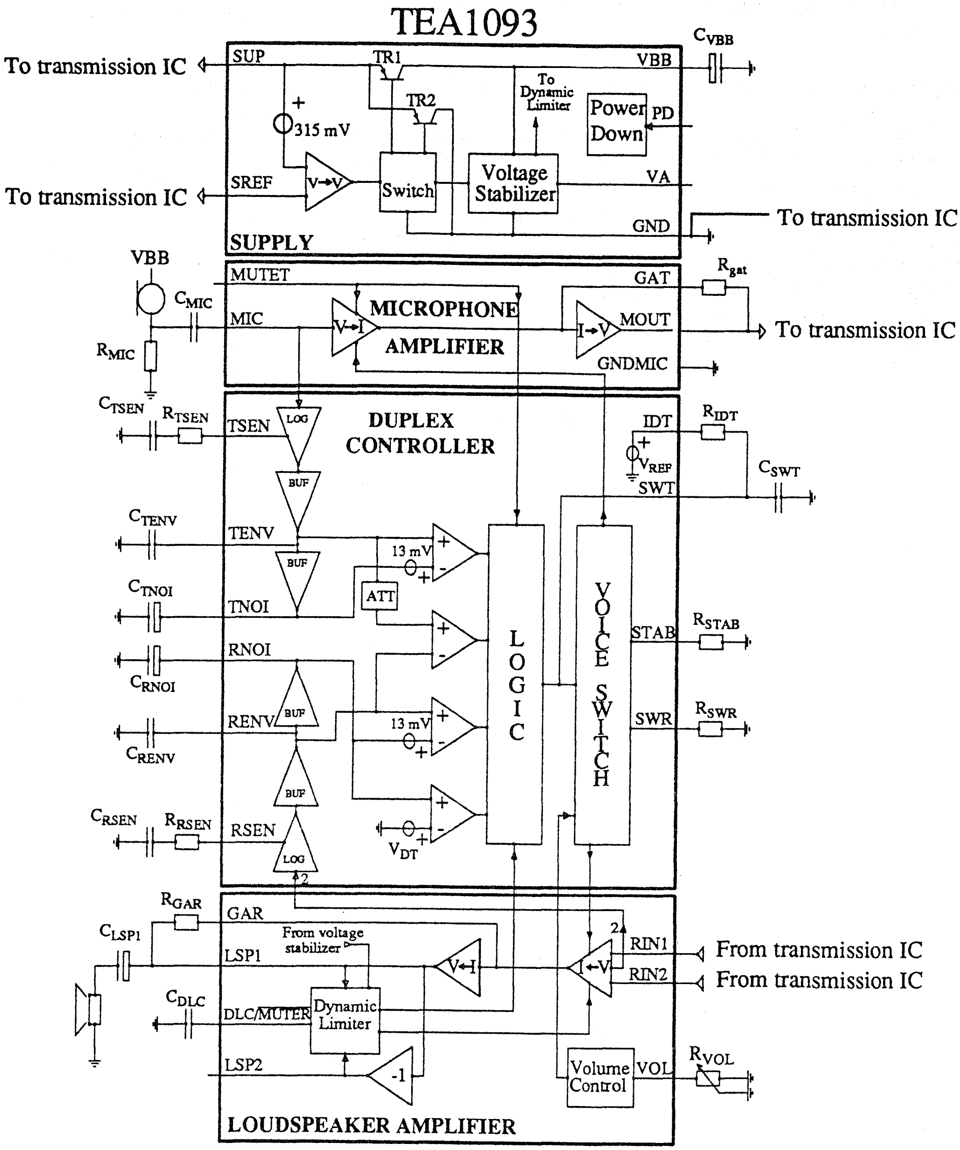
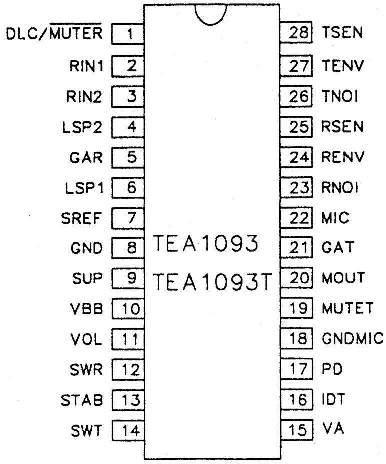


Fig.2 Block diagram of TEA1093



Pin	Name	Description
01	DLC/$\overline{MUTER}$	Dynamic limiter timing adjustment, receiver channel mute input
02	RIN1	Receiver amplifier input 1
03	RIN2	Receiver amplifier input 2
04	LSP2	Loudspeaker amplifier output 2
05	GAR	Receiver gain adjustment
06	LSP1	Loudspeaker amplifier output 1
07	SREF	Supply reference input
08	GND	Ground reference
09	SUP	Supply input
10	VBB	Supply output
11	VOL	Receiver volume adjustment
12	SWR	Switching range adjustment
13	STAB	Reference current adjustment
14	SWT	Switch-over timing adjustment
15	VA	VBB voltage adjustment
16	IDT	Idle-mode timing adjustment
17	PD	Power down input
18	GNDMIC	Ground reference for microphone amplifier
19	MUTET	Transmit channel mute input
20	MOUT	Microphone amplifier output
21	GAT	Microphone gain adjustment
22	MIC	Microphone input
23	RNOI	Receive noise envelope timing adjustment
24	RENV	Receive signal envelope timing adjustment
25	RSEN	Receive signal envelope sensitivity adjustment
26	TNOI	Transmit noise envelope timing adjustment
27	TENV	Transmit signal envelope timing adjustment
28	TSEN	Transmit signal envelope sensitivity adjustment

Fig.3 Pinning of the TEA1093

In figure 2 it can be seen that the IC consists out of four parts: the supply, the microphone amplifier, the loudspeaker amplifier and the duplex controller. These blocks will be shortly described below including the function of the external components. The detailed description will follow in chapter 3 on page 12.

Supply:

Via pins SUP and SREF the supply can be connected to the transmission IC, see chapter 3.1 on page 12. The resistor R_{SREF} connected between SUP and SREF then determines the bias current of the output stage of the transmission IC. A capacitor Csref connected to pin SREF is then needed for stability. A stabilized supply voltage is available between pins VBB and GND. This voltage can be increased by connecting a resistor Rva between pins VA and GND. The voltage can be decreased by connecting Rva between pins VA and VBB. Capacitor C_{VBB} serves as a buffer. By making pin PD high, the TEA1093 can be put in power down. The reference ground pin (GND) has to be connected the transmission pin SLPE in case of TEA106x or TEA111x and to pin VSS in case of the PCA1070.

Microphone amplifier:

The handsfree microphone signal is amplified from pin MIC to pin MOUT. The signal reference is pin GNDMIC, a 'clean ground' which has to be connected to GND. The sensitivity of the microphone can be set via Rmic, the signal is coupled in via Cmic. The gain of the amplifier can be set with Rgat. The amplifier can be muted by making pin MUTET high.

Loudspeaker amplifier:

A loudspeaker can be connected between output pin LSP1 and GND (single ended load) or between pin LSP1 and the inverted output pin LSP2 (bridge tied load). Capacitor Clsp1 is used to block DC. The gain from the symmetrical input RIN1 and RIN2 to the outputs can be set via resistor Rgar. The volume of the receive signal can be adjusted by means of a potentiometer Rvol. To minimize distortion of the receive signal a dynamic limiter is incorporated of which the timing can be set with the capacitor Cdlc. The amplifier can be muted by making pin DLC/$\overline{\text{MUTER}}$ low.

Duplex controller:

From both the transmit and receive signal, signal and noise envelopes are made. The transmit signal envelope is on pin TENV, the receive signal envelope on pin RENV. The transmit noise envelope is on pin TNOI and the receive envelope noise is on pin RNOI. The timing of the envelopes can be set by the capacitors Ctenv, Ctnoi, Crenv and Crnoi. The sensitivity of the envelope detectors can be set by means of the RC-combinations Rtsen with Ctsen for the transmit envelopes and by Rrsen with Crsen for the receive envelopes. The resistor sets the sensitivity and the capacitor blocks the DC-component. Also a high-pass filter is created.

The logic determines to which mode (Tx, Rx or Ix-mode) the set has to switch over. The timing for switching to the Tx and the Rx-mode is set with capacitor Cswt. The timing for switching to the Ix-mode is set by the combination of Cswt and Ridt. The switching range is set by the resistor Rswr. Resistor Rstab has a fixed value.

3. Description of the TEA1093

This chapter describes in detail the four blocks of the handsfree circuit TEA1093: the supply (3.1 on page 12), the microphone amplifier (3.2 on page 16), the loudspeaker amplifier (3.3 on page 20) and the duplex controller (3.4 on page 28). For each block the principle of operation is described and its adjustments and performance are discussed.

All values given in this chapter are typical and at room temperature unless otherwise stated.

3.1 Supply block

In the first paragraph of this chapter the principle of operation of the supply is described as well as its adjustments, performance and start-up behaviour. In the second paragraph, the power down function is discussed.

3.1.1 Supply

Principle of operation

In figure 4 the block diagram of the supply of the TEA1093 is depicted together with the supply interconnection with the TEA106x/TEA111x.

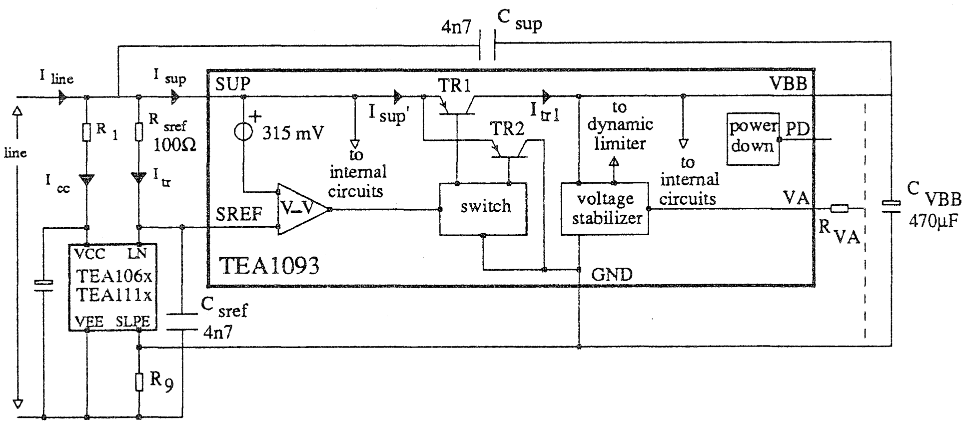


Fig.4 Supply arrangement TEA1093

The supply block of the TEA1093 splits up the line current into a constant part for the transmission circuit and the current I_{SUP}. A current switch controls transistors TR1 and TR2 such that the maximum available current is diverted to VBB. TR2 provides a constant impedance between SUP and SREF in case the voltage on SUP becomes below VBB. This can occur if a large speech signal is present on SUP. The voltage stabilizer generates a DC voltage between VBB and GND.

By connecting pin SUP to the line and pin SREF to pin LN of the TEA106x/TEA111x, a bias current for the output stage of the TEA106x/TEA111x (I_{tr}) is forced into pin LN of:

$I_{tr} = V(SUP\text{-}SREF)/R_{SREF}$.

With V(SUP-SREF) ≈ 315 mV and R_{SREF} = 100 Ω, the current I_{tr} will be approximately 3.15 mA. With a current I_{cc} of approximately 1 mA, the current I_{SUP} flowing into pin SUP follows as:

$$I_{SUP} = I_{line} - I_{cc} - I_{tr} \approx I_{line} - 4 \text{ mA}.$$

Around 0.4 mA of this current is used to supply internal circuitry. The rest of the current (I_{SUP}') is processed by the current switch of the supply block.

In order to create a sending signal on the line, the TEA106x/TEA111x modulates the current through the slope-resistor R9. In applications of the TEA106x/TEA111x without the TEA1093, this results in a modulated current through pin SLPE and thus LN. In applications of the TEA106x/TEA111x with the TEA1093, the current through LN is kept constant and therefore cannot be modulated. However, because the TEA1093 is connected between pin SLPE of the TEA106x/TEA111x and the line, the current through the TEA1093 will be modulated. As a result, the transmission characteristics are not affected with respect to the applications of the TEA106x/TEA111x without the TEA1093.

When the voltage difference between SUP and VBB is larger than 240 mV, the internal current I_{SUP}' mainly flows via transistor TR1 to the circuitry connected to VBB. The current is used to supply all internal circuitry which is connected to VBB and to power the loudspeaker and possible peripherals. All excessive current is conducted to GND by the voltage stabilizer.
When, due to a signal on the line, the voltage difference between SUP and VBB is smaller than 240 mV, the current I_{SUP}' is mainly diverted to GND via transistor TR2. This prevents transistor TR1 from saturating. The take-over between TR1 and TR2 is performed smoothly. As a result of these switching characteristics, the distortion on the line due to the supply of the TEA1093 can be neglected.
When transistor TR2 is conducting, almost no current is available via TR1 for the circuitry connected to VBB. To maintain operation, a buffer capacitor C_{VBB} is connected between VBB and GND.

The ratio of the currents I_{tr1} and I_{SUP}' is called the efficiency of the supply. It gives an indication of which part of the available current flows to the circuitry connected to VBB. The efficiency is mainly determined by the DC-voltage difference between pins SUP and VBB. The DC-voltage on SUP is determined by the PTT requirements and is set via the TEA106x/TEA111x. The DC-voltage on VBB is nominal 3.6V and can be adjusted by resistor Rva. The efficiency of the supply as a function of a continuous signal on the line and with the voltage difference between SUP and VBB as a parameter is depicted in figure 5. The efficiency does not reach 100% due to base current loss in transistor TR1. The measurements for this figure are done at a current I_{SUP} = 20.4 mA thus with $I_{SUP}' \approx$ 20 mA. At higher currents the efficiency will slightly decrease due to a higher base current loss in transistor TR1. From figure 5 it can be concluded that a correct DC-setting is essential for an optimum current management.

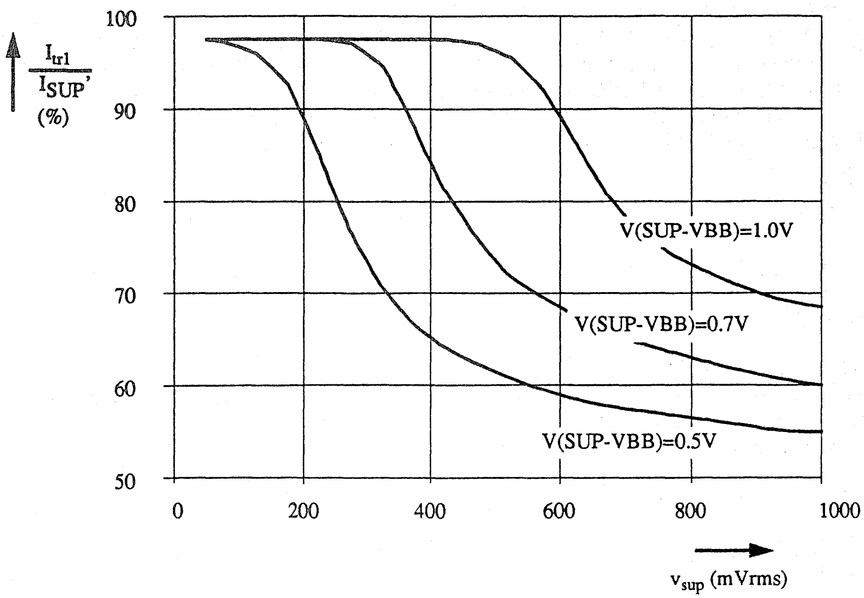


Fig.5 Efficiency of the supply

Adjustments and performance

The constant current I_{tr}, biassing the output stage of the TEA106x/TEA111x, can be adjusted with R_{SREF}. The advised value of 100 Ω for R_{SREF} results in a current of I_{tr} = 3.15 mA. Lower values for R_{SREF} increase I_{tr} and thus decrease the available current for the TEA1093 resulting in reduced performance of the loudspeaker amplifier. Higher values for Rsref have the reverse effect. However, a higher value for R_{sref} also decreases the DC regulator function of the TEA106x/TEA111x and the stability of the supply of the TEA1093. It is advised not to make R_{sref} higher than 150 Ω.

The capacitor C_{SREF}, connected between SREF and VEE of the TEA106x/TEA111x, is needed for stability of the supply part and has a preferred value of 4.7 nF. Another capacitor of 4.7 nF, Csup, for stability is needed between SUP and VBB. A coil of 150 µH, L_{SUP} in series with SUP, is advised for improved stability in the special case that the line has an inductive behaviour. With this coil the real part of the input impedance, between LN and VEE, is positive between 0 and 40 MHz, which means that it can not oscillate. All this three components are shown in figure 28 on page 50. In normal cases L_{SUP} need not to be used.

The maximum DC input current at SUP which the TEA1093 can handle is 140 mA. The maximum allowable voltage at SUP and SREF is 12V.

The nominal voltage on VBB with respect to GND is 3.6V. When a resistor Rva is connected between VA and GND, the voltage on VBB will be increased with respect to nominal. When a resistor Rva is connected between VA and VBB, the voltage on VBB will be decreased. The relation between Rva and VBB is defined in the device specification [Ref. 2].

It is advised to set VBB to a voltage that is 0.7V lower than the DC voltage at SUP. However, the maximum setting for VBB may be 0.4V below SUP. Higher VBB settings will drastically decrease the current efficiency of the supply as shown in figure 5. Lower VBB settings will result in less output voltage swing for the loudspeaker amplifier and thus lower output power. To ensure correct functioning, the voltage on VBB is preferably not to be adjusted lower than 3.2 Volts and not higher than $12V - I_{line,max} \times 20\Omega$. For more details like safe operating area and other limiting values, see the specification of the TEA1093 [Ref.2].
The voltage on VBB can also be set by an external supply, for instance a battery. More details of battery operation with the TEA1093 can be found in chapter 5 on page 50.

The voltage on pin VA is a band gap voltage of around 1.2V. It is temperature compensated and, when externally buffered, may be used as a reference voltage in the application when needed. The input current of the buffer must be smaller than 100nA in order not to affect the VBB setting.

Buffer capacitor C_{VBB} is used to buffer the momentary power gaps when transistor TR2 is conducting or during power-down. The advised value for C_{VBB} is 470µF. Larger values result in longer start-up times while smaller values can cause instability of the voltage stabilizer. For stable operation, it also advised not to use a capacitor with a high series resistance.

Start-up behaviour

During start-up of the TEA1093, the start-up circuit of the supply charges the capacitor C_{VBB} in three steps. This is depicted in figure 6.

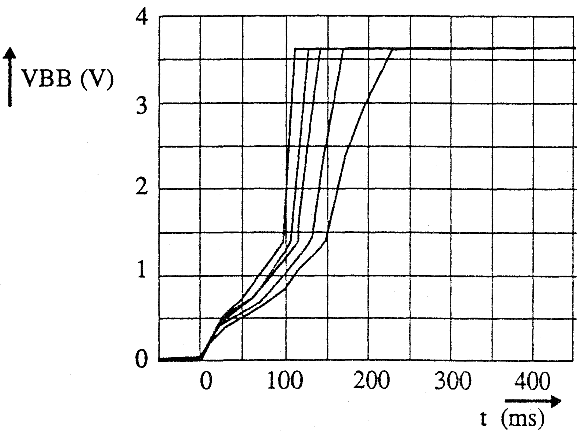


Fig.6 Start-up of the supply

Left to right: VBB at I_{SUP} = 60 mA, 50 mA, 40 mA, 30 mA, 20 mA

As can be seen in figure 6, the charge current is small in the beginning of the start-up phase. This is done to have the transmission circuit started-up quickly with the rest of the line current. The starter is stopped when the voltage on VBB is around 2.4V because at that time, the supply is working and is able to charge C_{VBB} up to its final value (3.6V nominal). The starter restarts when the voltage at VBB drops below 1.9V approximately.

The TEA1093 is also equipped with starters for the dynamic limiter capacitor (Chapter 3.3 on page 20) and the noise envelope capacitors (Chapter 3.4 on page 28). The start-up time of the complete set is strongly dependent on the application and the amount of line current.

3.1.2 Power down

In case of line breaks, occurring during pulse dialling or register recall (flash), the telephone set is without continuous power. In order to prevent relatively long start-up times and distortion of the dial pulse after a line break, the TEA1093 can be put into power down mode via pin PD. In power down mode, the current consumption is reduced and the TEA1093 is supplied from the charge in the buffer capacitor C_{VBB}.

When PD is a logic low, meaning the voltage on PD is lower than 0.3 Volts or pin PD is left open, the TEA1093 is in normal operation mode. The minimum operating current in this case is I_{SUP} = 5.5 mA. When PD is a logic high, meaning the voltage on PD is higher than 1.5 Volts, the TEA1093 is in power down mode. In that case the current consumption from SUP is 55µA, the current consumption from VBB is 400µA. The current which has to be sourced into pin PD when PD is high, is 5µA maximum.
The maximum allowable voltage on PD is VBB+0.4V, the minimum allowable voltage on PD is GND-0.4V.

3.2 Microphone amplifier block

In the first paragraph of this chapter the principle of operation of the microphone amplifier is described as well as its adjustments and performance. In the second paragraph, the mute transmit function is discussed.

3.2.1 Microphone amplifier.

Principle of operation

In figure 7 the block diagram of the microphone amplifier of the TEA1093 is depicted together with the interconnection with the TEA106x/TEA111x.

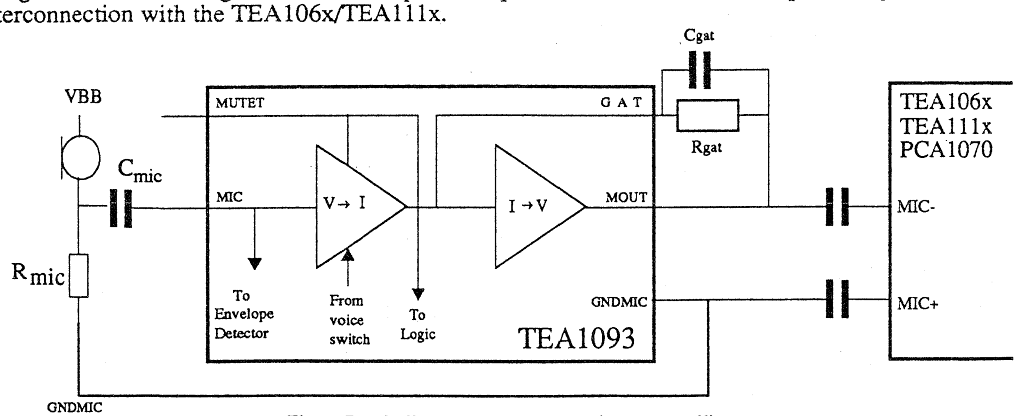


Fig.7 Block diagram of the microphone amplifier

As can be seen in figure 7, the microphone amplifier is referenced to pin GNDMIC instead of referenced to pin GND. This in order to prevent interference from other blocks within the TEA1093 (so called clean ground). The input and output signals of the microphone channel have to be referenced to GNDMIC. Pin GNDMIC itself has to be referenced to GND.

The input of the microphone amplifier is pin MIC. It is an a-symmetrical input well suited for electret microphones. Induced signals in the short wire between the microphone and pin MIC are assumed to be negligible. This in contrary with the handset microphone which is connected to the set via a long cord. The TEA106x/TEA111x family therefore has symmetrical microphone inputs.
The output of the microphone amplifier is pin MOUT. When interconnecting the TEA1093 and the TEA106x/TEA111x, pin MOUT is preferably connected to the TEA106x/TEA111x input pin MIC-. In that case the MIC+ pin of the TEA106x/TEA111x is connected to pin GNDMIC. It is advised not to reverse this interconnection.

As can be seen in figure 7, the microphone amplifier itself is built up out of two parts: a pre-amplifier and an end-amplifier. The gain of the pre-amplifier is determined by the duplex controller block, see chapter 3.4 on page 28. The gain of the end-amplifier is determined by the external feedback resistor Rgat.
The overall gain (Atx) of the microphone amplifier from input MIC to output MOUT in Tx-mode is given as:

$$Atx = 20 * \log (0.674 * Rgat / Rstab).$$

With Rstab being the resistor at STAB of 3.65kΩ.

Adjustments and performance

A handsfree microphone, referenced to GNDMIC, can be connected to the input MIC via a DC blocking capacitor Cmic. Together with the input impedance of pin MIC of 20kΩ, Cmic forms a first order high-pass filter.
The handsfree electret microphone can be supplied from VBB. However, during normal operation VBB will contain a small ripple, for instance due to large loudspeaker signals. Electret microphones with a small power supply rejection ratio should therefore be connected via an RC smoothing filter as depicted in figure 8.

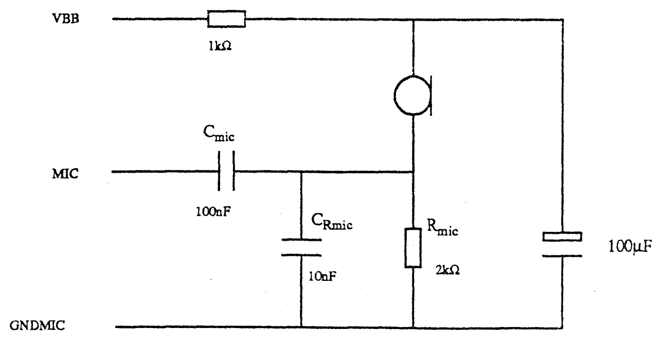


Fig.8 Supply arrangement for an electret microphone

As shown in figure 8, the RC smoothing filter is referenced to GNDMIC to have one ground reference for the whole microphone signal path. See chapter 4.4 on page 46 for microphone path layout rules. The sensitivity of the electret microphone is set via resistor Rmic. By placing a capacitor Crmic over this resistor a first order low-pass filter is formed for the microphone signal.

Via the resistor Rgat, the gain of the microphone amplifier can be adjusted from 5dB to 25dB to suit application specific requirements. With resistor Rgat=30.1kΩ the gain equals 15dB with a tolerance of ±2.5dB. Thus minimum Rgat=9.6kΩ and maximum Rgat=96.3kΩ.
Capacitor Cgat is applied in parallel with resistor Rgat to ensure stability of the microphone amplifier. Together with Rgat it also provides a first order low pass filter.

The input of the microphone amplifier can handle signals up to 18 mVrms with 2% total harmonic distortion. However the microphone input signal is also used by the duplex controller, see chapter 3.4. At 14 mVpeak at the input the positive part of the signal on pin TSEN starts clipping which might influence the switching behaviour. It is therefore advisory to keep the microphone input signal below this level.
The output drive capability at pin MOUT is 20µArms.

3.2.2 Noise calculations

In the application of the TEA1093 with a transmission circuit there are three mechanisms that affect the noise on the line:

1. noise floor of 1093-MIC input + gain from MIC to MOUT
2. attenuation circuit between MOUT and the MIC input of the transmission IC.
3. noise floor of the transmission IC + gain of the transmission IC

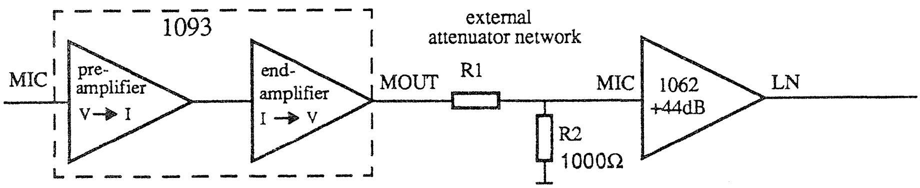


Fig.9 noise level calculation in TEA1039/TEA1062 combination.

Figure 9 shows the diagram from which the noise on LN can be calculated. It is the sum of all noise contributions in dBmp. The example is restricted to the TEA1062 with a gain of +44 dB, because just as example. Other situations can be calculated in a similar way. Ac and dc considerations are not considered in the slightly simplified schematic.

The attenuation circuit R1/R2 in figure 9 is externally to be added. It is in some applications required to adjust the right transmit gain and to affect the noise on the line. If R1/R2 are chosen carefully the noise from MOUT is attenuated. R2 has to be chosen that way that its own resistor noise can be neglected with respect to the noise floor of the TEA1062 MIC input.

Noise calculation can be done with the aid of table 1.

TABLE 1 Resistor noise values

a resistor of	produces a psophometrical noise level of:
1MΩ	-103 dBmp
100 kΩ	-113 dBmp
10 kΩ	-123 dBmp

The noise level of R2 is amplified by the TEA1062. From table 1 can be concluded that a resistor of 1000 Ω produces -133 dBmp. This is negligible with respect to the noise floor of the TEA1062 which is -121 dBmp, so if R2 = 1000 Ω it does not increase the noise level on LN.

Another example: if R2= 8.2kΩ the resistor noise equals -123.9 dBmp (table 1). This can not be neglected with respect to -121 dBmp and therefor noise on LN has to be calculated as follows: -123.9 dBmp corresponds with a noise voltage $e_1 = 0.775 \times 10^{-123.9/20}$. The bottom noise level of the TEA1062 is -121 dBmp, this corresponds with a noise voltage $e_2 = 0.775 \times 10^{-121/20}$. The sum of these noise voltages $e_3 = \sqrt{e_1^2 + e_2^2}$. This corresponds with a noise level of $20 \times \log e_3$ = -119.2 dBmp on LN. Table 2 summarizes a few examples of the complete noise calculation from MOUT to LN. Expessed in dB the noise levels can simply be added. The mentioned attenuations of R1/R2 are just arbitrary examples to show how calculation proceeds.

TABLE 2 Noise calculation

remarks	gain from MIC to MOUT (dB)	noise level on MOUT (dBmp)	attenuation by R1 and R2 (dB)	noise level on MIC of TEA1062 (dBmp)	bottom noise of TEA1062 (dBmp)	noise level on LN (dBmp)	1093-mode
1	15	- 100	0	-100	-121	-100+44= -56	transmit
2	5	-110	0	-110	-121	-110+44= -66	transmit
3	25	-90	-31	-121	-121	-121+44+3= -74	transmit
4	25	-110	-20	-130	-121	-121+44= -77	idle
5	25	-111	0	-111	-121	-111+44= -67	receive

Remarks:

1. Calculation starts with the device specification [Ref.2] that specifies a noise level of -100 dBmp at MOUT under the conditions that the gain is 15 dB and the TEA1093 is in transmit mode. The attenuation of R1/R2 is supposed to be 0 dB. The noise contribution of the TEA1062 is negligible with respect to -100 dBmp.
2. From remark 1 can be concluded that if the gain from MIC to MOUT is reduced by 10 dB, the noise on MOUT is also reduced by 10 dB. The minimum noise level on MOUT is -111 dBmp.
3. This example shows what happens if the attenuation of R1 and R2 results in a noise level that is in the order of magnitude of the noise level of the TEA1062. For minum noise on LN he attenuation of R1/R2 should be chosen below -31 dB.

4. In idle mode the noise from the TEA1093 MIC input is attenuated by half the switching range, nominally 20 dB (see chapter 3.4.3 on page 31 and figure 2 on page 9), so noise level on MOUT becomes -110 dBmp. In idle mode the noise on LN is in most cases determined by the TEA1062.
5. In receive mode the noise from the TEA1093 MIC input is attenuated by the full switching range, nominally 40 dB (see chapter 3.4.3 on page 31 and figure 2 on page 9), so the noise level on MOUT should be 40 dB lower resulting in -130 dBmp, but the bottom noise of the microphone end-amplifier is -111dBmp, therefore the noise on MOUT becomes -111dBmp.

An application example with the attenuation network between transmission IC and TEA1093 can be found in figure 29 on page 52. For the influence of the attenuator on the noise at the loudspeaker outputs, see the last paragraph of chapter 3.3.1. on page 24.

3.2.3 Mute transmit

During handsfree operation the microphone can be muted via MUTET so conversation cannot be heard by the other party. When the microphone amplifier is muted, automatically the TEA1093 switches over to Rx mode, see also chapter 3.4 on page 28.
When a logic high is applied to MUTET, meaning the voltage on MUTET is higher than 1.5 Volts, the microphone pre-amplifier is muted. The end-amplifier can still be used by applying a signal on GAT. The obtained gain reduction is 80dB. The current which has to be sourced into pin MUTET when MUTET is high, is 5μA maximum.
When MUTET is a logic low, meaning the voltage on MUTET is lower than 0.3 Volts or pin MUTET is left open, the microphone amplifier is not muted.
The maximum allowable voltage on MUTET is VBB+0.4V, the minimum allowable voltage on MUTET is GND-0.4V.

3.3 Loudspeaker amplifier block

The block diagram of the complete loudspeaker amplifier block is depicted in figure 10

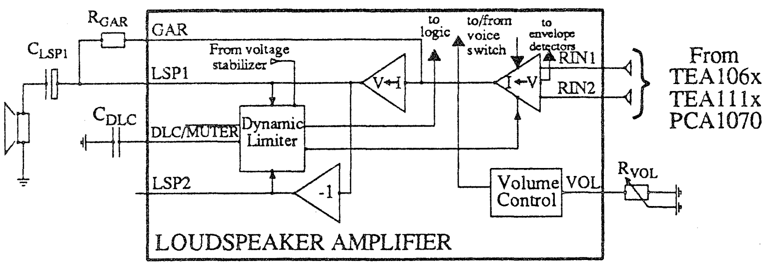


Fig.10 Principle of the loudspeaker amplifier

As can be seen in figure 10, the loudspeaker amplifier is built up out of three parts: the loudspeaker amplifier itself, volume control and dynamic limiter. In chapter 3.3.1 on page 21 the principle of operation of the loudspeaker amplifier is described as well as its adjustments and performance. In chapter 3.3.2 on page 24 the same items are discussed of the volume control part and in chapter 3.3.3 on page 25 of the dynamic limiter part. In chapter 4 on page 36 it is described how the receiver can be muted.

3.3.1 Loudspeaker amplifier

Principle of operation

As can be seen in figure 10, the input of the loudspeaker amplifier, pins RIN1 and RIN2, is symmetrical. A single, asymmetrical input is not applicable because the output signal of the speech transmission circuits of the TEA106x/TEA111x-family is referenced to VEE while the TEA1093 is referenced to SLPE. The input RIN1 can be connected to the earpiece output QR+ of the TEA106x/TEA111x. In that case, the other input RIN2, has to be connected to pin VEE of the TEA106x/TEA111x.
At the loudspeaker outputs LSP1 and LSP2, the amplified receiving signal from QR+ is available. The outputs are in opposite phase and referenced to GND. The outputs can drive a loudspeaker which is connected as a single ended load (SEL) or as a bridge tied load (BTL).

As can be seen in figure 10, the amplifier itself is built up out of two parts: a pre-amplifier and an end-amplifier with two outputs.
The gain of the pre-amplifier is determined by the duplex controller block, see chapter 3.4 on page 28. The gain of the end-amplifier is determined by the external feedback resistor Rgar. The signal on LSP1 is inverted by the second amplifier to pin LSP2.

When a loudspeaker is connected as a SEL, the overall gain of the loudspeaker amplifier (Arx) from inputs RIN1 and RIN2 to the output LSP1 or LSP2 in Rx-mode is given as:

$$Arx = 20 * \log (0.435 * Rgar / Rstab) \text{ dB} \quad \text{(SEL)}.$$

With Rstab being the resistor at STAB of 3.65kΩ. When a loudspeaker is connected as a BTL, the overall gain from inputs RIN1 and RIN2 to outputs LSP1 and LSP2 in Rx-mode is 6dB higher:

$$Arx = 20 * \log (0.870 * Rgar / Rstab) \text{ dB} \quad \text{(BTL)}.$$

Adjustments and performance

The input signal for the loudspeaker amplifier has to be coupled in via the capacitor Crin1 to block DC. Together with the input impedance of 20kΩ at RIN1, a first order high pass filter is introduced. When connecting the other input RIN2 to VEE, this also has to be done via a capacitor. This capacitor Crin2 preferably has the same value as Crin1 to obtain a good common mode rejection ratio. The input impedance of RIN2 is also 20kΩ.
The inputs can handle signals up to 390 mVrms with a total harmonic distortion of 2%. Because of this it is advised not to connect RIN2 to QR- but only to VEE. The inputs RIN1 and RIN2 are biassed at around zero volts with respect to GND. By applying a signal to the inputs, they can become negative. The protection on these pins however is made different from other pins which makes it possible to make RIN1 and RIN2 as low as -1.2V without damaging the circuit.

A loudspeaker can be connected to the TEA1093 as a single ended load (SEL) or as a bridge tied load (BTL). As a SEL it must be connected between LSP1 or LSP2 and GND. As a BTL it must be connected between LSP1 and LSP2. The outputs LSP1 and LSP2 are biassed at $\frac{1}{2}$ VBB. Therefore, in case of a SEL, a capacitor must be placed in series with the loudspeaker to block DC-current. Together with the impedance of the loudspeaker also a high pass filter is formed. In case of a BTL a DC-blocking capacitor is not essential but advisable.
BTL mode provides an impedance accommodation of a factor 4. Thus in case of BTL for instance a

loudspeaker of 100 Ω produces the same loudspeaker output power and has the same power consumption from the line as a 25Ω loudspeaker in SEL mode. So in line powered telephone sets the BTL function does not improve loudspeaker power. The available line current and line voltage determine the maximum loudspeaker output power.

In externally supplied telephone sets (answering machines etc.) BTL can be useful because in that case enough supply current is available to reach the maximum output voltage. If this maximum is upgraded by a factor 2 with the BTL function, the output power will be higher. See also chapter 4.1 on page 37.

Via the resistor Rgar, the gain of the loudspeaker amplifier can be adjusted from 3dB to 33dB in case of a SEL and from 9dB to 39dB in case of a BTL. With resistor Rgar=66.5kΩ, the gain equals 18dB with a SEL and 24dB with a BTL. The tolerance in both cases will be ±2.5dB.
A capacitor Cgar can be applied in parallel with resistor Rgar to provide a first order low pass filter.

In line powered applications, the output power of the TEA1093 loudspeaker amplifier is determined by the maximum output current and the maximum output voltage swing. The maximum output current is dependent on the line current, the current consumption of peripherals connected, the efficiency of the supply and the efficiency of the loudspeaker amplifier. The maximum output voltage swing is determined by the voltage on VBB.
As described in chapter 3.1 on page 12, a part of the line current is used for powering the TEA106x/TEA111x, the rest flows into the TEA1093 (I_{SUP}). As can be seen in figure 4 a part of this current is used to bias circuitry connected to SUP, the rest (I_{SUP}') flows into the current switch of the supply of the TEA1093. The part which becomes available at VBB (I_{tr1}) depends on the efficiency of the supply. This efficiency is depicted in figure 5. A part of the current I_{tr1} is used for biasing internal circuitry connected to VBB and for supplying peripherals, the rest can be used for powering the loudspeaker. Together with the efficiency of the loudspeaker amplifier this determines the maximum output current.
The voltage on VBB determines the maximum output voltage swing. For maximum voltage swing, the end-amplifier has a rail to rail output stage. The maximum voltage swing is dependent on the output current and the temperature, see also chapter 3.3.3 on page 25.

In figure 11, the maximum output power is depicted as a function of I_{SUP} being the current flowing into pin SUP.

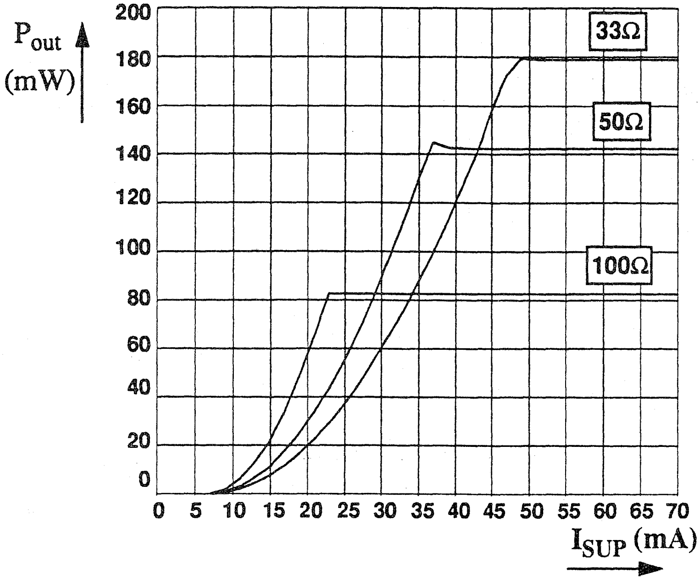


Fig.11 Maximum loudspeaker output power for three different loudspeaker impedances, at VBB=9V in SEL.

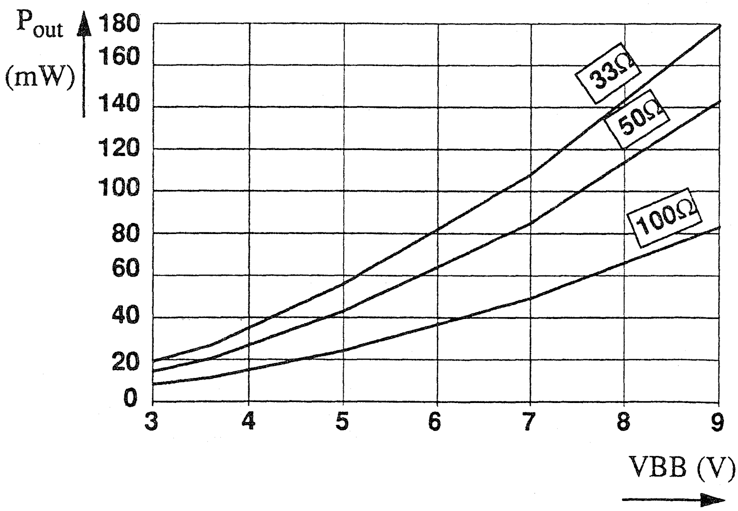


Fig.12 Maximum loudspeaker output power as a function of VBB

The output swing is limited by the supply voltage VBB, this is represented by the horizontal part of the curves in figure 11 for VBB = 9V. The parabola below this horizontal part represents the region where the maximum power is limited by I_{SUP} which on his turn depends on the available line current, the supply current of the transmission IC and the supply current of the peripherals:

$$I_{SUP} = I_{line} - I_{transmission\ IC} - I_{peripheral}$$

It can be seen that at low I_{SUP} the largest output power is obtained with a large loudspeaker impedance. But simultaneously it needs a high VBB (figure 12). If a lot of current is available, the largest output power is obtained with a low loudspeaker impedance.
In figure 12 the maximum loudspeaker output power is shown as a function of VBB. It can be seen that the largest output power is reached under all circumstances if VBB is as large as possible. It can also be seen that the lowest loudspeaker impedance has the largest output power, but it needs simultaneously more current (figure 11). The pros and cons of the several loudspeaker impedances should be weighed carefully.

Example:

Suppose a maximum loudspeaker power is required at I_{SUP} = 20 mA. From figure 11 can be read that with a loudspeaker impedance of 100 Ω an output power of 60 mW can be reached at 20 mA. In figure 12 it can be seen that this needs a VBB = 7.6V. If it is not allowed to adjust a VBB of 7.6V, because of the local maximum line voltage requirements, a lower loudspeaker impedance should be taken etc. See for a step by step procedure chapter 4.1 on page 37.

The noise level at the outputs LSP1 and LSP2 is 80µVrms at a gain of 18dB and with the inputs RIN1 and RIN2 shorted with 200Ω. However in an application with the TEA106x/TEA111x, the noise at the outputs is larger than 80µVrms. This is due to the fact that the noise generated in the transmit channel of the telephone set is fed back via the sidetone network to the receive channel and is amplified to the loudspeaker outputs.
The maximum noise level at the loudspeaker outputs can be perceived in receive mode. With for instance an electrical sidetone of -12dB, and an overall receive gain of 24dB, the noise level at the loudspeaker output will be 12dB higher than the bottom level of the sending noise on the line. With figure 7 on page 16 (-111dBmp at MOUT, transmit gain of the TEA106x/TEA111x of 44dB, thus -67dBmp on the line) this will lead to -55dBmp at the loudspeaker outputs if loudspeaker volume is maximum. This level can be reduced by placing an attenuation network between MOUT and MIC- as described in chapter 3.2.2 on page 18. Depending on the TEA106x/TEA111x used, the bottom level of the sending noise on the line can become lower than -80dBmp (44dB sending gain of the TEA106x/TEA111x) resulting in less than -68dBmp at the loudspeaker outputs.

3.3.2 Volume control

Principle of operation

Via the volume control block, the volume of the loudspeaker signal can be adjusted by the external potmeter Rvol connected to pin VOL. By turning the potentiometer, the gain of the loudspeaker pre-amplifier is varied. Volume control may not affect the transmit gain in transmit mode. To obtain this, the volume control acts upon the pre-amplifier via the duplex controller, see also chapter 3.4 on page 28.

Adjustments and performance

Out of pin VOL a current Ivol, set by Rstab, see chapter 3.4 on page 28, is flowing which is proportional to the absolute temperature (PTAT). At room temperature this current is around 10µA. Together with the resistance of the potmeter Rvol, the current Ivol creates a PTAT voltage on pin VOL. This PTAT voltage is processed by the volume control block. As a result, a temperature independent volume reduction in the loudspeaker signal of 3dB is obtained at approximately every 950Ω variation of the potentiometer Rvol. This means that a linear potentiometer can be used to control the volume logarithmically, thus in dB's. With the advised value for Rvol of 10kΩ, the maximum gain reduction of the volume control is more than 30dB. The maximum gain reduction however is limited by the switching range, see chapter 3.4 on page 28. When the resistor Rvol is zero (maximum loudspeaker gain), the receive gain is not influenced.

When digital volume control is desired this can be done as depicted in figure 13.

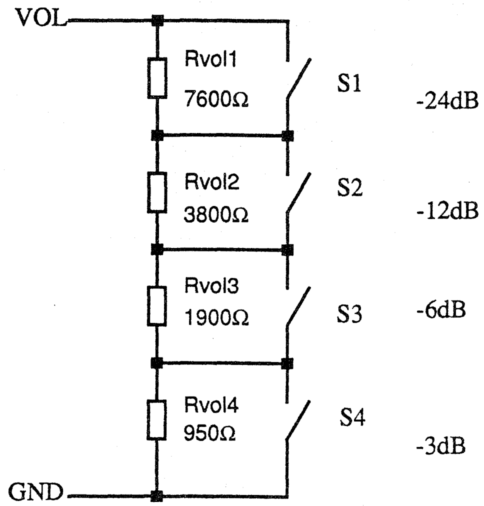


Fig.13 Digital volume control

With the 4 bit digital volume control of figure 13, 16 volume levels can be set via steps of 3dB (from 0dB to a maximum of 45dB of attenuation). The switches can be either MOSFET's or analogue switches, for instance the Philips HCT4066 type. It is advised not to use bipolar transistors as switches because of the saturation voltage of these devices.
When a voltage is applied to pin VOL to control the volume, preferably this voltage has to be a PTAT voltage source. If not, the obtained gain reduction is no longer temperature compensated.

3.3.3 Dynamic limiter

Principle of operation

The dynamic limiter minimizes the distortion of the output signal by reducing the gain of the loudspeaker pre-amplifier when the output signal starts clipping or when too low supply conditions are detected. The amount of gain reduction is determined by the voltage on pin DLC/$\overline{\text{MUTER}}$. This voltage is varied by charging and discharging the capacitor Cdlc. The combination of charge and discharge currents and the capacitance of Cdlc sets the timing of the dynamic limiter.

Clipping of the loudspeaker output signal occurs when the output transistors are driven into deep saturation. To prevent hard clipping, deep saturation has to be prevented. The saturation voltage of the output transistors strongly depends on the output current and the temperature. In the dynamic limiter of the TEA1093 these effects are taken into account to have maximum output swing under any condition. When the dynamic limiter detects the beginning of saturation, the capacitor Cdlc is discharged fast with a current of approximately 1 mA, see figure 14 left hand side. As a result, the gain of the loudspeaker amplifier is reduced. The loudspeaker amplifier stays in its reduced gain mode until the peaks of the loudspeaker signal no longer start to cause saturation. The gain then slowly returns to its normal value by charging Cdlc with a current of 1μA, see figure 14 right hand side. In this way it is ensured that always the maximum reachable output power can be obtained at low distortion

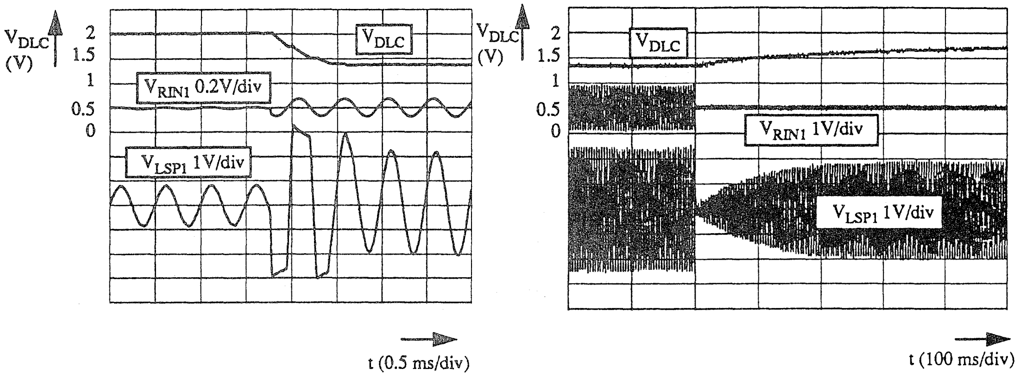


Fig.14 Action of the dynamic limiter clipping detector, attack time (left) and release time (right).

Under too low supply conditions, the gain of the loudspeaker amplifier is reduced in order to prevent the TEA1093 from malfunctioning. Only the gain of the loudspeaker amplifier is affected since it is considered to be the major power consuming part.

When the TEA1093 is running out of current, the supply voltage VBB decreases. In such a case, the gain of the loudspeaker amplifier is reduced slowly, see figure 15. This is done by discharging Cdlc with a small current of less than 100nA.

When the supply voltage continues to decrease and drops below the internal threshold of 2.75V, the gain is reduced immediately regardless of the voltage on pin DLC. However the capacitor Cdlc is also discharged to ensure that the gain remains reduced when the voltage on VBB raises again above the internal threshold. When the voltage on VBB has reached its set value, the gain is raised again by charging Cdlc with 1μA.

In case the TEA1093 runs out of current when raising the receive gain, the capacitor will not be charged further. This ensures maximum output power at low line currents.

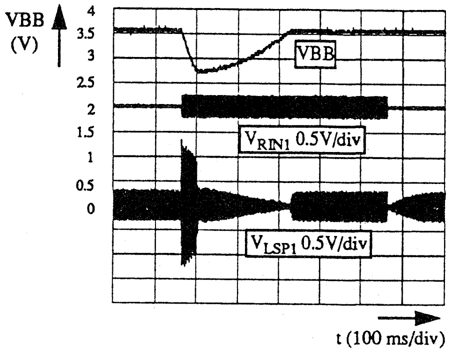


Fig.15 Action of the dynamic limiter supply detector

Adjustments and performance

The timing of the dynamic limiter is determined by the charge and discharge currents and by the capacitor Cdlc. The currents are internally fixed and cannot be changed. All charge and discharge time constants, and therefore the dynamic limiter timing, are proportional to the value of Cdlc. The only exception is when VBB drops below its threshold. In that case the gain is reduced immediately regardless of the voltage on Cdlc.
As a compromise between attack and release times of the dynamic limiter a capacitor of 470 nF is advised. Larger values will give a smoother (slower) response while smaller values may lead to more distortion at lower frequencies. It is advised not to use a capacitor with a high leakage current in order not to influence the behaviour of the dynamic limiter.
If only a faster attack time is desired, it is possible to connect a low ohmic resistor in series with Cdlc of maximum 100Ω. In that case the relatively high discharge currents (200 µA - 1 mA) will cause an instantaneous drop on pin DLC/$\overline{\text{MUTER}}$ when limiting action is needed. The instantaneous drop caused by the small charge current of 1µA will be negligible.

With Cdlc=470 nF, the attack time for the clipping detector is in the order of a few milliseconds. The attack time when the circuit runs out of current is in the order of several seconds. The attack time when the supply voltage VBB drops below the threshold of 2.75Volts is less than 1 millisecond. The release time in all cases is in the order of a few 10 milliseconds.

When the dynamic limiter is acting, in practice the distortion of the output stage will stay below 5%. The dynamic limiter does not limit the distortion of the input stage.

Start-up behaviour

When the TEA1093 is started up, the starter of the dynamic limiter charges the capacitor Cdlc with a current of approximately 80µA. The starter stops when the voltage at DLC/$\overline{\text{MUTER}}$ has reached a value of around 1.6V. Then a current of 1µA charges the capacitor further, up to a voltage of around 1.9V. At that point the voltage on DLC /$\overline{\text{MUTER}}$ is limited. The starter restarts when the voltage at DLC/$\overline{\text{MUTER}}$ drops below 200 mV.

3.3.4 Mute receive

The loudspeaker amplifier can be muted by making pin DLC/MUTER lower than 200 mV. As a result the gain of the loudspeaker amplifier is reduced with 80dB. Also the circuit is internally forced into Tx-mode.
Pin DLC/MUTER can be made low by placing a switch over the capacitor, for instance a simple transistor.
When the switch is open, nothing is influenced. When the switch is closed and DLC/MUTER is put below 200 mV, the loudspeaker pre-amplifier is muted. The end-amplifier can still be used by applying a signal on GAR. Because the starter is reactivated when the voltage on DLC/MUTER is smaller than 200 mV, the switch must be able to sink approximately 80μA. When the switch is opened again, the starter will recharge the capacitor Cdlc.
The minimum allowable voltage on DLC/MUTER is GND-0.4V.

3.4 Duplex controller block

In this chapter the principle of operation of the complete duplex controller will be discussed as well as its adjustments and performance. This will be done with the aid of figure 16, where the complete duplex controller is depicted.

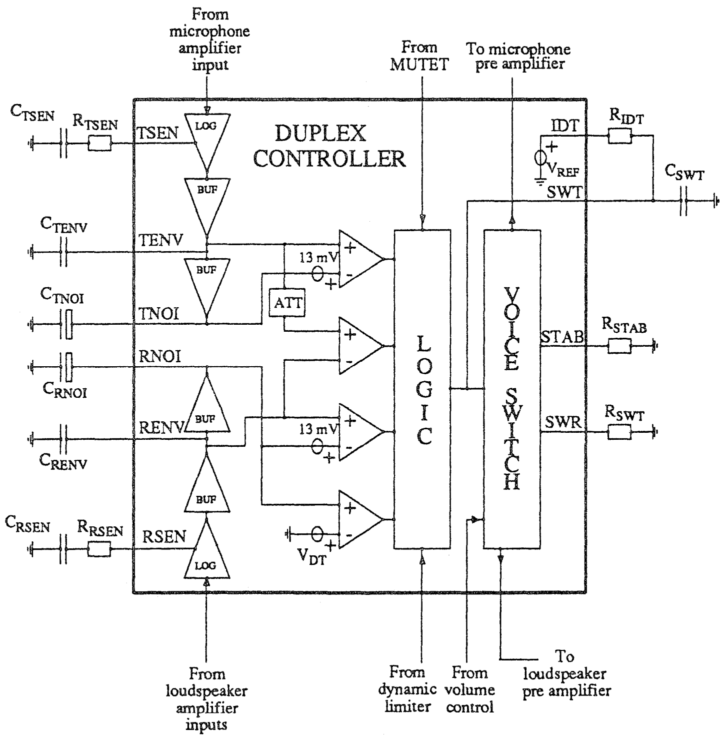


Fig.16 Principle of the duplex controller

As can be seen in figure 16, the duplex controller is built up out of signal and noise envelope detectors, decision logic and a voice switch.
The signal and noise envelope detectors determine the signal envelope and the noise envelope of both the transmit and receive signal. These envelopes are used by the decision logic to determine to which mode the set has to switch over (Tx, Rx or Ix-mode). The logic charges and discharges the capacitor Cswt and the resulting voltage on pin SWT controls the voice switch. The voice switch switches over the set between the three modes while keeping the loop gain constant.
In chapters 3.4.1 to 3.4.3, the principle of operation of the three parts is given. In chapter 3.4.4, the adjustments and performance of the complete duplex controller is given.

3.4.1 Signal and noise envelope detectors

The signal and noise monitors of the transmit and receive channel are globally the same. The principle of the detectors therefore will be explained with the help of one of them: the signal and noise envelope detector of the transmit channel.
The microphone signal on MIC is connected to the first stage of the detector, see figure 16. The first stage amplifies the microphone signal from MIC to TSEN with an internally fixed gain of 40dB. Via the RC combination RtsenCtsen the signal on TSEN is converted into a current. This conversion determines the sensitivity of the envelope detectors. The current is logarithmically compressed and internally converted to a voltage. This voltage thus represents the compressed microphone signal. At room temperature, an increase of the microphone signal with a factor of 2 will increase the signal envelope with 18 mV if the current through TSEN stays between 0.8µArms and 160µArms. Outside this region the compression is less accurate.
The compressed microphone signal is buffered by the second stage to pin TENV. Because the buffer can source maximum 120µA and sink maximum 1µA, the signal on TENV follows the positive peaks of the compressed signal. This is called the signal envelope. The time constants of the signal envelope are therefore determined by the combination of the internal current sources and the capacitor Ctenv.
The voltage on TENV is buffered by the third stage to pin TNOI. Because this buffer can source maximum 1µA and sink maximum 120µA, the signal on TNOI follows the negative peaks of the signal on TENV. This is called the noise envelope because it represents the background noise. The time constants of the noise envelope are therefore determined by the combination of the internal current sources and the capacitor Ctnoi.
Both capacitors Ctnoi and Crnoi are provided with a start circuit. During start up the capacitors are charged with approximately 40µA up to 1.9V. The starter will restart when the voltage on the capacitor drops below 0.9V.
As can be seen in figure 16, the principle of operation of the signal and noise envelope detectors of the receive channel is equal to that of the transmit channel. However, the gain of the first stage (inputs to pin RSEN) is 0dB and not 40dB as in case of the transmit channel.

The behaviour of the envelopes is illustrated in figure 17, where the signal and noise envelope of the receive channel (RENV, RNOI) are depicted together with the input signal on RIN1.

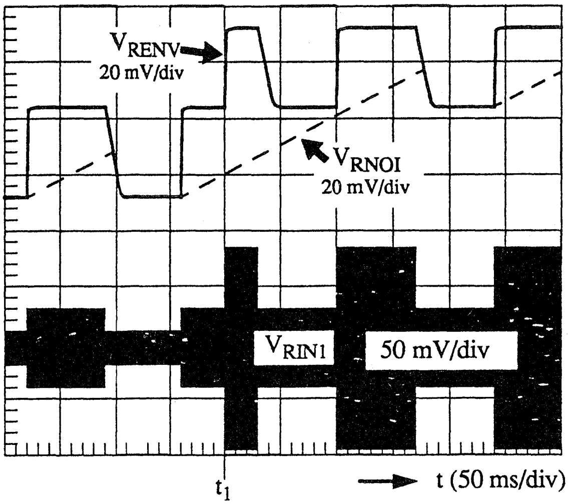


Fig.17 Typical behaviour of the signal and noise envelope detectors

In figure 17, the amplitude of the 1kHz input signal at RIN1 is modulated with 10dB and at moment t1 an extra 10dB is added. When, during modulation, the input signal is raised with 10dB, the signal envelope at RENV immediately follows. When the input signal drops with 10dB, the signal envelope drops less quick, thus reducing the influence of room echo. The noise envelope at RNOI slowly follows the signal envelope but never crosses it. When at t1 the extra 10dB is added, the signal envelope also increases but due to the logarithmic compression, the variation in the signal envelope due to the 10dB modulation is the same before and after t1.

3.4.2 Decision logic

The signal and noise envelopes of the transmit and receive signal are used by the decision logic to determine in which mode the TEA1093 has to be.
The output of the logic is a current source which charges or discharges the capacitor Cswt. If the logic determines Tx-mode, the capacitor Cswt is discharged with 10μA. When Rx-mode is determined, Cswt is charged with 10μA. When Ix-mode is determined, the current source is zero and the voltage on SWT becomes equal to the voltage on IDT via the resistor Ridt. The time constants of the duplex controller are therefore determined by the combination of the internal current source, the capacitor Cswt and the resistor Ridt.

As can be seen in figure 16, the envelopes are not used directly by the decision logic.
First, to have a clear choice between signal and noise, the signal is considered as speech when its envelope is more than 4.3dB above the noise envelope. At room temperature, this is equal to a voltage difference of 13 mV. This so called speech/noise threshold is implemented in both the transmit and receive channel. At the end of chapter 3.4.4 a way to increase this threshold is discussed.
Second, the signal on MIC contains both the signal of the talker using the set as well as the signal coming from the loudspeaker (acoustic coupling). In Rx-mode, the contribution of the loudspeaker overrules the contribution of the talker using the handsfree telephone set. As a result, the signal envelope on TENV is

mainly formed by the loudspeaker signal. To correct this, an attenuator is placed between TENV and the TENV/RENV comparator. This attenuation equals the attenuation applied to the microphone amplifier gain. Thus when the TEA1093 is in Rx-mode the attenuation equals the switching range.
Third, when a dial tone is present on the line, without measures this would be recognized as noise. This would happen because it is a signal with a constant level during a long period. As a result the TEA1093 would go to Ix-mode and the user of the set would hear the dial tone fade away. Therefore, a dial tone detector is incorporated which does not consider input signals as noise when they have a level higher than the dial tone level. The dial tone level is adjustable by Rrsen.
When these three corrections are made, the signal and noise envelopes are used by the comparators and the logic. As already explained the output of the logic is a current source. The relation between the current source and the output of the comparators is given in table 3. If, for instance, TENV>RENV (transmit signal is larger than receive signal) and TENV>TNOI (transmit signal more than 4.3dB larger than noise level), then the output current will be -10μA.

TABLE 3 Truth table of the decision logic

MODE	transmit (True = 1)	receive	receive dial tone	idle	idle
Comparator TENV > TNOI	1	x	x	0	x
Comparator TENV > RENV	1	0	0	1	0
Comparator RENV > RNOI	x	1	x	x	0
Comparator RNOI > Vdt	x	x	1	x	0
Output current to Cswt	**-10 μA**	**+10 μA**	**+10 μA**	**0 μA**	**0 μA**

When MUTET is made high, see chapter 3.2.3 on page 20, the output current is forced to be 10μA, which forces the TEA1093 into Rx-mode and mutes the microphone amplifier. When pin DLC/$\overline{\text{MUTER}}$ is made lower than 200 mV, see chapter 3.3.4 on page 28, the output current is forced to be -10μA which forces the set into Tx-mode and mutes the loudspeaker amplifier. When both MUTET is made high and DLC/$\overline{\text{MUTER}}$ is made low the output current is forced to be -10μA and both channels are muted.
The voltage on pin SWT is internally limited to IDT+400 mV and IDT-400 mV.

3.4.3 Voice switch

With the voltage on SWT, the voice switch regulates the gain of the microphone pre-amplifier and the loudspeaker pre-amplifier in such a way that the sum of the transmit and receive gain is kept constant. This is done to keep the loop gain of a handsfree telephone set constant, see also the introduction chapter 1. The switch-over behaviour of the voice switch will be described with the aid of figure 18.

When the voltage on SWT is more than 180 mV below the voltage on IDT, the TEA1093 is fully switched to Tx-mode (gain of the microphone amplifier is at its maximum and the gain of the loudspeaker amplifier is at its minimum). When the voltage on SWT is more than 180 mV above the voltage on IDT, the TEA1093 is fully switched to Rx-mode (gain of the microphone amplifier is at its minimum and the gain of the loudspeaker amplifier is at its maximum). The TEA1093 is considered to be in Ix-mode when the voltage on SWT equals the voltage on IDT. When the capacitor Cswt is charged or discharged, the voltage on SWT varies and as a result the voice switch will smoothly switch over between the modes.

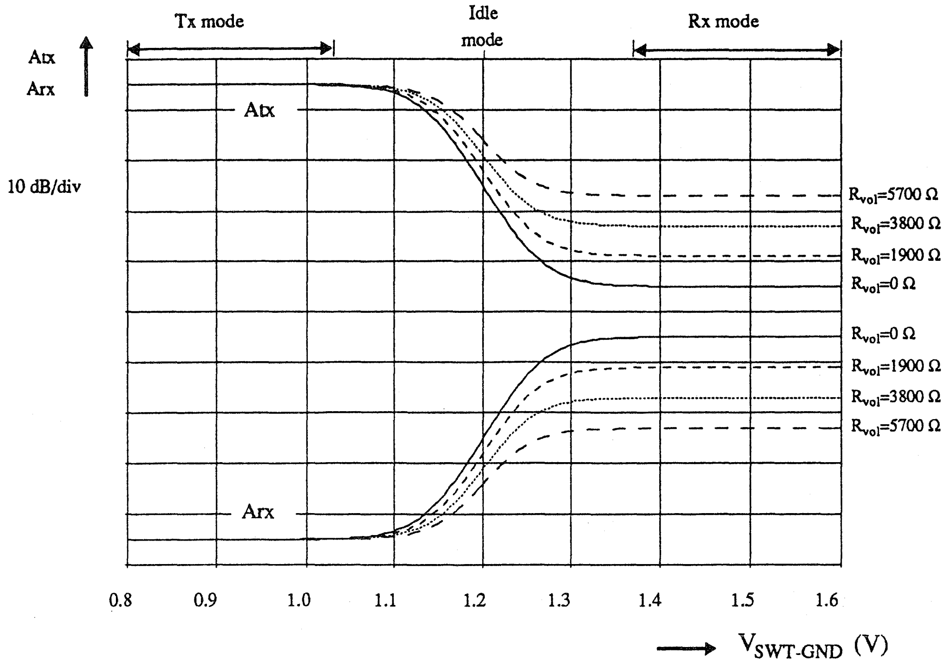


Fig.18 Switch-over behaviour

The difference between the maximum and minimum gain of the loudspeaker or microphone pre-amplifier is called the switching range. This range is determined by the ratio of Rswr and Rstab, see chapter 3.4.4 on page 33. Both Rswr and Rstab set internally used reference currents which are proportional to absolute temperature (PTAT).

As already stated in chapter 3.3 on page 20, the volume control does not directly act upon the loudspeaker amplifier gain but via the voice switch. As a result, the loop gain of the handsfree set is kept constant when the volume of the loudspeaker signal is adjusted. The voice switch however, is designed such that the volume control has no influence in Tx-mode. Therefore during transmit, the gain of the microphone amplifier of the TEA1093 is not affected. In the extreme case, which is not plotted in figure 18, that the volume of the loudspeaker signal is reduced with the switching range, the TEA1093 virtually does not switch over. It also follows from this plot that when it was possible to have a volume reduction larger than the switching range, the gain of the loudspeaker amplifier would be smaller in Rx-mode than in Tx-mode. To avoid this, the volume control range of the TEA1093 can not be made larger than the switching range.

3.4.4 Adjustments and performance

The adjustment of the duplex controller has to be performed according the following recipe:

1. Determine switching range
2. Determine dial tone detector level
3. Determine sensitivity
4. Determine timings.

Ad 1. Determine switching range

The switching range Asw is determined by the ratio of the two resistors Rstab and Rswr according:

$$Asw = 20 * \log (Rswr / Rstab) \text{ (in dB)}.$$

The resistor Rstab has to be taken 3.65kΩ. The resistor Rswr can be varied between 3.65kΩ and 1.5MΩ resulting in a switching range between 0dB and 52dB. With Rswr is 365kΩ, the switching range is set to 40dB typical.

The switching range is calculated out of the loop gain (Aloop). In a handsfree application the loop gain has to be smaller than 1 (equivalent to 0dB) and can be calculated as follows:

$$Aloop = Atx1093 + Atx_{transmission} + Ast + Arx_{transmission} + Arx1093 + Aac - Asw \text{ (in dB)}.$$

With:

Atx1093	= sending gain of the TEA1093 (MIC to MOUT)
$Atx_{transmission}$	= sending gain of the transmission IC (MIC+/- to LN)
Ast	= electrical sidetone
$Arx_{transmission}$	= receive gain of the transmission IC (LN to QR+)
Arx1093	= receive gain of the TEA1093 (RIN1/2 to LSP1/2)
Aac	= electro-acoustic coupling from loudspeaker to microphone (LSP1/2 to MIC)
Asw	= switching range.

For safety, the switching range Asw has to be chosen such that the maximum loop gain is far below 0dB (between -10dB and -20dB). Therefore, in calculations the worst case Ast and Aac have to be taken. The electrical sidetone is the difference (in dB's) between the wanted receive signal on pin IR of the TEA106x/TEA111x and the unwanted part of the transmit signal on pin IR while having an equal signal level on pin LN for both the transmit and the receive signal. The electrical sidetone is dependent on line length and frequency. The worst case sidetone can be found by measuring the sidetone over the telephony band for several line lengths. The acoustic coupling is dependent on the environment of the telephone set. A way of determining the worst case acoustic coupling is to move a hand to the set as if pushing a button.
When automatic line loss compensation (AGC) is used, the transmit and receive gain of the TEA106x/TEA111x are reduced at high line currents (short lines), see [Ref.1]. This will reduce the loop gain at high line currents and makes the use of a smaller switching range possible.
If a certain minimum volume control range is required, the switching range must not be chosen smaller than the required volume control range.
In chapter 4.3 on page 44, a method for measuring the required switching range is given.

Ad 2. Determine dial tone detector level

The dial tone detector level is determined by the value of Rrsen according:

$$Vdialtone = 12.7\mu * Rrsen \text{ (in Vrms)}.$$

With an Rrsen of 10kΩ, the dial tone detector level will be 127 mVrms. This means, a continuous signal on the inputs RIN1 and RIN2 larger than 127 mVrms will be recognized as a dial tone.

Ad 3. Determine sensitivity

The sensitivity is set by Rrsen and Rtsen. The resistor Rrsen is already determined by the dial tone detector level. It must however be checked if the chosen value for Rrsen is a practical one. The reason for this is the dynamic range of the logarithmic compressor. A 'linear' compression is guaranteed when the currents flowing through pin RSEN are between 0.8μArms and 160μArms. This means that at nominal receiving signals the current through RSEN is preferably around 11μArms. This gives a maximum dynamic range of plus and minus 23dB. The same counts for pin TSEN.

The resistor Rtsen has to be chosen in such a way that both channels have the same priority for the duplex controller. This can be obtained by choosing Rtsen according:

$$20 * \log(Rtsen) = 20 * \log(Rrsen) - Atx1093 - Atx_{transmission} - Ast - Arx_{transmission} + Atsen + \frac{1}{2} * Aloop \text{ (in dB)}.$$

With Atsen = internal gain from MIC to TSEN = 40dB.

In this relation, the maximum loop gain and the worst case sidetone are used, see also Ad1. If it is preferred to give the transmit channel priority above the receive channel, Rtsen has to be made smaller. When the opposite is the case, Rtsen has to be made larger. With respect to the calculated setting, resistor Rtsen can be varied with plus and minus $\frac{1}{2}$ * Aloop (in dB's). In chapter 4.3 on page 44, a method for measuring the required Rtsen is given.

The capacitors Ctsen and Crsen form a first order high pass RC-filter with Rtsen and Rrsen respectively to reduce the influence of low frequency bumps on the switching behaviour. It is advised to choose the capacitors Ctsen and Crsen such that the corner frequency of the RC-filters are equal.

When the calculated sensitivity setting is implemented, subjective tests with a real telephone line will be necessary to come to the optimal sensitivity setting.

Ad 4. Determine timings.

The timings which can be set are: signal envelope timing and noise envelope timing for both channels, switch over timing and idling timing.

The signal envelope timing is set by the capacitors Ctenv and Crenv. Because of the logarithmic compression between TSEN and TENV respectively RSEN and RENV, the timing can be expressed in dB/ms. At room temperature the following relation counts:

$$\text{Timing} \approx \frac{I}{3 \times C} \qquad \text{(in dB/ms)}.$$

With I = charge or discharge current from pin TENV, RENV, TNOI or RNOI (in A)
C = timing capacitor Ctenv, Crenv, Ctnoi or Crnoi (in F).

With the advisory signal envelope timing capacitors Ctenv and Crenv of 470 nF, the maximum attack-timing of the signal envelopes will be around 85dB/ms (I=120μA). This is enough to track normal speech. The release-timing will be 0.7dB/ms (I=1μA). This is enough to smoothen the signal envelope and to eliminate the influence of room-echoes on the switching behaviour.

With the advisory noise envelope timing capacitors Ctnoi and Crnoi of 4.7μF, the attack-timing of the noise envelopes will be 0.07dB/ms (I=1μA). This is small enough to track background noise and not to be influenced by speech bursts. The maximum release-timing will be 8.5dB/ms (I=120μA). This is enough to track the signal envelope during release because the signal envelope release timing is 0.7dB/ms which is a factor smaller.
It is advised to choose the signal envelope timing and the noise envelope timing of both channels equal for optimum operation of the duplex controller. To have clearly determined timings, it is advised not to use capacitors with a high leakage current.

The switch-over timing is determined by the value of the switch-over capacitor Cswt. The idling timing is determined by the combination of Cswt and the idling resistor Ridt.
If the output current of pin SWT is Iswt, a voltage difference over Cswt can be obtained according:

$$\delta Vswt/t = Iswt / Cswt \text{ (mV/ms)}.$$

With the advised value for Cswt of 220 nF, the obtained voltage difference is 45 mV/ms. The switch-over time is dependent on the voltage difference which has to be generated on SWT. Suppose the set is in full Tx-mode, then the voltage on SWT will be
V(IDT)-400 mV, see figure 18. To reach Rx-mode a voltage difference of 580 mV must be generated to end up at a voltage of V(IDT)+180 mV. So in this case the switch over time will be 13ms. When the set is in Ix-mode the voltage on SWT equals the voltage on IDT. In that case Rx-mode or Tx-mode will be reached within 4ms.
The idling timing is determined by an RC-time constant. It is supposed that Ix-mode is reached when a time (Tidt) is passed of:

$$Tidt = 4 * Ridt * Cswt.$$

With the advised value for Ridt of 2.2MΩ, an idling time of around 2 seconds is obtained. To have a clearly determined idling timing, it is advised not to use a capacitor with a high leakage current.

Miscellaneous

When a handsfree telephone set is used at one end of the subscriber line and a conventional set at the other end, the user of the conventional set will think that the line is 'dead' when the handsfree set stays in receive mode while no signal on the line is present. This is avoided when the handsfree set switches over to a so called idle mode. This mode is incorporated in the TEA1093 and is placed exactly in between the transmit and receive mode. When it is desired to have an idle mode which is closer to transmit than receive mode, the circuit of figure 19 can be applied.

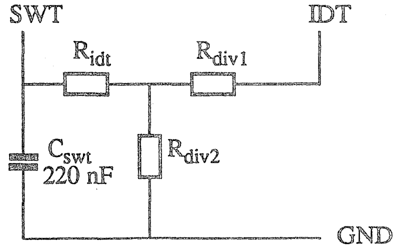


Fig.19 circuit for shifting the idle mode

With the circuit of figure 19, in idle mode, the voltage on SWT will not go to the voltage on IDT but to the voltage on IDT minus the voltage drop over Rdiv1. The voltage drop over Rdiv1 determines the shift of the idle mode (in dB's). This shift can be read from figure 18, when the voltage drop over Rdiv1 is taken as the x-axis value. The voltage on IDT is approximately 1.2V, so with for instance Rdiv1=33kΩ and Rdiv2=1MΩ the shift will be approximately 10dB. When dimensioning the resistor divider, it is advised not to choose Rdiv2 smaller than 1MΩ in order to limit the current drawn from IDT. By connecting Rdiv2 to VBB instead of to GND, the idle mode is shifted towards the receive mode.

In noisy environments, like offices, a handsfree set can show a popping behaviour in idle-mode (unwanted switching over from Ix to Tx-mode). This can be caused for instance by footsteps in the corridor. In the TEA1093, this popping behaviour is reduced by the implemented speech/noise threshold of 4.3dB. However, when a larger threshold is desired this can be achieved by connecting a resistor Rtnoi in series with Ctnoi.
When there is only noise present at the input of the envelope detector, the voltages on pins TENV and TNOI are equal. When, suddenly, a signal is present, the voltage on TENV will increase. Without Rtnoi, the voltage on TNOI will increase slowly because of the charging of Ctnoi by the 1μA internal current source. When a resistor Rtnoi is placed in series with Ctnoi, under the same conditions, this 1μA current source will cause a voltage jump on TNOI. This jump determines the shift of the speech/noise threshold. As depicted in figure 16, at room temperature, the 4.3dB threshold equals 13 mV. A resistor Rtnoi in series with Ctnoi will add an extra voltage to this threshold of 1μA*Rtnoi. When for instance a resistor of 10kΩ is chosen, the speech/noise level is increased up to 23 mV which is equal to 7.6dB at room temperature. The new speech/noise threshold is slightly dependent on temperature and the spread of the internal current source and therefore not as accurate as the internal 4.3dB. It is advised not to use a resistor larger than 15kΩ.

4. Application hints

During design phase:

The measurements on an application of the TEA1093, for instance to test the dial tone level or to check a gain, are normally done with the aid of sine waves. In case of pure sine waves instead of speech signals, the envelope detectors and noise detectors generate the same output after a certain time. So idle mode is then the result. To facilitate these measurements, the TEA1093 can be forced into transmit mode by activating DLC/$\overline{\text{MUTER}}$ (= make logic low). It can be forced into receive mode by activating MUTET (= make logic high). See specification for the levels.

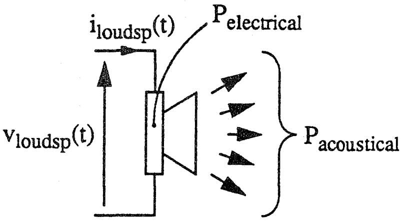
4.1 Adjust optimum loudspeaker output power

To adjust the maximum loudspeaker output power follow the steps described in chapter 4.1.1 on page 37 up to chapter 4.1.5 on page 42.

4.1.1 A few basic principles:

What determines the loudspeaker power?

Suppose:



The electrical loudspeaker power, $P_{electrical}$, equals:

$P_{electrical} = v_{loudsp}(t)\ i_{loudsp}(t)$

where:

$v_{loudsp}(t)$ = ac voltage over loudspeaker

$i_{loudsp}(t)$= ac current through loudspeaker

The acoustical loudspeaker power, $P_{acoustical}$, equals:

$P_{acoustical} = \eta\ P_{electrical}$

where:
η= characteristic sensitivity of the loudspeaker

The maximum acoustical loudspeaker power is achieved if:

η= maximum (see chapter 4.1.2 on page 38)

$v_{loudsp}(t)$= maximum (see chapter 4.1.3 on page 39)

$i_{loudsp}(t)$= maximum (see chapter 4.1.4 on page 42)

<u>and: loudspeaker resistance is optimized.</u> (see chapter 4.1.5 on page 42)

Optimizing loudspeaker resistance means: choose a loudspeaker resistance value that provides the maximum $v_{loudsp}(t)$ if the maximum $i_{loudsp}(t)$ is running.

4.1.2 Choosing a high characteristic sensitivity η

Compare the characteristic sensitivity of the following Philips speech loudspeakers (source Philips loudspeaker data book DC04 of 1990 [Ref 4]):

TABLE 4 Comparison loudspeaker sensitivities

type	characteristic sensitivity η	loudspeaker diameter	acoustical output power with reference to P_{out}
AD2071/Z	86 dB/W/m	2,5 inch	2 P_{out}
WD 20621/YL	83 dB/W/m	2 inch	P_{out}
WD 01621/YL	82 dB/W/m	1,5 inch	0.8 P_{out}

Difference between 86 dB and 83 dB is 3dB thus a difference in power of a factor 2.
This means that if a loudspeaker with a characteristic sensitivity of 86 dB produces the same acoustical power as a loudspeaker with 83 dB, the first one needs only half the electrical power.

From table 4 the conclusion could be drawn that the loudspeaker with the largest diameter produces the most acoustic output power, but this is not true because the case of the telephone set has a lot of influence on the acoustic output power. The combination of the several loudspeakers and the case should be checked on acoustic output power and compared. The highest characteristic sensitivity will always give the highest acoustic output power.

Conclusion: Take the loudspeaker with the best acoustic results.

4.1.3 Maximising ac voltage over the loudspeaker $v_{loudsp}(t)$

The voltage $v_{loudsp}(t)$ over the loudspeaker is maximum if the loudspeaker amplifier supply voltage (=VBB) is maximum. From figure 20 the upper limit of VBB can be determined.

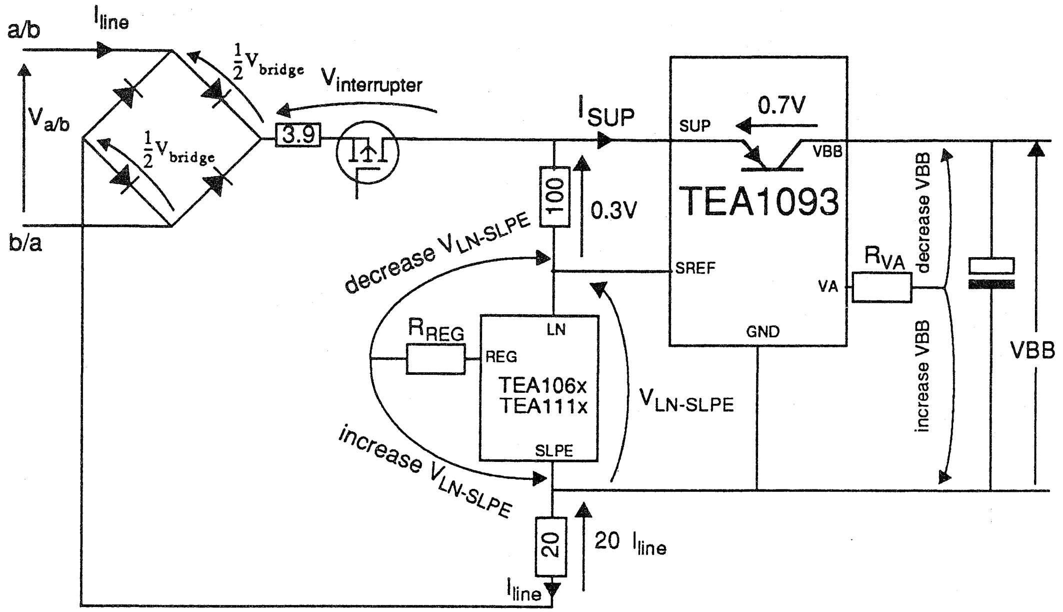


Fig.20 adjusting maximum VBB

In figure 20 only the items that have something to do with the dc adjustment are shown. It can be derived that:

$$V_{a/b} = V_{bridge} + V_{interrupter} + 0.3V + V_{LN-SLPE} + 20\Omega \times I_{line}$$

And:

$$V_{LN-SLPE} + 0.3V - 0.7V = VBB$$

For maximum VBB follow these steps (steps 4 and 5 are optional):

1. Find out what the maximum allowed Va/b is. This is normally specified at one certain line current. It can be derived from the DC current mask of the appropriate country.
2. Adjust R_{reg} (figure 20) that way that $V_{LN\text{-}SLPE}$ causes the maximum Va/b (first equation). If R_{reg} is connected to LN, $V_{LN\text{-}SLPE}$ decreases with decreasing R_{reg}, if R_{reg} is connected to SLPE, VLN-SLPE increases with decreasing R_{reg}. Without R_{reg} $V_{LN\text{-}SLPE}$ is its nominal value.

3. Adjust R_{VA} that way that between SUP and VBB remains a voltage difference of 0.7V (second equation). If R_{VA} is connected to VBB, VBB decreases with decreasing R_{VA} and if R_{VA} is connected to GND (=SLPE) VBB increases with decreasing R_{VA}.
4. $V_{interrupter}$ is in the example I_{line} x $R_{interrupter}$. $V_{interuper}$ needs to be as small as possible for saving power. A FET interrupter causes in most cases the smallest voltage drop. A Darlington causes a voltage drop of about 1 V. A bipolar (high voltage) transistor needs a high base current, thus spoils power too. Advise: use a FET.

Example of a FET interrupter:

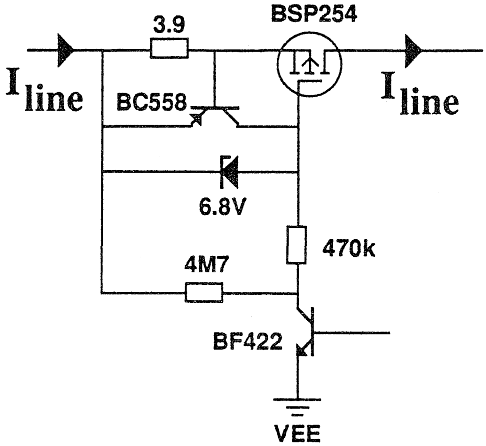


Fig.21 Example of FET interrupter

The dc resistance of the FET is approximately 10 Ω. So at 25 mA the power loss in the interrupter is:

In case of FET:	$(25\ mA)^2$ x 13.9Ω = 8.7 mW
In case of Darlington:	25 mA x 1 V = 25 mW

This is a difference of 16.3 mW for the loudspeaker. Compared with a loudspeaker power of 25 mW, which is a reasonable value, 16.3mW is a lot.

5. The power loss in the diode bridge (V_{bridge} x I_{line}) can be reduced by a "FET bridge". A FET bridge is shown below.

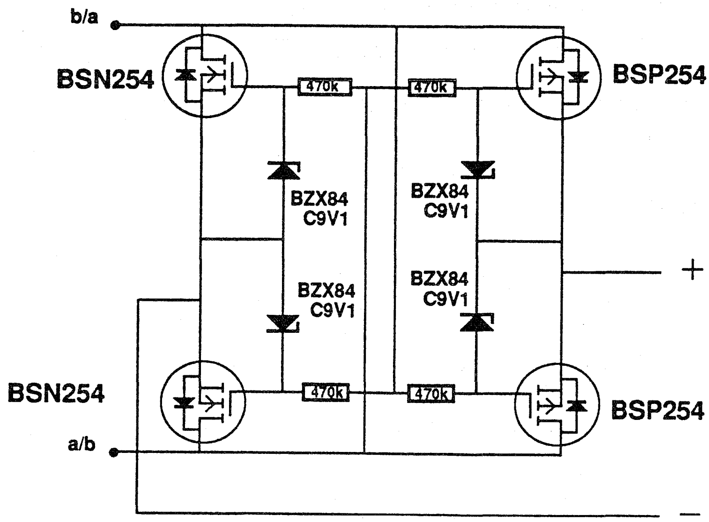


Fig.22 Increasing loudspeaker power with the aid of a rectifying FET bridge.

In the figure below is shown the forward voltage drop of a diode bridge and the FET bridge. I_{line} x voltage difference between diode and FET bridge equals the power that can be gained by using the FET bridge and is reflected by ΔP.This power can directly be added to the loudspeaker. Note that the FET's need to be high voltage FET's.

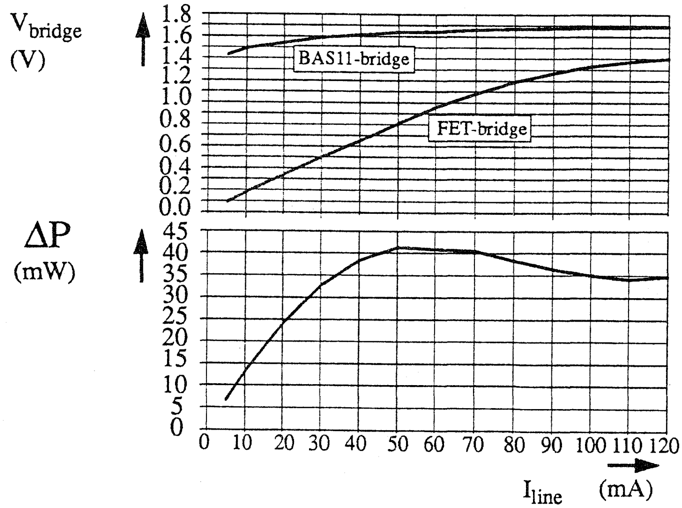


Fig.23 Power gain ΔP if rectifying FET bridge is used, versus line current.

Example:

Suppose in a practical set the line current is 20 mA, the loudspeaker impedance is 50 Ω and the supply currents through peripherals and transmission IC are in total 7 mA, then I_{SUP} = 13 mA. If we refer to figure 11, the loudspeaker power is about 7 mW. If the FET bridge is used the output power should be 7+25=32 mW, because ΔP at 20 mA (figure 23) is 25 mW. The result is an increase of 6.6 dB.

Example:

Suppose the line current is 40 mA, the loudspeaker impedance is 50 Ω and the supply currents through peripherals and transmission IC are in total 7 mA, then I_{SUP} = 33 mA. If we refer to figure 10, the loudspeaker power is about 120 mW. If the FET bridge is adjusted the output power should be 120+38mW=158mW, because ΔP at 40 mA is 38 mW. The result is an increase of 1.2 dB.

To achieve that this gain of power benefits the loudspeaker power, VBB has to be adjusted higher according step 1 up to 3 and the loudspeaker impedance has to be adjusted according chapter 4.1.5 on page 42.

So conclusion of these two examples is that the FET bridge has most effect if line current is small. Practically most problems concerning low loudspeaker volume occur in the low current area, so the FET bridge can be interesting in case of long lines.

4.1.4 Maximising ac current through the loudspeaker $i_{loudsp}(t)$

A built-in dynamic limiter controls the voltage VBB. If VBB becomes too low, for instance in case of a low line current, the dynamic limiter decreases the loudspeaker gain as result of which less power is taken from VBB and VBB will increase up to its adjusted value. Loudspeaker power is less now.

So the dynamic limiter is a tool that allows us to draw always as maximum loudspeaker power as possible, without disturbing operation of the TEA1093 and TEA106x/TEA111x -combination.

It is obvious that the currents used for peripherals like LED's, conducting transistors, microcontroller etc. can not be used to supply VBB and therefore they will cause that the dynamic limiter is activated earlier.

Thus: Use as less current as possible to supply peripherals.

4.1.5 Determine optimum loudspeaker impedance

The loudspeaker impedance is an important parameter and therefore it has to be chosen carefully. If $v_{loudsp}(t)$ and $i_{loudsp}(t)$ are maximised as indicated before, the loudspeaker impedance has to be chosen so that if the maximum $i_{loudsp}(t)$ is running, the maximum $v_{loudsp}(t)$ can be reached.

Example:

Suppose the loudspeaker resistance is 0Ω, if maximum $i_{loudsp}(t)$ is running the power is still 0W.

Example:

Suppose the loudspeaker resistance is 300Ω, maximum $i_{loudsp}(t)$ is 20 mA_{rms}, VBB is adjusted at 6V.
If maximum $i_{loudsp}(t)$ is running the amplitude of $v_{loudsp}(t)$ is:

$$20 \times 10^{-3} \times 300 \times \sqrt{2} = 8.5\,V$$

The amplitude of $v_{loudsp}(t)$ is limited by $\frac{1}{2}$VBB (approximately). So this case requires VBB=17V. Thus the maximum $i_{loudsp}(t)$ can not run.

A practical procedure to find the optimum loudspeaker resistance:

In most cases a choice has to be made out of the following loudspeaker resistance values: 25Ω, 32Ω, 50Ω, 110Ω. So the accuracy of the procedure need to be in the order of magnitude of 10%.

1. Determine the dc current that is available for the loudspeaker.
   This can simply be measured by decreasing the line current of a telephone set till VBB is going to drop. Now all line current is used to supply the telephone set.
   Suppose this current is I_p.
2. The optimum loudspeaker resistance is approximately:

$$R_{loudspeaker,optimum} \approx \frac{\frac{1}{2}VBB - 0.3V}{\pi \cdot (I_{line} - I_p)}$$

Example:

Suppose under point 1 is found: I_p = 15 mA.

Suppose under point 2 is determined that the loudspeaker power is to low at a line current of 25 mA: I_{line} = 25 mA.

Suppose VBB= 5V.

The formula under point 3 delivers:

$$R_{loudspeaker,optimum} \approx \frac{\frac{1}{2} \cdot 5V - 0.3V}{\pi \cdot (25mA - 15mA)} = 70\Omega$$

If the loudspeaker resistance is rounded to below up to 50 Ω, the possibility remains for higher line currents to create a larger loudspeaker power.

4.2 A practical way to adjust the switching behaviour.

1. Start with the nominal component values as advised in figure 28 on page 50.
2. Replace Rrsen and Rtsen by a variable resistor, or better by a resistor bank.The minimum value for Rrsen and Rtsen is specified as 1.2 kΩ at VBB=3.6V, or to be more precise the sum of the maximum peak currents through Rtsen and Rrsen has to remain below 1.2 mA. Take for Ctsen and Crsen a value of 1 μF, only during this measurement.
3. Find out what the minimum specified dial tone level is, for the country where the telephone set is to be used. Adjust Rrsen that way that the set is able to receive the dial tone without going into idle mode. If the set goes into idle mode Rrsen has to be adjusted at a lower value. From now on Rrsen may only be changed to a lower value.
4. Do a switching test, simply by two people calling each other. Let the set switch very often for instance by a so called count test: person A says 'one' person B (at the other side) says 'two', person A says 'three' and so on. Adjust Rtsen that way that a good transmit/receive balance is obtained. If the set is more likely to receive then Rtsen should get a lower value, if the set is more likely to transmit Rrsen should get a lower value.
5. If Rtsen and Rrsen are determined, Ctsen and Crsen can be determined. The two RC combinations Rrsen/Crsen and Rtsen/Ctsen form high passes. This means that the cut-off frequency has to be chosen below the frequencies that are intended to cause switching. If for instance a cut-off frequency of 1 kHz is adjusted, the frequencies above 1 kHz determine mainly the switching. Normally a cut-off frequency of about 150 Hz is adjusted to have a 1 dB attenuation at 300 Hz. Take care not to attenuate the dial tone.
6. Determine if the switching time is fast enough. Switching time, from receive to transmit and from transmit to receive is determined by Cswt. In the standard application 220 nF is used, the value depends on the taste of the set maker. A compromise has to be chosen between a quiet slow switching behaviour and a fast but restless switching behaviour. It is not comfortable if the set switches in the pauses between every word that is spoken, if only one person is talking. Practical values are between 220 nF and 80 nF.

4.3 Determine the switching range

This chapter describes how both the switching range can be adjusted. The switching range depends on the closed loop gain, that can be measured with the aid of figure 24. For determining the loop gain the worst-case acoustical coupling and sidetone must be adjusted.

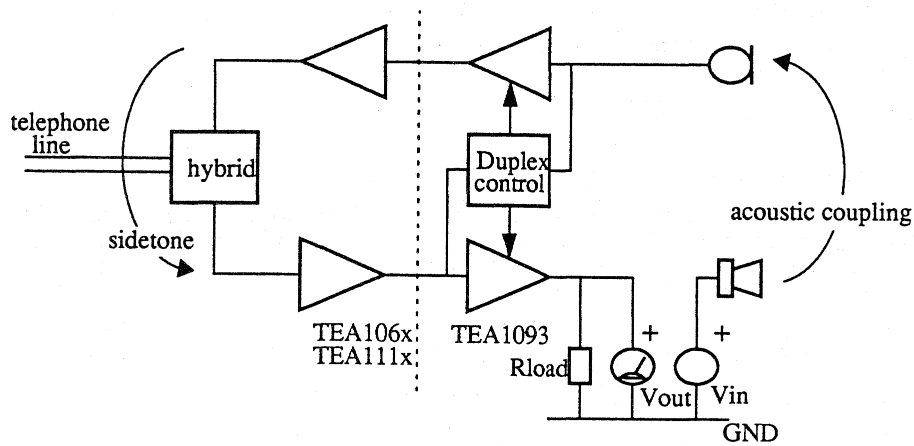


Fig.24 Simplified circuit diagram of measurement set up for measuring the loop gain

The loudspeaker is disconnected from the set and an electrical signal is applied to it. The acoustical signal coming from the loudspeaker is coupled to the microphone (Aac), transmitted to the line ($Atx1093 + Atx_{transmission}$), returns via sidetone (Ast) and is amplified again to the loudspeaker output of the TEA1093 ($Arx_{transmission} + Arx1093$). The output is loaded with a loudspeaker equivalent (Rload). The total gain of this loop is reduced with the switching range (Asw).
The gain from the loudspeaker connection (vin) to the loudspeaker equivalent (vout) is the loop gain (Aloop), given as:

$$20 * \log (vout/vin) = Aloop,$$

$$Aloop = Aac + Atx1093 + Atx_{transmission} + Ast + Arx_{transmission} + Arx1093 - Asw.$$

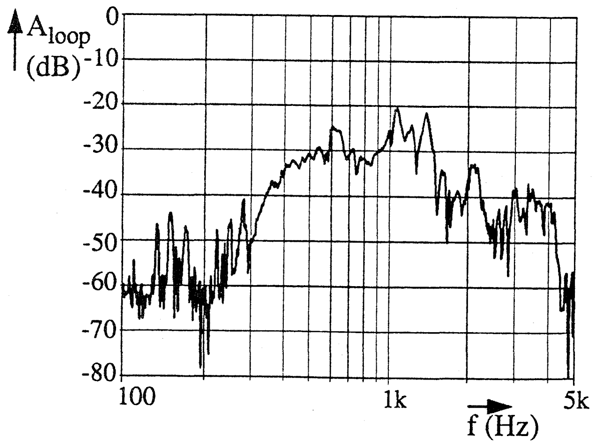


Fig.25 Example of loop gain versus frequency, according test set up of figure 24

4.4 A few lay out rules

4.4.1 Hints against ground bouncing

The TEA1093 is equipped with two ground pins: GND and GNDMIC. This enables the user to keep the sensitive grounds clean. The loudspeaker for instance uses a lot of power so the loudspeaker ground trace should not use (a part of) the microphone trace. The best performance is achieved if the following is done (where point 1 and 2 are most important):

1. Connect C_{VBB} to GND and uses this as ground "star point". Connect the loudspeaker ground trace by an unbranched trace to this star point. Connect the star point to SLPE.
2. Connect GNDMIC to Rswr, Rstab, Cswt, Rvol, Rmic and the star point.
3. Connect Ctsen, Crsen, Ctenv, Crenv, Ctnoi, Crnoi together and then to the star point.

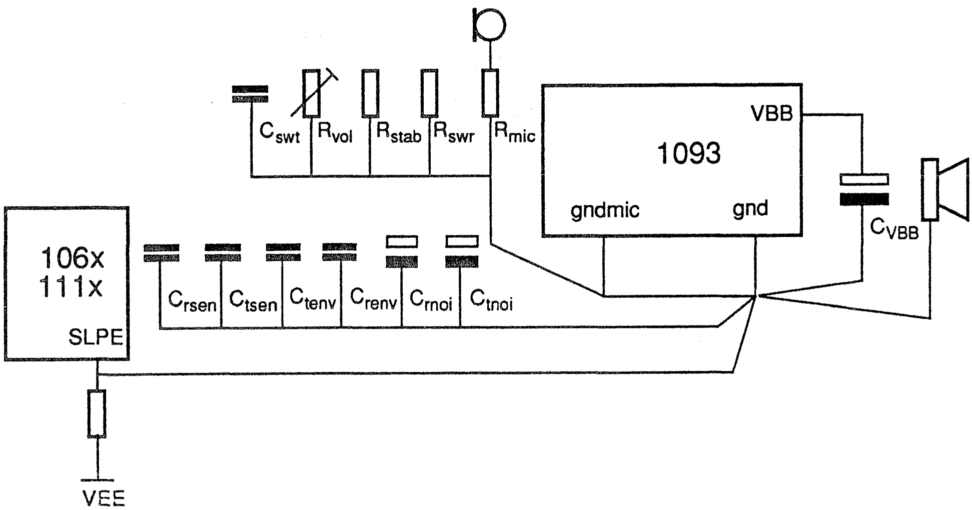


Fig.26 optimal layout of the ground traces

4.4.2 Hints to improve stability

1. It is important for stability to place Csup as close as possible to the TEA1093.
2. Place C_{SREF} close to the TEA1093

4.5 Application cookbook

In this chapter the procedure for making a basic application with a speech-transmission circuit of the TEA106x/TEA111x-family and the handsfree circuit TEA1093 will be given. With the aid of figure 28 on page 50, the design flow is given as a number of steps which should be made. As far as possible for every step also the components involved and their influence on every step are given. The preferred value is given between brackets { }.
More information on the setting of the TEA1093 can be found in the chapters given at every step. More information on the setting of the TEA106x/TEA111x-family see data handbook IC03, [Ref.5] and chapter 7 on page 64.

The application of figure 28 is an application of the TEA1093 handsfree circuit together with the TEA106x/TEA111x speech-transmission circuit. Switches are incorporated to switch over between handsfree operation and handset operation. A microcontroller and an interrupter are not incorporated. In figure 28, only a few components have a fixed value. These are the zener diode of 12Volts protecting the TEA106x/TEA111x and the resistors R5 and Rstab setting reference currents for the TEA106x/TEA111x and TEA1093 respectively. All other values will follow from the cookbook of figure 27 on page 49.

Worked out examples of applications of the TEA1093, following the cookbook are discussed in chapter 5 on page 50.

STEP	ADJUSTMENT (see figure 28 on page 50)
DC-settings. First adjust the DC setting of the TEA106x/TEA111x to the local PTT requirements and adjust VBB to have optimum output power, or follow the steps in chapter 4.1 on page 37. Background information in chapter 3.1 on page 12.	
Voltage LN-SLPE DC-slope Supply point VCC Artificial inductor Transmit current I_{tr} Voltage VBB Stability	R17; 315 mV is in series with LN R9 {20Ω}, R10 (also current protection) C1 {100μF} C3 {4.7μF} Rsref {100Ω}; I_{tr}=315 mV/Rsref (A) Rva; {V(SUP-VBB)=0.7 (V)} C_{SREF} {4.7 nF}, Csup {4.7 nF}, C_{VBB} {470μF}, if the line load is inductive: L_{SUP} {150 uH}.
Impedance and sidetone: After setting the set impedance, the sidetone has to be optimized for mean line length and line type. Also AGC can be chosen.	
Set impedance Sidetone AGC	Z1 (complex impedance allowed); R10 is in series R2, R3, R8, R11, R12, C12 R6
TEA106x/TEA111x Microphone and earpiece amplifiers, see chapter 7 on page 64: After the sensitivity of the microphone is adjusted, the gain can be adjusted to the desired value. Also the frequency curve can be set. The same counts for the earpiece.	
Sensitivity microphone Microphone gain Frequency curve and stability Earpiece gain Frequency curve and stability	R20,R21; microphone dependent R7; depends on TEA106x/TEA111x used C6; low pass with R7+3.5kΩ C20; stability for C20=10*C6 C8, C9, R14; high pass with input impedance R4; depends on TEA106x/TEA111x used C4; low pass with R4 C7; stability for C7=10*C4 C5; high pass with input impedance IR C2; high pass with earpiece impedance
TEA1093 Microphone amplifier, see chapter 3.2 on page 16: After the sensitivity of the microphone is adjusted, the gain can be adjusted to the desired value. Also the frequency curve can be set.	
Sensitivity microphone Transmit gain Frequency curve and stability	Rmic Rgat; Atx=20log(0.674*Rgat/Rstab) (dB) 9.6kΩ < Rgat < 96.3kΩ Cgat; low pass with Rgat, also stability Crmic; low pass with Rmic Cmic; high pass with input impedance (20kΩ)

STEP	ADJUSTMENT (see figure 28 on page 50)
TEA1093 Loudspeaker amplifier, see chapter 4.1 on page 37 background information in chapter 3.3 on page 20. Choose in case of line supplied telephone sets: SEL and in externally supplied sets if the loudspeaker amplifier output voltage limits the output power: BTL. Also the frequency curve can be set. The volume control potentiometer can be chosen as well as the dynamic limiter timing to have a minimal distortion.	
Receive gain	Rgar; SEL: Arx=20log(0.435*Rgar/Rstab) (dB) BTL: 6 dB higher than SEL 11.7kΩ < Rgar < 370kΩ
Frequency curve and stability	Cgar; low pass with Rgar Crin1; high pass with input impedance (20kΩ) Crin2 {Crin1} Clsp1; high pass with loudspeaker impedance
Volume control	Rvol {10kΩ}; 3dB reduction for each 950Ω
Dynamic limiter timing	Cdlc {470 nF}
TEA1093 Duplex controller, see chapter 3.4 on page 28 and chapter 4.3 on page 44: When all gains are adjusted the switching range can be determined by measuring the loop gain. Then the dial tone detector level can be set as well as the sensitivities of the duplex controller. Finally the timings of the envelopes and the switching are adjusted.	
Switching range	Loop gain: Aloop=Atx1093+$Atx_{transmission}$+Ast+ $Arx_{transmission}$+Arx1093+Aac-Asw<0 (dB) Choose Asw with safety margin of 10-20 (dB) Adjust Rswr; Asw=20log(Rswr/Rstab) (dB), with Rstab is fixed to 3.65kΩ
Dial tone	Rrsen; Vdialtone=12.7µA*Rrsen (Vrms)
Sensitivity	Note: Rrsen ≥ 1.5 kΩ Rtsen; For equal sensitivities of Tx and Rx: 20log(Rtsen)=20log(Rrsen)-Atx1093-$Atx_{transmission}$ -Ast-$Arx_{transmission}$+40+$\frac{1}{2}$ Aloop (dB) Note: Rtsen ≥ 1.5 kΩ.
Signal Envelope	Ctsen; high pass with Rtsen Crsen; high pass with Rrsen Ctenv {470 nF}, Crenv {470 nF};
Noise envelope	Maximum attack: 120µ/(3*Cenv) (dB/ms) Release: 1µ/(3*Cenv) (dB/ms) Ctnoi {4.7µF}, Crnoi {4.7µF};
Switch-over timing	Attack: 1µ/(3*Cnoi) (dB/ms)
Idle mode timing	Maximum Release: 120µ/(3*Cnoi) (dB/ms) Cswt {220 nF}; |ΔVswt|/t=10µ/Cswt (mV/ms) Ridt {2.2MΩ}; time constant: 4*Ridt*Cswt

Fig.27 Steps in the design flow of the TEA106x/TEA111x+TEA1093

5. Application examples

In this chapter some application examples of the TEA1093 handsfree circuit are given. In all examples only the essential elements are given. For instance no ringer or interrupter is included. The setting of the examples is made by following the cookbook of chapter 4.5 on page 47. The lower corner frequencies are chosen between 200Hz and 300Hz and the higher corner frequencies around 4kHz. Components which

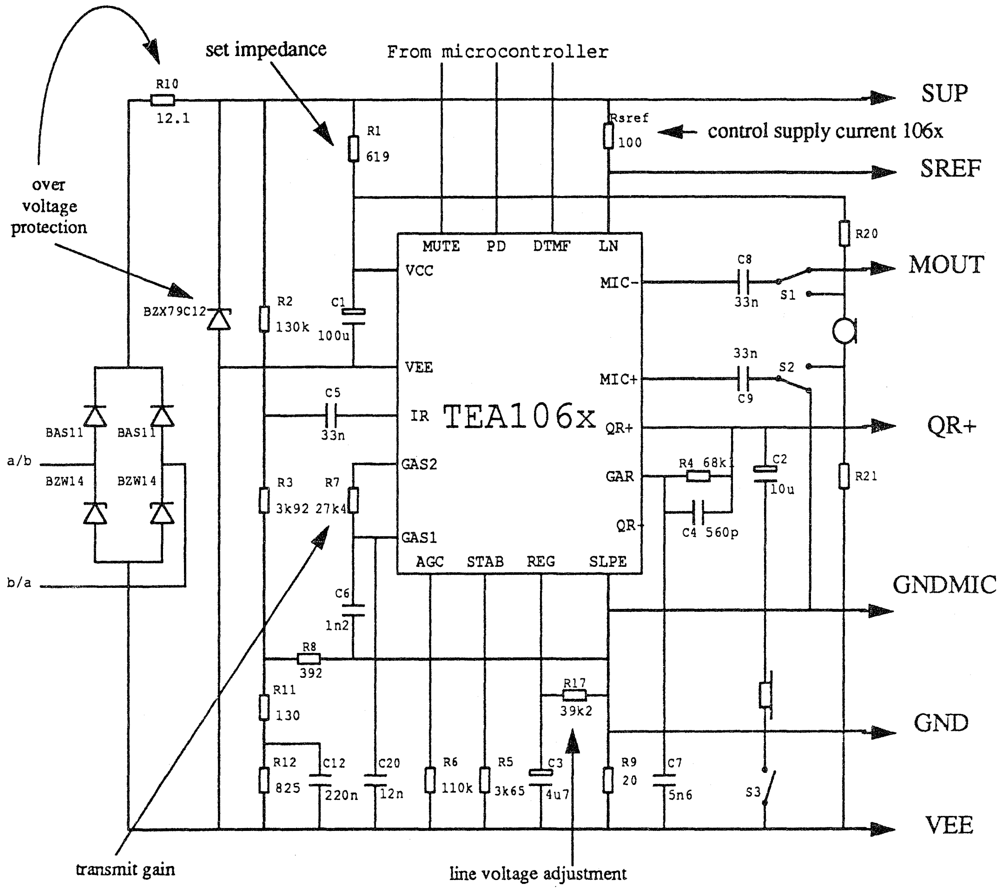


Fig.28 Basic application of the TEA106x and TEA1093, handsfree combined with handset configuration.

are not mentioned have the advised value. In chapter 7 on page 64 the gains and DC-settings of the TEA106x/TEA111x family can be found.

Figure 28 gives the basic handsfree application of the TEA1093 together with a speech transmission circuit of the TEA106x family. This basic application only incorporates handsfree telephony.

DC-settings: R17 is chosen 39.2kΩ which makes 4.5V to 5.0V after the bridge at 15 mA oflinecurrent, depending on the TEA106x used, see chapter 7 on page 64. Resistor R10 is 12Ω for current protection.

Impedance and sidetone: The set impedance is made approximately 600Ω for the telephony band with R1 and R10. The optimized sidetone bridge for this impedance in combination with a 5km cable of 0.5mm diameter copper twisted pair (176 Ω, 38 nF per km) is

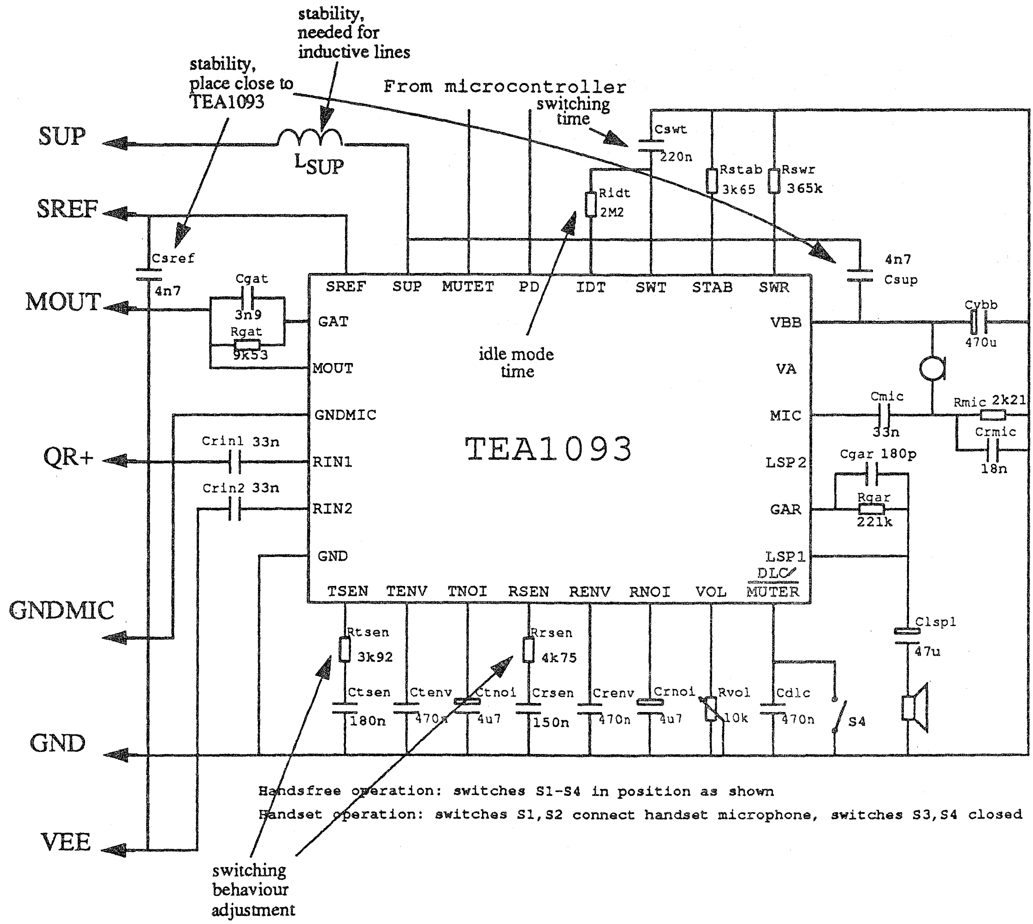


Fig. 28 Basic application of the TEA106x and TEA1093, handsfree combined with handset configuration.

taken from [Ref.1]. Also AGC is set with R6=110kΩ. This gives optimum result for an exchange of 48V and 600Ω in combination with the 0.5mm diameter cable (1.2dB attenuation per km).

TEA106x amplifiers: The microphone gain is set to the lowest value with R7=27.4kΩ (44dB for TEA1060/2/4/7/8). The receive gain is set to its maximum value of -1dB with R4=100kΩ (-32dB attenuation by the anti sidetone-bridge and 31dB gain from IR to QR+ for TEA1062/4/7).

TEA1093 amplifiers: The transmit gain is set to 5dB with Rgat=9.53kΩ. The receive gain is set to 25dB with Rgar=150kΩ. This results in an overall receive gain from line to loudspeaker output LSP1 of 24dB.

TEA1093 duplex controller: By measuring the loop gain following chapter 4.3 on page 44, the switching range can be determined. The loop gain is, of course, very dependent on the acoustic coupling between loudspeaker and microphone. With the set used for the measurements in chapter 4.3 on page 44, a switching range of 40dB (Rswr=365kΩ) leads to a gain margin of 18dB. The dial tone detector level was chosen 89 mVrms on the inputs RIN1 and RIN2 (meaning 100 mVrms on the line). This results in Rrsen=6.81kΩ.
When using the sensitivity measurement result of chapter 4.3 on page 44, 20log(vtsen/vrsen)=4dB, the resistor Rtsen follows out of:
20log(Rtsen/Rrsen)=4dB-(18dB/2)=-5dB or Rtsen=3.92kΩ.

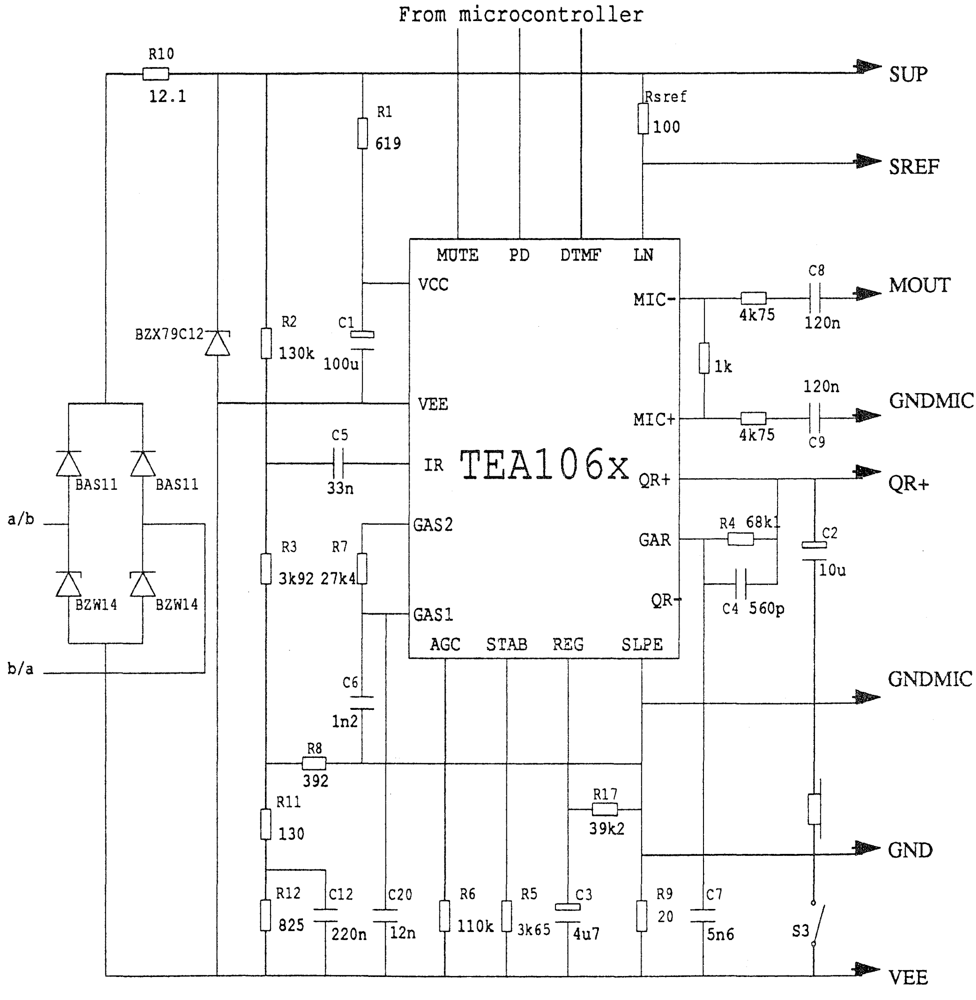
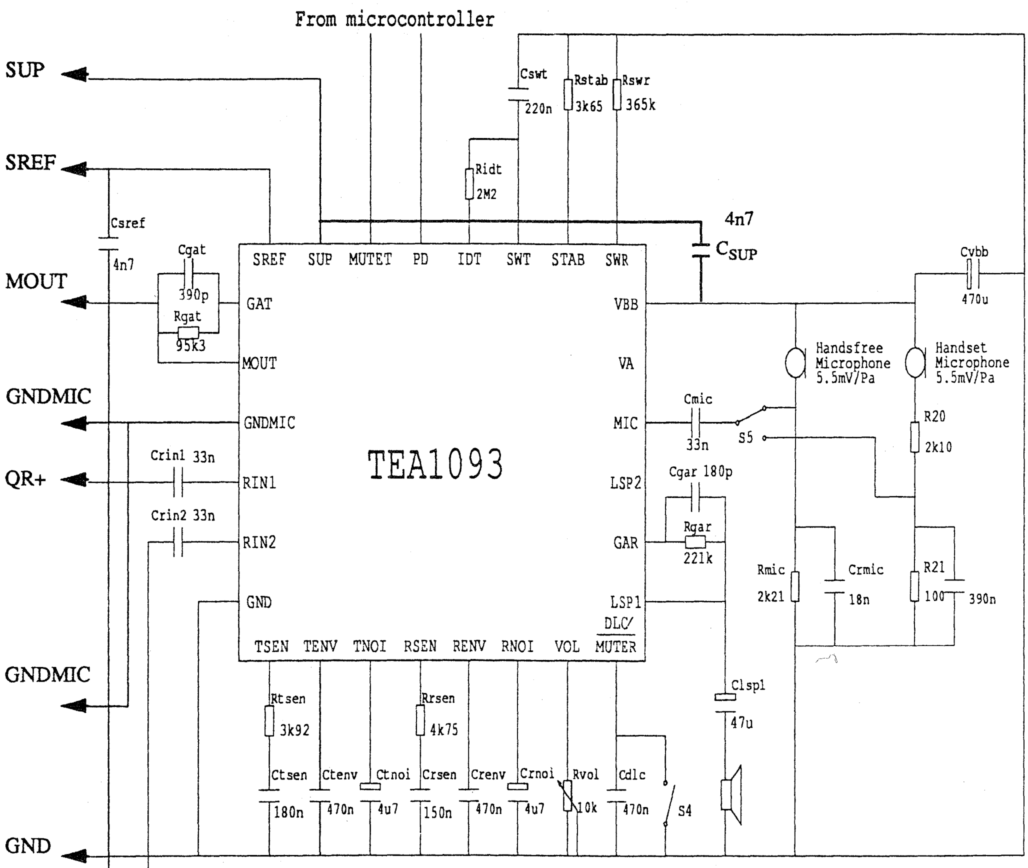


Fig.29 Handsfree/handset/switched listening-in application of the TEA1093

100 mVrms on the line). This results in Rrsen=6.81kΩ.
When using the sensitivity measurement result of chapter 4.3 on page 44, 20log(vtsen/vrsen)=4dB, the resistor Rtsen follows out of:
20log(Rtsen/Rrsen)=4dB-(18dB/2)=-5dB or Rtsen=3.92kΩ.

Figure 29 gives the handsfree/handset application of the TEA1093. This application incorporates both handsfree and handset operation. As an example the receive gain of the TEA106x is made 4dB lower with respect to figure 28. The receive gain of the TEA1093 is raised with 4dB to have the same overall receive gain. To detect a dial tone of 100 mVrms at the line, due to the lowered receive gain of the TEA106x, the resistor Rrsen is changed into 4.75kΩ. As can be seen in the formulas of chapter



Handsfree operation: switches S3-S5 in position as shown
Handset operation: switch S5 connects handset microphone, switches S3,S4 closed
Listening-in operation: switch S5 connects handset microphone, switch S3 closed, switch S4 opened

Figure 29 Handsfree/handset/switched listening-in application of the TEA1093

3.4.4 on page 33, Rtsen does not need to be changed. The switches can be either mechanical or electronic ones.
Figure 29 gives the handsfree/handset/listening-in application. In listening-in operation an acoustical loop is formed by the loudspeaker and the handset microphone. Without measures this loop can cause howling when the handset is close to the loudspeaker. This situation is quite similar to the handsfree howling problem, see figure 1 on page 7. Therefore, the TEA1093 can also be used to prevent howling

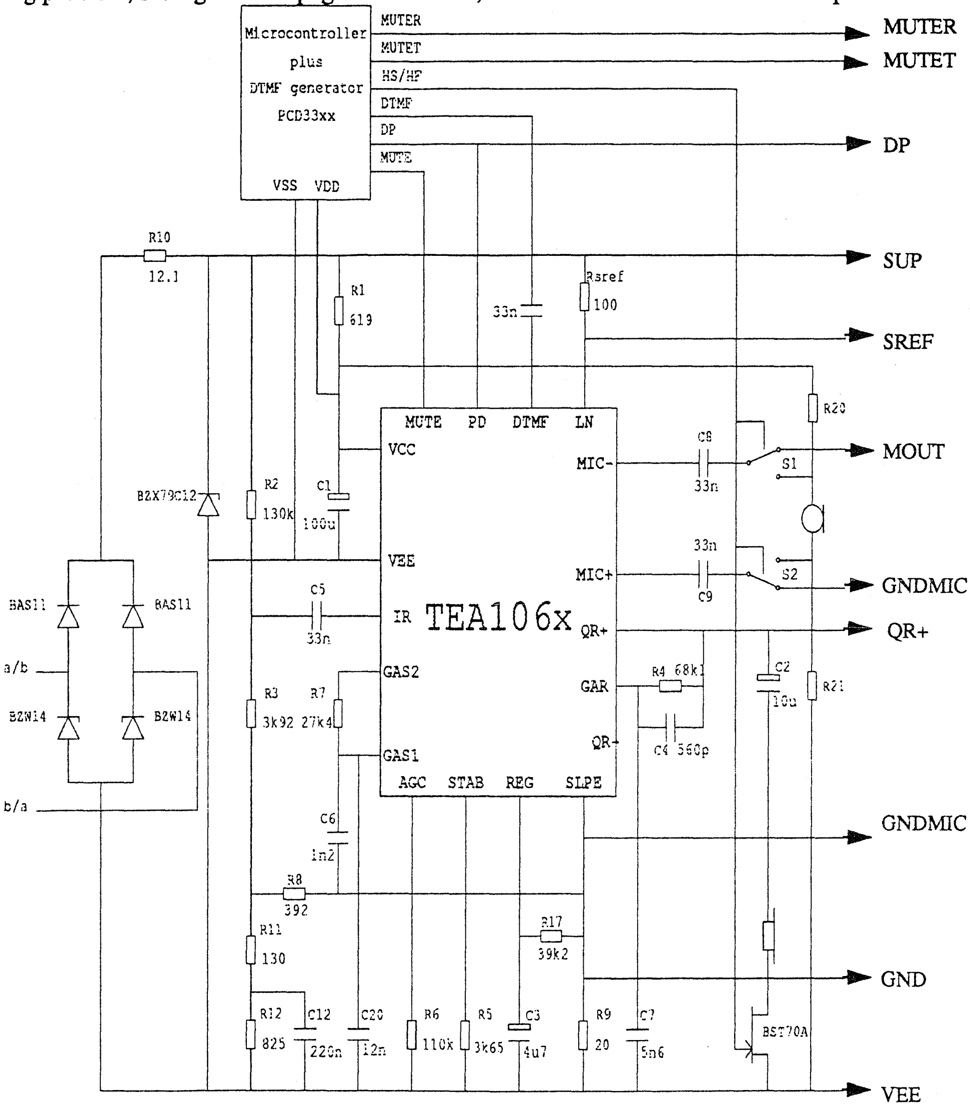
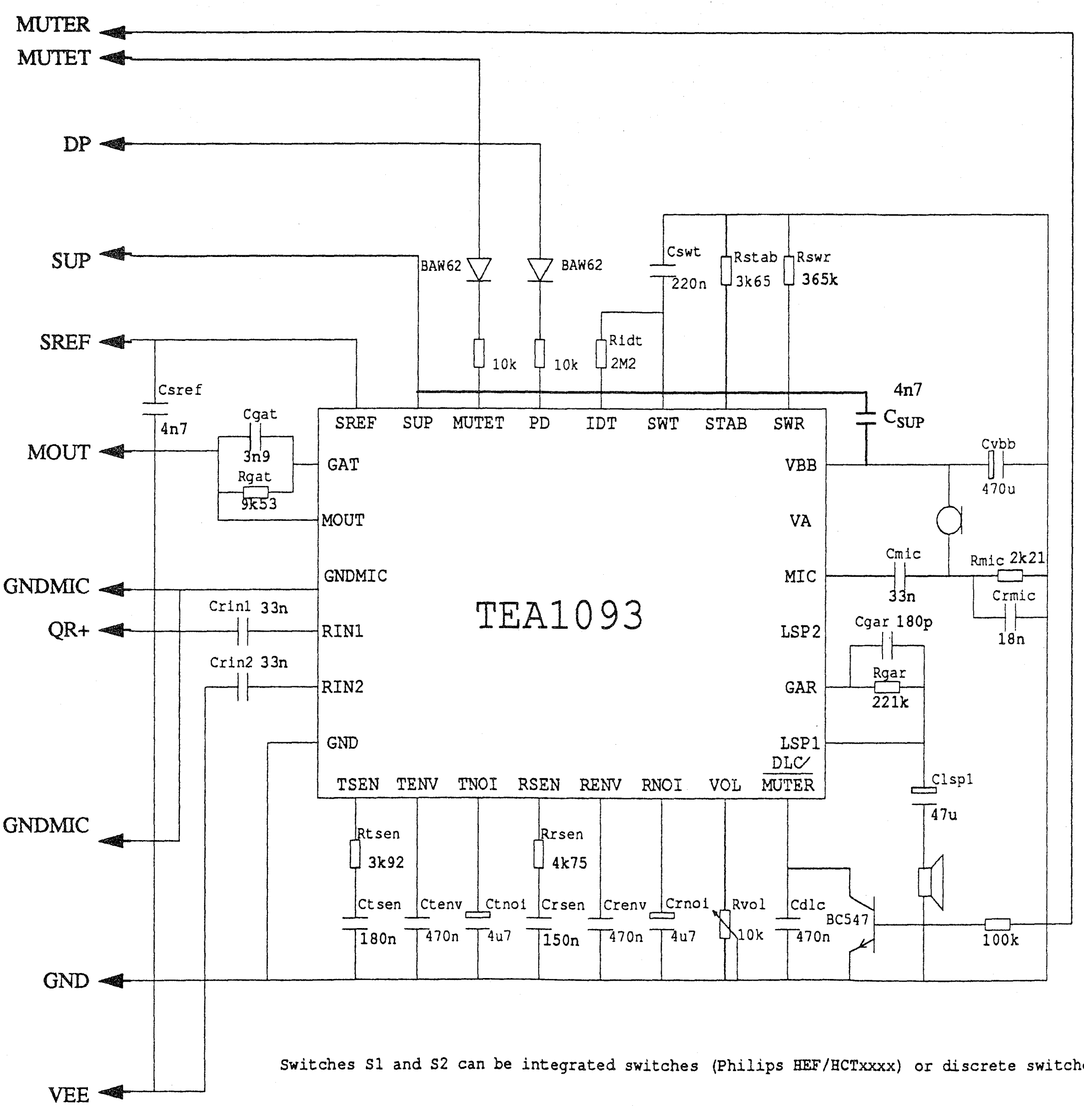


Fig.30 Application with microcontroller interfacing

in listening-in operation. This is done by connecting the handset microphone to the TEA1093 instead of connecting it directly to the TEA106x. The switching is now introduced in the listening-in loop. This switching has no effect on the earpiece level. Switch-over between the different modes is possible by the external switches.
The setting of the application is quite similar to the previous application examples. The receive channel is not changed. In the transmit channel an attenuator of 20dB is placed between the TEA1093 and the



Switches S1 and S2 can be integrated switches (Philips HEF/HCTxxxx) or discrete switches (JFET)

Figure 30. Application with microcontroller interfacing

TEA106x. This is done to reduce the sending noise on the line in receive mode, see chapter 3.2.1. To end up with the same overall transmit gain as in the previous examples, the transmit gain of the TEA1093 itself is raised with 20dB. To have a correct transmit gain during handset mode, the sensitivity of the handset microphone has been adjusted (R20, R21). In practice there will be no need for changing the switching range for listening-in operation with respect to handsfree operation.

Figure 30 gives an example of how the microcontroller can be connected to the application. The microcontroller is supplied between VCC and VEE. As a result the output signals of the microcontroller are referenced to VEE. All inputs of the TEA106x are referenced to VEE as well, so no extra interface is needed here. However, the inputs of the TEA1093 are referenced to GND (thus SLPE). Therefore the logic signals MUTET and DP need an interface. This is done via a diode and a resistor. The diode prevents currents from flowing out of the TEA1093 into the microcontroller when the logic outputs are low (thus VEE).The resistor
limits the current in case the voltage on VCC is higher than the voltage on VBB.

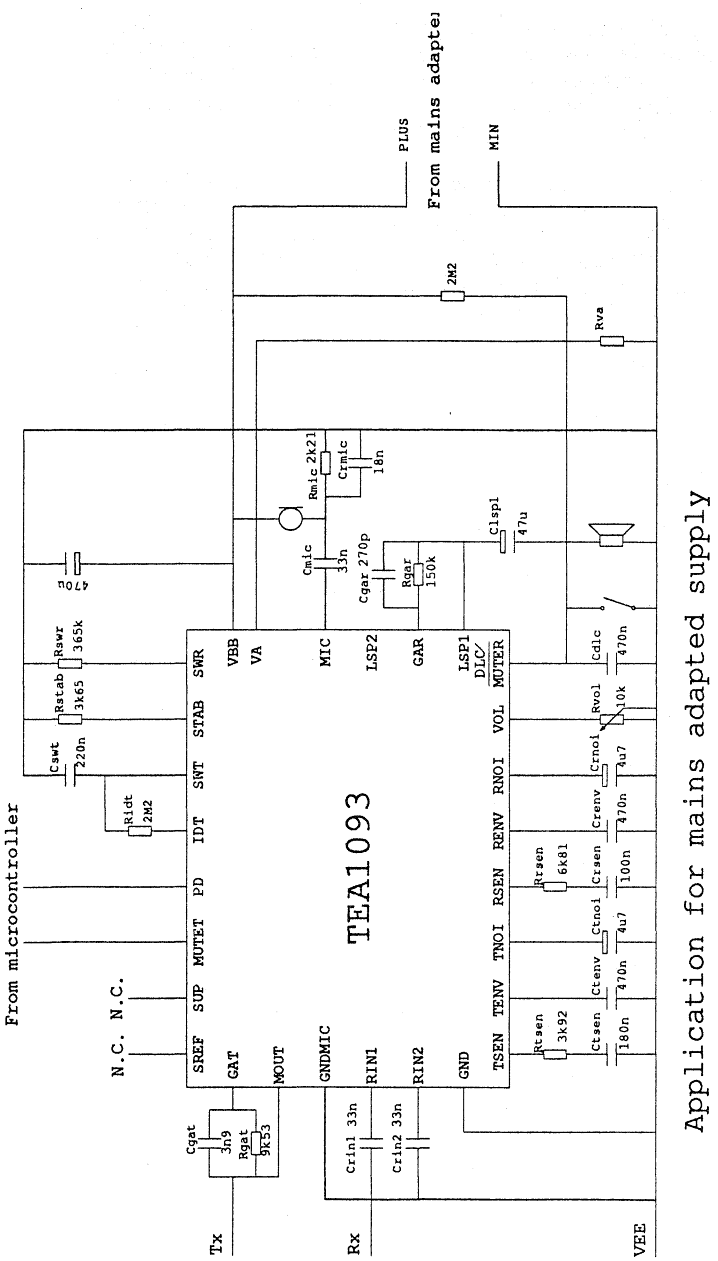


Fig.31 Application for mains adapted supply

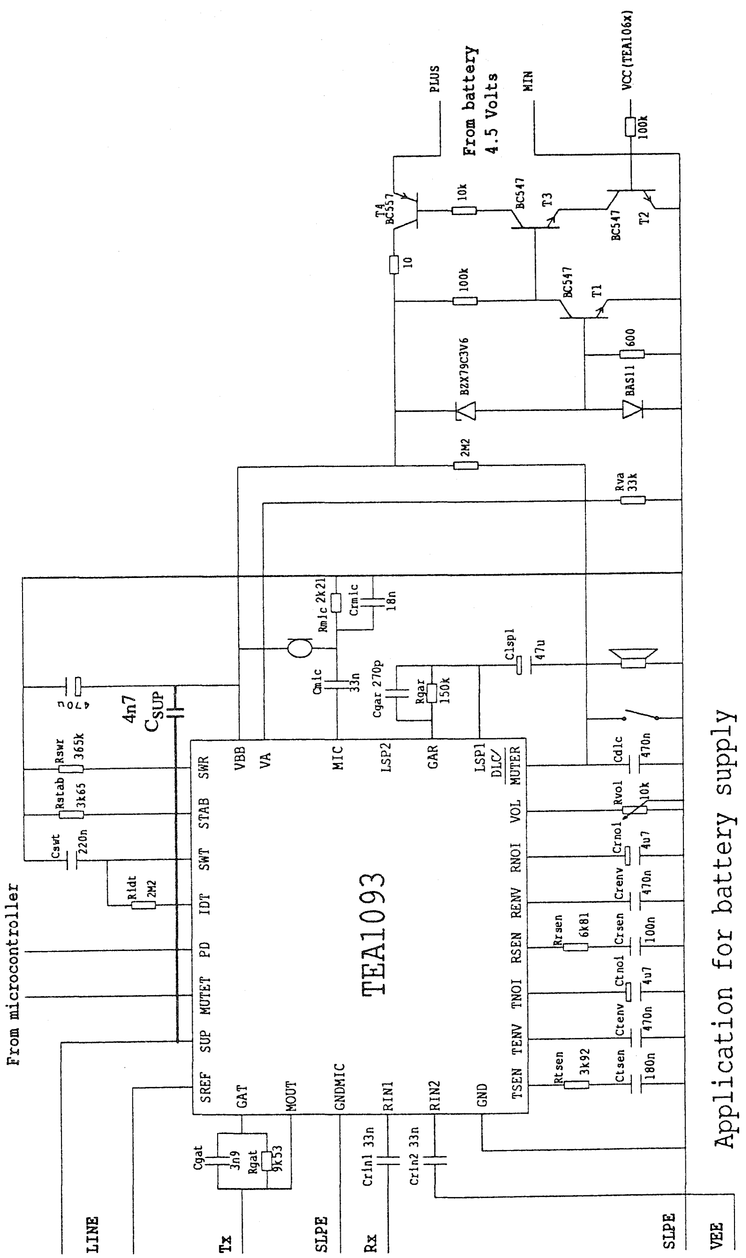


Fig.32 Application for battery supply

Figures 31 and 32 give an example of how to connect a mains adapted supply and a battery supply. In the application for mains adapted supply it is supposed that the TEA1093 is never powered from the line. Therefore, the supply input pins SUP and SREF are not connected and the ground GND is connected to VEE of the TEA106xTEA111x. With respect to the basic application, 2 resistors are added: Rva and one between VBB and DLC/$\overline{\text{MUTER}}$.
When a voltage is applied between VBB and GND, this voltage must be smaller than the voltage on VBB set by Rva (see [Ref.2] for correlation between RVA and VBB). If this is not the case, the voltage stabilizer of the TEA1093 will try to sink away all the current coming from the supply in order to reach its adjusted value. However, because VBB will never reach its set value in this set up, the current through the voltage stabilizer is zero. As a result, the capacitor Cdlc will be discharged to reduce the receive gain, see chapter 3.3.3. To prevent this, a resistor of 2.2MΩ is connected between VBB and DLC/$\overline{\text{MUTER}}$ to cancel the influence of the discharge current.
In the application for battery supply it is supposed that the TEA1093 is powered mainly from the line and from the battery when needed. Therefore, the supply input pins SUP and SREF and the ground GND are connected in the standard way. As with the application for mains adapted supply, resistor Rva and the resistor between VBB and DLC/$\overline{\text{MUTER}}$ are added. Between the battery and the TEA1093 an interface circuitry is placed. This circuitry takes care that the battery supplies the TEA1093 only when needed (low line currents, signal peaks) and that the battery is disconnected when enough current is available.
When approximately 1 mA or more is flowing through the zener and the 600Ω resistor, transistor T1 conducts and transistor T4 is non conducting. As a result the batteries are electrically disconnected from the TEA1093. The diode over the resistor of 600Ω is added to protect the current through transistor T1. At lower currents transistor T1 is non conducting and transistor T4 is conducting, provided that VCC is high. As a result the batteries supply the TEA1093 via the protection resistor of 10Ω. When VCC is low (meaning no telephone line is connected) the batteries are electrically disconnected from the TEA1093. The voltage on VBB is determined by the voltage over the zener and the diode.

6. Electromagnetic compatibility

With respect to electromagnetic compatibility (EMC) no common European or international specification yet exists. Also the measurement methods differ and are not always reproducible. At the application laboratory in Eindhoven (PCALE) the German current injection method is used (VDE 0878 part 200). It is a reliable method of measuring and is highly reproducible. The method is described in [Ref.3]. The hints for EMC of the TEA106x/TEA111x-TEA1093 combination given in chapter 6.2 on page 62 are based on this method. The same counts for the hints of the printed circuit board design given in chapter 6.1 on page 59.

6.1 Printed circuit board

In the current injection method, radio frequency (RF) signal currents enter the telephone set at the a/b wires and leave the set via any capacitive coupling to ground. Normally, in a telephone set the handset has the largest capacitance to ground and thus the main part of the RF signal current flows to ground via the handset. However, in handsfree operation the handset is not used. The RF signal current then flows to ground via the ground plane of the printed circuit board (PCB) and partly via the wires of the loudspeaker and microphone. Therefore, a proper PCB layout is essential for good EMC.

The best case is to create a ground plane on the PCB. The RF signals entering the PCB should be decoupled immediately to this ground plane. Preferably this ground plane is homogeneous and is not cut into parts by interconnection wires. To reach this a double layered PCB, with interconnection wires on one side of the board and a ground plane on the other side, is required. When interconnection wires within the ground plane are inevitable, the continuity of the ground plane should be restored. This is done by cross coupling these interconnection wires by jumpers and wires at the interconnect side of the PCB. In this way RF signal currents can flow freely over the ground plane.

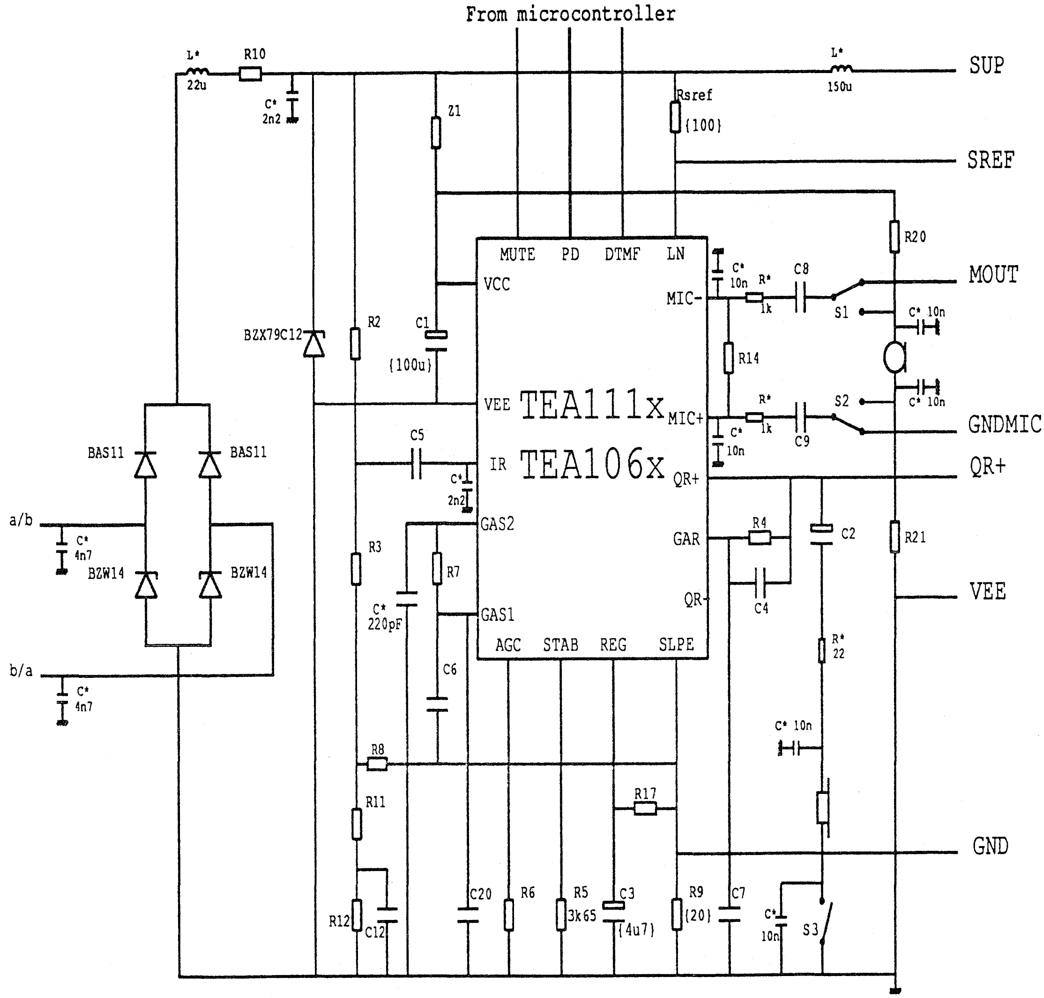
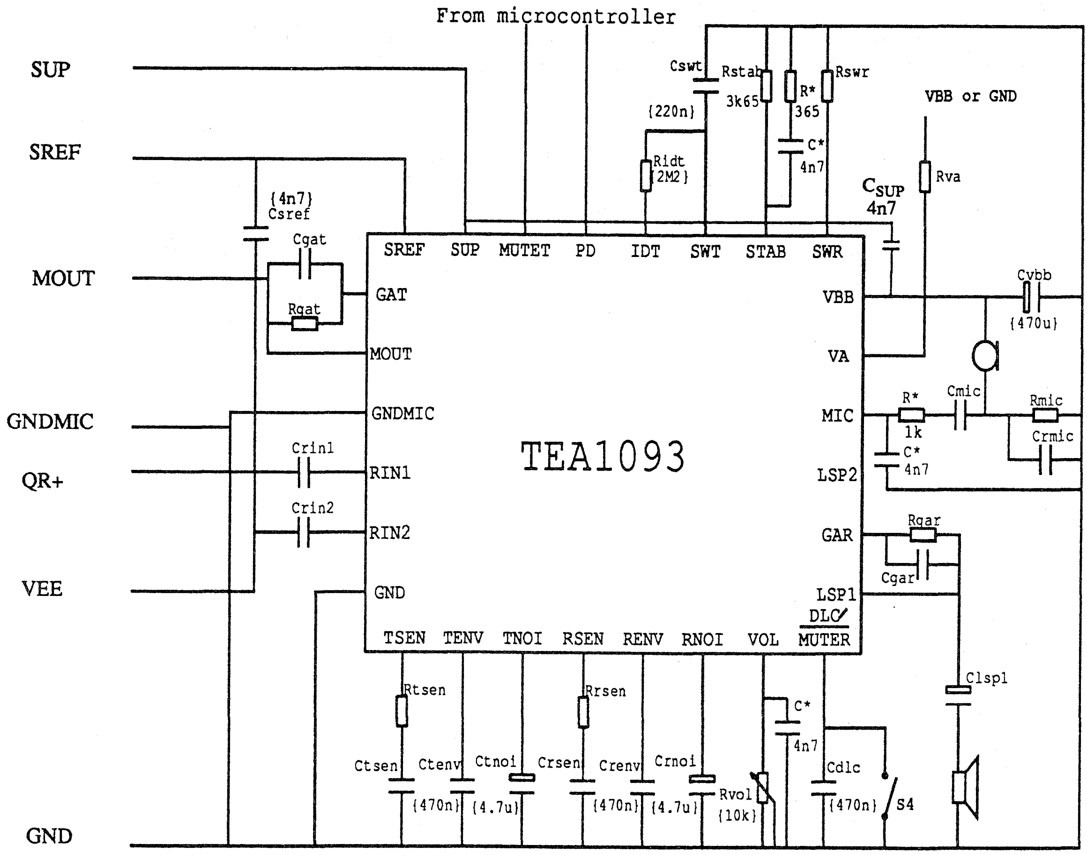


Fig.33 Proposal for EMC of the TEA106x plus TEA1093 application

Another measure is to keep the length of the wires between the different components as short as possible. Of course this measure is especially important for those wires which interconnect one or more RF signal sensitive parts, in particular the wire connected to SLPE which is also the reference for the TEA1093. As any current, RF signal currents prefer to flow to ground via the lowest ohmic path. Therefore, it



Components with value without brackets means fixed value.

Components with value between brackets means advised value.

Components without value have application specific values.

Components marked * are meant for EMC. Their advised value is given.

Handsfree operation: switches S1-S4 in position as shown

Handset operation: switches S1,S2 connect handset microphone, switches S3,S4 close

Figure 33. Proposal for EMC of the TEA106x plus TEA1093 application

should be noted that a wire of 10mm corresponds to an inductor of 10nH.
The wires to the handsfree microphone and the loudspeaker should be kept as short as possible as well. Normally, this is not a problem since loudspeaker, microphone and PCB are all within the same cabinet.

6.2 TEA106x-TEA1093 combination

All measures in this paragraph are based upon EMC experience and measurements at the TEA106x-TEA1093 application of figure 33 on page 60. This figure only gives the essential EMC components and it is just an example for a solution for better EMC performance since other applications and or different PCB layouts will demonstrate different behaviour. This chapter should therefore be interpreted as a guidance to reach proper EMC design.

This chapter is split into two parts. In the first part the standard measures for the TEA106x are given. In the second part it is explained how to improve the EMC behaviour of the TEA1093. When, after the proposed measures are taken, the EMC behaviour has to be optimized, it is advised to start with the transmit direction of the telephone set and to optimize the receive part thereafter. This because, due to sidetone, signals demodulated by the transmit channel will be seen in the receiver channel.
In the text below it is supposed that the printed circuit board on which the TEA106x-TEA1093 is built, is provided with a ground plane which is connected to VEE. It is preferred to place the components meant for EMC as close as possible to the pins, except when otherwise stated.
For more details on the EMC performance of the TEA106x, see [ref.3].

EMC behaviour can be improved by using the TEA111x instead of the TEA106x (see [Ref.9]). Also the PCA1070 has better EMC performance than the TEA106x (see [Ref.8]).

TEA106x

RF signals entering the printed circuit board at the a/b wires should be decoupled to the ground plane via capacitors, preferably placed as close as possible to the a/b connection. Since these capacitors will be in parallel with the set impedance, their value is limited. In practice, a total capacitance of 10 nF between the a/b wires may be applied without degrading the balance return loss.
Between the a/b wires and the ground plane two capacitors of 4.7 nF each are placed. Also a coil of 22μH can be placed in series with the line.

RF signals entering the inputs of the transmit channel (MIC+, MIC-) will be demodulated and amplified to the line. Because of the high gain (in the order of 50dB) the transmit channel is very sensitive to RF signals. Therefore, decoupling at the inputs is essential.
The inputs MIC+ and MIC- can best be decoupled by 2 capacitors of 10 nF connected to the ground plane. Also series resistors can be applied which in combination with the capacitors form low pass filters towards the inputs. Resistors of 1kΩ are advised which reduce the gain setting with less than half a dB, depending on the input impedance of pins MIC+ and MIC-. This can be compensated by adapting the transmit gain adjustment resistor R7 of the TEA106x.
In case a handset microphone is connected to the TEA106x microphone inputs via a long cord, extra decoupling is needed. This can be done by adding two capacitors of 10 nF each, placed between the cord connection and the ground plane, preferably as close as possible to the cord connection.

RF signals entering the inputs of the receive channel (IR) will be demodulated and amplified to the earpiece and, via the TEA1093, to the loudspeaker. Therefore, decoupling at the input is essential.
At the input IR of the TEA106x a capacitor of 2.2 nF connected to the ground plane is advised.
In case an earpiece is connected to the TEA106x via a long cord, extra decoupling is needed. This can

be done by adding two capacitors of 10 nF each, placed between the cord connection and the ground plane, preferably as close as possible to the cord connection. To prevent the remaining RF signals from entering the earpiece output stage of the TEA106x via the QR pin, and thus the loudspeaker amplifier inputs of the TEA1093, a series resistor should be applied to create a high ohmic path. The value of this resistor is dependent on the type of earpiece capsule used. When a dynamic capsule of 150Ω is used, a resistors of 22Ω is advised. This will reduce the gain setting with 1.19dB. This can be compensated by adapting the receive gain adjustment resistor R4 of the TEA106x.

Besides these essential measures some additional measures can be taken. A capacitor between GAS2 and the ground plane of 100pF can improve EMC in the transmit direction. A series combination of a resistor of 365Ω and a capacitor of 4.7 nF connected between STAB and the ground plane can improve EMC for both transmit and receive direction.

TEA1093

In series with pin SUP a coil of 150 μH can be used if the telephone line has inductive behaviour. The reference current source of the TEA1093 must have a high immunity because it sets the transmit and receive gains. Therefore, sometimes pin STAB has to be decoupled. This can be done with a series combination of a resistor of 365Ω and a capacitor of 4.7 nF connected between STAB and GND.

RF signals entering the input of the transmit channel (MIC) will be demodulated and amplified to the line. Because of the high gain (in the order of 50dB) the transmit channel is very sensitive to RF signals. Therefore, decoupling at the input is essential.
At the input MIC a capacitor of 4.7 nF connected to GND, and not to the ground plane, is advised. Also a series resistor can be applied which, in combination with the capacitor, will form a low pass filter towards the MIC input. A resistor of 1kΩ is advised which reduces the gain setting with 0.42dB. This can be compensated by adapting the resistor Rgat.
In case a handset microphone is connected to the TEA1093 microphone inputs, for instance with the application of figure 29, extra decoupling is needed. This can be done in the same way as in case the handset is connected to the TEA106x (thus extra decoupling to the ground plane and not to GND).

RF signals entering the volume control pin VOL will modulate the receive gain. Therefore, the pin VOL has to be decoupled. This can be done by connecting a capacitor of 4.7 nF between VOL and GND.

Besides these essential measures some additional measures can be taken. When the capacitor C_{VBB} is not placed very close to pin VBB, a small capacitor of 4.7 nF connected close to pin VBB will improve EMC behaviour.
To improve the EMC behaviour in the receive channel, the inputs RIN1 and RIN2 can be decoupled. This can be done by connecting 2 capacitors of 10 nF each between the inputs and the ground plane. Also series resistors can be applied which in combination with the capacitors form low pass filters towards the inputs. Resistors of 1kΩ are advised which reduce the gain setting with 0.42dB. This can be compensated by adapting the resistor Rgar of the TEA1093.
To improve EMC behaviour in the transmit channel, pin GAT can be decoupled to GND via a capacitor with the same value as capacitor Cgat. The influence of this capacitor on the transmission parameters can be neglected.

7. TEA106x quick reference data

	DC-CHARACTERISTICS (with slope resistance R9=20Ω)		
Member	V(LN-VEE) (in V) at I_{line}=15 mA	V(LN-VEE) (in V) at R(REG-LN)=68kΩ	V(LN-VEE) (in V) at R(REG-SLPE) (in Ω)
TEA1060	4.45 ± 0.20	3.80 + 0.25/-0.30	5.0 ± 0.30 at 39k
TEA1062	4.00 + 0.25/-0.45	3.50 ±....	4.5 ± .... at 39k
TEA1064B	3.50 ± 0.25	---	4.4 ± 0.35 at 20k
TEA1067	3.90 ± 0.25	3.40 ± 0.30	4.5 ± 0.30 at 39k
TEA1068	4.45 ± 0.25	3.80 ± 0.30	5.0 ± 0.35 at 39k

	SENDING GAIN		RECEIVE GAIN (from IR to QR+)	
Member	Setting range (in dB)	Gain (in dB) with R7=68kΩ	Setting range (in dB)	Gain (in dB) with R4=100kΩ
TEA1060	44 - 60	52 ± 1	17 - 33	25 ± 1
TEA1062	44 - 52	52 ± 1.5	20 - 31	31 ± 1.5
TEA1064B	44 - 52	52 ± 1	20 - 39	31 ± 1
TEA1067	44 - 52	52 ± 1	20 - 39	31 ± 1
TEA1068	44 - 60	52 ± 1	17 - 33	25 ± 1

	SENDING NOISE	
Member	Noise (in dBmp) with R7=68kΩ	Noise (in dBmp) with sending gain of 44dB
TEA1060	-70	-78
TEA1062	-69	-77
TEA1064B	-72	-80
TEA1067	-72	-80
TEA1068	-72	-80

For more data see data handbook [Ref.1]. For information about the transmission IC's TEA111x and PCA1070 is referred to [Ref.8], [Ref.9] and [Ref10].

8. List of abbreviations and definitions

Aac	Electro-acoustic coupling (electrically measured)
AGC	Automatic line loss compensation of the TEA106x/TEA111x
Aloop	Loop gain of a handsfree telephone set (must be < 0dB)
Arx	Receive gain of the TEA1093 (3dB to 39dB)

$Arx_{transmission}$	Receive gain of the transmission IC: TEA106x, TEA111x or PCA1070
ArxTEA1093	Receive gain of the TEA1093 (3dB to 39dB)
Ast	Electrical sidetone
Asw	Switching range (0dB to 52dB)
Atsen	Gain from MIC to TSEN of 40dB
Atx	Gain of the transmit channel of the TEA1093 (5dB to 25dB)
$Atx_{transmission}$	Gain of the transmit channel of the transmission IC: TEA106x, TEA111x or PCA1070.
AtxTEA1093	Gain of the transmit channel of the TEA1093 (5dB to 25dB)
BTL	Bridge tied load (loudspeaker between LSP1 and LSP2)
Cdlc	Dynamic limiter timing capacitor (470 nF advised)
Cgat	Capacitor over Rgat
Clsp1	DC blocking capacitor for loudspeaker
Cmic	Coupling capacitor at microphone input
Crenv	Capacitor determining the receive signal envelope (470 nF advised)
Crin1,Crin2	Couple capacitors at receiver input
Crmic	Capacitor over Rmic
Crnoi	Capacitor determining the receive noise envelope (4.7µF advised)
Crsen	DC-blocking capacitor of receive sensitivity setting
C_{SREF}	Stability capacitor of supply (4.7 nF advised)
Cswt	Switch over timing capacitor (220 nF advised)
Ctenv	Capacitor determining the transmit signal envelope (470 nF advised)
Ctnoi	Capacitor determining the transmit noise envelope (4.7 µF advised)
Ctsen	DC-blocking capacitor of transmit sensitivity setting
C_{VBB}	Buffer capacitor of supply (470 µF advised)
dBmp	0 dBmp equals 1 milli Watt in 600Ω, psophometrically weighted
DLC/$\overline{\text{MUTER}}$	Dynamic limiter timing adjustment and receiver channel mute pin
DP	Dial pulse, output pin on microcontroller
δVswt	Voltage difference on SWT
EMC	Electro Magnetic Compatibility: the collective noun for the susceptibility and the radiation of a circuit/apparatus.
GAR	Receiver gain adjustment pin
GAT	Microphone gain adjustment pin
GND	Ground reference pin
GNDMIC	Ground reference pin for microphone amplifier
HCT4053	Philips IC containing 3, 2-channel analogue switches
HCT4066	Philips IC containing 4 analogue switches
Icc	Current into pin VCC of TEA106x/TEA111x
IDT	Idle-mode timing adjustment pin
I_{line}	Line current
IR	Receive input pin of the TEA106x/TEA111x
I_{SUP}	Current flowing into pin SUP
I_{SUP}'	I_{SUP} minus supply current for internal circuitry connected to SUP
I_{SWT}	Current through pin SWT (typical 10µA)
I_{tr}	Bias current through output stage of the TEA106x/TEA111x (3 mA advised)
I_{tr1}	Collector current of the current switch transistor TR1
Ix-mode	Idle-mode: the mode which is halfway Tx-mode and Rx-mode.
LN	Positive line terminal pin of the TEA106x/TEA111x
LSP1,LSP2	Loudspeaker amplifier output pins
MIC	Microphone input pin
MIC+, MIC-	Microphone inputs pins on the TEA106x/TEA111x
MOSFET	Metal oxide field effect transistor

MOUT	Microphone amplifier output pin
MUTET	Transmit channel mute input pin
PCA1070	CMOS Transmission IC for high end telephone sets with programmable characteristics.
PCALE	Product concept & application laboratory Eindhoven
PCB	Printed circuit board
PD	Power down input pin
Power Down	Reduced current consumption mode during pulse dialling or flash
PTAT	Proportional to absolute temperature
PTT	Telephone company
QR,QR+	Telephone earpiece output on TEA106x/TEA111x
Rdiv1,Rdiv2	Resistors in voltage divider shifting the idle mode
RIN1,RIN2	Receiver amplifier input pins
RENV	Receive signal envelope timing adjustment pin
Rgar	Resistor setting receive (loudspeaker) gain (66.5kΩ for 18dB SEL)
Rgat	Resistor setting transmit (microphone) gain (30.1kΩ for 15dB)
Ridt	Resistor setting idle mode timing (2.2MΩ advised)
RF	Radio frequencies
Rload	Loudspeaker equivalent load resistor used for measurements
Rmic	Resistor setting the microphone sensitivity
RNOI	Receive noise envelope timing adjustment pin
Rrsen	Resistor setting sensitivity of the receive envelopes
RSEN	Receive signal envelope sensitivity adjustment pin
Rsref	Resistor determining I_{tr}
Rstab	Resistor setting an internally used PTAT current (3.65kΩ)
Rswr	Resistor determining switching range
Rtnoi	Resistor for increasing speech/noise threshold
Rtsen	Resistor setting sensitivity of transmit envelopes
Rva	Resistor setting voltage on VBB
Rvol	Volume control potentiometer (10kΩ advised)
Rx-mode	Receive-mode: the gain of the loudspeaker amplifier is at its maximum and the gain of the microphone amplifier is reduced
SEL	Single ended load (loudspeaker connected LSP1 or LSP2 to GND)
SLPE	DC slope pin on TEA106x/TEA111x
SREF	Supply reference input pin
STAB	Reference current adjustment pin
SUP	Supply input pin
Switching range	The difference between the maximum and the minimum gain in a channel as a result of the switching of the duplex controller (Asw)
SWR	Switching range adjustment pin
SWT	Switch-over timing adjustment pin
TEA106x	IC of the TEA106x speech transmission family: TEA1060/61, TEA1062, TEA1063, TEA1064, TEA1065, TEA1066, TEA1067, TEA1068.
TEA111x	IC of the TEA111x speech transmission family: TEA1112, TEA1112A, TEA1113, TEA1118.
TENV	Transmit signal envelope timing adjustment pin
THD	Total harmonic distortion
Tidt	Idle mode timing
TNOI	Transmit noise envelope timing adjustment pin
TR1, TR2	Switching transistors in the supply block
TSEN	Transmit signal envelope sensitivity adjustment pin
Tx-mode	Transmit-mode: the gain of the microphone amplifier is at its maximum and the

	gain of the loudspeaker amplifier is reduced
VA	VBB voltage adjustment pin
VBB	Supply output pin
VCC	Supply pin of the TEA106x/TEA111x
Vdialtone,Vdt	Dial tone detector level at the inputs RIN1 and RIN2
VEE	Reference pin on TEA106x/TEA111x
VOL	Receiver volume adjustment pin
Voltage stabilizer	Part of the IC that stabilizes the supply voltage on VBB

9. References

[Ref.1] Philips Semiconductors Data Handbook
Semiconductors for telecom systems - IC03
Philips Semiconductors, 1993

[Ref.2] data sheet
TEA1093 Handsfree IC
code number: 9397 750 00135

[Ref.3] Measures to meet EMC requirements for TEA1060-family speech transmission circuits
Mart Coenen, Klaas Wortel
PCALE report number: ETT89016

[Ref.4] Philips loudspeaker data book DC04, 1990.
code number: 9398 162 60011

[Ref.5] TEA1060 family versatile speech/transmission IC's for electronic telephone sets
Designers' guide by P.J.M. Sijbers
July 1987, 12NC 939834110011

[Ref.6] Application of the versatile speech/transmission circuit TEA1064 in full electronic telephone sets by F. van Dongen and P.J.M. Sijbers
PCALE report number: ETT89009

[Ref.7] Application of the speech-transmission circuit TEA1062
P.T.J. Biermans
PCALE report number: ETT89008

[Ref.8] High-end telephones with PCA1070, TEA1093 and PCD3755A
Report nr ETT/AN94069
Authors: J.C.F.van Loon, P.A.M. v.d. Sande, P.J.M.Sijbers

[Ref.9] Application of the TEA1112 and TEA1112A transmission circuits.
report number: AN95050
Authors: F.Courtois and F.van Dongen

[Ref.10] User Manual for the OM4757 Demonstration Board PCD3332-3/TEA1064B-1062/TEA1093-1094. Author: F.van Dongen

APPLICATION NOTE Nr	ETT/AN94004
TITLE	Application of the TEA1094 handsfree circuit
AUTHOR	C. H. Voorwinden
DATE	March 1994

Table of Contents:

Chapter 1. Introduction

The TEA1094 is a circuit which, in combination with a member of the TEA106x transmission circuits, offers a handsfree function. It incorporates a microphone amplifier, a loudspeaker amplifier and a duplex controller with signal and noise monitors on the transmit and the receive channel. In contrarary with the Philips handsfree circuit TEA1093, the TEA1094 has no integrated supply. This makes the TEA1094 most suitable for applications with mains adapted supply, such as cordless telephones and answering machines.
The function of the handsfree circuit will be illustrated with the help of figure 1.1.

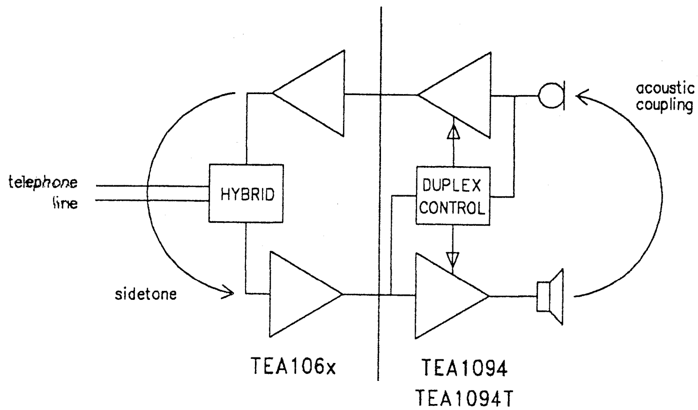


Figure 1.1: Handsfree telephone set principle

The left side of fig. 1.1 shows a principle diagram of a part of the TEA106x circuit by means of a receiving amplifier for the earpiece, a transmit amplifier for the microphone and a hybrid. The right side of fig. 1.1 shows a principle diagram of a part of the TEA1094 handsfree circuit by means of a microphone amplifier, a loudspeaker amplifier and a duplex controller.

As can be seen from fig. 1.1, a closed loop is formed via the amplifiers, the anti side-tone network and the acoustic coupling between loudspeaker and microphone of the handsfree circuit. When the loopgain is higher than one, the set starts howling. In a full-duplex application, this would be the case. To avoid howling, the loop-gain has to be much lower than one and therefore has to be decreased. This is done by the duplex controller.

The duplex controller of the TEA1094 monitors the signal and noise on both the transmit and the receive channel in order to detect which channel contains the 'largest' signal. As a result the duplex controller reduces the gain of the channel which contains the 'smallest' signal. This is done such that the sum of the transmit and receive gain remains constant.

As a result, the circuit can be in three stable modes to be referred to throughout this report:

1. Transmit mode (Tx-mode): the gain of the microphone amplifier is at its maximum and the gain of the loudspeaker amplifier is reduced.
2. Receive mode (Rx-mode): the gain of the loudspeaker amplifier is at its maximum and the gain of the microphone amplifier is reduced.
3. Idle mode (Ix-mode): the gain of the amplifiers are halfway their maximum and reduced value.

The difference between the maximum gain and the reduced gain is called the switching range.

This report gives a detailed description of the TEA1094 and its application with the TEA106x-family. The description is given by means of the block diagram of the TEA1094 (ch.2) and by discussing every detail of the sub-blocks (ch.3). The application is discussed by giving a guideline for application (the application cookbook ch.4) and by giving a number of worked-out applications including listening-in and cordless (ch.5). Also EMC aspects are discussed (ch.6). The appendices contain a measurement setup for the electro-acoustical adjustment of the TEA1094 handsfree application (A), quick reference data of the TEA106x family (B), a list of abbreviations (C) and application diagrams of the TEA1094 (D).

Chapter 2. Block diagram

In this chapter the block diagram of the TEA1094 is shown by means of figure 2.1. The pinning of the TEA1094 is given by means of figure 2.2. Also a short description of the block diagram is given including the function of the external components.

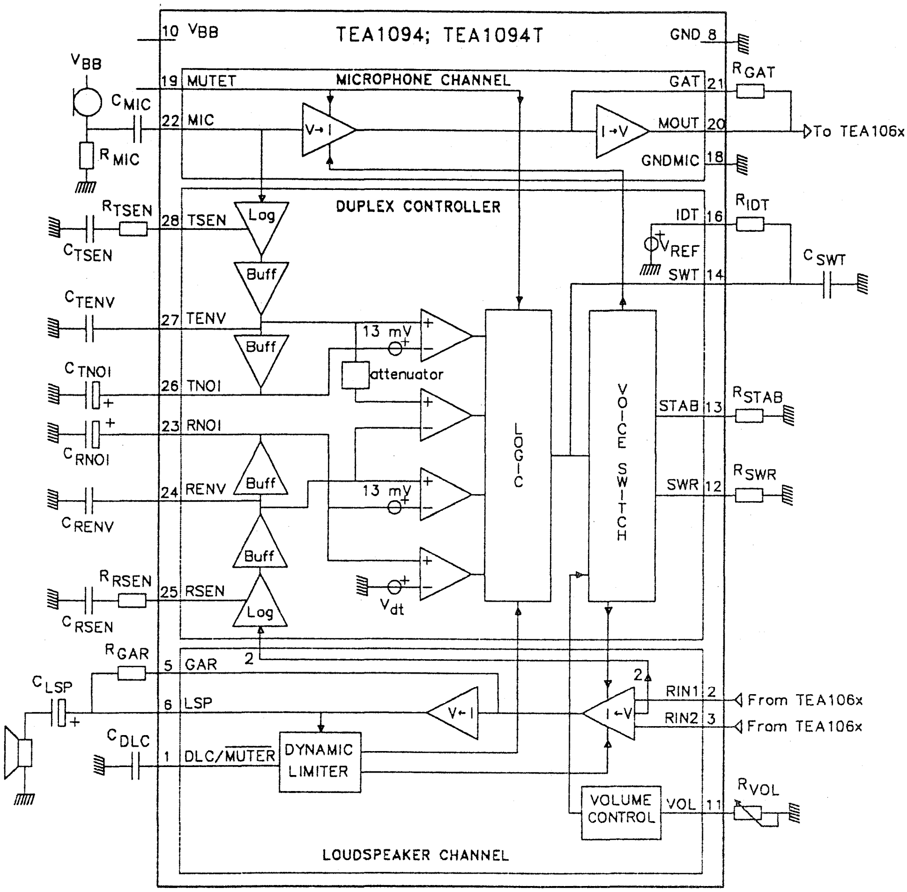
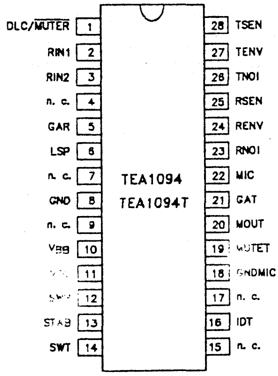


Figure 2.1: Block diagram of TEA1094



Pin	Name	Description
01	DLC/$\overline{MUTER}$	Dynamic limiter timing adjustment, receiver channel mute input
02	RIN1	Receiver amplifier input 1
03	RIN2	Receiver amplifier input 2
04	N.C.	Not connected
05	GAR	Receiver gain adjustment
06	LSP	Loudspeaker amplifier output
07	N.C.	Not connected
08	GND	Ground reference
09	N.C.	Not connected
10	VBB	Supply output
11	VOL	Receiver volume adjustment
12	SWR	Switching range adjustment
13	STAB	Reference current adjustment
14	SWT	Switch-over timing adjustment
15	N.C.	Not connected
16	IDT	Idle-mode timing adjustment
17	N.C.	Not connected
18	GNDMIC	Ground reference for microphone amplifier
19	MUTET	Transmit channel mute input
20	MOUT	Microphone amplifier output
21	GAT	Microphone gain adjustment
22	MIC	Microphone input
23	RNOI	Receive noise envelope timing adjustment
24	RENV	Receive signal envelope timing adjustment
25	RSEN	Receive signal envelope sensitivity adjustment
26	TNOI	Transmit noise envelope timing adjustment
27	TENV	Transmit signal envelope timing adjustment
28	TSEN	Transmit signal envelope sensitivity adjustment

Figure 2.2: Pinning of the TEA1094

In figure 2.1 it can be seen that the IC consists out of four parts: the supply, the microphone amplifier, the loudspeaker amplifier and the duplex controller. These blocks will be shortly described below including the function of the external components. The detailed description will follow in chapter 3.

Supply:

The circuit is supplied from pin VBB. A supply can be connected between VBB and GND.

Microphone amplifier:

The handsfree microphone signal is amplified from pin MIC to pin MOUT. The signal reference is pin GNDMIC, a 'clean ground' which has to be connected to GND. The sensitivity of the microphone can be set via Rmic, the signal is coupled in via Cmic. The gain of the amplifier can be set with Rgat. The amplifier can be muted by making pin MUTET high.

Loudspeaker amplifier:

A loudspeaker can be connected between output pin LSP and GND. Capacitor Clsp is used to block DC. The gain from the symmetrical input RIN1 and RIN2 to the output can be set via resistor Rgar. The volume of the receive signal can be adjusted by means of a potentiometer Rvol. To minimize distortion of the receive signal a dynamic limiter is incorporated of which the timing can be set with the capacitor Cdlc. The amplifier can be muted by making pin DLC/$\overline{MUTER}$ low.

Duplex controller:

From both the transmit and receive signal, signal and noise envelopes are made. The transmit signal envelope is on pin TENV, the receive signal envelope on pin RENV. The transmit noise envelope is on pin TNOI and the receive envelope noise is on pin RNOI. The timing of the envelopes can be set by the capacitors Ctenv, Ctnoi, Crenv and Crnoi. The sensitivity of the envelope detectors can be set by means of the RC-combinations Rtsen with Ctsen for the transmit envelopes and by Rrsen with Crsen for the receive envelopes. The resistor sets the sensitivity and the capacitor blocks the DC-component. Also a high-pass filter is created.
The logic determines to which mode (Tx,Rx or Ix-mode) the set has to switch over. The timing for switching to the Tx and the Rx-mode is set with capacitor Cswt. The timing for switching to the Ix-mode is set by the combination of Cswt and Ridt. The switching range is set by the resistor Rswr. Resistor Rstab has a fixed value.

Chapter 3. Description of the TEA1093

This chapter describes in detail the four blocks of the handsfree circuit TEA1093: the supply (3.1), the microphone amplifier (3.2), the loudspeaker amplifier (3.3) and the duplex controller (3.4). For each block the principle of operation is described and its adjustments and performance are discussed.

All values given in this chapter are typical and at room temperature unless otherwise stated. For more details of the TEA1094 specification, see [Ref.4].

Chapter 3.1. Supply block

Principle of operation

The TEA1093 handsfree circuit, see [Ref.2], has an integrated supply which effectively stabilizes a supply voltage and powers the loudspeaker out of the linecurrent. The TEA1094 has no integrated supply. This makes the TEA1094 most suitable for applications where the handsfree circuit is supplied from an external voltage source.

In figure 3.1.1, the supply arrangement of a TEA1094 with a TEA106x is shown.

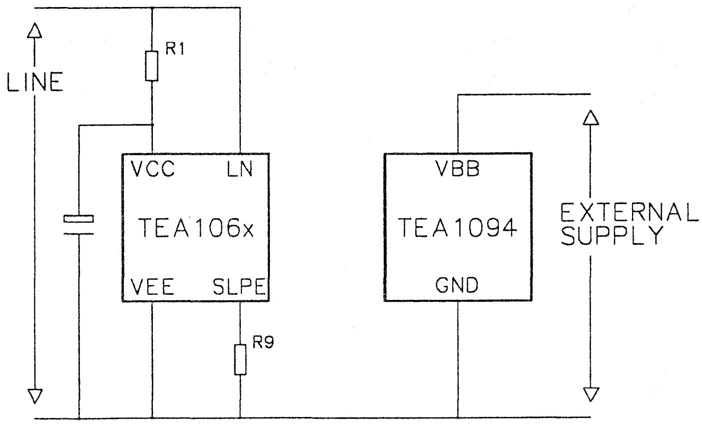


Figure 3.1.1: Supply arrangement TEA1094

As can be seen, the linecurrent flows through the TEA106x, while the TEA1094 is supplied from an external voltage source. The voltage source is connected between VBB and GND, while GND is connected to VEE of the TEA106x. In this way one reference for all signals is created.
It is also possible to connect the reference GND of the TEA1094 to pin SLPE of the TEA106x. In this way two references are created, SLPE and VEE. This makes the interconnection between the TEA1094 and the TEA106x less simple. This is also done in the line-powered applications of the TEA1093. Please refer to [Ref.5] for more details.

In case a line-powered handsfree application is made, preferably the TEA1093 is used because of its integrated supply. However, when the TEA1094 is used, the circuit can be supplied from the line via a large coil connected between the line and VBB. At VBB a capacitor has to be connected to serve as a reservoir. In this application, the TEA1094 has to be referenced to SLPE in order not to influence the transmission characteristics.

In figure 3.1.1 no galvanic seperation is drawn. When such is needed, the seperation can be done in the supply part or in the signal interfacing. In the supply part the galvanic seperation can be made in the mains-adapter. In the signal interfacing between the TEA1094 and the TEA106x, the galvanic seperation can be done either with transformers or opto-couplers. In chapter 5, application proposals which include galvanic seperation are given for a cordless telephone with handsfree base and for a telephone with answering machine.

Adjustments and performance

The voltage which can be applied between VBB and GND may vary between 3.3V and 12V. In case a lower voltage is applied, the circuit will strongly reduce the loudspeaker outputlevel, see chapter 3.3.3. The 12V is an absolute maximum rating. When a supply is used which, during transitions, can generate higher voltages, a 12V protection zener has to be applied between VBB and GND.

The current consumption of the TEA1094 is specified as Ibb=3.8mA at Vbb=5V. At higher voltages the current consumption slightly increases up to Ibb=4.8mA at Vbb=12V.

Chapter 3.2. Microphone amplifier block

In the first paragraph of this chapter the principle of operation of the microphone amplifier is described as well as its adjustments and performance. In the second paragraph, the mute transmit function is discussed.

3.2.1. Microphone amplifier.

Principle of operation

In figure 3.2.1 the block diagram of the microphone amplifier of the TEA1094 is depicted together with the interconnection with the TEA106x.

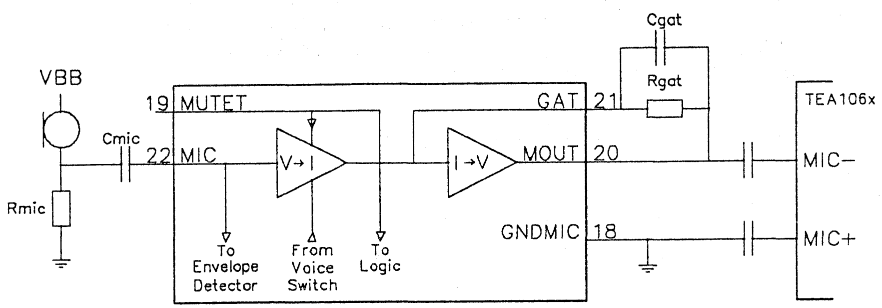


Figure 3.2.1: Block diagram of the microphone amplifier

As can be seen in figure 3.2.1, the microphone amplifier is referenced to pin GNDMIC instead of referenced to pin GND. This in order to prevent interference from other blocks within the TEA1094 (so called clean ground). The input and output signals of the microphone channel have to be referenced to GNDMIC. Pin GNDMIC itself has to be referenced to GND.

The input of the microphone amplifier is pin MIC. It is an a-symmetrical input well suited for electret microphones. Induced signals in the short wire between the microphone and pin MIC are assumed to be negligible. This in contrary with the handset microphone which is connected to the set via a long cord. The TEA106x family therefore has symmetrical microphone inputs.
The output of the microphone amplifier is pin MOUT. When interconnecting the TEA1094 and the TEA106x, pin MOUT is preferably connected to the TEA106x input pin MIC-. In that case the MIC+ pin of the TEA106x is connected to pin GNDMIC. It is advised not to reverse this interconnection.

As can be seen in figure 3.2.1, the microphone amplifier itself is built up out of two parts: a pre-amplifier and an end-amplifier. The gain of the pre-amplifier is determined by the duplex controller block, see chapter 3.4. The gain of the end-amplifier is determined by the external feedback resistor Rgat.

The overall gain (Atx) of the microphone amplifier from input MIC to output MOUT in Tx-mode is given as:

$$Atx = 20 * \log (0.674 * Rgat / Rstab).$$

With Rstab being the resistor at STAB of 3.65kΩ.

Adjustments and performance

A handsfree microphone, referenced to GNDMIC, can be connected to the input MIC via a DC blocking capacitor Cmic. Together with the input impedance of pin MIC of 20kΩ, Cmic forms a first order high-pass filter.
The handsfree electret microphone can be supplied from VBB. However, during normal operation VBB will contain a small ripple, for instance due to large loudspeaker signals. Electret microphones with a small power supply rejection ratio should therefore be connected via an RC smoothing filter as depicted in figure 3.2.2.

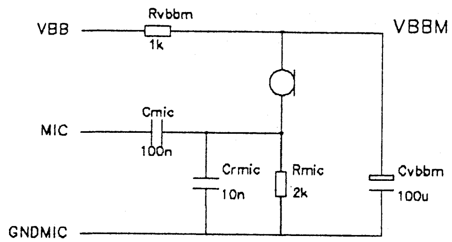


Figure 3.2.2: Supply arrangement for an electret microphone

As shown in figure 3.2.2, the RC smoothing filter is referenced to GNDMIC to have one ground reference for the whole microphone signal path. On the printed circuit board lay-out GNDMIC can best be connected with a separate wire to pin VEE of the TEA106x to reduce ground interference to a minimum.
The sensitivity of the electret microphone is set via resistor Rmic. By placing a capacitor Crmic over this resistor a first order low-pass filter is formed for the microphone signal.

Via the resistor Rgat, the gain of the microphone amplifier can be adjusted from 5dB to 25dB to suit application specific requirements. With resistor Rgat=30.1kΩ the gain equals 15dB with a tolerance of ±2dB.
Capacitor Cgat is applied in parallel with resistor Rgat to ensure stability of the microphone amplifier. Together with Rgat it also provides a first order low pass filter.

The input of the microphone amplifier can handle signals up to 18mVrms with 2% total harmonic distortion. However the microphone input signal is also used by the duplex controller, see chapter 3.4. At 10mVpeak at the input the positive part of the signal on pin TSEN starts clipping which might influence the switching behaviour. It is therefore advisory to keep the microphone input signal below this level.
The output drive capability at pin MOUT is 20μArms.

The output noise at MOUT of the TEA1094 is -100dBmp (psophometrically weighted P53-curve) at a gain of 15dB. With a sending gain of the TEA106x of 44dB the noise level on the line will be -56dBmp. The noise level of the TEA1094 is linear with the transmit gain. This means that at a transmit gain of 25dB the noiselevel at MOUT will be -90dBmp.
In transmit mode, the noise level will be at its maximum. In idle mode and receive mode, the noise level at MOUT will be lower because the contribution of the preamplifier is reduced then. However, the bottom level of the sending noise at MOUT is determined by the end-amplifier and is about -110dBmp, independent on gain.
The bottom level of the sending noise on the line is determined either by the TEA106x itself or by the noise at MOUT increased with the transmit gain of the TEA106x, whichever is largest. When, in receive mode, the noise on the line is determined by the noise at MOUT, it can be reduced by placing an attenuator between output MOUT and input MIC-. The attenuation has to be compensated by increasing the transmit gain of the TEA1094 with the same amount. Eventually, with enough attenuation, the bottom level of the sending noise on the line will be determined by the TEA106x itself. The use of an attenuator will not influence the amount of noise in transmit mode.
An application example with an attenuation network can be found in chapter 5. For the influence of the attenuator on the noise at the loudspeaker outputs, see chapter 3.3.1.

3.2.2. Mute transmit

During handsfree operation the microphone can be muted via MUTET so conversation cannot be heard by the other party. When the microphone amplifier is muted, automatically the TEA1094 switches over to Rx mode, see also chapter 3.4.
When a logic high is applied to MUTET, meaning the voltage on MUTET is higher than 1.5 Volts, the microphone pre-amplifier is muted. The end-amplifier can still be used by applying a signal on GAT. The obtained gain reduction is 80dB. The current which has to be sourced into pin MUTET when MUTET is high, is 5µA maximum.
When MUTET is a logic low, meaning the voltage on MUTET is lower than 0.3 Volts or pin MUTET is left open, the microphone amplifier is not muted.
The maximum allowable voltage on MUTET is VBB+0.4V, the minimum allowable voltage on MUTET is GND-0.4V.

Chapter 3.3. Loudspeaker amplifier block

The block diagram of the complete loudspeaker amplifier block is depicted in figure 3.3.1.

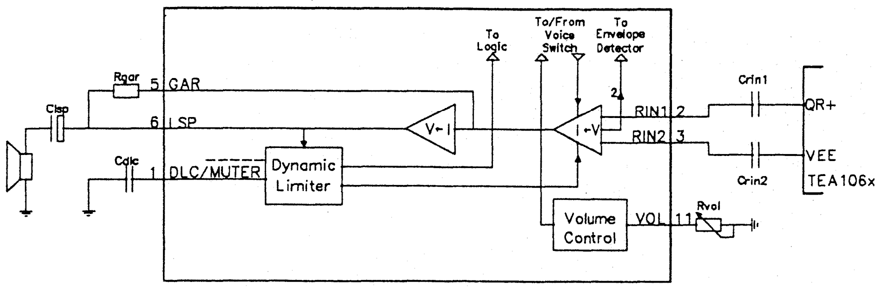


Figure 3.3.1: Principle of the loudspeaker amplifier

As can be seen in figure 3.3.1, the loudspeaker amplifier is built up out of three parts: the loudspeaker amplifier itself, volume control and dynamic limiter. In the first paragraph of this chapter the principle of operation of the loudspeaker amplifier is described as well as its adjustments and performance. In the second paragraph the same items are discussed of the volume control part and in the third paragraph of the dynamic limiter part. In paragraph 4 it is described how the receiver can be muted.

3.3.1. Loudspeaker amplifier

Principle of operation

As can be seen in figure 3.3.1 the input of the loudspeaker amplifier, pins RIN1 and RIN2, is symmetrical. The input RIN1 can be connected, via a capacitor, to the earpiece output QR+ of the TEA106x. When the TEA1094 is referenced to VEE, the other input RIN2 can be left open. In all other cases, RIN2 has to be connected, via a capacitor, to pin VEE of the TEA106x.
At the loudspeaker output LSP, the amplified receiving signal from QR+ is available. The output can drive a loudspeaker which is connected, via a capacitor, to GND.

As can be seen in figure 3.3.1, the amplifier itself is built up out of two parts: a pre-amplifier and an end-amplifier. The gain of the pre-amplifier is determined by the duplex controller block, see chapter 3.4. The gain of the end-amplifier is determined by the external feedback resistor Rgar. The overall gain of the loudspeaker amplifier (Arx) from inputs RIN1 and RIN2 to the output LSP in Rx-mode is given as:

$$Arx = 20 * \log (0.435 * Rgar / Rstab) \text{ dB}.$$

With Rstab being the resistor at STAB of 3.65kΩ.

Adjustments and performance

The input signal for the loudspeaker amplifier has to be coupled in via the capacitor Crin1 to block DC. Together with the input impedance of 20kΩ at RIN1, a first order high pass filter is introduced. When connecting the other input RIN2 to VEE, this also has to be done via a capacitor. This capacitor Crin2 preferably has the same value as Crin1 to obtain a good common mode rejection ratio. The input impedance of RIN2 is also 20kΩ.
The inputs can handle signals up to 390mVrms with a total harmonic distortion of 2%. Because of this it is advised not to connect RIN2 to QR- but only to VEE. The inputs RIN1 and RIN2 are biassed at around zero volts with respect to GND. By applying a signal to the inputs, they can become negative. The protection on these pins however is made different from other pins which makes it possible to make RIN1 and RIN2 as low as -1.2V without damaging the circuit.

A loudspeaker can be connected to the TEA1094 between LSP and GND. For the output is biassed at Vbb/2, a capacitor must be placed in series with the loudspeaker to block DC-current. Together with the impedance of the loudspeaker also a high pass filter is formed. Via the resistor Rgar, the gain of the loudspeaker amplifier can be adjusted from 3dB to 33dB. With resistor Rgar=66.5kΩ, the gain equals 18dB with a tolerance of ±2dB. A capacitor Cgar can be applied in parallel with resistor Rgar to provide a first order low pass filter.

In mains-adapted powered applications, the output power of the TEA1094 loudspeaker amplifier is limited by the maximum output current and the maximum output voltage swing. The maximum output current is dependent on the current the supply can deliver. The maximum output voltage swing is determined by the voltage supplied on VBB. For maximum voltage swing, the end-amplifier has a rail to rail output stage. The maximum voltage swing is dependent on the output current and the temperature, see also chapter 3.3.3.

In figure 3.3.2, the maximum output power in an 8Ω, 25Ω and 50Ω loudspeaker is depicted as a function of the voltage at VBB.

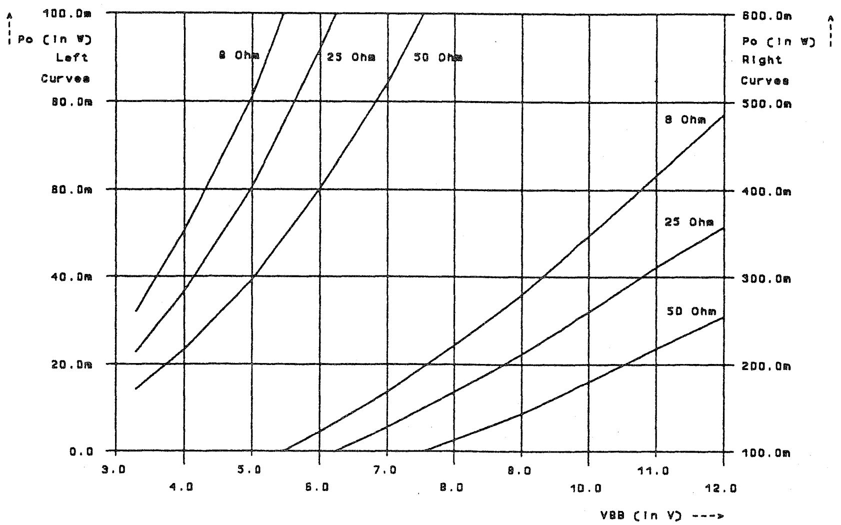


Figure 3.3.2: Maximum output power in the loudspeaker

The absolute maximum possible output power is limited by the maximum possible output current of the loudspeaker amplifier end-stage which is around 500mA peak. It is advised not to exceed this value since it might damage the output stage transistors.

The noise level at the output LSP is 80µVrms at a gain of 18dB and with the inputs RIN1 and RIN2 shorted with 200Ω. However in an application with the TEA106x, the noise at the outputs is larger than 80µVrms. This is due to the fact that the noise generated in the transmit channel of the telephone set is fed back via the sidetone network to the receive channel and is amplified to the loudspeaker outputs.
The maximum noise level at the loudspeaker outputs can be perceived in receive mode. With for instance an electrical sidetone of -12dB, and an overall receive gain of 24dB, the noise level at the loudspeaker output will be 12dB higher than the bottom level of the sending noise on the line. With the figures given in chapter 3.2.1 (-110dBmp at MOUT, transmit gain of the TEA106x of 44dB, thus -66dBmp on the line) this will lead to -54dBmp at the loudspeaker outputs. This level can be reduced by placing an attenuation network between MOUT and MIC- as described in chapter 3.2.1. Depending on the TEA106x used, the bottom level of the sending noise on the line can become lower than -77dBmp (44dB sending gain of the TEA106x) resulting in less than -65dBmp at the loudspeaker outputs.

3.3.2. Volume control

Principle of operation

Via the volume control block, the volume of the loudspeaker signal can be adjusted by the external potmeter Rvol connected to pin VOL. By turning the potentiometer, the gain of the loudspeaker pre-amplifier is varied. Volume control may not affect the transmit gain in transmit mode. To obtain this, the volume control acts upon the pre-amplifier via the duplex controller, see also chapter 3.4.

Adjustments and performance

Out of pin VOL a current Ivol, set by Rstab, see chapter 3.4, is flowing which is proportional to the absolute temperature (PTAT). At roomtemperature this current is around 10μA. Together with the resistance of the potmeter Rvol, the current Ivol creates a PTAT voltage on pin VOL. This PTAT voltage is processed by the volume control block. As a result, a temperature independent volume reduction in the loudspeaker signal of 3dB is obtained at approximately every 950Ω variation of the potentiometer Rvol.
This means that a linear potentiometer can be used to control the volume logarithmical, thus in dB's. With the advised value for Rvol of 10kΩ, the maximum gain reduction of the volume control is more than 30dB. The maximum gain reduction however is limited by the switching range, see chapter 3.4. When the resistor Rvol is zero, the receive gain is not influenced.

When digital volume control is desired this can be done as depicted in figure 3.3.3.

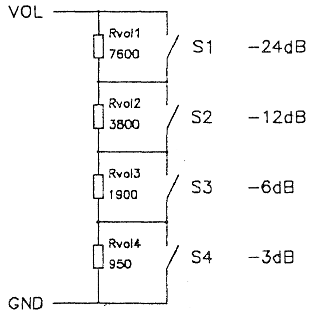


Figure 3.3.3: Digital volume control

With the 4 bit digital volume control of figure 3.3.3, 16 volume levels can be set via steps of 3dB (from 0dB to a maximum of 45dB of attenuation). The switches can be either MOSFETs or analogue switches, for instance the Philips HCT4066 type. It is advised not to use bipolar transistors as switches because of the saturation voltage of these devices. When a voltage is applied to pin VOL to control the volume, preferably this voltage has to be a PTAT voltage source. If not, the obtained gain reduction is no longer temperature compensated.

3.3.3. Dynamic limiter

Principle of operation

The dynamic limiter minimizes the distortion of the output signal by reducing the gain of the loudspeaker pre-amplifier when the output signal starts clipping or when too low supply conditions are detected. The amount of gain reduction is determined by the voltage on pin DLC/$\overline{MUTER}$. This voltage is varied by charging and discharging the capacitor Cdlc. The combination of charge and discharge currents and the capacitance of Cdlc sets the timing of the dynamic limiter.

Clipping of the loudspeaker output signal occurs when the output transistors are driven into deep saturation. To prevent hard clipping, deep saturation has to be prevented. The saturation voltage of the output transistors strongly depends on the output current and the temperature. In the dynamic limiter of the TEA1094 these effects are taken into account to have maximum output swing under any condition.
When the dynamic limiter detects the beginning of saturation, the capacitor Cdlc is discharged fast with a current of approximately 1mA, see figure 3.3.4 left hand side. As a result, the gain of the loudspeaker amplifier is reduced. The loudspeaker amplifier stays in its reduced gain mode until the peaks of the loudspeaker signal no longer start to cause saturation. The gain then slowly returns to its normal value by charging Cdlc with a current of 1µA, see figure 3.3.4 right hand side. In this way it is ensured that always the maximum reachable output power can be obtained at low distortion.

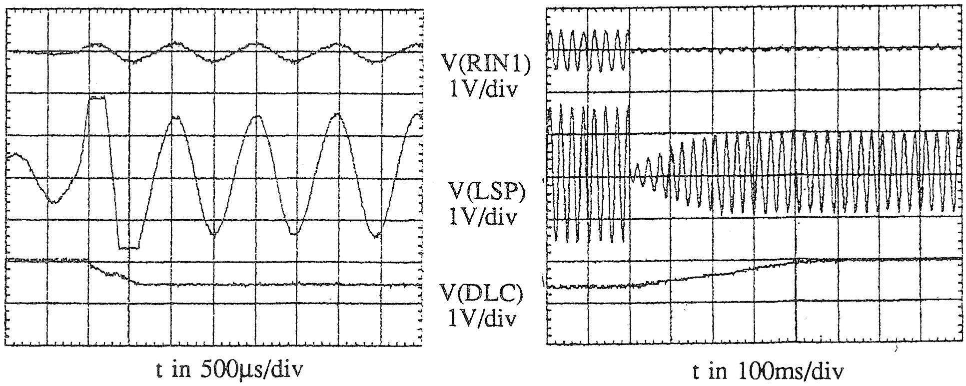


Figure 3.3.4: Action of the dynamic limiter clipping detector

Under too low supply conditions, the gain of the loudspeaker amplifier is reduced in order to prevent the TEA1094 from malfunctioning. Only the gain of the loudspeaker amplifier is affected since it is considered to be the major power consuming part. When the supply voltage drops below the internal threshold of 2.9V, the gain is reduced immediately regardless of the voltage on pin DLC. However, the capacitor Cdlc is also discharged to ensure stable operation of the limiter.

Adjustments and performance

The timing of the dynamic limiter is determined by the charge and discharge currents and by the capacitor Cdlc. The currents are internally fixed and cannot be changed. All charge and discharge time constants, and therefore the dynamic limiter timing, are proportional to the value of Cdlc. The only exception is when VBB drops below its threshold. In that case the gain is reduced immediately regardless of the voltage on Cdlc.
As a compromise between attack and release times of the dynamic limiter a capacitor of 470nF is advised. Larger values will give a smoother (slower) response while smaller values may lead to more distortion at lower frequencies. It is advised not to use a capacitor with a high leakage current in order not to influence the behaviour of the dynamic limiter.
If only a faster attack time is desired, it is possible to connect a low ohmic resistor in series with Cdlc of maximum 100Ω. In that case the relatively high discharge currents (200µA-1mA) will cause an instantaneous drop on pin DLC/$\overline{MUTER}$ when limiting action is needed. The instantaneous drop caused by the small charge current of 1µA will be negligible.

With Cdlc=470nF, the attack time for the clipping detector is in the order of a few milliseconds. The attack time when the circuit runs out of current is in the order of several seconds. The attack time when the supply voltage VBB drops below the threshold of 2.75Volts is less than 1 millisecond. The release time in all cases is in the order of a few 10 milliseconds.

When the dynamic limiter is acting, in practice the distortion of the output stage will stay below 5%. The dynamic limiter does not limit the distortion of the input stage.

Start-up behaviour

When the TEA1094 is started up, the starter of the dynamic limiter charges the capacitor Cdlc with a current of approximately 80µA. The starter stops when the voltage at DLC/$\overline{MUTER}$ has reached a value of around 1.6V. Then a current of 1µA charges the capacitor further, up to a voltage of around 1.9V. At that point the voltage on DLC/$\overline{MUTER}$ is limited. The starter restarts when the voltage at DLC/$\overline{MUTER}$ drops below 200mV.

3.3.4. Mute receive

The loudspeaker amplifier can be muted by making pin DLC/$\overline{MUTER}$ lower than 200mV. As a result the gain of the loudspeaker amplifier is reduced with 80dB. Also the circuit is internally forced into Tx-mode.
Pin DLC/$\overline{MUTER}$ can be made low by placing a switch over the capacitor, for instance a simple transistor.
When the switch is open, nothing is influenced. When the switch is closed and DLC/$\overline{MUTER}$ is put below 200mV, the loudspeaker pre-amplifier is muted. The end-amplifier can still be used by applying a signal on GAR. Because the starter is reactivated when the voltage on DLC/$\overline{MUTER}$ is smaller than 200mV, the switch must be able to sink approximately 80µA. When the switch is opened again, the starter will recharge the capacitor Cdlc.
The minimum allowable voltage on DLC/$\overline{MUTER}$ is GND-0.4V.

Chapter 3.4. Duplex controller block

In this chapter the principle of operation of the complete duplex controller will be discussed as well as its adjustments and performance. This will be done with the aid of figure 3.4.1, where the complete duplex controller is depicted.

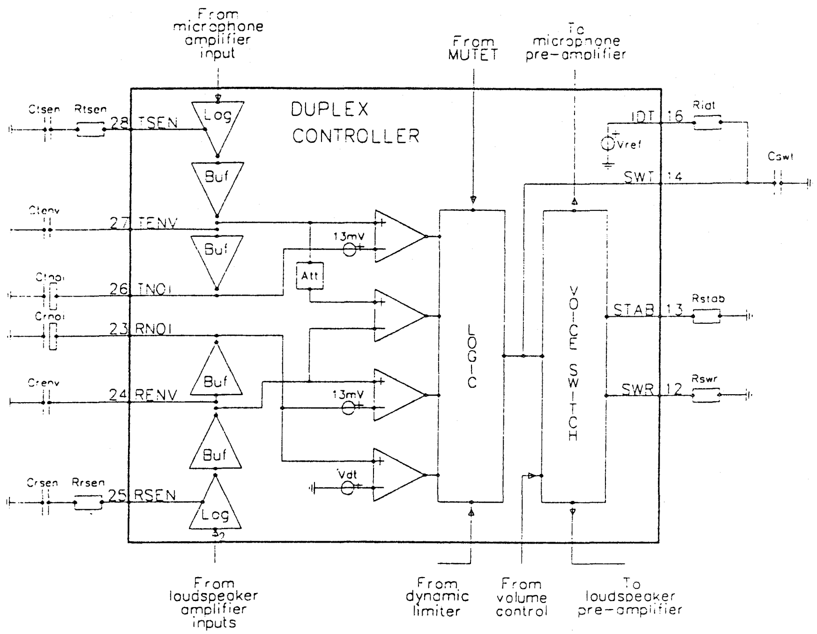


Figure 3.4.1: Principle of the duplex controller

As can be seen in figure 3.4.1, the duplex controller is built up out of signal and noise envelope detectors, decision logic and a voice switch.
The signal and noise envelope detectors determine the signal envelope and the noise envelope of both the transmit and receive signal. These envelopes are used by the decision logic to determine to which mode the set has to switch over (Tx, Rx or Ix-mode). The logic charges and discharges the capacitor Cswt and the resulting voltage on pin SWT controls the voice switch. The voice switch switches over the set between the three modes while keeping the loopgain constant.
In paragraphs 3.4.1 to 3.4.3, the principle of operation of the three parts is given. In paragraph 3.4.4, the adjustments and performance of the complete duplex controller is given.

3.4.1. Signal and noise envelope detectors

The signal and noise monitors of the transmit and receive channel are globally the same. The principle of the detectors therefore will be explained with the help of one of them: the signal and noise envelope detector of the transmit channel.
The microphone signal on MIC is connected to the first stage of the detector, see figure 3.4.1. The first stage amplifies the microphone signal from MIC to TSEN with an internally fixed gain of 40dB. Via the RC combination RtsenCtsen the signal on TSEN is converted into a current. This conversion determines the sensitivity of the envelope detectors. The current is logarithmically compressed and internally converted to a voltage. This voltage thus represents the compressed microphone signal. At roomtemperature, an increase of the microphone signal with a factor of 2 will increase the signal envelope with 18mV if the current through TSEN stays between 0.8µArms and 160µArms. Outside this region the compression is less accurate.
The compressed microphone signal is buffered by the second stage to pin TENV. Because the buffer can source maximum 120µA and sink maximum 1µA, the signal on TENV follows the positive peaks of the compressed signal. This is called the signal envelope. The time constants of the signal envelope are therefore determined by the combination of the internal current sources and the capacitor Ctenv.
The voltage on TENV is buffered by the third stage to pin TNOI. Because this buffer can source maximum 1µA and sink maximum 120µA, the signal on TNOI follows the negative peaks of the signal on TENV. This is called the noise envelope because it represents the background noise. The time constants of the noise envelope are therefore determined by the combination of the internal current sources and the capacitor Ctnoi.
Both capacitors Ctnoi and Crnoi are provided with a start circuit. During startup the capacitors are charged with approximately 40µA up to 1.9V. The starter will restart when the voltage on the capacitor drops below 0.9V.
As can be seen in figure 3.4.1, the principle of operation of the signal and noise envelope detectors of the receive channel is equal to that of the transmit channel. However, the gain of the first stage (inputs to pin RSEN) is 0dB and not 40dB as in case of the transmit channel.

The behaviour of the envelopes is illustrated in figure 3.4.2, where the signal and noise envelope of the receive channel (RENV, RNOI) are depicted together with the input signal on RIN1.

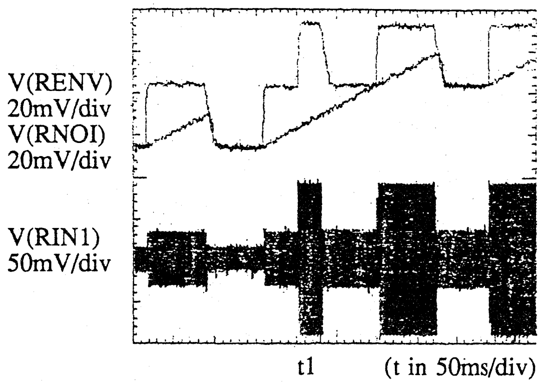


Figure 3.4.2: Typical behaviour of the signal and noise envelope detectors

In figure 3.4.2, the amplitude of the 1kHz input signal at RIN1 is modulated with 10dB and at moment t1 an extra 10dB is added. When, during modulation, the input signal is raised with 10dB, the signal envelope at RENV immediately follows. When the input signal drops with 10dB, the signal envelope drops less quick, thus reducing the influence of room echo. The noise envelope at RNOI slowly follows the signal envelope but never crosses it. When at t1 the extra 10dB is added, the signal envelope also increases but due to the logarithmic compression, the variation in the signal envelope due to the 10dB modulation is the same before and after t1.

3.4.2. Decision logic

The signal and noise envelopes of the transmit and receive signal are used by the decision logic to determine in which mode the TEA1094 has to be.
The output of the logic is a current source which charges or discharges the capacitor Cswt. If the logic determines Tx-mode, the capacitor Cswt is discharged with 10μA. When Rx-mode is determined, Cswt is charged with 10μA. When Ix-mode is determined, the current source is zero and the voltage on SWT becomes equal to the voltage on IDT via the resistor Ridt. The time constants of the duplex controller are therefore determined by the combination of the internal current source, the capacitor Cswt and the resistor Ridt.

As can be seen in figure 3.4.1, the envelopes are not used directly by the decision logic. First, to have a clear choice between signal and noise, the signal is considered as speech when its envelope is more than 4.3dB above the noise envelope. At room temperature, this is equal to a voltage difference of 13mV. This so called speech/noise threshold is implemented in both the transmit and receive channel. At the end of paragraph 3.4.4 a way to increase this threshold is discussed.
Second, the signal on MIC contains both the signal of the talker using the set as well as the signal coming from the loudspeaker (acoustic coupling). In Rx-mode, the contribution of the loudspeaker overrules the contribution of the talker using the handsfree telephone set. As a result, the signal envelope on TENV is mainly formed by the loudspeaker signal. To correct this, an attenuator is placed between TENV and the TENV/RENV comparator. This attenuation equals the attenuation applied to the microphone amplifier gain. Thus

when the TEA1094 is in Rx-mode the attenuation equals the switching range.
Third, when a dial tone is present on the line, without measures this would be recognized as noise. This would happen because it is a signal with a constant level during a long period. As a result the TEA1094 would go to Ix-mode and the user of the set would hear the dial tone fade away. Therefore, a dial tone detector is incorporated which does not consider input signals as noise when they have a level higher than the dial tone level. The dial tone level is adjustable by Rrsen.
When these three corrections are made, the signal and noise envelopes are used by the comparators and the logic. As already explained the output of the logic is a current source. The relation between the current source and the output of the comparators is given in the table of figure 3.4.3. If, for instance, TENV>RENV (transmit signal is larger than receive signal) and TENV>TNOI (transmit signal more than 4.3dB larger than noise level), then the output current will be -10μA.

Comparator TENV / TNOI	1	x	x	0	x
Comparator TENV / RENV	1	0	0	1	0
Comparator RENV / RNOI	x	1	x	x	0
Comparator RNOI / Vdt	x	x	1	x	0
Output current	-10μA	+10μA	+10μA	0μA	0μA

Figure 3.4.3: Truth table of the decision logic

When MUTET is made high, see paragraph 3.2.2, the output current is forced to be 10μA, which forces the TEA1094 into Rx-mode and mutes the microphone amplifier. When pin DLC/$\overline{MUTER}$ is made lower than 200mV, see paragraph 3.3.4, the output current is forced to be -10μA which forces the set into Tx-mode and mutes the loudspeaker amplifier. When both MUTET is made high and DLC/$\overline{MUTER}$ is made low the output current is forced to be -10μA and both channels are muted.
The voltage on pin SWT is internally limited to IDT+400mV and IDT-400mV.

3.4.3 Voice switch

With the voltage on SWT, the voice switch regulates the gain of the microphone pre-amplifier and the loudspeaker pre-amplifier in such a way that the sum of the transmit and receive gain is kept constant. This is done to keep the loopgain of a handsfree telephoneset constant, see also the introduction chapter 1. The switch-over behaviour of the voice switch will be described with the aid of figure 3.4.4.

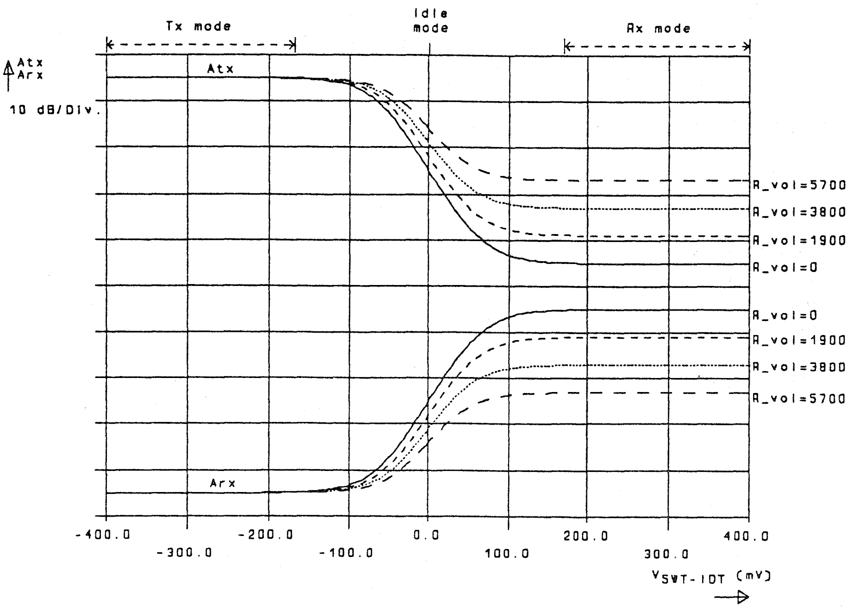


Figure 3.4.4: Switch-over behaviour

When the voltage on SWT is more than 180mV below the voltage on IDT, the TEA1094 is fully switched to Tx-mode (gain of the microphone amplifier is at its maximum and the gain of the loudspeaker amplifier is at its minimum). When the voltage on SWT is more than 180mV above the voltage on IDT, the TEA1094 is fully switched to Rx-mode (gain of the microphone amplifier is at its minimum and the gain of the loudspeaker amplifier is at its maximum). The TEA1094 is considered to be in Ix-mode when the voltage on SWT equals the voltage on IDT. When the capacitor Cswt is charged or discharged, the voltage on SWT varies and as a result the voice switch will smoothly switch over between the modes.

The difference between the maximum and minimum gain of the loudspeaker or microphone pre-amplifier is called the switching range. This range is determined by the ratio of Rswr and Rstab, see paragraph 3.4.4. Both Rswr and Rstab set internally used reference currents which are proportional to absolute temperature (PTAT).

As already stated in chapter 3.3, the volume control does not directly act upon the loudspeaker amplifier gain but via the voice switch. As a result, the loopgain of the handsfree set is kept constant when the volume of the loudspeaker signal is adjusted. The voice switch however, is designed such that the volume control has no influence in Tx-mode. Therefore during transmit, the gain of the microphone amplifier of the TEA1094 is not affected. In the extreme case, which is not plotted in figure 3.4.4, that the volume of the loudspeaker signal is reduced with the switching range, the TEA1094 virtually does not switch over. It also follows from this plot that when it was possible to have a volume reduction larger than the switching range, the gain of the loudspeaker amplifier would be smaller in Rx-mode than in Tx-mode. To avoid this, the volume control range of the TEA1094 can not be made larger than the switching range.

3.4.4. Adjustments and performance

The adjustment of the duplex controller has to be performed according the following recipe:

1. Determine switching range
2. Determine dial tone detector level
3. Determine sensitivity
4. Determine timings.

Ad 1. Determine switching range

The switching range Asw is determined by the ratio of the two resistors Rstab and Rswr according:

Asw = 20 * log (Rswr / Rstab) (in dB).

The resistor Rstab has to be taken 3.65kΩ. The resistor Rswr can be varied between 3.65kΩ and 1.5MΩ resulting in a switching range between 0dB and 52dB. With Rswr is 365kΩ, the switching range is set to 40dB with a tolerance of ±3.5dB.

The switching range is calculated out of the loopgain (Aloop). In a handsfree application the loopgain has to be smaller than 1 (equivalent to 0dB) and can be calculated as follows:

Aloop = Atx1094 + Atx106x + Ast + Arx106x + Arx1094 + Aac - Asw (in dB).

With Atx1094 = sending gain of the TEA1094 (MIC to MOUT)
Atx106x = sending gain of the TEA106x (MIC+/- to LN)
Ast = electrical sidetone
Arx106x = receive gain of the TEA106x (LN to QR+)
Arx1094 = receive gain of the TEA1094 (RIN1/2 to LSP)
Aac = electro-acoustic coupling from loudspeaker to microphone (LSP to MIC)
Asw = switching range.

For safety, the switching range Asw has to be chosen such that the maximum loop gain is far below 0dB (between -10dB and -20dB). Therefore, in calculations the worst case Ast and Aac have to be taken.
The electrical sidetone is the difference (in dB's) between the wanted receive signal on pin IR of the TEA106x and the unwanted part of the transmit signal on pin IR while having an equal signal level on pin LN for both the transmit and the receive signal. The electrical sidetone is dependent on linelength and frequency. The worst case sidetone can be found by measuring the sidetone over the telephonyband for several linelengths. The acoustic coupling is dependent on the environment of the telephoneset. A way of determining the worst case acoustic coupling is to move a hand to the set as if pushing a button.
When automatic line loss compensation (AGC) is used, the transmit and receive gain of the TEA106x are reduced at high line currents (short lines), see [Ref.1]. This will reduce

the loopgain at high line currents and makes the use of a smaller switching range possible. If a certain minimum volume control range is required, the switching range must not be chosen smaller than the required volume control range.
In appendix A, a method for measuring the required switching range is given.

Ad 2. Determine dial tone detector level

The dial tone detector level is determined by the value of Rrsen according:

$$Vdialtone = 12.7\mu * Rrsen \text{ (in Vrms).}$$

With an Rrsen of 10kΩ, the dial tone detector level will be 127mVrms. This means, a continuous signal on the inputs RIN1 and RIN2 larger than 127mVrms will be recognized as a dial tone.

Ad 3. Determine sensitivity

The sensitivity is set by Rrsen and Rtsen. The resistor Rrsen is already determined by the dial tone detector level. It must however be checked if the chosen value for Rrsen is a practical one. The reason for this is the dynamic range of the logarithmic compressor. A 'linear' compression is guaranteed when the currents flowing through pin RSEN are between 0.8μArms and 160μArms. This means that at nominal receiving signals the current through RSEN is preferably around 11μArms. This gives a maximum dynamic range of plus and minus 23dB. The same counts for pin TSEN.

The resistor Rtsen has to be chosen in such a way that both channels have the same priority for the duplex controller. This can be obtained by choosing Rtsen according:

$$20 * \log(Rtsen) = 20 * \log(Rrsen) - Atx1094 - Atx106x - Ast - Arx106x + Atsen + \tfrac{1}{2} * Aloop \text{ (in dB).}$$

With Atsen = internal gain from MIC to TSEN = 40dB.

In this relation, the maximum loopgain and the worst case sidetone are used, see also Ad1. If it is preferred to give the transmit channel priority above the receive channel, Rtsen has to be made smaller. When the opposite is the case, Rtsen has to be made larger. With respect to the calculated setting, resistor Rtsen can be varied with plus and minus ½ * Aloop (in dB's).
In appendix A, a method for measuring the required Rtsen is given.

The capacitors Ctsen and Crsen form a first order high pass RC-filter with Rtsen and Rrsen respectively to reduce the influence of low frequency bumps on the switching behaviour. It is advised to choose the capacitors Ctsen and Crsen such that the corner frequency of the RC-filters are equal.

When the calculated sensitivity setting is implemented, subjective tests with a real telephone line will be necessary to come to the optimal sensitivity setting.

Ad 4. Determine timings.

The timings which can be set are: signal envelope timing and noise envelope timing for both channels, switchover timing and idling timing.

The signal envelope timing is set by the capacitors Ctenv and Crenv. Because of the logarithmic compression between TSEN and TENV respectively RSEN and RENV, the timing can be expressed in dB/ms. At roomtemperature the following relation counts:

Timing $\approx$ I / (3 * C) (in dB/ms).
With I = charge or discharge current from pin TENV, RENV, TNOI or RNOI (in A)
C = timing capacitor Ctenv, Crenv, Ctnoi or Crnoi (in F).

With the advisory signal envelope timing capacitors Ctenv and Crenv of 470nF, the maximum attack-timing of the signal envelopes will be around 85dB/ms (I=120µA). This is enough to track normal speech. The release-timing will be 0.7dB/ms (I=1µA). This is enough to smoothen the signal envelope and to eliminate the influence of room-echoes on the switching behaviour.

With the advisory noise envelope timing capacitors Ctnoi and Crnoi of 4.7µF, the attack-timing of the noise envelopes will be 0.07dB/ms (I=1µA). This is small enough to track background noise and not to be influenced by speech bursts. The maximum release-timing will be 8.5dB/ms (I=120µA). This is enough to track the signal envelope during release because the signal envelope release timing is 0.7dB/ms which is a factor smaller.
It is advised to choose the signal envelope timing and the noise envelope timing of both channels equal for optimum operation of the duplex controller. To have clearly determined timings, it is advised not to use capacitors with a high leakage current.

The switch-over timing is determined by the value of the switch-over capacitor Cswt. The idling timing is determined by the combination of Cswt and the idling resistor Ridt.
If the output current of pin SWT is Iswt, a voltage difference over Cswt can be obtained according:

δVswt/t = Iswt / Cswt (mV/ms).

With the advised value for Cswt of 220nF, the obtained voltage difference is 45mV/ms. The switch-over time is dependent on the voltage difference which has to be generated on SWT. Suppose the set is in full Tx-mode, then the voltage on SWT will be V(IDT)-400mV, see figure 3.4.4. To reach Rx-mode a voltage difference of 580mV must be generated to end up at a voltage of V(IDT)+180mV. So in this case the switch over time will be 13ms. When the set is in Ix-mode the voltage on SWT equals the voltage on IDT. In that case Rx-mode or Tx-mode will be reached within 4ms.
The idling timing is determined by an RC-time constant. It is supposed that Ix-mode is reached when a time (Tidt) is passed of:

Tidt = 4 * Ridt *Cswt.

With the advised value for Ridt of 2.2MΩ, an idling time of around 2 seconds is obtained. To have a clearly determined idling timing, it is advised not to use a capacitor with a high leakage current.

Miscellaneous

When a handsfree telephone set is used at one end of the subscriber line and a conventional set at the other end, the user of the conventional set will think that the line is 'dead' when the handsfree set stays in receive mode while no signal on the line is present. This is avoided when the handsfree set switches over to a so called idle mode. This mode is incorporated in the TEA1094 and is placed exactly in between the transmit and receive mode. When it is desired to have an idle mode which is closer to transmit than receive mode, the circuit of figure 3.4.5 can be applied.

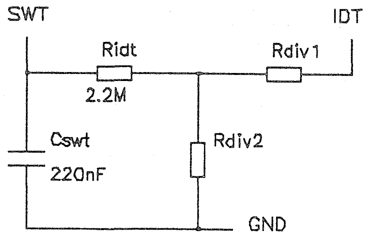


Figure 3.4.5: Circuitry for shifting the idle mode

With the circuit of figure 3.4.5, in idle mode, the voltage on SWT will not go to the voltage on IDT but to the voltage on IDT minus the voltage drop over Rdiv1. The voltage drop over Rdiv1 determines the shift of the idle mode (in dB's). This shift can be read from figure 3.4.4, when the voltage drop over Rdiv1 is taken as the x-axis value. The voltage on IDT is approximately 1.2V, so with for instance Rdiv1=33kΩ and Rdiv2=1MΩ the shift will be approximately 10dB. When dimensioning the resistor divider, it is advised not to choose Rdiv2 smaller than 1MΩ in order to limit the current drawn from IDT. By connecting Rdiv2 to VBB instead of to GND, the idle mode is shifted towards the receive mode.

In noisy environments, like offices, a handsfree set can show a popping behaviour in idle-mode (unwanted switching over from Ix to Tx-mode). This can be caused for instance by footsteps in the corridor. In the TEA1094, this popping behaviour is reduced by the implemented speech/noise threshold of 4.3dB. However, when a larger threshold is desired this can be achieved by connecting a resistor Rtnoi in series with Ctnoi.
When there is only noise present at the input of the envelope detector, the voltages on pins TENV and TNOI are equal. When, suddenly, a signal is present, the voltage on TENV will increase. Without Rtnoi, the voltage on TNOI will increase slowly because of the charging of Ctnoi by the 1μA internal current source. When a resistor Rtnoi is placed in series with Ctnoi, under the same conditions, this 1μA current source will cause a voltage jump on TNOI. This jump determines the shift of the speech/noise threshold.
As depicted in figure 3.4.1, at roomtemperature, the 4.3dB threshold equals 13mV. A resistor Rtnoi in series with Ctnoi will add an extra voltage to this threshold of 1μA*Rtnoi. When for instance a resistor of 10kΩ is chosen, the speech/noise level is increased up to 23mV which is equal to 7.6dB at roomtemperature. The new speech/noise threshold is slightly dependent on temperature and the spread of the internal current source and therefore not as accurate as the internal 4.3dB. It is advised not to use a resistor larger than 15kΩ.

Chapter 4. Application cookbook

In this chapter the procedure for making a basic application with a speech-transmission circuit of the TEA106x-family and the handsfree circuit TEA1094 will be given. With the aid of figure D.1 in appendix D, the design flow is given as a number of steps which should be made. As far as possible for every step also the components involved and their influence on every step are given. The preferred value is given between brackets { }. More information on the setting of the TEA1094 can be found in the chapters given at every step. More information on the setting of the TEA106x-family see data handbook [Ref.1] and appendix B.

The application of figure D.1 is an application of the TEA1094 handsfree circuit together with the TEA106x speech-transmission circuit. Switches are incorporated to switch over between handsfree operation and handset operation. A microcontroller and an interruptor are not incorporated.
As can be seen in figure D.1, only a few components have a fixed value. These are the zener diode of 12Volts protecting the TEA106x and the resistors R5 and Rstab setting reference currents for the TEA106x and TEA1094 respectively. All other values will follow from the cookbook of figure 4.1.

Worked out examples of applications of the TEA1094, following the cookbook are discussed in chapter 5.

STEP	ADJUSTMENT
DC-settings: Adjust the DC setting of the TEA106x to the local PTT requirements.	
Voltage LN-SLPE DC-slope Supply point VCC Artificial inductor	R17 R9 {20Ω}, R10 (also current protection) C1 {100μF} C3 {4.7μF}
Impedance and sidetone: After setting the set impedance, the sidetone has to be optimized for mean linelength and linetype. Also AGC can be chosen.	
Set impedance Sidetone AGC	Z1; R10 is in series R2, R3, R8, R11, R12, C12 R6
TEA106x Microphone and earpiece amplifiers, see appendix B: After the sensitivity of the microphone is adjusted, the gain can be adjusted to the desired value. Also the frequency curve can be set. The same counts for the earpiece.	
Sensitivity microphone Microphone gain Frequency curve and stability Earpiece gain Frequency curve and stability	R20,R21; microphone dependent R7; depends on TEA106x used C6; low pass with R7+3.5kΩ C20; stability for C20=10*C6 C8, C9, R14; high pass with input impedance R4; depends on TEA106x used C4; low pass with R4 C7; stability for C7=10*C4 C5; highpass with input impedance IR C2; highpass with earpiece impedance
TEA1094 Microphone amplifier, see chapter 3.2: After the sensitivity of the microphone is adjusted, the gain can be adjusted to the desired value. Also the frequency curve can be set.	
Sensitivity microphone Transmit gain Frequency curve and stability	Rmic Rgat; Atx=20log(0.674*Rgat/Rstab) (dB) Cgat; low pass with Rgat, also stability Crmic; low pass with Rmic Cmic; high pass with input impedance (20kΩ)

STEP	ADJUSTMENT
TEA1094 Loudspeaker amplifier, see chapter 3.3: The loudspeakergain can be adjusted to the desired value. Also the frequency curve can be set. The volume control potentiometer can be chosen as well as the dynamic limiter timing to have a minimal distortion.	
Receive gain Frequency curve and stability Volume control Dynamic limiter timing	Rgar; Arx=20log(0.435*Rgar/Rstab) (dB) Cgar; low pass with Rgar Crin1; high pass with input impedance (20kΩ) Crin2 {Crin1} Clsp; high pass with loudspeaker impedance Rvol {10kΩ}; 3dB reduction for each 950Ω Cdlc {470nF}
TEA1094 Duplex controller, see chapter 3.4 and appendix A: When all gains are adjusted the switching range can be determined by measuring the loopgain. Then the dialtone detector level can be set as well as the sensitivies of the duplex controller. Finally the timings of the envelopes and the switching are adjusted.	
Switching range Dial tone Sensitivity Signal Envelope Noise envelope Switch-over timing Idle mode timing	Loopgain: Aloop=Atx1094+Atx106x+Ast+ Arx106x+Arx1094+Aac-Asw<0 (dB) Choose Asw with safety margin of 10-20 (dB) Adjust Rswr; Asw=20log(Rswr/Rstab) (dB), with Rstab is fixed to 3.65kΩ Rrsen; Vdialtone=12.7µA*Rrsen (Vrms) Rtsen; For equal sensitivities of Tx and Rx: 20log(Rtsen)=20log(Rrsen)-Atx1094-Atx106x-Ast -ArxTEA106x+40+½Aloop (dB) Ctsen; high pass with Rtsen Crsen; high pass with Rrsen Ctenv {470nF}, Crenv {470nF}; Maximum attack: 120µ/(3*Cenv) (dB/ms) Release: 1µ/(3*Cenv) (dB/ms) Ctnoi {4.7µF}, Crnoi {4.7µF}; Attack: 1µ/(3*Cnoi) (dB/ms) Maximum Release: 120µ/(3*Cnoi) (dB/ms) Cswt {220nF}; |ΔVswt|/t=10µ/Cswt (mV/ms) Ridt {2.2MΩ}; timeconstant: 4*Ridt*Cswt

Figure 4.1: Steps in the design flow of the TEA106x+TEA1094

Chapter 5. Application examples

In this chapter some application examples of the TEA1094 handsfree circuit are given. In all examples only the essential elements are given. For instance no ringer or interruptor is included. The setting of the examples is made by following the cookbook of chapter 4. The lower corner frequencies are chosen between 200Hz and 300Hz and the higher corner frequencies around 4kHz. Components which are not mentioned have the advised value. In appendix B the gains and DC-settings of the TEA106x family can be found.

Figure D.2 gives the basic handsfree application of the TEA1094 together with a speech transmission circuit of the TEA106x family. This basic application only incorporates handsfree telephony.

DC-settings: R17 is chosen 39.2kΩ which makes 4.5V to 5.0V after the bridge at 15mA of line current, depending on the TEA106x used, see appendix B. Resistor R10 is 12Ω for current protection.

Impedance and sidetone: The set impedance is made approximately 600Ω for the telephony band with R1 and R10. The optimized sidetone bridge for this impedance in combination with a 5km cable of 0.5mm diameter copper twisted pair (176Ω, 38nF per km) is taken from [Ref.1]. Also AGC is set with R6=110kΩ. This gives optimum result for an exchange of 48V and 600Ω in combination with the 0.5mm diameter cable (1.2dB attenuation per km).

TEA106x amplifiers: The microphone gain is set to the lowest value with R7=27.4kΩ (44dB for TEA1060/2/4/7/8). The receive gain is set to its maximum value of -1dB with R4=100kΩ (-32dB attenuation by the anti sidetone-bridge and 31dB gain from IR to QR+ for TEA1062/4/7).

TEA1094 amplifiers: The transmit gain is set to 5dB with Rgat=9.53kΩ. The receive gain is set to 25dB with Rgar=150kΩ. This results in an overall receive gain from line to loudspeaker output LSP of 24dB.

TEA1094 duplex controller: By measuring the loopgain following appendix A, the switching range can be determined. The loopgain is, of course, very dependent on the acoustic coupling between loudspeaker and microphone. With the set used for the measurements in appendix A, a switching range of 40dB (Rswr=365kΩ) leads to a gainmargin of 18dB. The dialtone detector level was chosen 89mVrms on the inputs RIN1 and RIN2 (meaning 100mVrms on the line). This results in Rrsen=6.81kΩ.

When using the sensitivity measurement result of appendix A, 20log(vtsen/vrsen)=4dB, the resistor Rtsen follows out of 20log(Rtsen/Rrsen)= 4dB-(18dB/2)=-5dB or Rtsen=3.92kΩ.

Figure D.3 gives the handsfree/handset application of the TEA1094. This application incorporates both handsfree and handset operation. As an example the receive gain of the TEA106x is made 4dB lower with respect to figure D.2. The receive gain of the TEA1094 is raised with 4dB to have the same overall receive gain. To detect a dial tone of 100mVrms at the line, due to the lowered receive gain of the TEA106x, the resistor Rrsen is changed into 4.75kΩ. As can be seen in the formulas of chapter 3.4.4, Rtsen does not need to be changed.
The switches can be either mechanical or electronic ones.

Figure D.4 gives the handsfree/handset/listening-in application. In listening-in operation an acoustical loop is formed by the loudspeaker and the handset microphone. Without measures this loop can cause howling when the handset is close to the loudspeaker. This situation is quite similar to the handsfree howling problem, see figure 1.1. Therefore, the TEA1094 can also be used to prevent howling in listening-in operation. This is done by connecting the handset microphone to the TEA1094 instead of connecting it directly to the TEA106x. The switching is now introduced in the listening-in loop. This switching has no effect on the earpiece level. Switch-over between the different modes is possible by the external switches.
The setting of the application is quite similar to the previous application examples. The receive channel is not changed. In the transmit channel an attenuator of 20dB is placed between the TEA1094 and the TEA106x. This is done to reduce the sending noise on the line in receive mode, see chapter 3.2.1. To end up with the same overall transmit gain as in the previous examples, the transmit gain of the TEA1094 itself is raised with 20dB. To have a correct transmit gain during handset mode, the sensitivity of the handset microphone has been adjusted (R20, R21). In practice there will be no need for changing the switching range for listening-in operation with respect to handsfree operation.

Figure D.5 gives a blockdiagram of a cordless telephone with handsfree base-unit. The line-interface is supplied by the telephoneline, while the handsfree and cordless part are mains supplied. The galvanic seperation between the two parts is done with opto-couplers but it can also be done with transformers. The signals which are coupled are the transmit (Tx) and receive (Rx) signals, the dial-pulse signal (DP), the ring detect signal (Rdet) and the on/off-hook information (Hook). The ground of the line-interface part is VEE, while GND is the ground of the rest of the base.
In a cordless telephone with handsfree-base, it must be possible to transmit the signals of the handset to both the line and the base (intercom). Also, it must be possible to have a handsfree conversation with the far-end user. To make this possible, a switch-block is depicted. The mode of the set (thus the mode of the switches, the TEA1094 and the handset) is defined by the microcontroller (uC). To save opto-couplers, DTMF dialling is done via the MIC+/- inputs of the TEA106x.

Figure D.6 gives a blockdiagram of a telephoneset with answering machine. The line-interface and the handset are supplied by the telephoneline, while the handsfree and answering machine part are mains supplied. The galvanic seperation between the two parts is done with opto-couplers but it can also be done with transformers. The signals which are coupled are the transmit (Tx) and receive (Rx) signals, the dial-pulse signal (DP), the ring detect signal (Rdet) and the on/off-hook information (Hook). Also a signal defining handset or handsfree/answering mode is coupled (MODE) which sets the block with switches. The ground of the line-interface part is VEE, while GND is the ground of the rest of the set.
In a telephone with answering machine, it must be possible to transmit an outgoing message to both the line and the loudspeaker. Also, it must be possible to record incoming messages as well as to have a handsfree conversation with the far-end user. To make this possible, a switch-block is depicted. The mode of the set (thus the mode of the switches, the TEA1094 and the handset) is defined by the microcontroller (uC). To save opto-couplers, DTMF dialling is done via the MIC+/- inputs of the TEA106x.

Chapter 6. Electromagnetic compatibility

With respect to electromagnetic compatibility (EMC) no common European or international specification yet exists. Also the measurement methods differ and are not always reproducible. At the application laboratory in Eindhoven (PCALE) the German current injection method is used (VDE 0878 part 200). It is a reliable method of measuring and is highly reproducible. The method is described in [Ref.3]. The hints for EMC of the TEA106x-TEA1094 combination given in the second paragraph of this chapter are based on this method. The same counts for the hints of the printed circuit board design given in the first paragraph.

6.1. Printed circuit board

In the current injection method, radio frequency (RF) signal currents enter the telephone set at the a/b wires and leave the set via any capacitive coupling to ground. Normally, in a telephone set the handset has the largest capacitance to ground and thus the main part of the RF signal current flows to ground via the handset. However, in handsfree operation the handset is not used. The RF signal current then flows to ground via the groundplane of the printed circuit board (PCB) and partly via the wires of the loudspeaker and microphone. Therefore, a proper PCB layout is essential for good EMC.

The first measure to be taken is to create a groundplane on the PCB. The RF signals entering the PCB should be decoupled immediately to this groundplane. Preferably this groundplane is homogeneous and is not cut into parts by interconnection wires. To reach this a double layered PCB, with interconnection wires on one side of the board and a groundplane on the other side, is a minimum. When interconnection wires within the groundplane are inevitable, the continuity of the groundplane should be restored. This is done by cross coupling these interconnection wires by jumpers and wires at the interconnect side of the PCB. In this way RF signal currents can flow freely over the groundplane.

Another measure is to keep the length of the wires between the different components as short as possible. Of course this measure is especially important for those wires which interconnect one or more RF signal sensitive parts, in particular the wire connected to SLPE which is also the reference for the TEA1094. As any current, RF signal currents prefer to flow to ground via the lowest ohmic path. Therefore, it should be noted that a wire of 10mm corresponds to an inductor of 10nH.
The wires to the handsfree microphone and the loudspeaker should be kept as short as possible as well. Normally, this is not a problem since loudspeaker, microphone and PCB are all within the same cabinet.

6.2. TEA106x-TEA1094 combination

All measures in this paragraph are based upon EMC experience and measurements at the TEA106x-TEA1094 application of figure D.7. This figure only gives the essential EMC components and it is just an example for a solution for better EMC performance since other applications and or different PCB layouts will demonstrate different behaviour. This paragraph should therefore be interpreted as a guidance to reach proper EMC design.

This paragraph is split into two parts. In the first part the standard measures for the

TEA106x are given. In the second part it is explained how to improve the EMC behaviour of the TEA1094. When, after the proposed measures are taken, the EMC behaviour has to be optimized, it is advised to start with the transmit direction of the telephoneset and to optimize the receive part thereafter. This because, due to sidetone, signals demodulated by the transmit channel will be seen in the receiver channel.
In the text below it is supposed that the printed circuit board on which the TEA106x-TEA1094 is built, is provided with a groundplane which is connected to VEE. It is preferred to place the components meant for EMC as close as possible to the pins, except when otherwise stated.
For more details on the EMC performance of the TEA106x, see [Ref.3].

TEA106x

RF signals entering the printed circuit board at the a/b wires should be decoupled to the groundplane via capacitors, preferably placed as close as possible to the a/b connection. Since these capacitors will be in parallel with the set impedance, their value is limited. In practice, a total capacitance of 10nF between the a/b wires may be applied without degrading the balance return loss.
Between the a/b wires and the groundplane two capacitors of 4.7nF each are placed. Another capacitor of 2.2nF is placed between pin SUP of the TEA1094 and the groundplane. Also a coil of 22μH is placed in series with the line.

RF signals entering the inputs of the transmit channel (MIC+, MIC-) will be demodulated and amplified to the line. Because of the high gain (in the order of 50dB) the transmit channel is very sensitive to RF signals. Therefore, decoupling at the inputs is essential. The inputs MIC+ and MIC- can best be decoupled by 2 capacitors of 10nF connected to the groundplane. Also series resistors can be applied which in combination with the capacitors form low pass filters towards the inputs. Resistors of 1kΩ are advised which reduce the gain setting with less than half a dB, depending on the input impedance of pins MIC+ and MIC-. This can be compensated by adapting the transmit gain adjustment resistor R7 of the TEA106x.
In case a handset microphone is connected to the TEA106x microphone inputs via a long cord, extra decoupling is needed. This can be done by adding two capacitors of 10nF each, placed between the cord connection and the groundplane, preferably as close as possible to the cord connection.

RF signals entering the inputs of the receive channel (IR) will be demodulated and amplified to the earpiece and, via the TEA1094, to the loudspeaker. Therefore, decoupling at the input is essential.
At the input IR of the TEA106x a capacitor of 2.2nF connected to the groundplane is advised.
In case an earpiece is connected to the TEA106x via a long cord, extra decoupling is needed. This can be done by adding two capacitors of 10nF each, placed between the cord connection and the groundplane, preferably as close as possible to the cord connection. To prevent the remaining RF signals from entering the earpiece outputstage of the TEA106x via the QR pin, and thus the loudspeaker amplifier inputs of the TEA1094, a series resistor should be applied to create a high ohmic path. The value of this resistor is dependent on the type of earpiece capsule used. When a dynamic capsule of 150Ω is used, a resistors of 22Ω is advised. This will reduce the gain setting with 1.19dB. This can be compensated by adapting the receive gain adjustment resistor R4 of the TEA106x.

Besides these essential measures some additional measures can be taken. A capacitor between GAS2 and the groundplane of 100pF can improve EMC in the transmit direction. A series combination of a resistor of 365Ω and a capacitor of 4.7nF connected between STAB and the groundplane can improve EMC for both transmit and receive direction.

TEA1094

The reference for the TEA1094 is GND, connected to VEE. When GND is connected to SLPE, this reference has to be decoupled to obtain a good EMC behaviour. This can be achieved by using a coil-capacitor combination between SLPE, GND and VEE. It is advised to use a coil of a value between 5µH and 33µH and a parasitic series resistance smaller than 5Ω. The capacitor has an advised value of 4.7nF.
The reference current source of the TEA1094 must have a high immunity because it sets the transmit and receive gains. Therefore, pin STAB must be decoupled. This can be done with a series combination of a resistor of 365Ω and a capacitor of 4.7nF connected between STAB and GND.

RF signals entering the input of the transmit channel (MIC) will be demodulated and amplified to the line. Because of the high gain (in the order of 50dB) the transmit channel is very sensitive to RF signals. Therefore, decoupling at the input is essential.
At the input MIC a capacitor of 4.7nF connected to GND, is advised. Also a series resistor can be applied which, in combination with the capacitor, will form a low pass filter towards the MIC input. A resistor of 1kΩ is advised which reduces the gain setting with 0.42dB. This can be compensated by adapting the resistor Rgat.
In case a handset microphone is connected to the TEA1094 microphone inputs, for instance with the application of figure D.4, extra decoupling is needed. This can be done in the same way as in case the handset is connected to the TEA106x (thus extra decoupling to the groundplane).

RF signals entering the volume control pin VOL will modulate the receive gain. Therefore, the pin VOL has to be decoupled. This can be done by connecting a capacitor of 4.7nF between VOL and GND.

Besides these essential measures some additional measures can be taken. When the stabilized supply is not connected very close to pin VBB, a small capacitor of 4.7nF connected close to pin VBB will improve EMC behaviour.
To improve the EMC behaviour in the receive channel, the inputs RIN1 and RIN2 can be decoupled. This can be done by connecting 2 capacitors of 10nF each between the inputs and the groundplane. Also series resistors can be applied which in combination with the capacitors form low pass filters towards the inputs. Resistors of 1kΩ are advised which reduce the gain setting with 0.42dB. This can be compensated by adapting the resistor Rgar of the TEA1094.
To improve EMC behaviour in the transmit channel, pin GAT can be decoupled to GND via a capacitor with the same value as capacitor Cgat. The influence of this capacitor on the transmission parameters can be neglected.

Chapter 7. References

[Ref.1] Philips Semiconductors Data Handbook
Semiconductors for telecom systems - IC03
Philips Semiconductors, 1993

[Ref.2] TEA1093 Handsfree IC
Final device specification
May 1993, version 3.0

[Ref.3] Measures to meet EMC requirements for TEA1060-family speech transmission circuits
M. Coenen, K. Wortel
PCALE reportnumber: ETT89016

[Ref.4] TEA1094 Handsfree IC
Tentative device specification
February 1994, version 2.1

[Ref.5] Application of the TEA1093 handsfree circuit
C.H. Voorwinden, K. Wortel
PCALE reportnumber: ETT/AN93015

Extended application information on the speech transmission circuits can be found in:

1. TEA1060 family versatile speech/transmission ICs for electronic telephone sets
Designers' guide by P.J.M. Sijbers
July 1987, 12NC 939834110011

2. Application of the versatile speech/transmission circuit TEA1064 in full electronic telephone sets by F. van Dongen and P.J.M. Sijbers
PCALE reportnumber: ETT89009

3. Application of the speech-transmission circuit TEA1062
P.T.J. Biermans
PCALE reportnumber: ETT89008

More information on the Philips cordless telephony products can be found in:

4. CT1 Evaluation board OM4735
J. Soree
PCALE reportnumber: ETT/AN93014

Appendix A. Measurement setup

This appendix describes how both the loopgain and the sensitivity setting can be measured. These measurements give a starting point for obtaining the final values which however, only can be reached by performing subjective tests.

For determining the loopgain the worst-case acoustical coupling and sidetone must be measured. This can be done with the measurement setup of figure A.1.

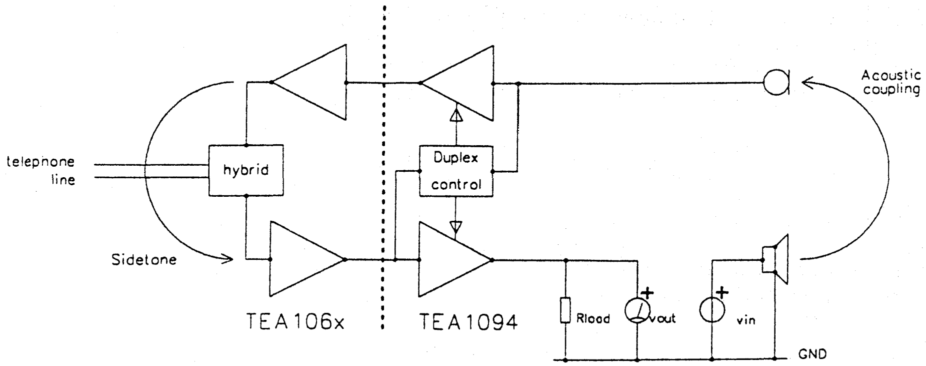


Figure A.1: Simplified measurement setup for measuring the loopgain

The loudspeaker is disconnected from the set and an electrical signal is applied to it. The acoustical signal coming from the loudspeaker is coupled to the microphone (Aac), transmitted to the line (Atx1094 + Atx106x), returns via sidetone (Ast) and is amplified again to the loudspeaker output of the TEA1094 (Arx106x + Arx1094). The output is loaded with a loudspeaker equivalent (Rload). The total gain of this loop is reduced with the switching range (Asw).
The gain from the loudspeaker connection (vin) to the loudspeaker equivalent (vout) is the loopgain (Aloop), given as:

$$20 * \log (vout/vin) = Aloop,$$
$$Aloop = Aac + Atx1094 + Atx106x + Ast + Arx106x + Arx1094 - Asw.$$

When this measurement is done with a vin of 100mVrms and between 100Hz and 10kHz, this will result in a curve such as displayed in figure A.2.

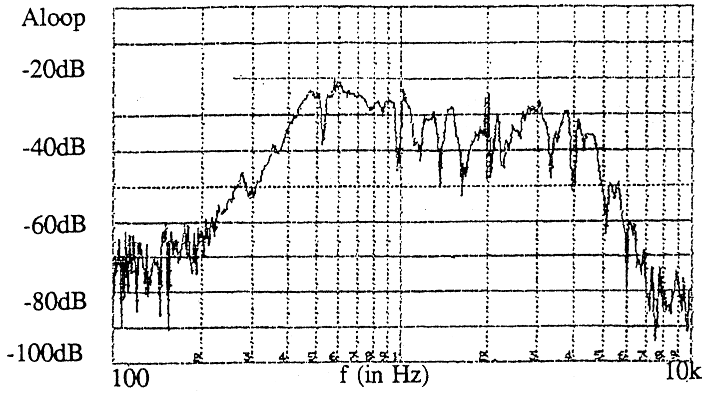


Figure A.2: Example of loopgain versus frequency

The highest value of the loopgain gives an indication of the gain margin.

The measurement of the loopgain Aloop has to be done with fixed transmit and receive gains, a fixed switching range and worst case acoustic coupling and sidetone.
The transmit and receive gains of the TEA106x and TEA1094 are supposed to be set, the switching range however, can only be set after the measurement. Therefore, during measurement the switching range Asw must be set to a certain value. An advised value for measurements is 40dB. This means that the resistor Rswr must be set to 365kΩ. The mode of the TEA1094 is not of influence on the measurement.
The worst case acoustical coupling can be found by moving around the set. A fairly good way of determining it, is keeping a hand close to the set as if pushing a button.
The worst case sidetone can be found by measuring the loopgain for several linelengths, for instance every kilometer, while maintaining the worst case acoustical coupling.
When AGC is used, the loopgain for short lines, smaller than 5km, will be lower than without AGC. As an effect, the switching range for a set with AGC can be made smaller than for a set without AGC.
The measurement combining both worst case sidetone and acoustical coupling gives the worstcase loopgain and thus the gain margin. The switching range now can be adapted to fit the gain margin desired.

When the correct switching range is set, the sensitivity setting can be determined.
For determining the sensitivity, calculations can be done as described in chapter 3.4.4. It is also possible measuring the correct sensitivity setting. This can be done with the measurement setup of figure A.3.

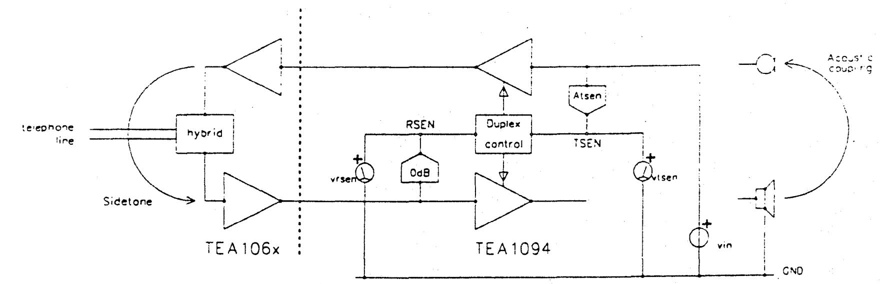


Figure A.3: Simplified measurement setup for determining sensitivity

Both the loudspeaker and microphone are disconnected from the set. An input signal (vin) is applied to the microphone input. This signal is amplified to pin TSEN (Atsen). It is also transmitted to the line, returns via the sidetone, amplified by the TEA106x to the TEA1094 receive inputs and buffered to pin RSEN. The ratio of the signals on pin TSEN (vtsen) and RSEN (vrsen) is given as:

20 * log (vtsen/vrsen) = Atsen - Atx1094 - Atx106x - Ast - Arx106x.

When this measurement is done with a vin of 1mVrms and between 100Hz and 10kHz, this will result in a curve such as displayed in figure A.4.

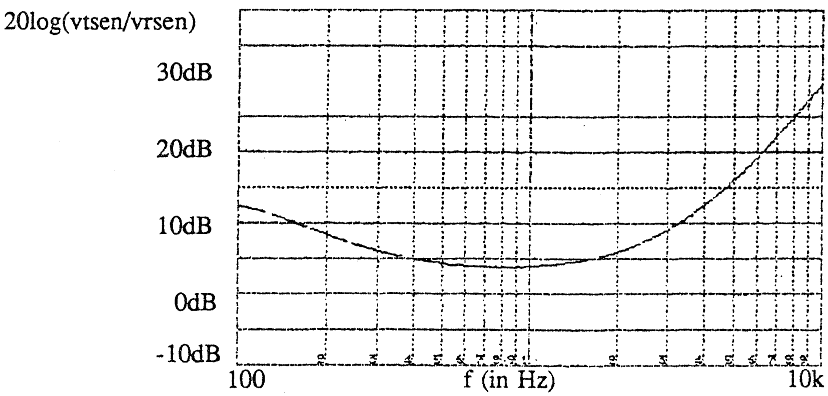


Figure A.4: Example of 20log(vtsen/vrsen) versus frequency

This is a smooth curve because the acoustical coupling is not in it. During this measurement the set must be in Tx mode. This can be done by making DLC/$\overline{MUTER}$ lower than 200mV. This will mute the receiver, and puts the set into Tx mode, but will not effect the signal on RSEN.

Out of the measured ratio of vtsen and vrsen it follows, see also chapter 3.4.4, Ad 3:

$$20 * \log(Rtsen/Rrsen) = 20 * \log(vtsen/vrsen) + Aloop/2.$$

This means that the optimal Rtsen is dependent on frequency because both the ratio of vtsen and vrsen as well as Aloop are frequency dependent. However, this can be simplified by using the lowest value of 20*log(vtsen/vrsen) and the highest value of Aloop to obtain the correct Rtsen. This will lead to the same answer as will be obtained by performing the calculations of chapter 3.4.4, Ad 3.

The value for Rtsen found, is a starting point for obtaining the final value which only can be reached by subjective tests with a real telephone line.

Appendix B. TEA106x quick reference data

	DC-CHARACTERISTICS (with slope resistance R9=20Ω)		
Member	V(LN-VEE) (in V) at Iline=15mA	V(LN-VEE) (in V) at R(REG-LN)=68kΩ	V(LN-VEE) (in V) at R(REG-SLPE) (in Ω)
TEA1060	4.45 ± 0.20	3.80 + 0.25/-0.30	5.0 ± 0.30 at 39k
TEA1062	4.00 + 0.25/-0.45	3.50 ±	4.5 ± .... at 39k
TEA1064B	3.50 ± 0.25	---	4.4 ± 0.35 at 20k
TEA1067	3.90 ± 0.25	3.40 ± 0.30	4.5 ± 0.30 at 39k
TEA1068	4.45 ± 0.25	3.80 ± 0.30	5.0 ± 0.35 at 39k

	SENDING GAIN		RECEIVE GAIN (from IR to QR+)	
Member	Setting range (in dB)	Gain (in dB) with R7=68kΩ	Setting range (in dB)	Gain (in dB) with R4=100kΩ
TEA1060	44 - 60	52 ± 1	17 - 33	25 ± 1
TEA1062	44 - 52	52 ± 1.5	20 - 31	31 ± 1.5
TEA1064B	44 - 52	52 ± 1	20 - 39	31 ± 1
TEA1067	44 - 52	52 ± 1	20 - 39	31 ± 1
TEA1068	44 - 60	52 ± 1	17 - 33	25 ± 1

	SENDING NOISE	
Member	Noise (in dBmp) with R7=68kΩ	Noise (in dBmp) with sending gain of 44dB
TEA1060	-70	-78
TEA1062	-69	-77
TEA1064B	-72	-80
TEA1067	-72	-80
TEA1068	-72	-80

For more data see datahandbook [Ref.1].

Appendix C. List of abbreviations and definitions

Aac	Electro-acoustic coupling (electrically measured)
AGC	Automatic line loss compensation of the TEA106x
Aloop	Loopgain of a handsfree telephone set (must be < 0dB)
Arx	Receive gain of the TEA1094 (3dB to 39dB)
ArxTEA106x	Receive gain of the TEA106x
ArxTEA1094	Receive gain of the TEA1094 (3dB to 39dB)
Ast	Electrical sidetone
A[illegible]	Switching range (0dB to 52dB)
A[illegible]en	Gain from MIC to TSEN of 40dB
Atx	Gain of the transmit channel of the TEA1094 (5dB to 25dB)
AtxTEA106x	Gain of the transmit channel of the TEA106x
AtxTEA1094	Gain of the transmit channel of the TEA1094 (5dB to 25dB)
BTL	Bridge tied load (loudspeaker between LSP1 and LSP2)
Cdlc	Dynamic limiter timing capacitor (470nF advised)
Cgat	Capacitor over Rgat
Clsp	DC blocking capacitor for loudspeaker
Cmic	Coupling capacitor at microphone input
Crenv	Capacitor determining the receive signal envelope (470nF advised)
Crin1,Crin2	Couple capacitors at receiver input
Crmic	Capacitor over Rmic
Crnoi	Capacitor determining the receive noise envelope (4.7µF advised)
Crsen	DC-blocking capacitor of receive sensitivity setting
Cswt	Switchover timing capacitor (220nF advised)
Ctenv	Capacitor determining the transmit signal envelope (470nF advised)
Ctnoi	Capacitor determining the transmit noise envelope (4.7µF advised)
Ctsen	DC-blocking capacitor of transmit sensitivity setting
dBmp	0 dBmp equals 1 milliWatt in 600Ω, psophometrically weighted
DLC/$\overline{MUTER}$	Dynamic limiter timing adjustment and receiver channel mute pin
DP	Dial pulse, output pin on microcontroller
δVswt	Voltage difference on SWT
EMC	Electro Magnetic Compatibility: the collective noun for the susceptibility and the radiation of a circuit/apparatus.
GAR	Receiver gain adjustment pin
GAT	Microphone gain adjustment pin
GND	Ground reference pin
GNDMIC	Ground reference pin for microphone amplifier
HCT4053	Philips IC containing 3, 2-channel analogue switches
HCT4066	Philips IC containing 4 analogue switches
IDT	Idle-mode timing adjustment pin
IR	Receive input pin of the TEA106x
Iswt	Current through pin SWT (typical 10µA)
Ix-mode	Idle-mode: the mode which is halfway Tx-mode and Rx-mode.
LN	Positive line terminal pin of the TEA106x
LSP	Loudspeaker amplifier output pin
MIC	Microphone input pin
MIC+, MIC-	Microphone inputs pins on the TEA106x
MOSFET	Metal oxide field effect transistor
MOUT	Microphone amplifier output pin

MUTET	Transmit channel mute input pin
PCALE	Product concept & application laboratory Eindhoven
PCB	Printed circuit board
Power Down	Reduced current consumption mode during pulse dialling or flash
PTAT	Proportional to absolute temperature
PTT	Telephone company
QR,QR+	Telephone earpiece output on TEA106x
RIN1,RIN2	Receiver amplifier input pins
RENV	Receive signal envelope timing adjustment pin
Rgar	Resistor setting receive (loudspeaker) gain (66.5kΩ for 18dB SEL)
Rgat	Resistor setting transmit (microphone) gain (30.1kΩ for 15dB)
Ridt	Resistor setting idle mode timing (2.2MΩ advised)
RF	Radio frequencies
Rload	Loudspeaker equivalent load resistor used for measurements
Rmic	Resistor setting the microphone sensitivity
RNOI	Receive noise envelope timing adjustment pin
Rrsen	Resistor setting sensitivity of the receive envelopes
RSEN	Receive signal envelope sensitivity adjustment pin
Rstab	Resistor setting an internally used PTAT current (3.65kΩ)
Rswr	Resistor determining switching range
Rtnoi	Resistor for increasing speech/noise threshold
Rtsen	Resistor setting sensitivity of transmit envelopes
Rvol	Volume control potentiometer (10kΩ advised)
Rx-mode	Receive-mode: the gain of the loudspeaker amplifier is at its maximum and the gain of the microphone amplifier is reduced
SLPE	DC slope pin on TEA106x
STAB	Reference current adjustment pin
Switching range	The difference between the maximum and the minimum gain in a channel as a result of the switching of the duplex controller (Asw)
SWR	Switching range adjustment pin
SWT	Switch-over timing adjustment pin
TEA106x	IC of the TEA106x speech transmission family: TEA1060/61, TEA1062, TEA1063, TEA1064, TEA1065, TEA1066, TEA1067, TEA1068.
TENV	Transmit signal envelope timing adjustment pin
THD	Total harmonic distortion
Tidt	Idle mode timing
TNOI	Transmit noise envelope timing adjustment pin
TSEN	Transmit signal envelope sensitivity adjustment pin
Tx-mode	Transmit-mode: the gain of the microphone amplifier is at its maximum and the gain of the loudspeaker amplifier is reduced
VBB	Supply output pin
VCC	Supply pin of the TEA106x
Vdialtone,Vdt	Dialtone detector level at the inputs RIN1 and RIN2
VEE	Reference pin on TEA106x
VOL	Receiver volume adjustment pin

Appendix D. Figures

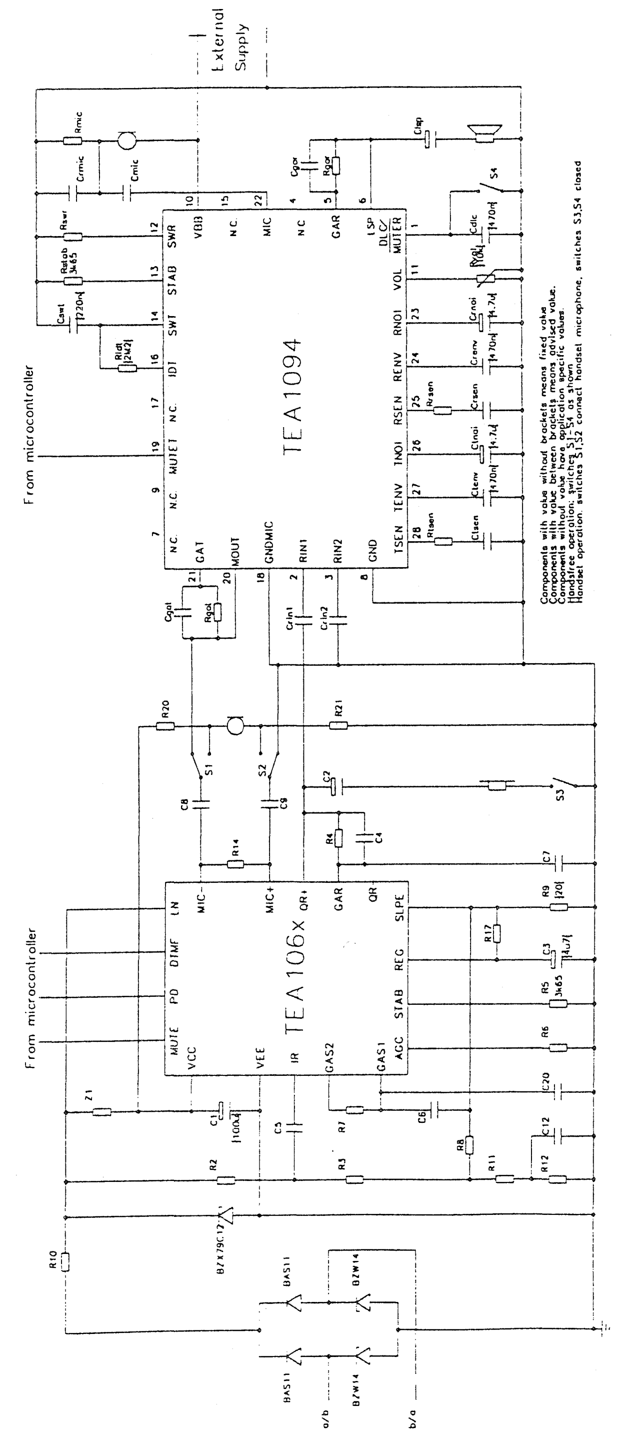


Figure D.1: Application TEA106x plus TEA1094

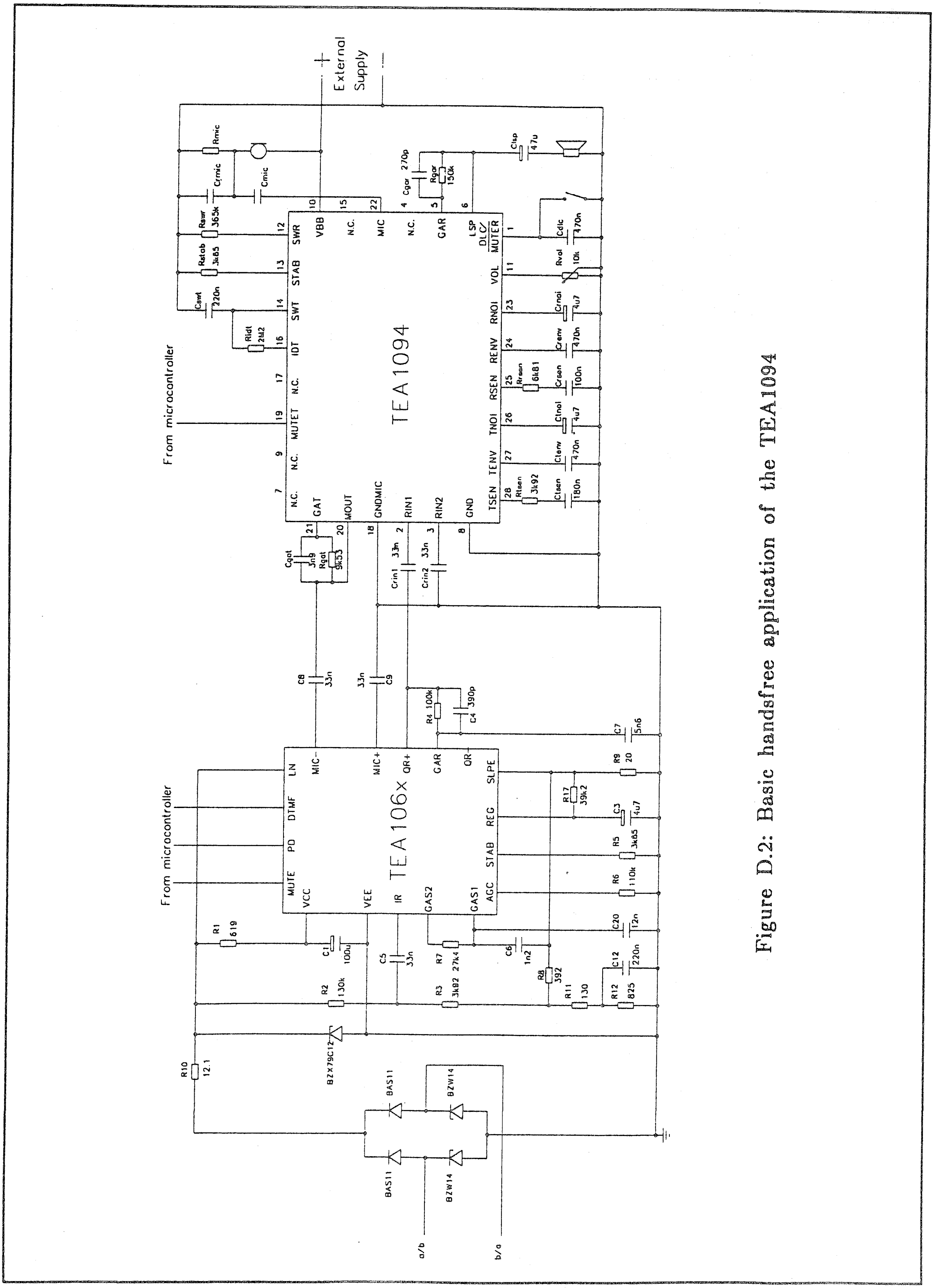


Figure D.2: Basic handsfree application of the TEA1094

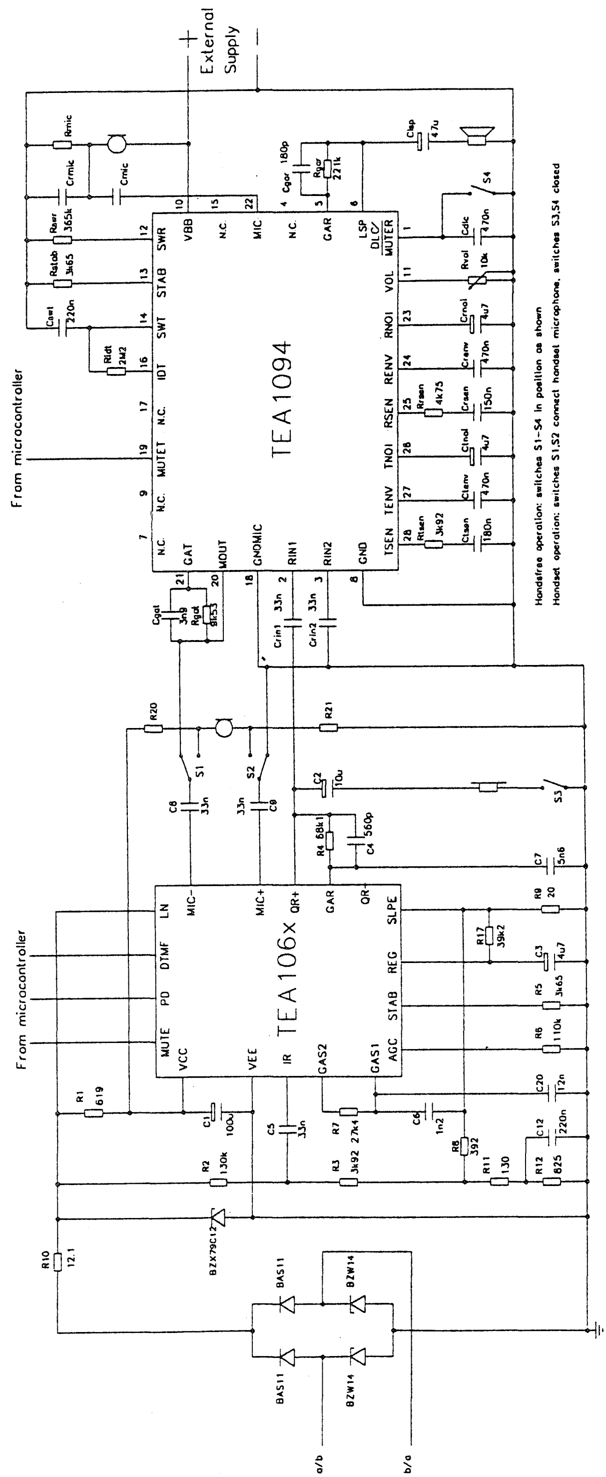


Figure D.3: Handsfree/handset application of the TEA1094

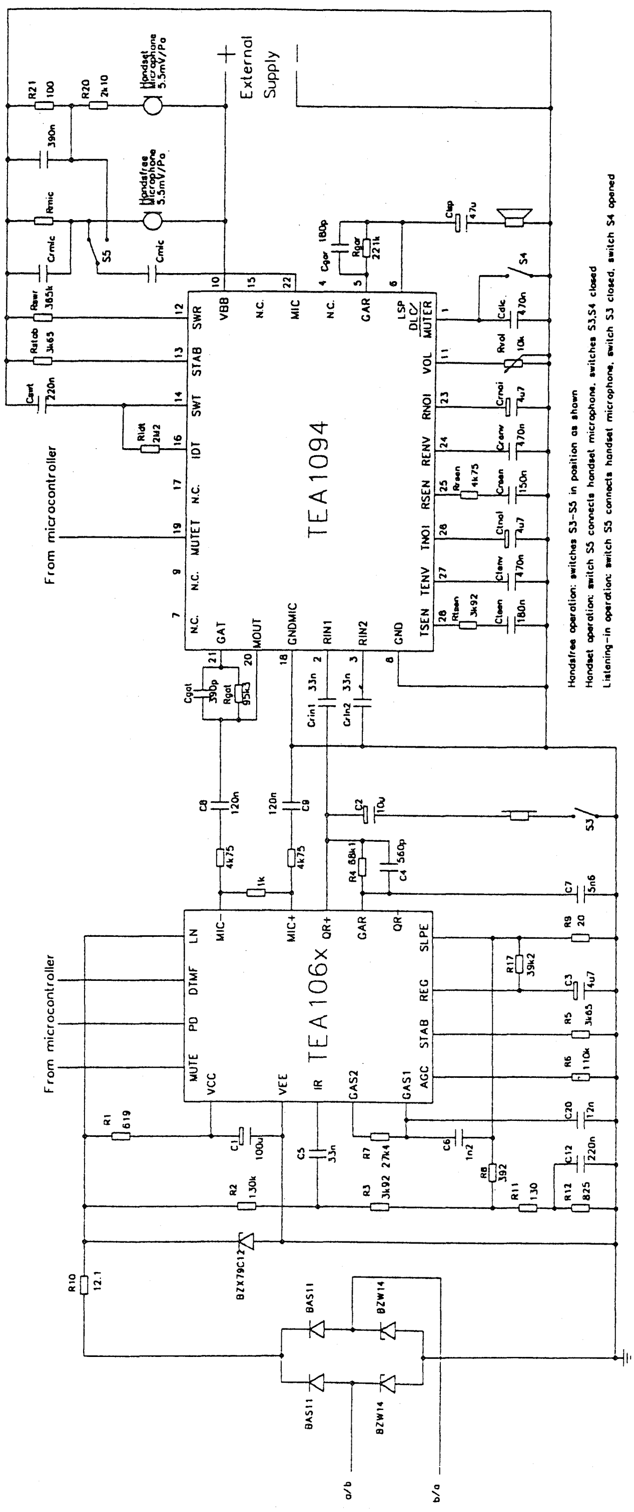


Figure D.4: Handsfree/handset/listening-in application of the TEA1094

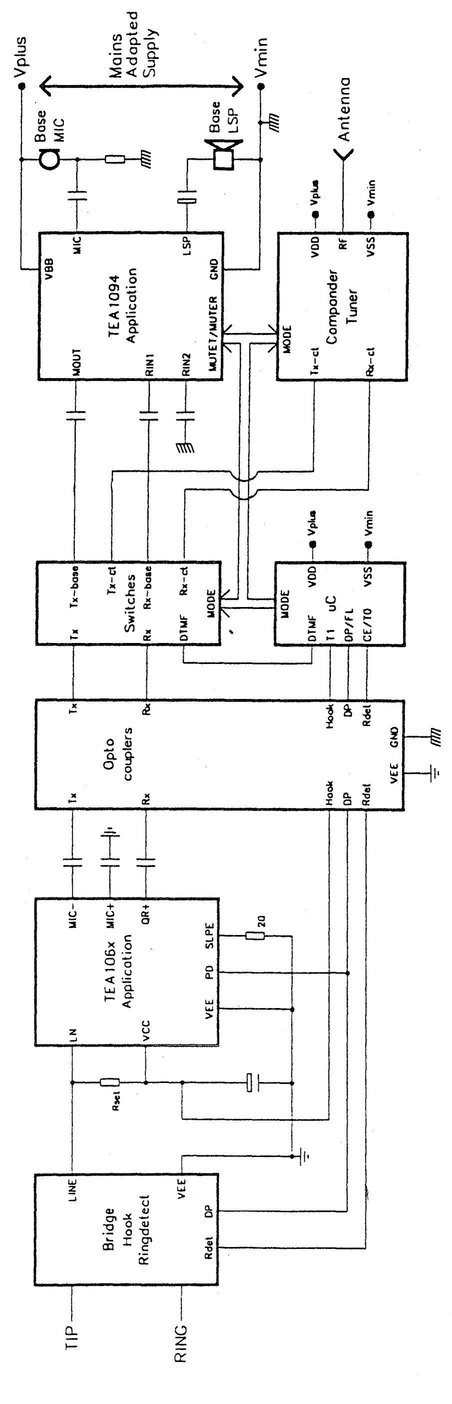


Figure D.5: Application proposal for cordless telephone with handsfree base

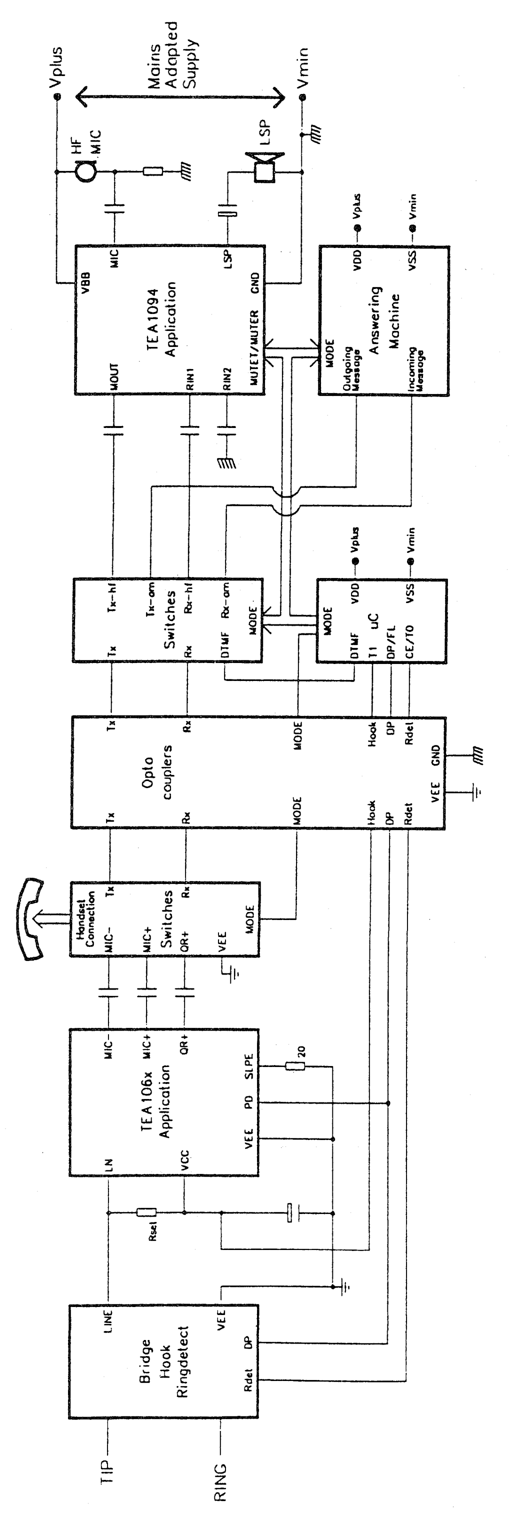


Figure D.6: Application proposal for corded telephone with answering machine

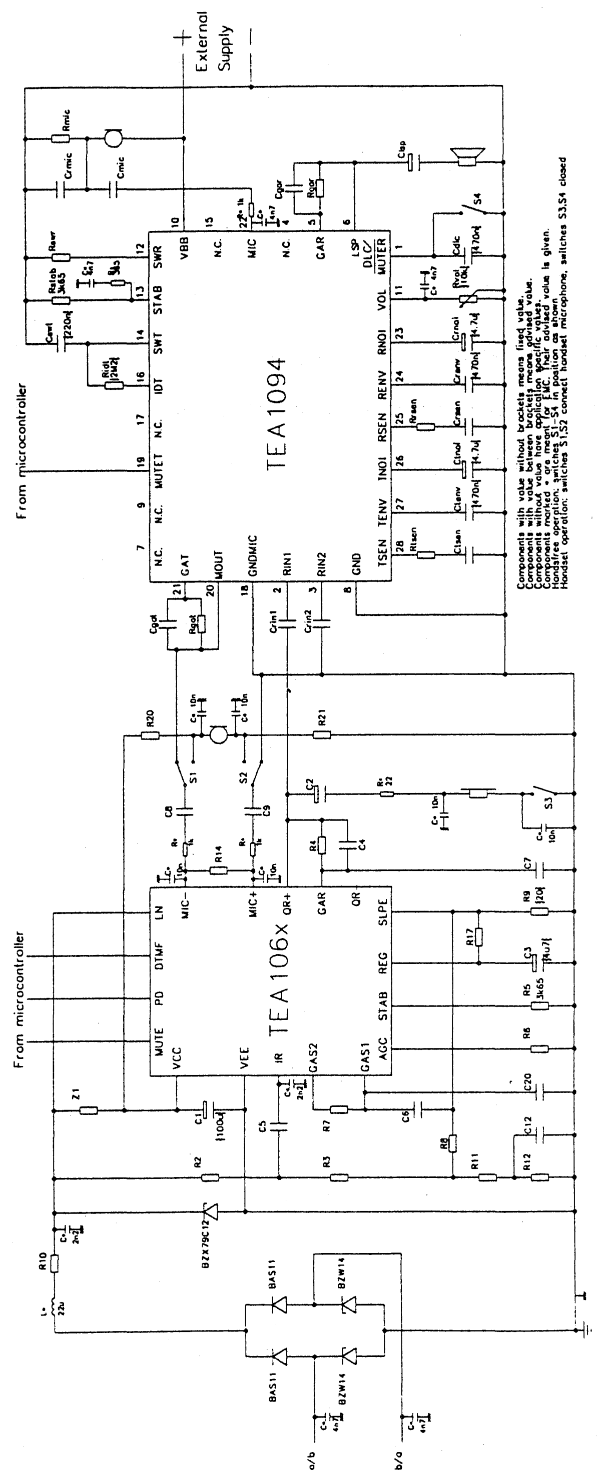


Figure D.7: Proposal for EMC of the TEA106x plus TEA1094 application

APPLICATION NOTE Nr	CTT/AN95083
TITLE	TEA1095 voice switched speakerphone IC
AUTHOR	J. M. Malaurie
DATE	September 1995

CONTENTS

1. INTRODUCTION

The TEA1095 is a circuit which, in combination with TEA1096 (transmission circuit with built-in loudspeaker amplifier), offers a handsfree function. It incorporates a microphone amplifier, a volume control of the receive channel and a duplex controller with signal and noise monitor on the transmit and receive channel. In contrary with the Philips handsfree circuits TEA1093 and TEA1094, the TEA1095 has neither integrated supply nor loudspeaker amplifier. This makes the TEA1095 more flexible for implementation in applications with external loudspeaker amplifier, external supply, such as cordless telephones and answering machines.

The function of the handsfree application will be illustrated with the help of fig 1.1.

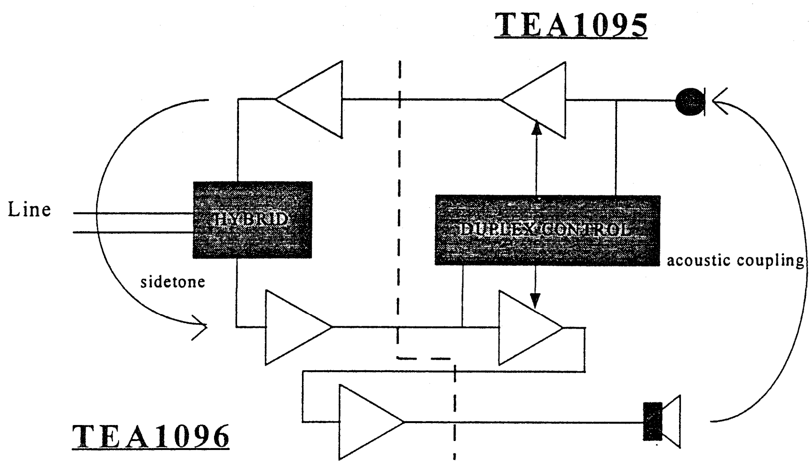


Fig.1.1 Handsfree telephone set principle

The left side of fig 1.1 shows a principle diagram of a part of the TEA1096 circuit by means of a receiving amplifier, a transmit amplifier, the loudspeaker amplifier and the hybrid. The right side of fig 1.1 shows a principle diagram of a part of the TEA1095 circuit by means of the microphone amplifier and the duplex controller.

As can be seen from fig 1.1, a closed loop is formed via the amplifiers, the antisidetone network and the acoustic coupling between loudspeaker and microphone. When the loop-gain is higher than one, the set starts howling. In a full-duplex application, this would be the case. To avoid howling, the duplex controller reduces the loop-gain to a value much lower than one.

The duplex controller of the TEA1095 monitors the signal and noise on both the transmit and the receive channel in order to detect which channel contains the 'largest' signal. As a result, the duplex controller reduces the gain of the channel which contains the smallest signal. This is done such that the sum of the transmit and the receive gains remains constant.

As a result, the circuit can be in three stable modes to be referred to throughout this report::

1. Transmit mode (Tx-mode): the gain of the microphone amplifier is at its maximum and the gain of the receive path (to loudspeaker amplifier) is reduced.

2. Receive mode (Rx-mode): the gain of the receive path is at its maximum and the gain of the microphone amplifier is reduced.
3. Idle mode (Ix-mode): the gain of the microphone amplifier and of the receive path are halfway their maximum and reduced values.

The difference between the maximum gain and the reduced gain is called the switching range.

This report gives a detailed description of the TEA1095 and its application with the TEA1096. The description is given by means of the block diagram of the TEA1095 (&2) and by discussing every detail of the sub-blocks (&3). The application is discussed by giving a guideline for application (the application cookbook &4) and by giving an application example (&5). EMC aspects are also discussed (&6). The appendices contain a measurement setup for the electro-acoustical adjustment of the TEA1095 handsfree application (A), a list of abbreviations (B) and application diagrams of the TEA1095 (fig. C1, C2).

2. BLOCK DIAGRAM

In this chapter, the block diagram of the TEA1095 is shown by means of fig.2.1. The pinning of the TEA1095 is given by means of fig.2.2. A short description of the block diagram is given including the function of the external components.

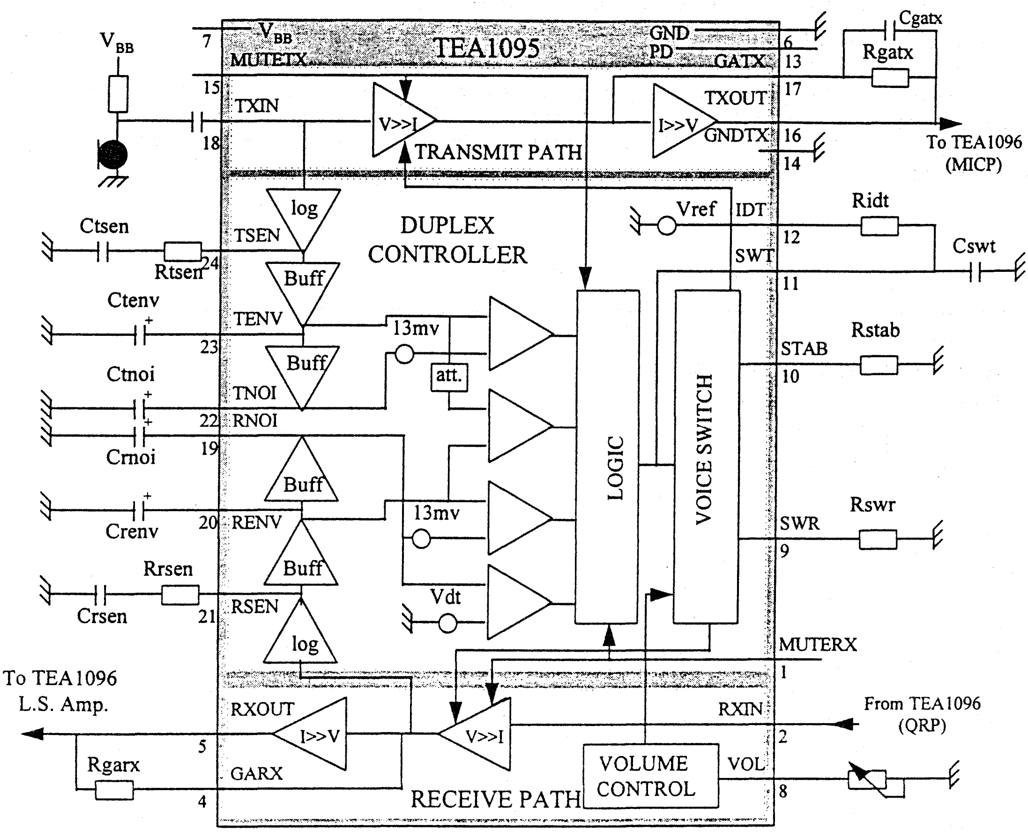


Fig. 2.1 Block diagram of TEA1095

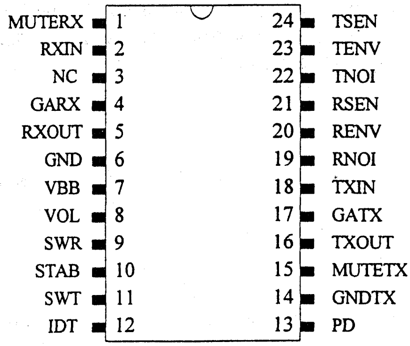


Fig. 2.2 Pinning of TEA1095

PIN	NAME	DESCRIPTION
1	MUTERX	Receive channel mute input
2	RXIN	Receive channel input
3	NC	Not connected
4	GARX	Receive gain adjustment
5	RXOUT	Receive channel output
6	GND	Ground reference
7	VBB	Positive supply input
8	VOL	Receive channel volume adjustment
9	SWR	Switching range adjustment
10	STAB	Reference current adjustment
11	SWT	Switching timing adjustment
12	IDT	Idle-mode timing adjustment
13	PD	Power-down input
14	GNDTX	Ground reference for microphone amplifier
15	MUTETX	Transmit channel mute input
16	TXOUT	Transmit channel output
17	GATX	Microphone gain adjustment
18	TXIN	Microphone amplifier input
19	RNOI	Receive noise envelope timing adjustment
20	RENV	Receive signal envelope timing adjustment
21	RSEN	Receive signal envelope sensitivity adjustment
22	TNOI	Transmit noise envelope timing adjustment
23	TENV	Transmit signal envelope timing adjustment
24	TSEN	Transmit signal envelope sensitivity adjustment

In fig.2.1 it can be seen that the IC consists out of four parts: the supply, the microphone amplifier, the receive path and the duplex controller. These blocks will be shortly described below including the function of the external components. The detailed description will follow in chapter 3.

Supply :

The circuit is supplied between pins VBB and GND. The TEA1095 can be switched into a low power consumption mode with the pin PD.

Microphone amplifier :

The handsfree microphone signal is amplified from pin TXIN to pin TXOUT. The signal reference is GNDTX, a "clean ground" which has to be connected to GND. The input TXIN has to be coupled to the microphone by means of a capacitor. The gain of the amplifier can be set with Rgatx. This amplifier can be muted by making pin MUTETX high.

Receive path :

The receive signal is amplified from pin RXIN to pin RXOUT. The input RXIN has to be coupled by means of a capacitor. The gain of the amplifier can be set with Rgarx, and the volume of the receive signal can be adjusted by means of the potentiometer connected between input VOL and GND. This channel can be muted by making pin MUTERX high.

Duplex controller :

From both the transmit and receive signal, signal and noise envelopes are made. The transmit signal envelope is on pin TENV and the receive one on pin RENV. The transmit noise envelope is on pin TNOI and the receive one on pin RNOI. The timing of the envelopes can be set by the capacitors Ctenv, Ctnoi, Crenv and Crnoi. The sensitivity of the envelope detectors can be set by means of the RC combinations Rtsen with Ctsen for the transmit envelope and Rrsen with Crsen for the receive one. The resistors set the sensitivity and the capacitors block the DC-component, creating also high-pass filters.

The logic determines to which mode (Tx, Rx or Ix-mode) the set has to switch over. The timing for switching to the Tx or the Rx -mode is determined with the capacitor Cswt. The timing for switching to the Ix-mode is set by the combination Cswt and Ridt. The switching range is determined by the resistor Rswr. Resistor Rstab has a fixed value.

3. DESCRIPTION OF THE TEA1095

This chapter describes in detail the four blocks of the duplex controller circuit TEA1095: the supply (3.1), the microphone amplifier (3.2), the receive path (3.3) and the duplex controller (3.4). For each block the principle of operation is described and its adjustments and performances are discussed.

All values given in this chapter are typical and at room temperature unless otherwise stated. For more details of the TEA1095 specification, see TEA1095 device specification.

3.1 Supply block

Principle of operation

As opposite to TEA1093 which includes an integrated supply which stabilizes a supply voltage out of the line current and as the TEA1094, the TEA1095 has no integrated supply. This makes the TEA1095 most suitable for applications where the handsfree controller is supplied from an external voltage source.

In fig. 3.1.1, different supply arrangements with TEA1095 are shown.

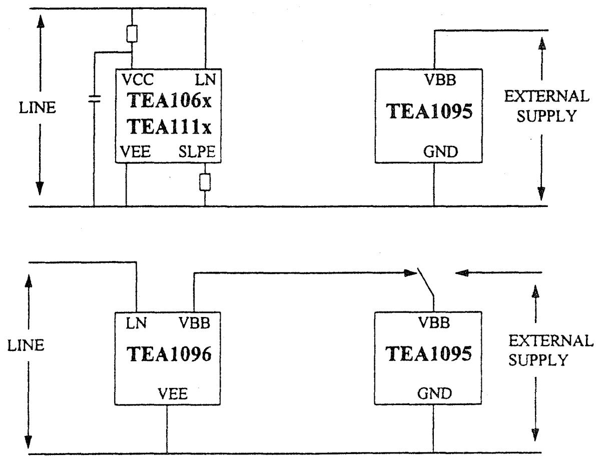


Fig. 3.1.1 Supply arrangement examples with TEA1095

As can be seen, the line current flows through the TEA106x or the TEA1096, while the TEA1095 is supplied from an external voltage source. Nevertheless, as can be seen in the bottom example, TEA1096 is able to provide this supply, leading to a line-powered very integrated solution. The common reference for all signals is GND on TEA1095 and VEE on TEA106x or TEA1096 side.

In case of a line-powered application using TEA106x or TEA111x, the TEA1095 as well as an external loudspeaker amplifier can be supplied via a large coil (or TEA1081) connected between line and VBB. At VBB, a

capacitor has to be connected to serve as reservoir. The TEA1095 and the loudspeaker amplifier have to be referenced at SLPE in order not to influence the transmission characteristics.
In fig.3.1.1 no galvanic insulation is drawn. When such is needed, the insulation can be done in the supply part or in the signal interfacing. In the supply part, the galvanic insulation can be done in the main-adaptor. In the signal interfacing between the TEA1095 and the TEA106x, the galvanic insulation can be done either with transformers or with opto-couplers.
The power consumption of the TEA1095 can be dramatically reduced when its functionalities are not required by making pin PD high.

Adjustments and performances

The voltage which can be applied between VBB and GND may vary between 2.9 V and 12 V. The 12 V is an absolute maximun rating. When a supply is used which may generate, during transients, a higher voltage, an appropriate protection device must be applied between VBB and GND in order to control this voltage.
The current consumption of the TEA1095 is typically 2.7 mA at VBB = 5 V. At higher voltages, the current consumption slightly increases.
In power-down mode, the current consumption is typically reduced to 140 μA at VBB = 5 V.

3.2 Microphone amplifier block

In the first paragraph of this chapter, the principle of operation of the microphone amplifier is described as well as its adjustments and performance. In the second paragraph, the mute transmit function is described.

3.2.1 Microphone amplifier

Principle of operation
In fig. 3.2.1 the block diagram of the microphone amplifier of the TEA1095 is depicted together with the interconnection with the TEA1096.

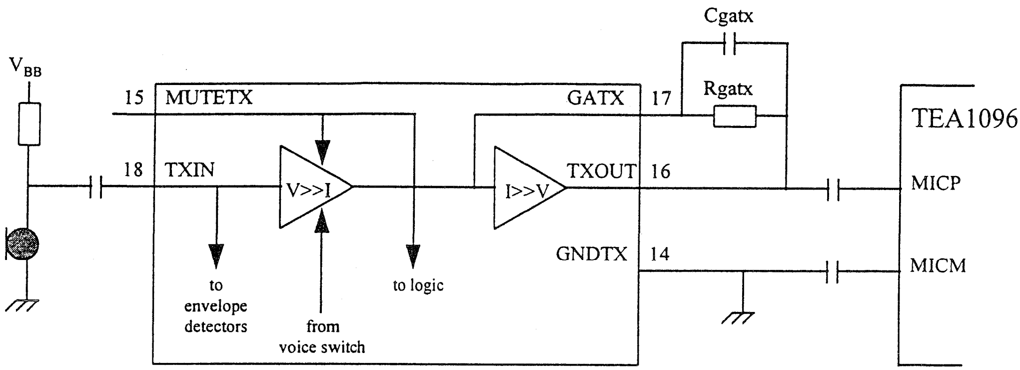


Fig. 3.2.1 Block diagram of the microphone path

As can be seen in fig.3.2.1, the microphone amplifier is referenced to pin GNDTX instead of being referenced to GND. This is in order to prevent interference from other blocks of the TEA1095 or of the application, GNDTX is called a clean ground. The input and output signals of the microphone channel have to be referenced to GNDTX. Pin GNDTX itself has to be referenced to GND.

The input of the microphone amplifier is pin TXIN. It is an assymmetrical input well suited for electret microphones. Induced signals in the short wire between the microphone and pin TXIN are assumed to be negligible. This in contrary with the handset microphone which is connected via the handset cord. The TEA106x/TEA111x families as well as the TEA1096 have symmetrical microphone inputs.

The output of the microphone amplifier is pin TXOUT. When interconnecting the TEA1095 and the TEA1096, pin TXOUT is preferably connected to the TEA1096 input MICP, in that case, pin MICM of the TEA1096 is connected to pin GNDTX.

As can be seen in fig.3.2.1, the microphone amplifier is built up out of two parts: a preamplifier and an end-amplifier. The gain of the preamplifier is determined by the duplex controller block, see & 3.4. The gain of the end-amplifier is determined by the external feedback resistor Rgatx.

The overall gain (Atx) of the microphone amplifier from input TXIN to output TXOUT in TX-mode is given as:

$$Atx = 20 * \log (0.72 * Rgatx / Rstab).$$

With Rstab being the resistor at pin STAB of 3.65 kΩ.

Adjustments and performances

A handsfree microphone, referenced to GNDTX, can be connected to the input TXIN via a DC blocking capacitor Ctxi. Together with the input impedance of pin TXIN of 20 kΩ, this capacitor form a first order high-pass filter which can be used to adjust the transmit curve.

The handsfree electret microphone can be supplied from VBB via a resistor. However, during normal operation, VBB may contain a small riple, and due to poor power supply rejection of electret microphones it is advised to add an RC smoothing filter in the feeding part, as shown in fig.3.2.2.

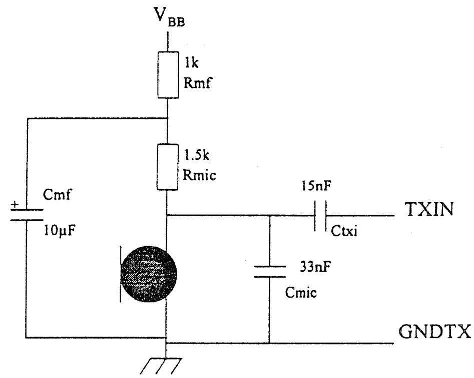


Fig. 3.2.2 Supply arrangement for electret microphone

As shown in fig.3.2.2, the RC smoothing filter is referenced to GNDTX in order to have one good reference for the whole microphone signal path. On the printed circuit board lay-out GNDTX can be connected with a separate wire to pin VEE of the TEA1096 in order to reduce ground interference as much as possible.

The sensitivity of the electret microphone is set via resistor Rmic. By putting a capacitor Cmic in parallel with the microphone, a first order low-pass filter is formed for the microphone signal in order to adjust the transmit curve.

Via the resistor Rgatx, the gain of the microphone amplifier can be adjusted from -15 to +25 dB to suit application specific requirements. With the resistor Rgatx = 30.1 kΩ, the gain equals typically 15.5 dB.

Capacitor Cgatx is applied in parallel with resistor Rgatx to ensure stability of the microphone amplifier, it also provides a first order low-pass filter for the adjustment of the transmit curve.

The input of the microphone amplifier can handle signals up to 18 mVrms with 2% total harmonic distortion. However, the microphone input signal is also used by the duplex controller, see &3.4. At 10 mVpeak at the input, the positive part of the signal on pin TSEN starts clipping which might influence the switching behavior. It is therefore advisable to keep the microphone input signal below this level.

The output drive capability at pin TXOUT is 20 µArms.

The output noise at TXOUT of the TEA1095 is -100 dBmp (psophometrically weighted) at a gain of 15 dB. With a sending gain of the TEA1096 set at 35 dB (total handsfree transmit gain of 50 dB), the noise level on the line will be : -65 dBmp.

In TX-mode, the noise level will be at its maximum. In Ix-mode and Rx-mode, the noise at TXOUT will be lower because the contribution of the preamplifier is reduced. However, the bottom level of the sending noise at TXOUT is limited by the end-amplifier and is about -110 dBmp.

The bottom level of the sending noise on the line is determined either by the speech circuit (TEA1096 or TEA106x/TEA111x) or by the noise at TXOUT increased by the transmit gain of the speech circuit, whichever is largest. When in Ix-mode, the noise on the line is due to TXOUT noise level, it can be reduced by changing the distribution of the gains between TEA1095 and speech circuit : increasing the transmit gain of the TEA1095 and decreasing with the same value the gain on the speech circuit side by modification of the attenuation placed between TXOUT and MICP of the speech circuit.

3.2.2 Transmit mute

During handsfree operation, the microphone can be muted by making pin MUTETX high, so conversation cannot be heard by the other party. When the microphone amplifier is muted, automatically the TEA1095 switches over to the RX-mode, see also &3.4.

When a logic high is applied to MUTETX, meaning the voltage on MUTETX is higher than 1.5 V, the microphone preamplifier is muted. The end-amplifier can still be used by applying a current signal on GATX. The obtained gain reduction is 80 dB. The current which has to be sourced into pin MUTETX when high is typically 2.5 µA.

When MUTETX is logic low, meaning the voltage on MUTETX is lower than 0.3 V or pin MUTETX is left open, the microphone amplifier is not muted.

The maximum allowable voltage on pin MUTETX is VBB+0.4 V, the minimum allowable is GND-0.4 V.

3.3 Receive path block

The block diagram of the complete receive path block is depicted in fig.3.3.1

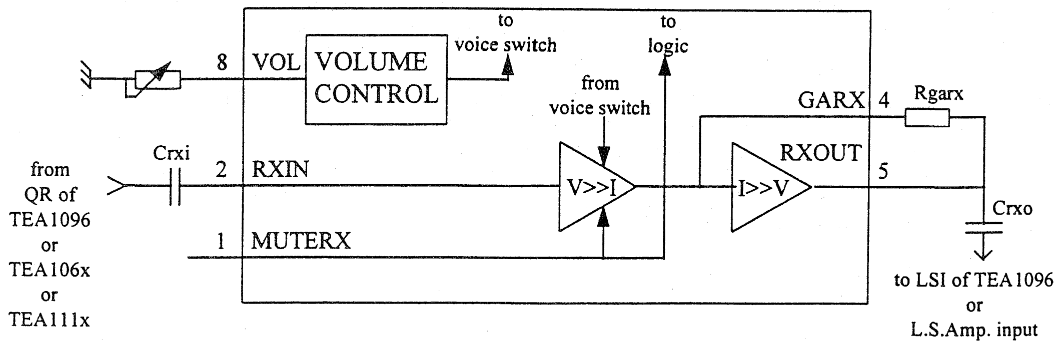


Fig.3.3.1 Principle of the receive path

As can be seen in fig.3.3.1, the receive channel is built up out of two parts: the receive path itself and the volume control. In the first paragraph of this chapter, the principle of operation of the receive path is described as well as its adjustments and performances. In the second paragraph, the same items are described for the volume control. In the third paragraph, the receive mute function is described.

3.3.1 Receive path

Principle of operation

As can be seen in fig.3.3.1, the input of the receive channel, pin RXIN is assymetrical and the signal has to be referenced to GND. The input RXIN can be connected, via a decoupling capacitor, to the earphone output QRP of the speech circuit (TEA1096 or TEA106x/TEA111x).

The output of the receive channel is pin RXOUT. When interconnecting the TEA1095 and the TEA1096, pin RXOUT is connected via a decoupling capacitor to input pin LSI of the TEA1096.

As can be seen in fig.3.3.1, the receive path itself is built up out of two parts: a preamplifier and an end-amplifier. The gain of the preamplifier is determined by the duplex controller block. The gain of the end-amplifier is determined by the external feedback resistor Rgarx.

The overall gain (Arx) of the receive path from input RXIN to output RXOUT is given as:

$$Arx = 20 * \log(0.46 * Rgarx / Rstab).$$

With Rstab being the resistor at STAB of 3.65 kΩ.

Adjustments and performances

The input signal for the receive channel has to be coupled in via the capacitor Crxi to block DC. Together with the input impedance of 20 kΩ at RXIN, a first order high-pass filter is introduced which can be used to adjust the receive curve and/or to reduce any low frequency unwanted signal coming from the line.

The input RXIN can handle signal up to 390 mVrms with a total harmonic distortion of 2%.

The input RXIN is biased at around 0V with respect to GND. By applying a signal to the input, it can become negative. The protection on this pin is made different from other pins which makes it possible to have RXIN as low as -1.2 V without damaging the circuit.

The output RXOUTcan be connected to the LSI input of TEA1096 or the input of any loudspeaker amplifier via a decoupling capacitor Crxo. The output is biased at 1.4 V referenced to GND. Together with the input impedance of the loudspeaker amplifier, Crxo forms a first order high-pass filter. With the resistor Rgarx, the gain of the receive path can be adjusted from -20 to +20 dB. The gain equals typically 6.3 dB with resistor Rgarx=16.2 kΩ. A capacitor Cgarx can be connected in parallel with Rgarx to provide a low pass filter which can be used to adjust the loudspeaker amplifier curve.

The output drive capability at pin RXOUT is 100µArms.

The noise level at the output RXOUT is -91 dBmp at a gain of 6 dB and with the input RXIN shorted with 200 Ω to GND.

3.3.2 Volume control

Principle of operation

Via the volume control block, the volume of the receive signal can be adjusted by the external potentiometer connected to pin VOL. By changing the potentiometer resistance, the gain of the preamplifier varies through the duplex controller. Volume control doesn't affect the transmit gain in Tx-mode.

Adjustments and performances

Out of pin VOL a current Ivol, set by Rstab, see &3.4, is flowing which is proportional to the absolute temperature (PTAT). At room temperature this current is around 10 µA. Together with the resistance of the potentiometer, the current Ivol creates a PTAT voltage on pin VOL. This PTAT voltage is processed by the volume control block, as a result, a temperature independent volume reduction of the output receive signal of 3 dB is obtained at approximately every increase of 950 Ω of the potentiometer resistance.

This means that a linear potentiometer can be used to control the volume logarithmically, thus in dB. With the advised value of 10 kΩ, the maximum gain reduction of the volume control is more than 30 dB. However, this maximum gain reduction is limited by the switching range, see &3.4. When the resistance of the potentiometer is zero, the receive gain is maximum in Rx-mode.

When digital volume control is desired, the switches can be either MOSFETs or analog switches with very low saturation voltage. Due to saturation voltage, it is advised not to use bipolar transistors as switches.

When a voltage is applied to pin VOL to control the volume, preferably this voltage has to be a PTAT voltage source. If not, the obtained gain varation is no longer temperature compensated.

3.3.3 Receive mute

During handsfree operation, the receive channel can be muted by making pin MUTERX high. As a result the receive signal is reduced by 80 dB, also the TEA1095 is internally forced into Tx-mode.

When a logic high is applied to MUTERX, meaning a voltage higher than 1.5 V, the receive preamplifier is muted. The end amplifier can still be used by applying a current signal on GARX. The obtained gain reduction is 80 dB when the volume control is set at maximum volume. The current which has to be sourced into pin MUTERX when high is typically 2.5 µA.

When MUTERX is logic low, meaning the voltage is lower than 0.3 V or pin MUTERX is left open, the receive path is muted.

The maximum allowable voltage on pin MUTERX is VBB+0.4 V, the minimum allowable is GND-0.4 V.

3.4 Duplex controller block

In this chapter, the principle of operation of the duplex controller will be described as well as its adjustments and performances. This will be done with the help of fig.3.4.1.

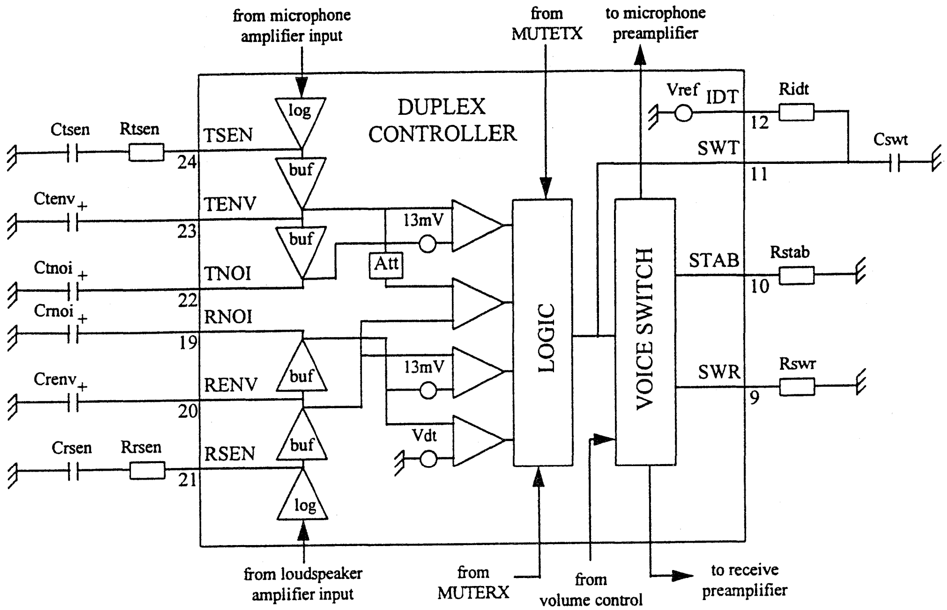


Fig. 3.4.1 Principle of the duplex controller

As can be seen in fig.3.4.1, the duplex controller is built up out of signal and noise envelope detector, decision logic and a voice switch.

The signal and noise envelope detectors determine the signal envelope and the noise envelope of both the transmit and receive signal. These envelopes are used by the decision logic to determine to which mode the TEA1095 has to switch over (Tx, Rx or Ix-mode). The logic charges and discharges the capacitor Cswt and the resulting voltage on pin SWT controls the voice switch. The voice switch switches over the TEA1095 between the three modes while keeping the loopgain constant.

In paragraphs 3.4.1 to 3.4.3, the principle of operation of the three parts is given. In paragraph 3.4.4, the adjustments and performances of the complete duplex controller are given.

3.4.1 Signal and noise envelope detectors

The signal and noise monitors of the transmit and receive channels are globally the same.

Therefore, the principle of the detectors will be explained with the help of one of them: the signal and noise detector of the transmit channel.

The microphone signal on pin TXIN is sent to the first stage of the detector, see fig.3.4.1. The first stage amplifies the microphone signal from pin TXIN to pin TSEN with an internal gain of 40 dB. Via the RC combination RtsenCtsen, the signal on TSEN is converted into a current. This conversion determines the sensitivity of the envelope detector. The current is logarithmically compressed and internally converted to a voltage which represents the compressed microphone signal. At room temperature, an increase of the microphone signal with a factor of 2 will increase the signal envelope with 18 mV if the current through TSEN stays between 0.8 and 160 μArms. Outside this region the compression is less accurate.

The compressed microphone signal is buffered by the second stage to pin TENV. As the buffer can source 120μA and sink 1μA, the signal on TENV follows the positive peaks of the compressed signal, this signal is called the signal envelope. The time constants of the signal envelope are therefore determined by the combination of the internal current sources and the capacitor Ctenv.

The voltage on TENV is buffered by the third stage to pin TNOI. As this buffer can source 1 μA and sink 120 μA, the signal on TNOI follows the negative peaks of the signal on TENV. This is called the noise envelope because it represents the background noise. The time constants of the noise envelope are determined by the combination of the internal current sources and the capacitor Ctnoi. Both capacitors Ctnoi and Crnoi are provided with a start-up circuit. During start-up the capacitors are charged with approximately 40 μA up to 1.9 V. The starter will restart when the voltage on the capacitors drops below 0.9 V.

As can be seen in fig.3.4.1, the principle of operation of the signal and noise envelope detectors of the receive channel is equal to the one of the transmit channel. However, the gain of the first stage (input to pin RSEN) is 0 dB instead of 40 db for the transmit channel, this is in order to compensate the level on TXIN which is not yet amplified.

The behavior of the envelopes is illustrated in fig.3.4.2 where the signal and noise envelope of one channel are depicted together with the input signal.

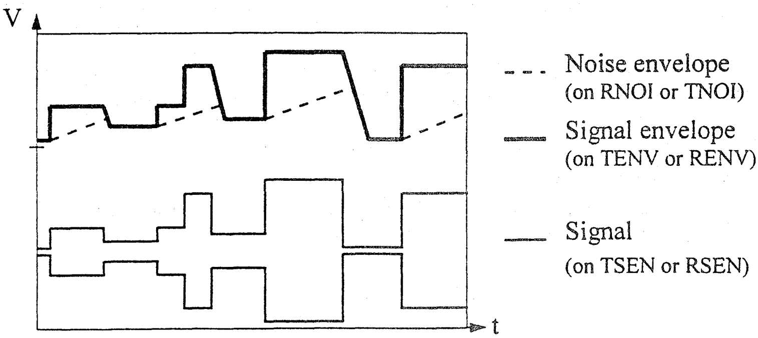


Fig. 3.4.2 Typical behavior of the signal and noise detectors

In fig.3.4.2 it is shown that when the input signal raises quickly, the envelope signal follows immediately and the noise envelope slowly follows the envelope signal. When the input signal decreases, the envelope signal follows immediatly but nevertheless less quickly than when it raises, the noise envelope follows immediatly the decrease of the envelope signal and never crosses it.

3.4.2 Decision logic

The signal and noise envelopes of the transmit and receive signal are used by the decision logic to determine in which mode the TEA1095 has to be.

The output of the logic is a current source which charges or discharges the capacitor Cswt at pin SWT. If the logic determines Tx-mode, the capacitor Cswt is discharged with 10 µA. When Rx-mode is determined, Cswt is charged with 10 µA. When Ix-mode is determined, the current source is zero and the voltage on SWT becomes equal to the voltage on pin IDT via the current provided through the resistor Ridt. The time constants of the duplex controller are therefore determined by the combination of the internal current sources, the capacitor Cswt and the resistor Ridt.

As can be seen in fig.3.4.1, the envelopes are not used directly by the decision logic.

First, to have a clear choice between signal and noise, the signal is considered as speech when its envelope is more than 4.3 dB above the noise envelope. At room temperature, this is equal to a voltage difference of 13mV. This so called speech/noise threshold is implemented in both the receive and the transmit channel. At the end of paragraph 3.4.4 a way to increase this threshold is discussed.

Second, the signal on TXIN contains both the signal of the local talker as well as the signal coming from the loudspeaker (acoustic coupling). In Rx-mode, the contribution of the loudspeaker overrules the contribution of the local talker. As a result, the signal envelope on TENV is mainly formed by the loudspeaker signal, to correct this, an attenuator is placed between TENV and the TENV/RENV comparator. The attenuation equals the attenuation applied to the microphone amplifier gain. Thus when the TEA1095 is in Rx-mode, the attenuation equals the switching range.

Third, when a dial tone is present on the line, without measures this would be recognized as noise after some delay because its level is constant. As a result, the TEA1095 would go to Ix-mode and the user of the set would hear the dial tone fade away. Therefore, a dial tone detector is incorporated which doesn't consider input signals as noise when they have a level higher than the dial tone level. The dial tone level, represented by Vdt in fig.3.4.1, is adjustable by Rsen.

When these three corrections are made, the signal and noise envelopes are used by the comparators and the logic. As already explained, the output of the logic is a current source. The relation between the current source and the output of the comparators is given in the table of fig.3.4.3. If for instance, TENV>RENV (transmit signal larger than receive signal) and TENV>TNOI (transmit signal more than 4.3 dB larger than noise level), then the output current will be -10 µA.

Comparator TENV/TNOI	1	x	x	0	x
Comparator TENV/RENV	1	0	0	1	0
Comparator RENV/RNOI	x	1	x	x	0
Comparator RNOI/Vdt	x	x	1	x	0
Output current	-10µA	+10µA	+10µA	0µA	0µA

Fig.3.4.3 Truth table of the detection logic

When pin MUTETX is made high, see paragraph 3.2.2, the output current is forced to be +10 µA, which forces the TEA1095 into Rx-mode and mutes the microphone amplifier. When pin MUTERX is made high, see paragraph 3.3.3, the output current is forced to be -10 μA, which forces the TEA1095 into Tx-mode and mutes the receive path. When both MUTETX and MUTERX are made high, both channels are muted.

The voltage on pin SWT is internally limited to IDT-0.4 V and IDT+0.4 V.

3.4.3 Voice switch

With the voltage on pin SWT, the voice switch regulates the gain of the microphone preamplifier and the receive channel preamplifier in such a way that the sum of the transmit and receive gain is kept constant. This is done to keep the loop gain of the handsfree telephone set constant, see also the introduction &1. The switch-over behavior of the voice switch will be described with the help of fig.3.4.4.

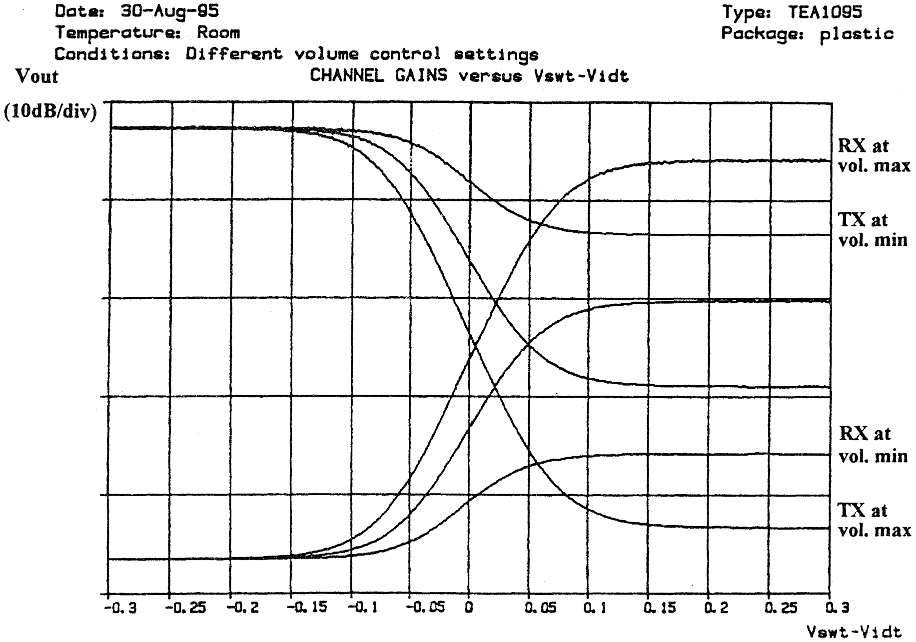


Fig. 3.4.4 Behavior of the voice switch

When the voltage on SWT is more than 180 mV below the voltage on IDT, the TEA1095 is fully switch to Tx-mode (gain of the transmit path at maximum and gain of the receive path at minimum). When the voltage on SWT is more than 180 mV above the voltage on IDT, the TEA1095 is fully switch to Rx-mode (gain of the receive path at maximum and gain of the transmit path at minimum). The TEA1095 is considered to be in Ix-mode when the voltage on SWT equals the voltage on IDT. When the capacitor Cswt is charged or discharged, the voltage on SWT varies and as a result the voice switch will smoothly switch over between the modes keeping the sum of the transmit and receive gains constant.

The difference between the maximum and the minimum gain of the receive or transmit preamplifiers is called the switching range. This range is determined by the ratio of Rswr and Rstab, see paragraph 3.4.4. Both Rswr and Rstab set internally used reference currents which are proportional to absolute temperature (PTAT).

As already stated in &3.3 the volume control acts upon the receive preamplifier via the control of the voice switch. As a result, the loop gain of the handsfree set is kept constant when the volume of the receive path is adjusted. However, the voice switch is designed such that the volume control has no influence in Tx-mode. In the extreme case, when the volume of the receive channel is reduced with the value of the switching range, the TEA1095 virtually does not switch over. In order to avoid inversion of the gain in Rx-mode, the volume control range of the TEA1095 cannot be larger than the switching range.

3.4.4 Adjustments and performances

The adjustment of the duplex controller has to be performed according to the following recipe:

1. Determine the switching range
2. Determine dial tone detector level
3. Determine sensitivity
4. Determine timings

Ad 1. Determine switching range

The switching range Asw is determined by the ratio of the two resistors Rswr and Rstab according to:

Asw (dB) = 20 * log (Rswr / Rstab)

The resistor Rstab has to be taken 3.65 kΩ. The value of the resistor Rswr can vary between 3.65 kΩ and 1.5 MΩ resulting in a switching range between 0 dB and 52 dB. With Rswr of 365 kΩ, the switching range is typically set to 40 dB.

The switching range is calculated out of the loop gain (Aloop). In a handsfree application, the loop gain has to be smaller than one (<0 dB) and can be calculated as follows:

Aloop = Atx1095 + Atx1096 + Ast + Arx1096 + Arx1095 + Als1096 + Aac - Asw.

with Atx1095 = sending gain of the TEA1095 (TXIN to TXOUT)

Atx1096 = sending gain of the TEA1096 (MIC to LN)

Ast = electrical sidetone

Arx1096 = receive gain of the TEA1096 (LN to QR)

Arx1095 = receive path gain of the TEA1095 (RXIN to RXOUT)

Als1096 = loudspeaker amplifier gain of the TEA1096 (LSI to QLS)

Aac = electro-acoustic coupling from loudspeaker to microphone (QLS to TXIN)

Asw = switching range

In this calculation, the worst case has to be taken for Ast and Aac. Furthermore, for safety, it is advised to choose Asw large enough to compensate spreads (margin from 10 to 15 dB).

The electrical sidetone is the difference (in dB) between the wanted receive signal on the TEA1096 and the unwanted part of the transmit signal received while having an equal signal level on pin LN for both the transmit and the receive signal. Ast is dependent of frequency and connecting conditions of the set (line lenght, line impedance).

The acoustic coupling is dependent on the environment of the telephone set, for the determination of Aac, the worst condition has to be searched.

If a certain minimum volume control range is required, the switching range must not be chosen smaller.

In appendix A, a method for measuring the required switching range, based on the above calculation, is given.

It is also possible to determine the switching range by experiments:

As for the calculation, it is necessary to identify what are the worst conditions for sidetone and acoustic coupling. In these worst conditions, Rswr can be adjusted in such a way that the handsfree telephone set is at the limit of howling. Then the determined value of Rswr must be increased in order to have a margin of 10 dB to 15 dB.

Handsfree behavior will be more comfortable for the user if the switching range is not too large. So, **it is advised to take care of the acoustic coupling between the loudspeaker and the microphone** which might come from the cabinet of the terminal itself.

Ad 2. Determine dial tone detector level

The dial tone detector level is determined by the value of Rrsen according to:

Vdialtone = 4.2 μA * Rrsen

With Rrsen of 10 kΩ, the dial tone detector level will be 42 mVrms. This means, a continuous signal on the input RXIN larger than 42 mVrms will be recognized as a dial tone.

Ad 3. Determine sensitivity

The sensitivity is set by Rrsen and Rtsen. The resistor Rrsen is already determined by the dial tone detector level. It must however be checked if the choosen value for Rrsen is a practical one for the dynamic range of the logarithmic compressor. The optimized range for the compression is when the current flowing through pin RSEN is between 0.8 to 160 μArms. This means that at nominal receiving signal the current through RSEN is preferably around 11 μArms. This gives a maximum dynamic range of plus and minus 23 dB.

The same counts for pin TSEN.

The resistor Rtsen has to be chosen in such a way that both channels have the same priority for the duplex controller. This can be obtained by choosing Rtsen according to:

20 * log (Rtsen) = 20 * log (Rrsen) - Atx1095 - Atx1096 - Ast - Arx1096 + Atsen + 1/2 Aloop

with Atsen = internal gain from TXIN to TSEN = 40 dB.

In this relation, the maximum loop gain and the worst case sidetone are used. If it is prefered to give the transmit channel priority above the receive channel, the value of Rtsen has to be chosen smaller. For the opposite, the value of Rtsen has to be chosen larger. With respect to the calculated setting, Rtsen and Rrsen can be varied with plus and minus 1/2 * Aloop (in dB).

In appendix A, a method for determining the required Rtsen, based on the above calculation, is given.

The capacitors Ctsen and Crsen form first order high-pass filters respectively with Rtsen and Rrsen to reduce influence of low frequencies on the switching behavior. It is suggested to choose the capacitors Ctsen and Crsen such that the cut-off frequencies of the filters are similar.

When the calculated sensitivity setting is implemented, subjective tests with real telephone lines will be necessary to come to the optimal sensitivity setting.

Once Rrsen is determined, it would also be possible to determine Rtsen only by experiments. In this case, subjective tests with different line conditions (attenuation, impedance, length) have to be carried-out until the optimal sensitivity setting is found.

Ad 4. Determine timings

The timings which can be set are : signal envelope timing and noise envelope timing for both channels, switch-over timing and idling timing.

The signal envelope timing is set by the capacitors Ctenv and Crenv. Because of the logarithmic compression between TSEN and TENV respectively RSEN and RENV, the timing can be expressed in dB/ms. At room temperature, the following relation counts :

Timing $\cong$ I / (3 * C) (in dB/ms)

With I = charge or discharge current from pin TENV, RENV, TNOI, RNOI

C = timing capacitor Ctenv, Crenv, Ctnoi, Crnoi

With the advisable signal envelope timing capacitors Ctenv and Crenv of 470 nF, the maximum attack-timing of the signal envelopes will be around 85 dB/ms (I=120 µA). This is enough to track normal speech. The release timing will be 0.7 dB/ms (I=1 µA). This is enough to smoothen the signal envelope and to eliminate the influence of room echoes on the switching behavior.

With the advisable noise envelope timing capacitors Ctnoi and Crnoi of 4.7 µF, the attack timing of the noise envelopes will be 0.07 dB/ms (I=1 µA). This is small enough to track background noise and not to be influenced by speech bursts. The maximum release timing will be 8.5 dB/ms (I=120 µA). This is enough to track the signal envelope during release because the signal envelope release timing is 0.7 dB/ms which is a factor smaller. It is advised to choose the signal envelope timing and the noise envelope timing of both channels equal for optimum operation of the duplex controller.

The switch-over timing is determined by the value of the switch-over capacitor Cswt. The idling timing is determined by the combination of Cswt and the idling resistor Ridt.

The output current of pin SWT is Iswt, a voltage difference over Cswt can be obtained according to :

δVswt/t = Iswt / Cswt (mV/ms).

With the advised value of 220 nF for Cswt, the obtained voltage difference is 45 mV/ms. The switch-over time is dependent on the voltage difference which has to be generated on pin SWT. Suppose the set is in full Tx-mode then the voltage on SWT will be V(IDT)-400 mV, see fig.3.4.4. To reach Rx-mode a voltage difference of 580 mV must be generated to end up a voltage of V(IDT)+180 mV. So in this case the switch-over time will be 13 ms. When the set is in Ix-mode, the voltage on SWT equals the voltage on IDT, in that case switching to Tx-mode or to Rx-mode requires a voltage generation of only 180 mV and they will be reached in 4ms.

The idling timing is determined by an RC time constant. It is supposed that Ix-mode is reached when a time (tidt) is elapsed:

tidt = 4 * Ridt * Cswt

With the advised value for Ridt of 2.2 MΩ, an idling time of around 2 seconds is obtained. To have a clearly determined idling timing, it is advised not to use a capacitor with a high leakage current.

When the calculated timing settings are implemented, subjective tests with real telephone lines will be necessary to be sure that the optimal timings have been set.

Miscellaneous

When a handsfree telephone set is used at one end of the line and a conventional set at the other end,the user of the conventional set may think that the line is cut when the handsfree set stays in receive mode while no signal on the line is present. This is avoided when the handsfree set switches over to idle mode. This mode is incorporated in the TEA1095 and is placed exactly at mid attenuation between transmit and receive mode. When it is desired to have an idle mode which is closer to transmit than receive mode, the circuit of fig.3.4.5 can be applied.

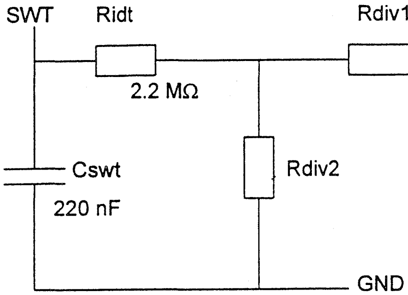


Fig.3.4.5 Circuit for shifting the idle mode

With the circuit of fig.3.4.5, in idle mode, the voltage on SWT will not go to the voltage on IDT but to the voltage on IDT minus the voltage drop over Rdiv1. The voltage drop over Rdiv1 determines the shift of the idle mode (in dB). This shift can be read from fig.3.4.4 , when the voltage drop over Rdiv1 is taken as the X-axis value. The voltage on IDT is approximately 1.2 V, so with for instance Rdiv1 = 33 kΩ and Rdiv2 = 1 MΩ, the shift will be about 10 dB. It is advised not to choose Rdiv2 lower than 1Mohms inorder to limit the current drawn from IDT. By connecting Rdiv2 to VBB instead of GND, the idle mode is shifted towards the receive mode.

In noisy environments, like offices, a handsfree set can show an unsteady behavior in idle mode (unwanted switching over from Ix to Tx-mode). In the TEA1095, this unsteady behavior is reduced by the implemented speech/noise threshold of 4.3 dB. However, when a larger threshold is required, this can be achieved by connecting a resistor Rtnoi in series with Ctnoi.

When there is only noise present at the input of the envelope detector, the voltages on pins TENV and TNOI are equal. When suddenly, a signal is present, the level on TENV will increase. Without Rtnoi, the voltage on TNOI will increase only slowly because of the charging current of 1 µA. When a resistor Rtnoi is placed in serie with Ctnoi, under the same conditions, this 1 µA current will cause a voltage jump on TNOI. This jump determines the shift of the speech /noise threshold. As depicted in fig.3.4.1, at room temperature, the 4.3 dB threshold equals 13 mV. A resistor Rtnoi in series with Ctnoi will add an extra voltage to this threshold of 1 µA * Rtnoi. When for instance, a resistor of 10 kΩ is chosen for Rtnoi, the speech/noise is increased to 23 mV which is equal to 7.6 dB at room temperature. The new speech/noise threshold is slightly dependent on temperature and on the spread of the internal current source and therefore less accurate than the internal 4.3 dB. It is advised not to use a resistor larger than 15 kΩ.

4. APPLICATION COOKBOOK

In this chapter, the procedure for making a basic application with a speech and listening-in IC TEA1096 and the handsfree duplex controller TEA1095 will be given. With the help of fig. C1 in appendix, the design flow is given as a number of steps which should be made. As far as possible for every step, the components involved and their influence on every step are given. The preferred values are given between brackets.

The application of fig. C1 is a basic application of the TEA1095 voice switched speakerphone IC with the TEA1096 speech and listening-in circuit.

As can be seen in fig. C1, only a few components have a fixed value. All other values will follow from the cookbook of fig.4.1.

Step	Adjustment
DC setting :	
Adjust the DC setting of the TEA1096 to the local PTT requirements.	
Voltage LN-VEE	
DC slope	Refer to [4] and/or to [12] (N° : ETT94001)
Supply point VCC	
Supply point VBB	
Artificial inductor	
Impedance and sidetone :	
After setting the required set impedance, the sidetone has to be optimized using the two sidetone netwoks in order to minimize the loop gain in all line conditions. AGC can be adjusted at that step.	
Set active impedance	Refer to [4] and/or to [12] (N° : ETT94001)
Sidetone	
AGC	
TEA1096 transmit and receive gains	
Transmit gain	Total transmit and receive gains have to be siplitted between TEA1096 and TEA1095. It is suggested to leave 15dB of transmit gain for the TEA1095 and to adjust the receive gain between LN and QRP at -3 dB. Ctxo and MICP input impedance form a high-pass filter. A capacitor in parallel with the transmit gain resistor (between TEA1096 pins 17 and 11) form a low-pass filter. A resistor bridge can be inserted between TXOUT of TEA1095 and MICP of TEA1096.
Receive gain	A capacitor in parallel with the receive gain resistor (between TEA1096 pins 26 and 25) form a low-pass filter. For more details, refer to [4] and/or to [12] (N° : ETT94001)
TEA1095 microphone amplifier (see & 3.2):	
After the sensitivity and the curve of the microphone are adjusted, the gain can be adjusted to the desired value	
Microphone sensitivity	Rmic sets the sensitivity. Together with Rmf, Rmic provides the polarisation of the electret.
Frequency curve	Cmic with Rmic and the output impedance of the electret form a low-pass filter. Ctxin with the 20 kΩ input impedance at TXIN form a high-pass filter.
Transmit gain and stability	Rgatx sets the microphone amplifier gain : Atx=20*log(0.72*Rgatx/Rstab) Cgatx may be necessary for stability and forms a low-pass filter with Rgatx.

Step	Adjustment
TEA1096 loudspeaker amplifier :	
The gain is fixed at 35.5 dB , a high-pass filter can be made, the dynamic limiter timing can be chosen.	
Frequency curve	Crxo and/or capacitor in series with loudspeaker can form high-pass filters
Dynamic limiter timing	Capacitor at pin DLL.
TEA1095 receive channel (see & 3.3) :	
The gain of the receive pass and the curve can be adjusted. The volume control range can be chosen.	
Receive gain	Rgarx : the total receive gain of the set is equal to receive gain of the TEA1096 (-3 dB ?), the gain of the loudspeaker amplifier (35.5 dB) and the gain set with Rgarx.
Receive curve	Crxi with the input impedance of 20 kΩ at pin RXI form a high-pass filter (a cut-off frequency between 100 and 200 Hz is advised),
Volume control	Cgarx in parallel with Rgarx forms a low-pass filter. A linear potentiometer of 10 kΩ is suggested (3 dB for each 950 Ω).
TEA1095 Duplex controller (see & 3.4) :	
When all gains are adjusted, the switching range can be determined. Then the dial tone detector level followed by the sensitivities can be set. Finally the timings of the envelopes and the switching are adjusted.	
Switching range	Loop gain : Aloop=Atx1095+Atx1096+Ast+Arx1096+Arx1095+Als1096+Aac-Asw < 0dB Choose Asw with safety margin Adjust Rswr : Asw=20log(Rswr/Rstab) with Rstab fixed at 3.65kΩ
Dial tone detector	Rrsen : Vdialtone=4.2 μA*Rsen
Sensitivities	Rtsen for balanced sensitivities between Tx and Rx Ctsen form a high-pass filter with Rtsen Crsen form a high-pass filter with Rrsen
Signal envelopes	Ctenv (0.47 μF), Crenv (0.47 μF) : maximum attack : 120 μ/(3*Cenv) (dB/ms), release : 1 μ/(3*Cenv) (dB/ms)
Noise envelopes	Ctnoi (4.7 μF), Crnoi (4.7 μF) : attack : 1 μ/(3*Ctnoi) (dB/ms), maximum release : 120 μ/(3*Ctnoi) (dB/ms)
Switch-over timing	Cswt (220 nF) : δVswt/t=10μ/Cswt (mV/ms)
Ix-mode timing	Ridt (2.2 MΩ) : time constant=4*Ridt*Cswt

Fig. 4.1 Steps in the design flow of the TEA1095 with TEA1096

5. APPLICATION EXAMPLE

In this chapter, a basic application example of the TEA1095 with the TEA1096 is depicted. Only the essential elements are given, for instance no bridge, no protection, no ringer, no line interruptor are included.

Fig. C2 gives the basic handsfree application of the TEA1095 together with the speech and listening-in circuit TEA1096. This application doesn't include handset telephony (which would have to be added by means of analog switches) but only basic handsfree telephony.

- **DC setting** : nominal DC characteristic and nominal V_{BB} of 3.6 V are chosen.

- **Impedance and sidetone** : the impedance is a complex one (220 Ω, 825 Ω//115 nF). A standard 600 Ω impedance would save six components.

- **TEA1096 transmit and receive gains** : the transmit gain of TEA1096 is set at 30.5dB by means of a resistor bridge (22 kΩ, 7.5 kΩ at microphone input combined with an internal gain of 43 dB set by 32.5 kΩ). The cut-off frequency of the high-pass filter is set at 110 Hz by 2*100 nF in series, the cut-off frequency of the low-pass filter is set at 7.2 kHz by 680 pF//32.5 kΩ.

The receive gain of TEA1096 is set at -3 dB by 82.5 kΩ. The cut-off frequency of the low-pass filter is set at 4.1 kHz by 470 pF//82.5 kΩ.

-**TEA1095 microphone amplifier** : the sensitivity of the electret microphone is set by Rmic=1.5 kΩ. The microphone power supply is filtered by Rmf, Cmf with a cut-off frequency of 16Hz. An electrical low-pass filter is provided with Cmic, in combination with the effect of the cabinet and the self curve of the electret, it adjusts the high frequency part of the transmit curve. A high-pass filter is formed with Ctxin and the 20 kΩ input impedance at TXIN, its cut-off frequency is set at 530 Hz (a higher cut-off frequency may be necessary to better minimize the room echo and/or to better compensate the line attenuation).

The gain of the microphone channel is set at 15.5 dB by Rgatx=30.1 kΩ, a low-pass filter with Cgatx=2.2 nF would provide a cut-off frequency of 2.4 kHz.

- **Loudspeaker amplifier** : nominal loudspeaker amplifier application is made (gain of 35.5 dB).

- **TEA1095 receive channel** : the gain of the receive path of the TEA1095 is set at -7 dB by Rgarx=3.65 kΩ (the total receive gain of the set is : 35.5 - 3 - 7 = 25.5 dB). A high-pass filter with a cut-off frequency of 120 Hz is formed by Crxin=68nF and the input impedance of 20 kΩ at RXIN. A low-pass filter with Cgarx=15 nF would provide a cut-off frequency of 2.9 kHz for the adjustment of the loudspeaker curve.

A 10kΩ linear potentiometer provides a volume control range of about 31dB.

- **Duplex controller** : after optimized acoustic adaptation (**reduction of acoustic coupling between loudspeaker and microphone** and good sidetone adaptation) the switching range is set at 40dB by Rswr=365k, including safety margin.

With Rrsen=10 kΩ, the dial tone detector level is set at 42 mV which means 60 mVrms on line.

With Crsen=100 nF, the cut-off frequency of the high-pass filter is 160 Hz.

Rtsen and Ctsen respectively equal Rrsen and Crsen in order to balance transmit and receive sensitivities.

With Ctenv and Crenv=0.47 µF, the max. attack time of the signal envelopes is set at 85 dB/ms.

With Ctnoi and Crnoi=4.7 µF, the max. attack time of the noise envelopes is set at 0.07 dB/ms.

The switch-over timing is nominally set by Cswt=220 nF and the idling timing by Ridt=2.2 MΩ.

Due to its large flexibility (range of adjustments and architecture), the TEA1095 can be implemented in a lot of environments, offering handsfree function with various speech interfaces and loudspeaker amplifiers, with or without external supply.

6. ELECTROMAGNETIC COMPATIBILITY

As no common international specification exists for RFI immunity, and as different assembly methods may lead to different solutions, only some advices can be provided.

It is advisable to take care of the impedance of the GND, the smallest is always the best. Even if it is required to separate low level microphone signals on GNDTX from high level signals (loudspeaker or others), GND and GNDTX traces must be as wide as possible.

Also, the connection of Rstab, Rswr, Rgatx and Rgarx has to be done with very short traces (specially STAB input which sets all the gains must be very immune).

VOL and TXI inputs may also be sensitive (RF signals entering these pins would be amplified). Rvol can be connected with short traces if possible or VOL input may be lightly decoupled by a capacitor to GND. Care has to be taken with the lay-out of the microphone amplifier, which is also helpfull for the noise, providing a good decoupling to GNDTX. A low-pass RC filter may be added at the input of the amplifier.

It is not allowed to put a capacitor directly between STAB and GND, only an RC network can be implemented if it helps (365 Ω, 4.7 nF).

RXIN input may also be decoupled to GND.

A low impedance capacitor in parallel with the electrolythic one between VBB and GND may help.

7. REFERENCES

[1] TEA1095 Voice Switched Speakerphone IC
Device specification

[2] Philips Semiconductors
SEMICONDUCTORS FOR TELECOM SYSTEMS - IC03 -

[3] TEA1094 Handsfree IC
Device specification

[4] TEA1096 / TEA1096A Line interface and Listening-in ICs
Device specification

[5] TEA1081 Supply circuit with power-down for telephone set peripherals
Device specification

[6] TEA1083 / TEA1083A Call progress monitor for line-powered telephone set
Device specification

[7] TEA1085 / TEA1085A Listening-in circuit for line-powered telephone set
Device specification

[8] TEA1112 / TEA1112A / TEA1113 Low voltage versatile telephone transmission circuits with dialler interface
Device specifications

[9] TEA1062 / TEA1062A Low voltage versatile telephone transmission circuits with dialler interface
Device specification

[10] TEA1064A / TEA1064B Low voltage versatile telephone transmission circuits with dialler interface and level dynamic limiting
Device specification

[11] Application of the TEA1094 handsfree circuit (ETT/AN94004)

[12] Application of the TEA1096 transmission and listening-in circuit (ETT94001)

[13] Measures to meet EMC requirements for TEA1060-family speech transmission circuits (ETT89016)

[14] Application of the speech transmission circuit TEA1062 (ETT89008)

[15] Application of the versatile speech/transmission circuit TEA1064 in full electronic telephone sets (ETT89009)

[16] Philips Semiconductors
Wirebound telecom
APPLICATIONS HANDBOOK

APPENDIX A MEASUREMENT SETUP

This appendix describes how the loop gain and the sensitivity setting can be measured. These measurements provide a base for obtaining of the final values which, however, can only be reached by performing subjective tests.

For determining the loop gain, the worst case for both acoustical coupling and sidetone must be considered. This can be done with the measurement setup of fig. A.1.

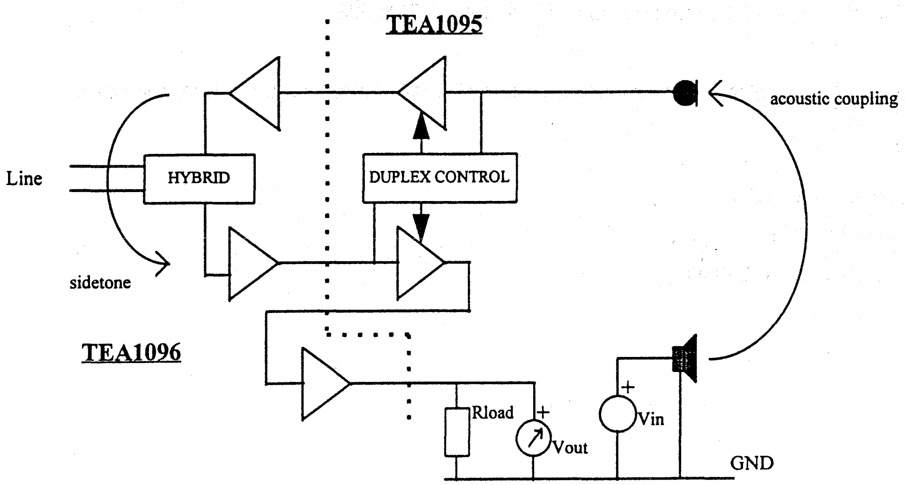


Fig. A. 1 Setup for loop gain measurement

The loudspeaker is disconnected from the set and an electrical signal is applied to it. The acoustical signal coming from the loudspeaker is coupled to the microphone (Aac), transmitted to the line (Atx1095 + Atx1096), returns via sidetone (Ast) and is amplified again to the loudspeaker output of the TEA1096 (Arx1096 + Arx1095 + Als1096). The output is loaded with a loudspeaker equivalent impedance (Rload), it has to be checked that the TEA1096 loudspeaker amplifier is not saturated and that its gain is not reduced by the dynamic limiter. The total gain of this loop is reduced with the switching range (Asw).

The gain from the loudspeaker connection (Vin) to the loudspeaker amplifier output (Vout) is the loop gain (Aloop), given as:

$$20 * \log (Vout / Vin) = Aloop$$

$$Aloop = Aac + Atx1095 + Atx1096 + Ast + Arx1096 + Arx1095 + Als1096 - Asw$$

This measurement has to be carried-out with a Vin level avoiding saturation of the loudspeaker amplifier (start with 100 mVrms) in the frequency range 200 Hz to 5000 Hz (100 to10000 Hz for high quality aparatus).

From the curve obtained, the maximum gain (<0 dB) indicates the gain margin.

Sidetone curve changes with line impedance, so, the Aloop curve will change in the same way. It is advisable to double-check with different line conditions (impedance and length) that the identified worst line condition is really the worst all over the frequency range.

Moving around the set also modifies acoustic coupling, so, it is advised to proceed as for sidetone.

The measurement combining both worst case conditions of sidetone and acoustical coupling gives the worst case loop gain and thus the gain margin.

All transmit and receive gains and curves are supposed to be set, the value of the switching range however, can only be finalized after the measurement. Therefore, during the measurement, the switching range must be set to a preliminary value. An advised preliminary value for measurement is 40 dB.

When the correct switching range is set, the sensitivity settings can be determined.

Rrsen is determined to get the correct tone detector level.

For determining the sensitivity, calculations can be done as described in &3.4.4. It is also possible measuring the correct sensitivity setting with measurement setup of fig. A. 2.

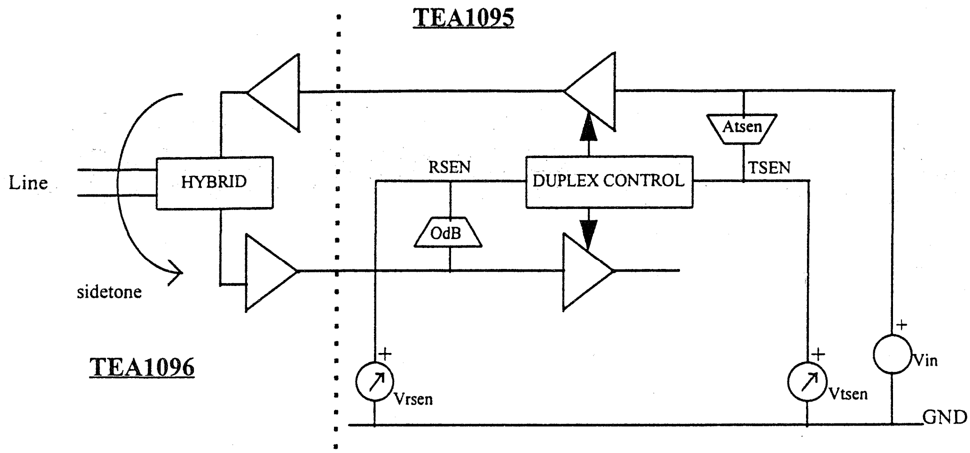


Fig. A. 2 Simplified measurement setup for determining sensitivity

Both the loudspeaker and microphone are disconnected from the set. An input signal (Vin) of 1 mVrms is applied to the microphone input. This signal is amplified to pin TSEN (Atsen), it is also transmitted to the line, returns via sidetone, is amplified by the TEA1096 to the TEA1095 pin RXIN and buffered to pin RSEN. The ratio of the signal on pin TSEN (Vtsen) and RSEN (Vrsen) is given as :

$$20 * \log (Vtsen/Vrsen) = Atsen - Atx1095 - Atx1096 - Ast - Arx1096$$

This measurement has to be done in nominal line and environment conditions in the frequency range 300 Hz to 3400 Hz. The TEA1095 must be forced in Tx-mode by making pin MUTERX high, the receive channel is muted but Vrsen is not affected.

Out of the measured ratio of Vtsen and Vrsen it follows :

$$20 * \log (Rtsen / Rrsen) = 20 * \log (Vtsen / Vrsen) + Aloop / 2 \qquad \text{(see \&3.4.4, Ad.3)}$$

Both the ratio of Vtsen and Vrsen and Aloop are frequency dependent, so the optimal value of Rtsen would be frequency dependent. However, this can be simplified by using the lowest value of 20*log(Vtsen/Vrsen) and the highest value of Aloop to obtain the correct Rtsen.

The value of Rtsen found is a starting point for obtaining the final value which can only be reached by subjective tests with real telephone conditions.

APPENDIX B LIST OF ABBREVIATIONS AND DEFINITIONS

Aac	Electro-acoustic coupling (electrically measured)
AGC	Automatic line loss compensation of the TEA1096
Aloop	Loop gain of a handsfree telephone set
Als1096	TEA1096 loudspeaker amplifier gain
Arx1095	Gain of the receive path of TEA1095
Arx1096	Receive gain of TEA1096
Ast	Sidetone gain
Asw	Switching range
Atsen	Gain from TXIN to TSEN of 40dB
Atx1095	Gain of the transmit path of TEA1095
Atx1096	Transmit gain of the TEA1096
Cgarx	Capacitor setting receive path amplifier low-pass filter
Cgatx	Capacitor setting microphone amplifier low-pass filter
Cmf	Microphone supply filter capacitor
Cmic	Microphone low-pass filter capacitor
Crenv	Capacitor determining the receive signal envelope
Crnoi	Capacitor determining the receive noise envelope
Crsen	DC blocking capacitor of receive sensitivity setting
Crxi	Receive input capacitor
Crxo	Receive output capacitor
Cswt	Switch-over timing capacitor
Ctenv	Capacitor determining the transmit signal envelope
Ctnoi	Capacitor determining the transmit noise envelope
Ctsen	DC blocking capacitor of transmit sensitivity setting
Ctxin	Microphone amplifier input capacitor
Ctxo	Transmit output capacitor
dBmp	dBm psophometrically weighted (0dBmp=1mW)
δVswt	Voltage difference on SWT
GARX	Receive gain adjustment pin
GATX	Transmit gain adjustment pin
GND	Ground reference pin

GNDTX	Ground reference pin for microphone signals
IDT	Idle-mode timing adjustment pin
Iswt	Output current through pin SWT (from logic)
Ix-mode	Idle mode
LN	Positive line terminal of TEA1096 or TEA106x
LSI	Loudspeaker amplifier input of TEA1096
MICP,MICM	Microphone input of TEA1096
MOSFET	Meta Oxide Field Effect Transistor
MUTERX	Receive channel mute
MUTETX	Transmit channel mute
NC	not connected
PD	Power-down input pin (reduced power consumption)
PTAT	Proportional to absolute temperature
PTT	Public telephone company
QRP	Earphone amplifier output of TEA1096
RENV	Receive signal envelope timing adjustment pin
RFI	Radio frequency interference
Rgarx	Resistor setting receive path amplifier gain
Rgatx	Resistor setting transmit path amplifier gain
Ridt	Resistor setting Ix-mode timing
Rload	Loudspeaker equivalent load resistor
Rmf	Microphone supply filter resistor
Rmic	Resistor setting microphone sensitivity
RNOI	Receive noise envelope timing adjustment pin
Rrsen	Resistor setting sensitivity of the receive envelopes
RSEN	Receive signal envelope sensitivity adjustment pin
Rslpe	Resistor setting slope of the DC characteristic of TEA1096
Rstab	Resistor setting an internally used PTAT current
Rswr	Resistor setting switching range
Rtnoi	Resistor increasing microphone speech/noise threshold
Rtsen	Resistor setting sensitivity of the transmit envelopes
Rvol	Volume control potentiometer
RXIN	Receive path input
RXOUT	Receive path output

Rx-mode	Receive mode
SLPE	DC slope pin of TEA1096 or TEA106x
STAB	Reference current pin
SWR	Switching range adjustment pin
SWT	Switch-over timing adjustment pin
TENV	Transmit signal envelope timing adjustment pin
Tidt	Idle mode timing
TNOI	Transmit noise envelope timing adjustment pin
TSEN	Transmit signal envelope sensitivity adjustment pin
TXIN	Microphone amplifier input
TXOUT	Transmit path output
Tx-mode	Transmit mode
V_{BB}	Positive supply input of TEA1095
Vdt, Vdialtone	Dial tone detector level
VEE	Reference pin on TEA1096
VOL	Volume adjustment pin

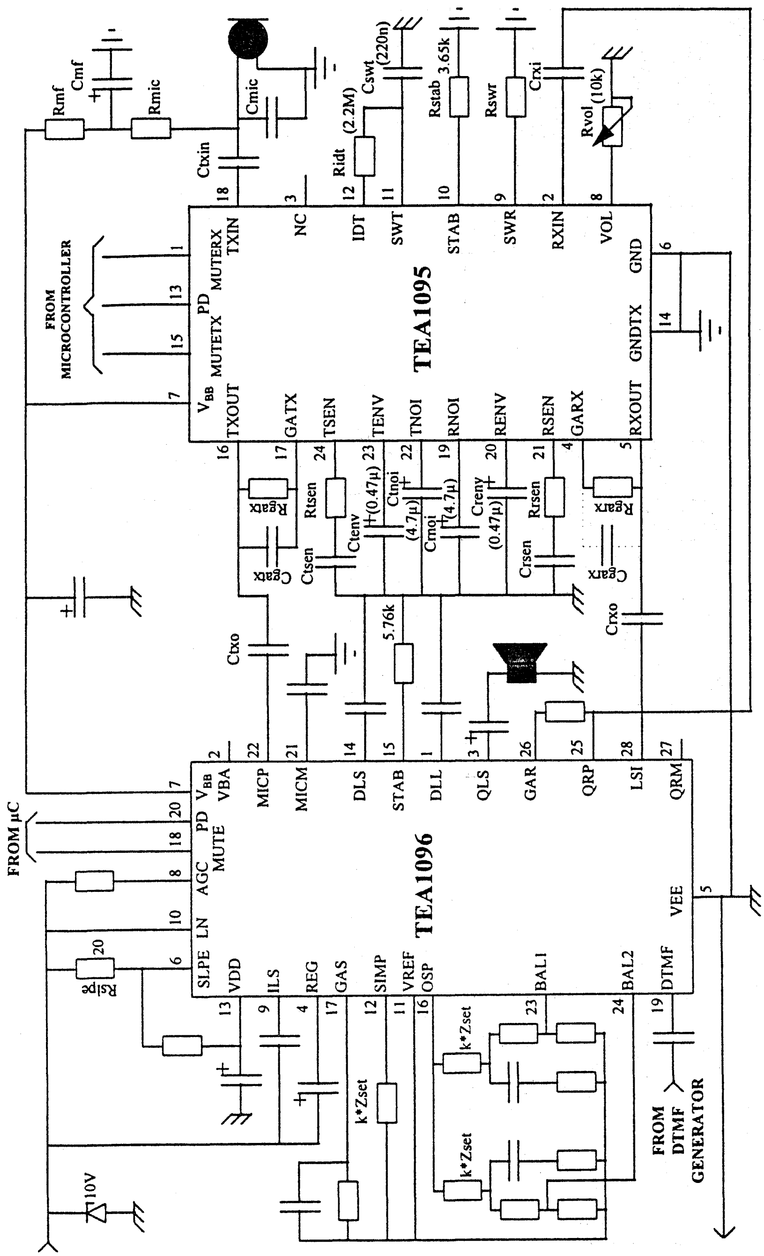


Fig. C1 Basic application of TEA1095 with TEA1096

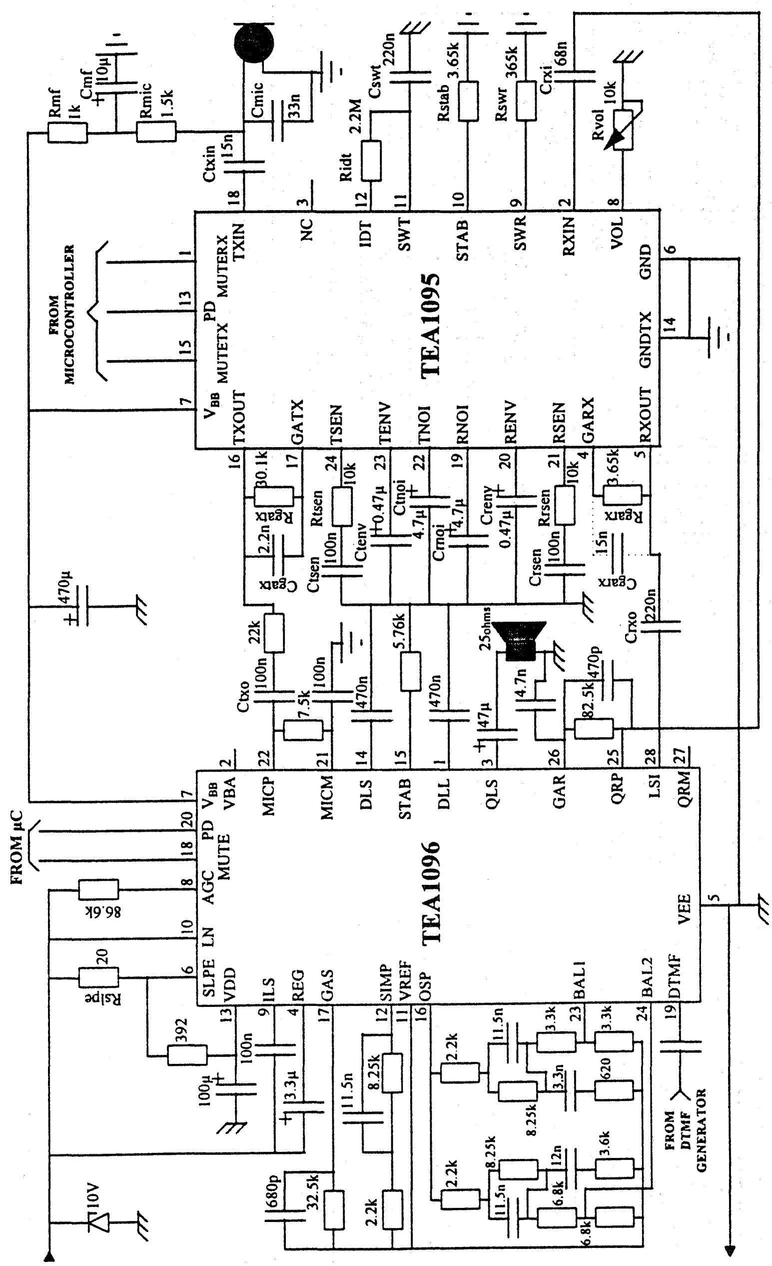


Fig. C2 Basic application of TEA1095 with TEA1096

4 DIALLERS/CONTROLLERS

CONTENTS

APPLICATION NOTE Nr	ETT8612
TITLE	Application of the PCD3310 bilingual dialler in electronic telephone sets
AUTHOR	J. C. F. van Loon
DATE	October 1986

CONTENTS

Figures and tables

1. Introduction

The PCD3310 is a dialling circuit, intended for those electronic telephone sets which have to be able to generate both pulse or/and dual tone multi frequency (DTMF) dialling procedures.

The switching from pulse to DTMF or visa versa can be done by either switching the PD/DTMF input to V_{SS} or V_{DD} by a switch or interconnection or via a special keyboard button.

Any standard matrix keyboard can be connected to the circuit for dialling in either PD or DTMF mode. Numbers of up to 23 digits can be retained in RAM for redial or note-pad facilities.
In order to handle access pause and its program problems, in both modes the Cursor method is implemented.

The adaption with the telephone line for supply and DTMF signal transfer and adjustment need a speech transmission circuit. Several combinations are possible, in this report the combination with the TEA1060 family has been chosen, due to its good performance and flexibility.
For line current interruptions a mosfet is used.

Figure 1 gives the block diagram of this application.

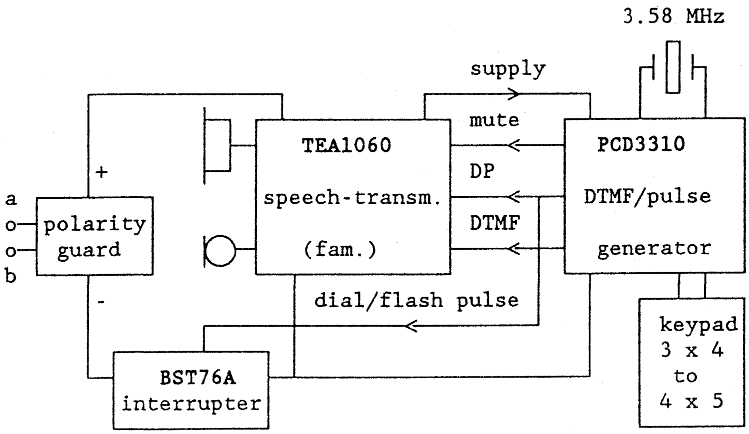


Fig.1. Block diagram of the PCD3310 and TEA1060 application

2. Description of the circuit

Detailed information about the PCD3310 dialling circuit can be found in the data sheets (see Ref.1 chapter 6).

2.1 Universal block diagram

The block diagram of the PCD3310 is shown is figure 2.

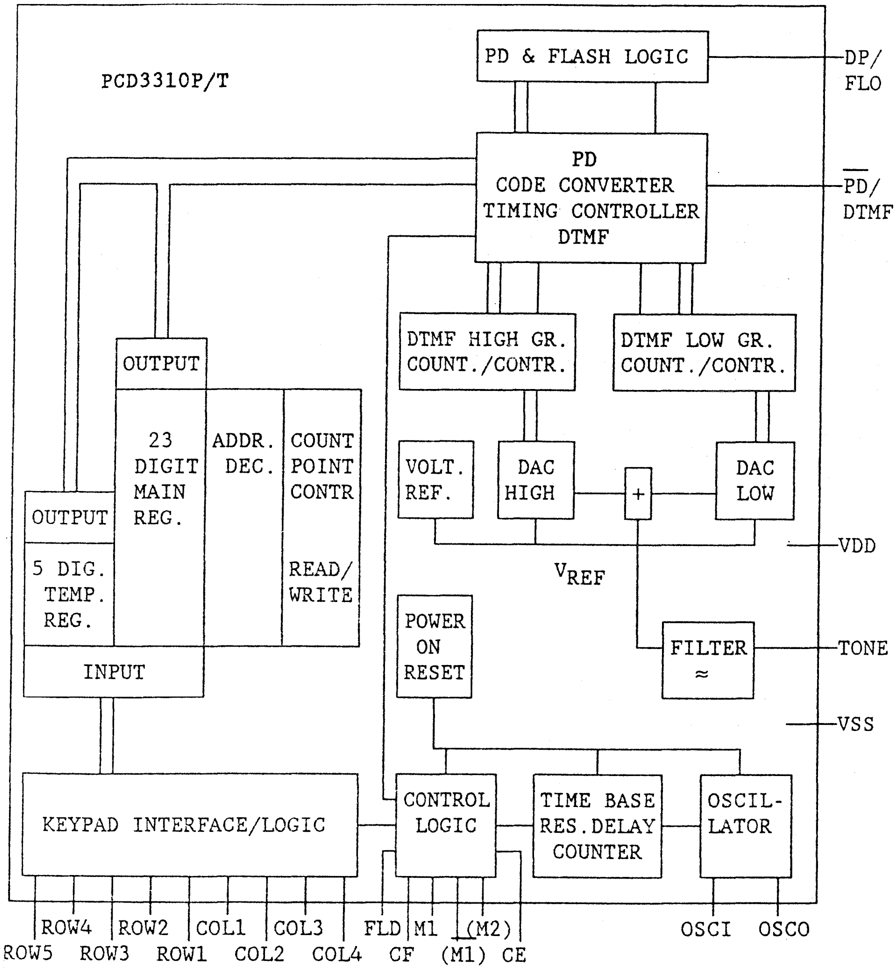


Fig.2. Block diagram of the PCD3310P/T

Features of this circuit are:

* Pulse and DTMF dial
* 23-digits capacity for redial operation (cursor method)
* Memory clear and electronic note-pad
* Mixed mode dialling; start with PD and end with DTMF dialling
* dual redial buffers for PABX and public calls
* Four extra functions keys; program, flash, redial, PD to DTMF (mixed dialling)
* DTMF timing:
  - manual dialling - minimum duration for bursts and pauses
  - redialling - calibrated timing
* On-chip voltage reference for supply and temperature independent tone output
* On-chip filtering for low output distortion (CEPT CS 203 compatible)
* On-chip oscillator uses low-cost 3,58 MHz (tv colour burst) crystal
* Uses standard single-contact or double-contact (common left open) keyboard
* Keyboard entries fully debounced
* Flash (register recall) output

2.2 Pinning of the PCD3310P and PCD3310T

This PCD3310 is available in two packages namely:

- The PCD3310P: a 20-lead dual in line shown in figure 3.
- The PCD3310T: a 28-lead mini-pack shown in figure 4.

OSCI	1		20	OSCO
PD/DTMF	2	P	19	Vdd
TONE	3	C	18	CE
Vss	4	D	17	M1
FLD	5	3	16	DP/FLO
ROW 5	6	3	15	CF
ROW 4	7	1	14	COL 4
ROW 3	8	0	13	COL 3
ROW 2	9	P	12	COL 2
ROW 1	10		11	COL 1

Fig.3. Pinning diagram of the PCD3310P

PINNING:

1 OSCI oscillator input
2 PD/DTMF select pin; pulse or DTMF dialling

3	TONE	single or dual tone frequency output
4	Vss	negative supply
5	FLD	flash duration control input/output
6	ROW 5	scanning row keyboard input/outputs
7	ROW 4	
8	ROW 3	
9	ROW 2	
10	ROW 1	
11	COL 1	sense column keyboard inputs with internal pull-ups
12	COL 2	
13	COL 3	
14	COL 4	
15	CF	330 Hz confidence tone output to provide audible feedback of key entries
16	DP/FLO	dialling pulse and flash output
17	M1	muting output
18	CE	chip enable input
19	Vdd	positive supply
20	OSCO	oscillator output

OSCI	1		28	OSCO
PD/DTMF	2		27	Vdd
TONE	3		26	CE
n.c.	4	P	25	n.c.
Vss	5	C	24	M1
n.c.	6	D	23	M1
n.c.	7	3	22	M2
FLD	8	3	21	DP/FLO
ROW 5	9	1	20	CF
ROW 4	10	0	19	COL 4
n.c.	11	T	18	n.c.
ROW 3	12		17	COL 3
ROW 2	13		16	COL 2
ROW 1	14		15	COL 1

Fig.4. Pinning diagram of the PCD3310T

PINNING:

1	OSCI	oscillator input
2	PD/DTMF	select pin; pulse or DTMF dialling
3	TONE	single or dual tone frequency output
4	n.c.	not connected
5	Vss	negative supply
6	n.c.	not connected
7	n.c.	not connected
8	FLD	flash duration control input/output

9	ROW 5	scanning row keyboard input/outputs
10	ROW 4	
11	n.c.	not connected
12	ROW 3	scanning row keyboard input/outputs
13	ROW 2	
14	ROW 1	
15	COL 1	sense column keyboard inputs with internal pull-ups
16	COL 2	
17	COL 3	
18	n.c.	not connected
19	COL 4	sense column keyboard input with internal pull-up
20	CF	330 Hz confidence tone output to provide audible feedback of key entries
21	DP/FLO	dialling pulse and flash output
22	M2	strobe; active HIGH during transmission
23	M1	inverted mute output
24	M1	muting output
25	n.c.	not connected
26	CE	chip enable input
27	Vdd	positive supply
28	OSCO	oscillator output

2.3 Mask options

In previously developed dialling circuit such as the PCD3320 family of pulse dialers the dialling characteristics can be changed by some pin options.
With the PCD3310 this is not possible.
However there is the possibility to make special mask versions.
The values which can be changed are given below.

- reset delay time(160 ms)
- flash time (100 ms + t_{FLRC})
- flash in both modes or DTMF mode only
- mute hold-over time (80ms)
- mute outputs (M1 or M1 or M2)
- DTMF tone burst time (70/70 ms)
- mark/space ratio (66/33ms)
- output configuration (open drain N or P or push-pull)
- reset state outputs

Requests for a special versions have to be addressed to our local sales office.

2.4 Keyboard inputs/outputs

The sense column inputs COL 1 to COL 4 and the scanning row outputs ROW 1 to ROW 5 of the PCD3310 are directly connected to the keyboard as shown in figure 5.

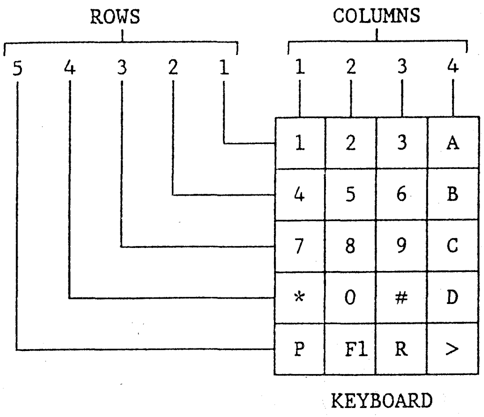


Fig.5. Keyboard organization

Row 5 of the keyboard has the following special function keys:

- P memory clear and programming (note-pad)
- F1 flash or register recall
- R redial
- > change of dial mode from PD to DTMF in mixed dialling mode

2.4.1 Other keyboards

The PCD3310 can work on several other kind of keyboards, such as a 3 by 4 and 3 by 5 matrix keyboard, depending on which features the dialling circuit has to fulfil.

A 3 by 5 matrix keyboard.
Here the buttons A,B,C,D and > are not available. Switch over from pulse dialling to DTMF dialling can be done by pressing the * or # key or by activating the PD/DTMF input.
Figure 6 gives this keyboard organization.

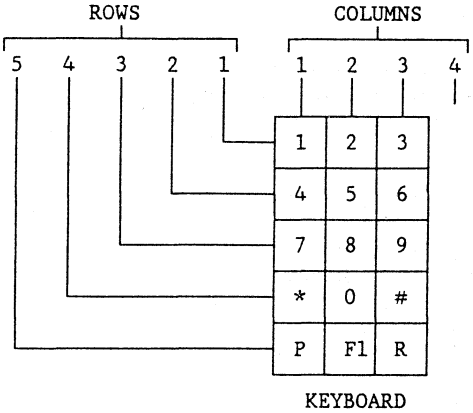


Fig.6. The 3 x 5 Keyboard organization diagram

A 3 by 4 matrix keyboard.
When using this keyboard configuration only two of the four function keys can be pressed, while the DTMF keys A, B, C, D, * and # have to be removed.
Figure 7 gives this keyboard organization the function keys used here are; R (redial) and P (memory clear and note-pad).

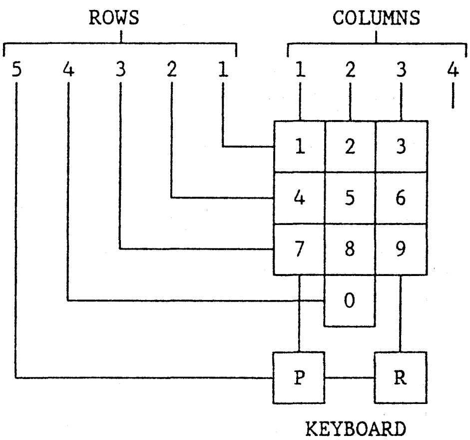


Fig.7. The 3 x 4 Keyboard organization diagram

2.5 Keyboard scanning principle

In principle the keyboard debounce procedure consists out of 2 parts, first the debounce detection time and second the real key detection. The debounce procedure for release keys is just the other way around.

First it tests if there will be a stable detection for the whole debounce time (about 12 ms).
The second test is to detect which key has been pressed. This last test is done with an active LOW and active HIGH detection respectively.

This debounce method will result in a relatively high immunity against interferences (disturbances), results will be given in the EMC part of this report.

A small disadvantage of this scanning method is that it is more complex to drive this circuit by the outputs of a micro-computer.

2.6 Flash time extension

Flash (or register recall) is activated by the FL key and can be used in DTMF and pulse dialling mode. Pressing the FL push button will produce a timed line-break of 100 ms at the DP/FLO output. During the conversation mode this flash pulse entry will act as a chip enable. This flash pulse duration (t_{FL})is calibrated and can be prolonged with an external resistor and capacitor connected to the FLD input output (see Fig.8).

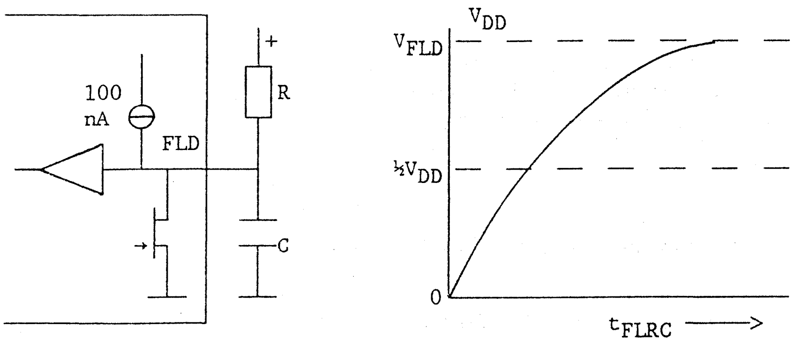


Fig.8. Flash pulse duration setting.

The total FLASH time = $t_{FL} + t_{FLRC}$
in which $t_{FLRC} \approx R * C$
Note I_{FLD} = typ. 100 nA

Pin FLD can also be used to control the line interrupt time (FLASH duration time) by holding FLD LOW.

Figure 9 gives the total FLASH time as a function of the resister value for three capacitor values.

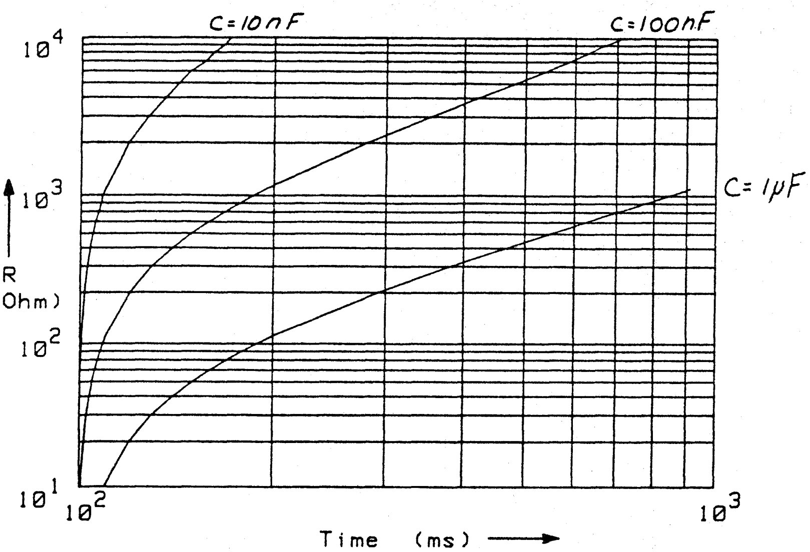


Fig.9. The FLASH time extension.

The flash pulse resets the read address counter (RAC). Later redial is possible (equal to a CE activation). The counter of the reset delay time is held during the period of t_{FL}.

2.7 Chip enable

The CE input enables the circuit and is used to initialize the IC.

CE=LOW provides the static standby condition. In this state the clock oscillator is disabled, all registers and logic are reset with the exceptions of those necessary for redial.

CE=HIGH activates the clock oscillator and the circuit changes from static standby condition to the conversation mode.

If the CE input is taken to a LOW level for more than the reset delay time (typ. 160 ms) an internal reset pulse will be generated and the IC will change to the static standby mode. This prevent a reset during pulse dialling or flash when also a line current break occurs.

Often the CE-pin is controlled by another positive voltage (not V_{DD}). In these cases it has to be noted that the voltage on any pin, thus also the CE pin, never may rise above the supply voltage + 0.7 V because then the input protection diodes of the IC come into conduction.

2.8 $\overline{PD}$/DTMF input/output

This $\overline{PD}$/DTMF pin has three operating modes.

Pulse mode

If $\overline{PD}$/DTMF = Vss the pulse dial mode is selected. Entries of non-numeric keys are neglected, they are not stored in the redial register or transmitted.

DTMF mode

If $\overline{PD}$/DTMF = Vdd the dual tone multi-frequency dialling mode is selected. Each non-function key corresponds to a combination of two tones, each one of four possible LOW and HIGH group frequencies. The harmonic contents fulfil the CEPT CS 203 recommendations.

Mixed mode

When the $\overline{PD}$/DTMF pin is open-circuit the mixed mode is selected. After activation of CE or flash the circuit starts as a pulse dialer and remains in this state until a non-numeric (A,B,C,D,*,#) or the ">" key is activated. Then the circuit changes over to DTMF dialling and remains there until flash or, after a static standby condition, CE is re-activated.
Only those digits dialled in pulse dialling position will be redialled as long as the total of dialled digits (pulse and DTMF) is lower than 23.

In mixed mode it is recommended to connect a capacitor of 1 nF between $\overline{PD}$/DTMF and ground for better EMC behaviour.

$\overline{PD}$/DTMF pin used as output

At mixed mode dialling the $\overline{PD}$/DTMF pin is a output which indicates the dialling status of the PCD3310.

After CE or flash the status is LOW (indicates pulse dialling), but as soon as a non-numeric key or the > button is activated this output becomes HIGH. This switch-over also happen if the PCD3310 is still sending dialling pulses.
After a flash the status is LOW again, but when CE becomes LOW (on-hook) the PD/DTMF output stays HIGH and will go LOW when CE becomes HIGH again (off-hook).

2.9 General output configuration

The configuration of output pins M1, $\overline{\text{M1}}$, M2, DP/FLO, CF, $\overline{\text{PD}}$/DTMF is given in figure 10.

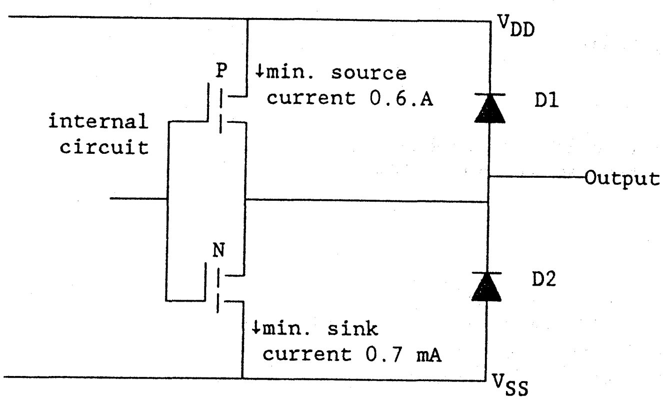


Fig.10. Output stage configuration

Diodes D1 and D2 are for protection against electro static discharges, but in spite of these precautions they can be damaged by accidental over-voltages. Therefore see first chapter "HANDLING MOS DEVICES" in the Philips Data books.

3. Dialling procedures

At dialling, redial, ect. this IC uses two registers:

* temporary register
* main register

and three counters:

* write address counter (WAC)
* temporary write address counter (TWAC)
* read address counter (RAC).

Figure 11 gives the complete memory map.

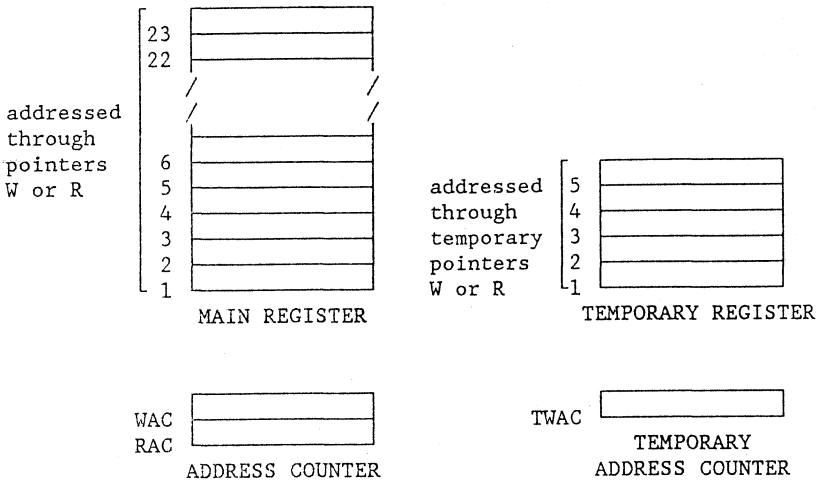


Fig.11. Program memory map.

After CE has risen to Vdd the oscillator starts running and the RAC is set to the first address. By entering the first valid digit, the TWAC will be set to the first address, the decoded digit will be stored in the register and the TWAC incremented to the next address. Any subsequent keyboard entry will be decoded and stored in the redial register after validation. The first 5 entries have no effect on the main register and its associated WAC. After the sixth valid digit is entered TWAC indicates an overflow condition. The data from the temporary register will be copied into the 5 least significant places of the main register and TWAC into the WAC. All following digits (including the sixth digit) will be stored in the main register (maximum of 23). If more than 23 digits are entered redial will be inhibited. If not more than 5 digits are entered only the temporary register and the associated TWAC are effected.

* In DTMF mode all non-function keys are valid
* In PD mode only numeric keys are valid

Simultaneously with their acceptance and in correspondence with the selected mode (PD, DTMF or mixed), the entries are transmitted as PD pulse-train or as DTMF frequencies in accordance with postal requirements. Non-numeric entries are neglected during pulse dialling, they are neither stored nor transmitted.

3.1 Redial

If "R" is the first keyboard entry the circuit starts redialling the contents of the temporary register. If the overflow flag of the TWAC was set in the previous dialling, the redialling continues in the main register. If the flag was not set, the number residing in the temporary register will only be redialled until the temporary read and write registers are equal.

Before pressing "R" a dialling sequence with up to 4 digits is possible. If the digits are equal to the corresponding ones in the main register, then redial starts in the main register until the last digit stored is transmitted.
This makes it possible to operate on a PABX exchange. First the access digits can be dialled manually (e.g. 0) wait for the access tone after which redialling can started by pressing the "R" button.

During redial keyboard entries (function or non-function) are not accepted.

No redial activity takes place if one of the following events occur:

* power-on reset
* memory clear ("P" without successive data entry)
* memory overflow (more than 23 valid data entries)

3.2 Dialling procedures at mixed mode dialling

After activation of CE or flash the circuit starts as a pulse dialer and remains in this state until a non-numeric (A,B,C,D,*,#) or the ">" key is activated. Then the circuit changes over to DTMF dialling and remains there until flash or, after a static standby condition, CE is re-activated.
In mixed mode dialling the first part entered (the pulse dialled part of the stored number) can be redialled.

3.3 Note pad programming and memory clear

The redial register can also be used as a note-pad. In conversation mode a number with up to 23 digits can be entered and stored for redialling. By activating the program key (P) the WAC and TWAC pointers are reset. This acts like a memory clear (redial is inhibited). Afterwards, by entering and storing any digits, redialling will be possible after flash or hook on and off.

During note-pad programming the numbers entered will neither be transmitted nor is the mute active, only the confidence tone is generated.

3.4 Single tone generation

When the DTMF mode is selected, single tones may be generated for test purposes (CE = HIGH). Each row and column has one corresponding frequency. High group frequencies are generated by connecting the column to Vss. Low group frequencies are generated by forcing the row to Vdd. The single tone frequency will be transmitted during activation time, but it is neither calibrated nor stored. Table 1 shows the frequencies for the single tones.

row/column	tone output frequency Hz (1)
row 1	697.90
row 2	770.46
row 3	850.45
row 4	943.23
col 1	1206.45
col 2	1341.66
col 3	1482.21
col 4	1638.24

(1) Tone output frequency with a 3.579545 MHz crystal.

Table 1 DTMF frequencies for single tones

4. Subscriber set architectures

When using the PCD3310 in a telephone set a special interface circuit is necessary to adapt to the telephone line.
Transmission requirements are fulfilled by the TEA1060 family of transmission circuits; characteristics and application information of this family can be found in data sheets (Ref.2 and Ref.3 chapter 6) and application reports (see Ref.4, and Ref.5 chapter 6).
The line current interruptions are done by an interface circuit using a BST76A MOST.
Figure 12 at page 26 shows the schematic circuit diagram of this application using our Philips BST76A mosfet.

4.1 Interrupter circuits

During pulse dialling and at DTMF dialling during flash the line current has to be interrupted. There are two circuits namely:
- The high performance MOST interrupter circuit.
- The low D.C. line voltage drop interrupter.

4.1.1 The MOST interrupter circuit

Figure 13 shows the line current interrupter circuit using a BST76A MOST.

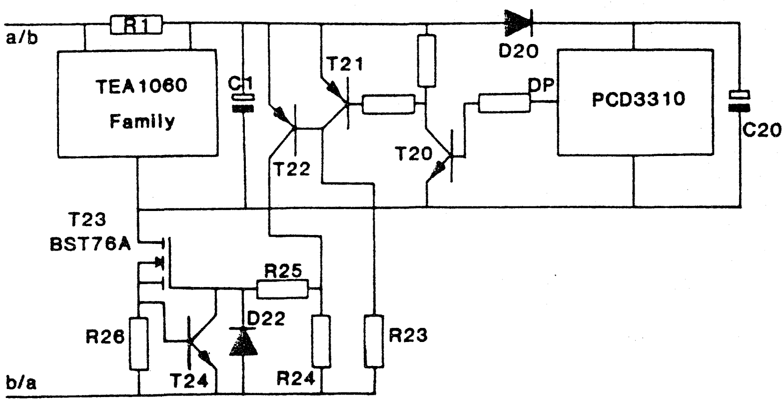


Fig.13. The line current interrupter circuit with BST76A MOST.

At the DP/FLO-output of the PCD3310 signals are available to interrupt the line current. Here DMOS-FET BST76A is the actual interrupter. The maximum current through the interrupter is limited by means of R26 and T24 to approximately Vbe/R26. Here a resistor of 3.9 Ohm is used for R26 resulting in a maximum current of about 130 mA. Zener-diode D22 and resistor R25 are required to limit the maximum gate-source voltage of T23. The level-shift function between Vss of the PCD3310 and the source voltage of T23 is performed by T22. R23 is the start-up resistor of the interrupter circuit. T20 and T21 convert the DP/FLO signals to the correct level and polarity.

4.1.2 Low voltage interrupter

Figure 14 shows the circuit diagram of an interrupter which operates at input voltages inside the polarity guard down to 1 V.

4.2 Supply

In applications where no extremely low D.C. voltage drop is necessary the PCD3310 can directly been powered by the supply point V_{CC} of the TEA1060/1061/1068 (capacitor C1) and TEA1067 with increased line voltage drop.

Capacitor C20 is used to back-up the redial-RAM in the on-hook position of the telephone set. The capacitance of C20 determines the back-up time. The back-up time is determined by the value of the capacitor and the total current at on-hook. This current is:

- standby current of PCD3310 +
- leakage current of the diode +
- leakage current of the capacitor

Because the leakage current of the capacitor gives the biggest contribution to its discharging, the quality of this capacitor determines for a big part the back-up time. In our application a capacitor of 100μF gives a back-up time of about 1 minute.

Diode D20 is used to avoid discharge of the back-up capacitor via the speech-circuit.
In some cases the minimum supply voltage of 2.5 V for the PCD3310 cannot be guaranteed under minimum line current conditions. Then it is preferred to use a Schottky type of diode for D20 (e.g. BAT85).
If this is still not sufficient then an increase of the voltage drop over the speech-circuit can be a solution (see Ref. 4 and 5 chapter 6).

However at low D.C. voltage drop and when parallel operation with other sets is necessary the TEA1060 has to be replaced by the TEA1067 which is special developed for parallel use. The line voltage of the TEA1067 has been decreased with 0.55 V with respect to the TEA1060. This results in reduced supply capabilities with respect to the TEA1060. At 2.9 volt at the supply point of the TEA1067 a minimum supply current of only 300 μA can be guaranteed. This is not enough to power the PCD3310 special not in DTMF mode.
Solutions for this problem are given below.

4.2.1 Supply with the TEA1067

The bottle neck for the supply problems of the TEA1060 family with low voltage drop is in the 620 Ω resistor connected between the pins LN and VCC of the TEA1060 (figure 12). It determines the supply capabilities of the TEA1060 family as well as the impedance of the circuit. A reduction of the resistance therefore results in improved supply capabilities but also in less good BRL-figures.

Several possibilities to improve the supply of the TEA1067 are given also those which meet the USA requirements (RS470). All of them have advantages and also disadvantages. The best choice depends on the requirements of both PTT and the telephone setmakers. These methods are given below:

(1) Increasing the line voltage. In cases where this is allowed the supply problems can be overcome simply by setting the voltage drop across the circuit to a higher value.

(2) In case only DTMF-dialling is used (without FLASH) no interrupter circuit is required and therefore no transients due to line current interruptions can occur. This makes it possible to realize protection with rugged low-voltage zener diodes (e.g. Philips BZW14 with a maximum voltage during transients of 28 V). At such low voltages the high voltage diodes required in the polarity guard normally (e.g. BAS11 can withstand 300 V) can be replaced by low-voltage Schottky barrier diodes (e.g. BAT86 can withstand 50 V) resulting in a lower voltage drop over the polarity guard. The voltage gain can be between 0.5 V and 1 V depending on the solution chosen.
It is possible now to increase the line voltage of the TEA1067 with 0.5 or 1.0 V thus increasing the supply capabilities of the TEA1067 with 0.8 mA or 1.6 mA.

(3) An alternative way to meet USA requirements (RS470) is increasing the line voltage into the conditionally acceptable region at the moments when this is allowed. The voltage is switched back into the acceptable region in those cases where this is required.

(4) The TEA1067 itself fulfills the balance return loss figures required with a large margin. Accepting a smaller margin by means of decreasing the A.C.-impedance will result in an increase of the supply capabilities.

(5) RC smoothing filter between LN and SLPE. For relatively small supply currents an RC-filter between LN (pin 1) and SLPE (pin 18) of the TEA1067 can be used to power peripherals.
Advantage of this method is that the internally generated reference voltage is used.
Disadvantage is that a higher line current is necessary for the same output swing of the transmit output stage on the line.
Futhermore a problem is that the TEA1067 and the PCD3310 do not have a common reference. Thus level shifters are necessary between the PCD3310 and the TEA1067.

(6) If the above described methods cannot be used, an inductor in parallel with R1 (the 620 Ω resistor) extends the supply possibilities.

There are two possibilities:

* use of a coil
* use of an electronic inductor:
    - Philips TEA1080 supply IC
    - (discrete) gyrator circuit

More detailed information and solutions can be found in a USA application report (see Ref.6 chapter 6) and in the application report TEA1067 (see Ref. 5 chapter 6).

4.2.2 Standby problems

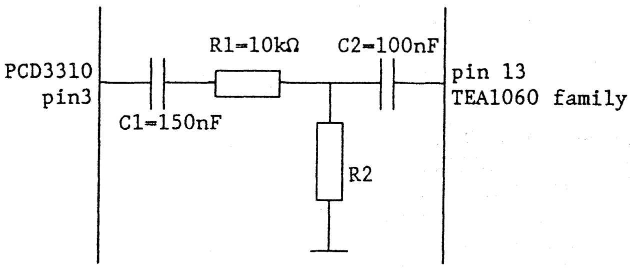
In the application diagram figure 12 two resistors R31 and R32 are in series with the DP and M1 output of the PCD3310. This is necessary because otherwise there is a possibility that at a HIGH on the given outputs the capacitor at Vcc of the TEA1060 is charged via the protection diodes on the TEA1060 inputs. Because also the CE input of the PCD3310 is connected to this Vcc point this input also stay high. This will result in a longer time before the circuit goes too the standby mode.

4.3 DTMF-interface circuit

At the TONE-output of the PCD3310 the DTMF-tones generated are available.
They are coupled to the TEA1060 via an attenuation network consisting of C21, R27, R28 and C14.
Their are considerations for the design of the coupling network between the DTMF generator PCD3310 and the transmission circuit TEA1060/1061/1067/1068.
The specifications of both devices are such that by simply adding the particular worst case values of tolerance the total variation of the DTMF level on the line amounts max. +/- 2.8 dB, while the CEPT recommendation allows +/- 2 dB maximum.
Three components determine the overall variation of the line voltage:

* the output voltage of the PCD3310
* the DTMF gain of the TEA1060 family
* the input resistance of the TEA1060 family (to a lesser extent)

It can be supposed that each of these parameters has a normal or Gaussian distribution which is clipped at a certain minimum and maximum values. A computer program was made to calculate the overall probability density function of the DTMF voltage on the line.
Taking for the spread of the parameters realistic but still conservative values and limiting the temperature range to -15 to +55 °C gives a DTMF line voltage variation within +/- 2 dB for 99% of the PCD3310 - TEA1060 family combinations. The attenuation of the coupling network is such that the deviation of the line voltage from the required DTMF level is symmetrical. The design formula is given in Fig. 16.



$$R2 = \frac{209.1}{20.7 / A1 - 30.8} \quad (k\Omega)$$

$$A1 = 10^{H/20}$$

$$H = 12.9 + Lh - A2 \qquad (dBm)$$

Lh = required level of High group component (dBm)

A2 = DTMF gain of TEA1060 family (dB)

Fig.16. Coupling network between PCD3310 and the TEA1060 family.

Typical start time of the DTMF signal on the line for 1 dB within its stabilized value is less than 3 ms, the D.C. line voltage settles within 10% of its final value in less than 7 ms typically.
Since the PCD3310 contains on-chip filters fulfilling the CEPT CS203 recommendation no additional filtering is required in this network.
More details can be found in a application report see Ref.7 chapter 6.

For an interface between the PCD3310 tone-out pin and the telephone line using discrete transistor instead of the TEA1060, see Ref.8 chapter 6.

4.4 Recommended crystal or resonator for the oscillator circuit

Quartz crystal: Philips 3.58 MHz type nr. 4322 143 04401.

Ceramic resonator: Murata*) 3.58 MHz type nr. CSA3.58MG310VA.

*) Murata manufacturing Co. Ltd, Kyoto Japan.

For more information, see Ref. 9 chapter 6.

4.5 Output M2

When the PCD3310 is used in combination with a transformer hybrid, an extra mute is necessary which will short-circuit the earpiece.
The PCD3310T (the 28 pins device) has this extra muting signal M2.

4.6 Confidence tone CF

Pin CF of the PCD3310 generates a 330 Hz signal which indicates that a button is pressed. This signal can directly drive a PXE plate with a capacitor value of 10 nF.

5. Electro Magnetic Immunity

The PCD3310 was tested on electrostatic discharges (ESD) and electro magnetic radiation (EM) immunity and compared with the PCD3315 and PCD3321 dialling circuits.

5.1. Electro static discharge (ESD)

ESD tests were executed on three test circuits, equipped with dialling circuits PCD3310, PCD3315 or PCD3321 respectively and each combined with the speech transmission circuit TEA1060.

The discharge finger of discharge probe was of type NSG 431 from Schaffner, which generates pulses with 5 ns rise time, 30 ns half amplitude time, charge capacitor of 150 pF and discharge resistor of 150 Ω. was placed successively on top of each key. After this actions good operation of the dialling circuits was tested.

It appeared that the keys with cap could withstand 21 kV, being the maximum voltage of the discharge probe.

Next discharge were made from the discharge finger to the metal contacts in the connectors of the keyboard. At 2 kV all dialling circuits refused good operation after this torture.

5.2. Electro magnetic radiation

The test circuits, each consisting of the dialling circuit PCD3310, PCD3315 or PCD3321 respectively and the speech transmission circuit TEA1060, were placed in the stripline ("jacky") together with the field-strength-meter EFS1.
The electro magnetic field-strength inside the jacky was increased until dialling or redialling was not longer possible. These field-strength values at frequencies ranging from 0.5 to 200 MHz are presented in table 2.

frequency (MHz)	PCD3310 (V/m)	PCD3315 (V/m)	PCD3321 (V/m)
0.5	> 100	> 100	> 100
1	> 100	65	60
2	> 100	75	60
4	> 100	> 100	35
6	> 100	75	14
8	27	20	6
10	50	45	20
15	45	45	20
20	50	55	30
30	65	70	80
40	45	75	19
50	> 100	35	19
75	25	27	26
100	37	40	26
150	9	7	18
200	4	13	12

Table 2 Maximum field-strength at complete frequency range

After reduction of the field-strength all circuits operated well again.

Furthermore the field-strength was increased to 300 V/m at all frequencies but after reduction of the field-strength none of the dialling circuits had lost memory information and all operated well.

6. List of references

1)-Elcoma data sheet PCD3310,
"Pulse and DTMF dialer with redial".

2)-Elcoma data sheets TEA1066, TEA1067 and TEA1068,
"Versatile telephone transmission circuit".

3)-Elcoma data handbook Integrated circuits Book IC03N
TEA1060 and TEA1061, "Versatile telephone transmission circuit".

4)-Elcoma Technical Publication 162,
"Versatile transmission ICs for electronic telephone sets",
by P.J.M. Sijbers.

5)-CAB-report ETT8606,
"Application of the low voltage versatile transmission circuit TEA1067",
by P.J.M. Sijbers

6)-CAB-report ETT8602,
"Supply of peripheral circuits with the TEA1067 speech circuit",
by J. v. Tiggelen.

7)-CAB-report ETT8503,
"The coupling network between the DTMF generator PCD3311/12 and the transmission circuit TEA1060/61,
by J. Mulder.

8)-CAB-report ETT8504,
"Line interface circuit for PCD3311/PCD3312",
by J. Geboers.

9)-CAB-report ETT8614,
"Specification of quartz crystals and PXE resonators for CMOS telephony ICs",
by J. Mulder.

Acknowledgement

Many thanks to Mr. Dorn, Mr. van der Hof, Mr. Mulder and Mr. van Tiggelen of the application laboratory CAB in Eindhoven for their contribution, in the form of circuit solutions and comments on this report.

J.C.F. van Loon.

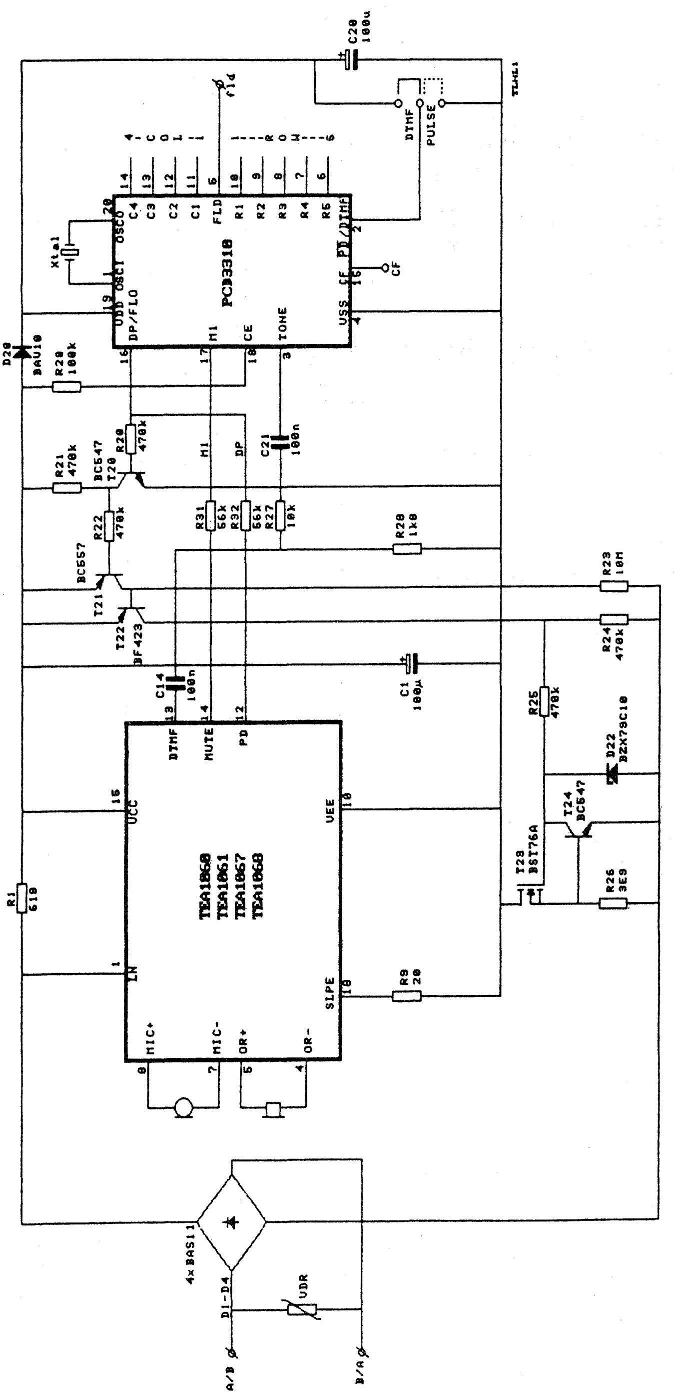


FIG.12 SCHEMATIC CIRCUIT DIAGRAM OF THE PCD3310/TEA1060-FAMILY APPLICATION

Appendix 1

PCD4410 :

On special request Philips developed an alternative dual dialling IC the PCD4410.

The PCD4410 has the following modifications with respect to the PCD3310:

- 3 x 4 keypad and flash, thus no A,B,C and D and no special row for function keys.
- There are three operating modes; pulse dialling, DTMF dialling and data transmission (DTMF). During data transmission mode a special register is used , so that the contents of the redial registers is not changed.
- The PCD4410 has three function keys:
  - flash: - 100 ms line break
  - #-key: - Redial; if first key after CE or flash
    - Program access pause in RAM
    - Ending access pause when redialling
    - DTMF-tone in data transmission mode
  - *-key: - Switching from dialling mode (DP or DTMF) to data transmission mode
    - DTMF-tone in data transmission mode
- The data transmission buffer is 8 digits.
- DTMF tone transmit / tone pause is 70 and 140 ms respectively.
- Inter-digit pause is 500 ms.
- The PCD4410 is mounted in a 18 lead DIL package.

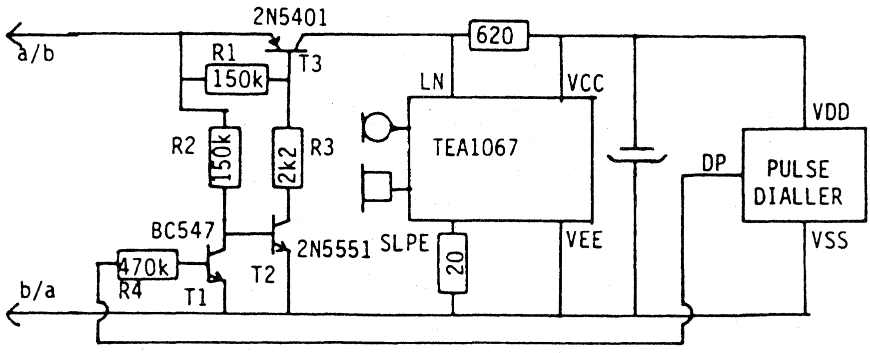


Fig.14. Circuit diagram of low voltage interrupter

The circuitry around T2 and T3 is commonly used already in telephone applications and needs no further explanation. The interface function between this interrupter and the pulse dialer is performed by transistor T1 and resistor R4. Using transistors of the type 2N5401 and 2N5551 allows operation up to 150 V. In case higher voltages can occur a voltage limiting device (e.g. a VDR) has to be used in front of the circuitry. No current limiting function is accomplished in this circuit.

In figure 15 the typical voltage drop over the interrupter (Vec of T3) is given as function of loop current using a 2.2 kΩ resistor for R3. Because resistor R3 is switched in parallel with the line impedance, its value has to be as high as possible. A lower resistance will lower the voltage drop at high line currents but reduces also the current which is left for the TEA1067.

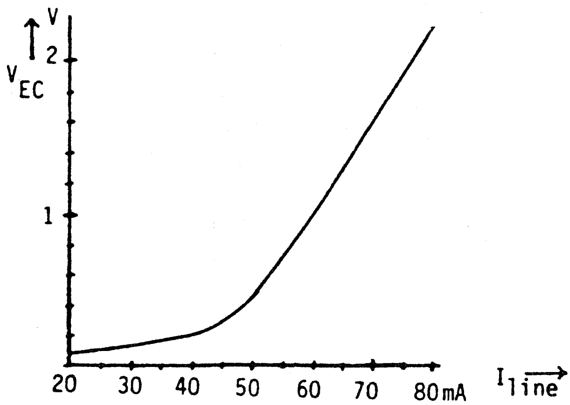


Fig.15. Voltage drop over the low voltage interrupter

5 MISCELLANEOUS

CONTENTS

APPLICATION NOTE Nr	ETT8710
TITLE	Specification of quartz and ceramic resonators for the PCD33XX and PCF84CXX integrated circuits
AUTHOR	J. Mulder
DATE	August 1987

Contents.

Contents

1. Introduction.

The oscillator of the PCD33XX and PCF84CXX was originally designed to operate with a quartz resonator. The excellent frequency accuracy and frequency stabilizing properties of quartz are well known. However, quartz resonators are rather expensive and their properties are not always needed. A less expensive device is a ceramic (or PXE) resonator. Its frequency accuracy and stabilisation is acceptable for a lot of cases but its electrical properties are rather poor compared to quartz. Simply replacing quartz by a standard ceramic resonator can give problems. Perhaps not for the typical situation but most probably for worst case conditions.
It turned out that a there could not be given a simple answer to the question: does the circuit oscillate or not? Although not simple we will try to make things not more complicated than they already are. In Chapter 2 we will start with an introduction on oscillators and resonators, but this part can be skipped without making the rest of the report incomprehensible. In Chapter 3, starting on page R11, we will give information on the practical oscillator circuit, a method to specify resonators, some examples, formulas to calculate start-up time and information on parasitic effects in resonators.

2. Introduction to oscillators and resonators.

2.1. The Colpitts oscillator.

Fig. 1 gives the basic circuit of the Colpitts oscillator.

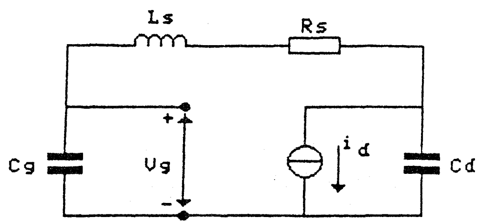


Fig.1 The Colpitts oscillator.

Its active part is a voltage controlled current source, so the current i_d depends on the voltage v_g in the following way:

$$i_d = g_m v_g$$

in which g_m is the so-called transconductance. The circuit consists further of the capacitors C_g and C_d and the inductance L_s with reactance $X_s = 2\pi f L_s$ and loss resistance R_s. The voltage controlled current source is supposed to be ideal: no frequency dependency and infinite input and output resistance. Under steady-state conditions the loop gain of an oscillator must be equal to 1, the loop phase must be 0. From the latter follows the requirement that the inductance must tune out the capacitance of C_g and C_d in series:

$$X_s = \frac{1}{2\pi f_{osc} C_t}$$

with: f_{osc}: the oscillation frequency

$$C_t = \frac{C_g C_d}{C_g + C_d}$$

The loss resistance R_s determines with g_m, C_g and C_d the loop gain. For a loop gain >= 1:

$$R_s <= \frac{g_m}{(2\pi f_{osc})^2 C_g C_d}$$

Another method to describe the loop gain condition is by using the inductance quality factor:

$$Q = \frac{X_s}{R_s} >= \frac{2\pi f_{osc} L_s}{R_s}$$

$$Q >= \frac{2\pi f_{osc}(C_g + C_d)}{g_m}$$

So we see that the required Q factor is proportional to the oscillation frequency and the total circuit capacitance but inversely proportional to the transconductance.

The oscillation frequency is inversely proportional to $\sqrt{L_s C_t}$, so small variations of L_s or C_t will be found back for about 50% in the frequency. Although the initial oscillation frequency can be trimmed to its desired value by L_s, C_g or C_d influences of supply voltage, temperature and ageing will make it difficult to have a frequency accuracy of better than 1%. The problems of frequency trimming and variation can be solved by replacing the inductance by a quartz or a ceramic resonator. Then the oscillator changes its name: Colpitts becomes Pierce.

2.2 The Pierce oscillator and the differences between quartz and ceramic resonators.

Replacing the inductance in a Colpitts oscillator by a quartz or a ceramic resonator (see Fig.2) does not change the oscillation conditions, at least not for practical values of transconductance. See Note 1 on page R23.

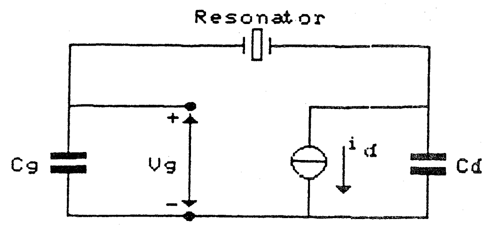


Fig.2 The Pierce oscillator

The oscillator will only work properly when the resonator simulates the required inductance L_s and quality factor Q. How does the electrical equivalent diagram of these resonators look like? You can see that in Fig.3. We use the normally accepted diagram but added an extra loss resistance R_0 to it. Especially for ceramic resonators this resistance can not be neglected. Although the same diagram holds for both quartz and ceramic resonators the values of the circuit components are quite different as can be seen in Table 1 on next page. The values are those of typical 3.58 MHz devices, on which examples in this report are based.

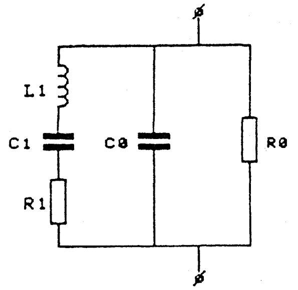


Fig.3 Resonator equivalent circuit.

	Quartz	Ceramic
L_1=	125 mH	440 μH
C_1=	15.9 fF	5 pF
R_1=	20 Ohm	8 Ohm
C_0=	5 pF	40 pF
R_0=	10 MOhm	500 kOhm

Table 1

The magnitude and phase of the impedance versus frequency are given in Figs.4 and 5. To illustrate the large differences between quartz and ceramic we plotted them on the same frequency scale. Although differences are obvious we see for both a frequency where the impedance has a minimum value, the resonance frequency f_r, and one for which it has a maximum, the anti-resonance frequency f_a.

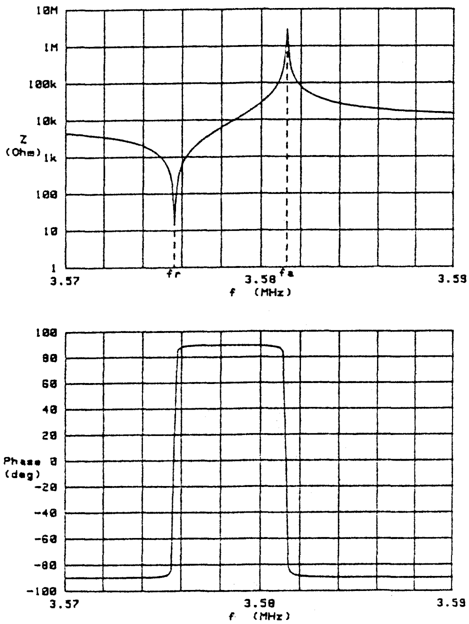


Fig.4 Impedance magnitude and phase versus frequency of 3.58 MHz quartz resonator.

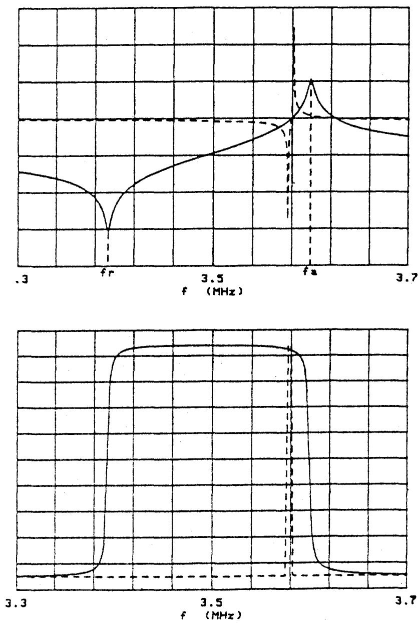


Fig.5 Ditto for ceramic (solid) and quartz (dashed) resonators.

In the corresponding phase plot we see that the impedance has the following character:

$$f < f_r\text{: capacitive}$$

$$f = f_r\text{: resistive}$$

$$f_r < f < f_a\text{: inductive}$$

$$f = f_a\text{: resistive}$$

$$f > f_a\text{: capacitive}$$

The series resonance frequency f_r follows directly from the equivalent circuit diagram:

$$f_r = \frac{1}{2\pi\sqrt{L_1 C_1}}$$

Anti-resonance occurs when L_1 resonates with C_1 and C_0 in series, so:

$$f_a = \frac{1}{2\pi\sqrt{L_1 C_s}}$$

with: $$C_s = \frac{C_0\, C_1}{C_0 + C_1}$$

For oscillation the impedance must be inductive so for the oscillation frequency f_{osc} it always holds that:

$$f_r < f_{osc} < f_a$$

But what will be the exact oscillation frequency for our ideal model? In the Colpitts oscillator we saw that at oscillation the reactance of the inductance tunes out the reactance of C_g and C_d in series. For the Pierce oscillator the situation is identical. The reactance of L_1 tunes out all other capacitive reactances provided the influence of R_0 can be neglected. So, starting at the left-hand side of L_1 (see Fig.6) we have C_0 paralleled by the series circuit of C_g and C_d with C_1 in series closing the loop.

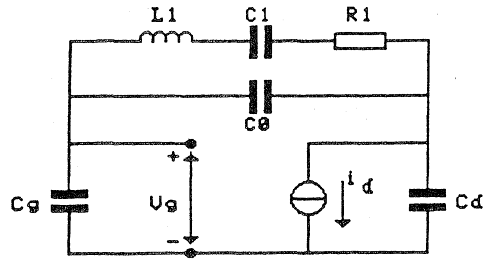


Fig 6 Pierce oscillator with resonator equivalent circuit.

The voltage controlled current source has no influence on the frequency. So:

$$f_{osc} = \frac{1}{2\pi\sqrt{L_1 C_{st}}}$$

with
$$C_{st} = \frac{C_1(C_0 + C_t)}{C_1 + C_0 + C_t}$$

and
$$C_t = \frac{C_g\, C_d}{C_g + C_d}$$

We illustrate the frequency stabilizing properties of quartz or ceramic resonators on the above given 3.58 MHz devices. Therefore we vary the external capacitances C_g and C_d by a factor of 10, from 6 to 60 pF. This would cause a frequency variation by a factor of $\sqrt{10}$ in the Colpitts oscillator. To find the oscillation frequency in the Pierce oscillator we plotted the absolute value of the capacitive load on the magnitude curves. See Figs.7 and 8 on next page. The circuit oscillates at the point where the lines intersect, provided $f_r < f < f_a$. The difference between quartz and ceramic is evident: frequency shift for quartz is 2.7 kHz or .08 %, for ceramic 73 kHz or 2.0 %. This example shows that, when using quartz, external capacitors for frequency adjustment are for most applications not needed. For ceramic resonators the load capacitance ($=C_t$) is more critical.

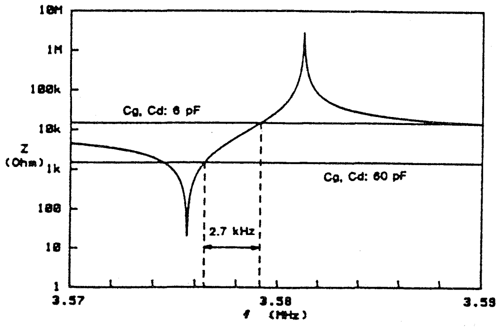


Fig.7 Frequency shift due to change of capacitive load for a quartz resonator

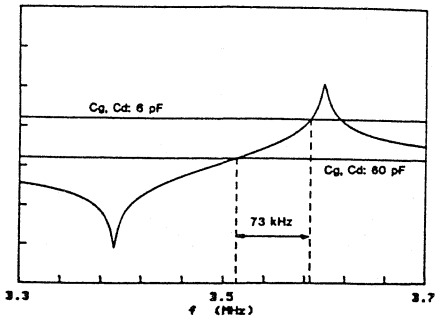


Fig.8 Ditto for ceramic resonator

Although the influence of load capacitance on the oscillation frequency for ceramic is significantly larger than for quartz the latter has still good frequency stabilizing properties. If we take $C_g = C_d = 6 \pm 1$ pF, being a more realistic value, then we will have a frequency variation of $\pm$ .1 %.

From the above we can derive that frequency accuracy and stability is now mainly determined by the resonator itself so that these properties must get full attention.
Another oscillation condition which has to be fulfilled is the quality factor Q. If we define the Q factor of the resonator as the ratio between series reactance and series resistance then it is not a simple parameter. We will illustrate this with Fig.9 on next page. The impedance of the ceramic resonator from Table 1 is plotted as a function of frequency in a complex plane. The impedance curve can be approximated by a circle which intersects the R_s axis two times: for $f = f_r$ with $R_s = R_r$ (which is not zero as it seems) and for $f = f_a$ with $R_s = R_a$.

R_r and R_a can be approximated by:

$$R_r = R_1$$

$$R_a = \frac{L_1 C_1}{R_1 C_0 (C_0 + C_1)} \text{ in parallel with } R_0.$$

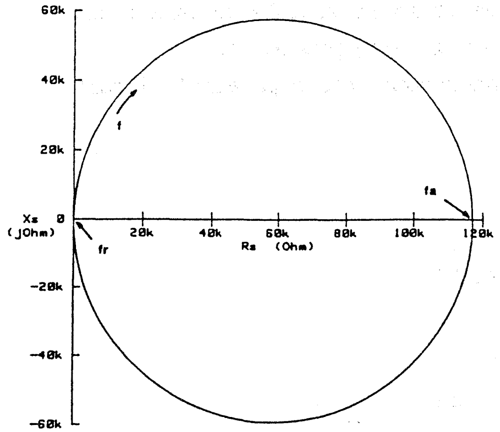


Fig.9 Real and imaginary components of ceramic resonator impedance versus frequency

From the figure we can draw some interesting conclusions. Starting with $f = f_r$ we have $R_s = R_1$ and $X_s = 0$, so a Q factor of zero. With increasing frequency we see a rapid increase of X_s and a slower one for R_s so, an increase of the Q factor. When X_s arrives at its maximum value, which is appr. $R_a/2$, the Q factor has dropped to one. Beyond this point it drops further until it reaches the value zero at $f = f_a$. The Pierce oscillator will always operate in the upper left quadrant of the impedance circle. The capacitive load reactance (C_g in series with C_d) must be sufficiently low to stay away from the maximum value of X_s above which no oscillation is possible. For this example we calculated $R_a = 117$ kOhm so maximum $X_s = R_a/2 = 58.5$ kOhm and therefore $C_g = C_d > 1.5$ pF.

2.3 From Colpitts to Pierce.

If we suppose that X_s and R_s are given how do we arrive at the circuit components of the resonator? As we know only two components and we must determine the value of five it is inevitable to take fixed values for three of them. We use the following procedure.

1) The series circuit X_s, R_s is transformed into a parallel circuit X_p and R_p.

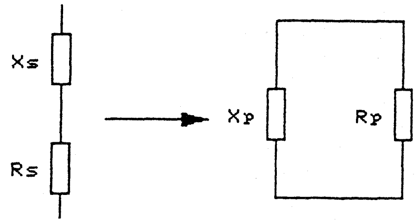


Fig.10 Series to parallel transformation.

The influence of the resonator parallel resistance R_0 and capacitance C_0 is subtracted from X_p and R_p. We suppose that the resonator manufacturer guarantees a minimum value of R_0 (not relevant for a quartz resonator) and a maximum value for C_0.

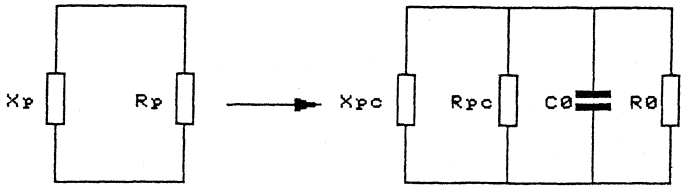


Fig.11 Correction for R_0 and C_0.

3) The corrected parallel circuit is transformed back into a series circuit X_1, R_1 with X_1 being the total reactance of L_1 and C_1.

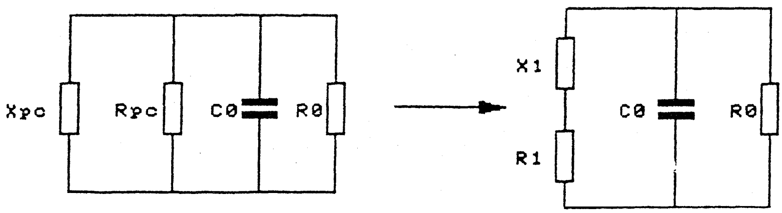


Fig.12 Parallel to series transformation.

We have now transformed the inductive reactance X_S with series resistance R_S into the equivalent circuit of a resonator. The only difference is that we do not yet have separate values for L_1 and C_1 but only for their total reactance. This means that they may be chosen arbitrarily as long as their sum equals X_1 at the oscillation frequency. In practice this freedom does not exists because L_1 and C_1 are values determined by the piezo-material.
With X_1 having the calculated value the oscillator will operate at its specified frequency. A more practical method is to give the resonators load capacitance. We can derive it from X_p by using the formula:

$$C_{load} = \frac{1}{2\pi f_{osc} |X_p|}$$

3. Specifying the resonator.

3.1 The PCD33XX and PCF84CXX oscillator.

Fig.13 shows the simplified circuit diagram of the oscillator.

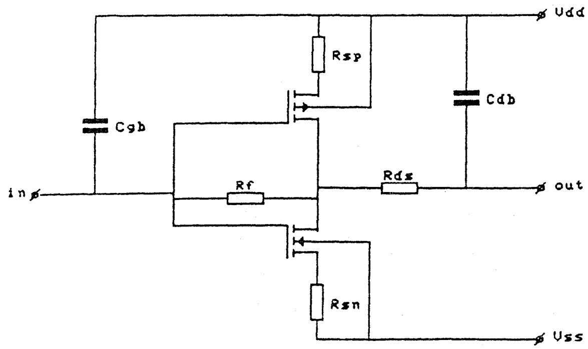


Fig.13. Simplified circuit diagram of oscillator.

It consists of an N and P-channel MOST inverter. An on-chip resistor R_f feeds the output voltage back to the input so that, at least for dc, both voltages are equal. Also drain currents are equal for both devices and this current is stabilized by source resistors R_{sp} and R_{sn}. On-chip capacitors with lead- and wiring-capacitance (1.5pF for both input and output) add up to a base level input capacitance C_{gb} of 7.0 pF and an output capacitance C_{db} of 7.8 pF. For ac both MOSTs are in parallel and act as a voltage controlled current source for small signal levels. The loop gain must be greater than 1 for quick start-up but must level off to 1 to stabilize the amplitude of the oscillator. This is automatically obtained by non-linearities of the MOSTs for larger amplitudes. As output resistance of the inverter will drop in this case, reducing the influence of output load capacitance, R_{ds} is added for isolation purposes. From measurements and computer calculations on the practical circuit we derived a simple calculation model. See Fig.14.

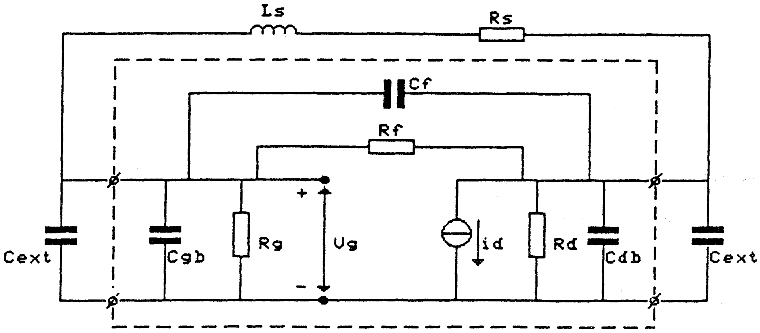


Fig.14 Calculation model of oscillator circuit.

The circuit was based on that of Fig.1 with parasitic elements added to it. Tests showed that this model can be used to calculate the external feedback network needed for oscillation. Although we used the same model we made a distinction in transconductance between the PCD33XX and PCF84CXX oscillators. For both types we did worst case calculations over the whole specified temperature and supply voltage range. As nearly all members of the PCD33XX "family" are designed for an oscillator frequency of 3.58 MHz we did calculations only at that frequency. For the PCF84CXX family we give the calculated results over the 1 to 10 MHz range.

3.2 The resonator.

From Figs. 18 to 28 at the back of this report the value of the external impedance between oscillator input and output can be derived. We give this impedance in a parallel configuration: an (absolute) reactance $|X_p|$ with a minimum quality factor Q_{min}. This facilitates the procedure for specifying the resonator.
Figs. 18 and 19 give $|X_p|$ and Q_{min} for a stand-alone PCD33XX oscillator at a frequency of 3.58 MHz. Figs. 20 and 21 give, for the same frequency, $|X_p|$ and Q_{min} if the output of the PCD33XX oscillator is loaded, via 3.9 pF, with the oscillator input of a PCD3311/12. See Note 2 on page R23. $|X_p|$ is plotted versus C_{ext}, the external capacitance. External capacitors are connected from oscillator input and output to ground and must both have the same value. For low values of C_{ext} two lines can be distinguished: the upper one for a supply voltage of 6.0 V, the lower one for 2.5 V, intermediate values can be found by linear interpolation.
For a stand-alone PCF84CXX in the frequency range 1 to 10 MHz we find $|X_p|$ in Fig.22 and Q_{min} in Figs.23 to 26 for different supply voltages. Also in Fig.22 we see for low values of C_{ext} two curves: the upper one for a supply voltage of 5.5 V, the lower one for 2.5 V.
Finally, in Figs. 27 and 28, we find $|X_p|$ and Q_{min} for the PCF84CXX oscillator at 3.58 MHz, loaded via 3.9 pF with the input of a PCD3311/12.
How do we use these curves?
First determine the minimum supply voltage (V_{dd}), the load capacitance (C_{load}) needed by the resonator to operate at its correct frequency and, for the PCF84CXX, the oscillation frequency (f_{osc}). The load capacitance (if needed) is specified by the resonator manufacturer.

Next calculate $|X_p|$, the reactance of the resonator at the oscillation frequency:

$$|X_p| = \frac{1}{2\pi f_{osc} C_{load}}$$

In Figs. 18, 20, 22 or 27 we find the value of C_{ext} which is needed to obtain the required load capacitance. If we don't use external capacitors we can read from the same figures for $C_{ext} = 0$ the value of $|X_p|$.

From Figs. 19 or 21 for the PCD33XX or Figs. 23 to 26 and 28 for the PCF84CXX we obtain the value for Q_{min}. The curves in Figs. 23 to 26 are given for $V_{dd} = 2.5$, 3.5, 4.5 and 5.5 V. Round off downwards for other supply voltages.

We have now found $|X_p|$ and Q_{min} and follow the same procedure as described in Chapter 2.3 to arrive at the components of the resonator equivalent circuit as shown in Fig.3.

1) Calculate R_p and subtract the influence of R_0 and C_0. R_0 is the parallel loss resistance of the resonator, C_0 is its parallel capacitance. The resonator manufacturer must give a minimum value for R_0 and a maximum value for C_0.

$$R_p = Q_{min}\, |X_p|$$

$$|X_0| = \frac{1}{2\pi f_{osc}\, C_0}$$

$$R_{pc} = 1/(1/R_p - 1/R_0)$$

$$|X_{pc}| = 1/(1/|X_p| + 1/|X_0|)$$

2) Transform R_{pc}, $|X_{pc}|$ into a series circuit.

$$Q_{pc} = R_{pc} \,/\, |X_{pc}|$$

$$R_{1max} = \frac{R_{pc}}{1 + Q_{pc}^2}$$

The resonator load capacitance, if not yet fixed in advance, follows from the parallel reactance $|X_p|$:

$$C_{load} = \frac{1}{2\pi f_{osc} |X_p|}$$

3.3 Examples.

Example 1.

We have a PCD3343 at 3.58 MHz, driving via 3.9 pF the oscillator input of a PCD3312. The minimum supply voltage is 2.5 V and there are no external capacitors. The minimum parallel loss resistance of the (ceramic) resonator specified by the manufacturer is R_0= 300 kOhm, maximum parallel capacitance C_0= 40 pF.

From the curves in Fig. 20 and 21 for C_{ext}= 0 pF and V_{dd}= 2.5 V we find: $|X_p|$ = 9.1 kOhm and Q_{min}= 7.6

$$R_p = 7.6 * 9100 = 69.2 \text{ kOhm}$$

C_0= 40 pF so $\quad |X_0| = \frac{1}{2 * \pi * 3.58E6 * 40E-12} = 1111$ Ohm

R_0= 300 kOhm so $\quad R_{pc} = 1/(1/69200-1/300E3) = 89.9$ kOhm

$$|X_{pc}| = 1/(1/9100+1/1111) = 990 \text{ Ohm}$$

$$Q_{pc} = 89900/990 = 90.8$$

$$R_{1max} = \frac{89900}{1 + 90.8^2} = 10.9 \text{ Ohm}$$

The load capacitance for the resonator is derived from $|X_p|$:

$$C_{load} = \frac{1}{2 * \pi * 3.58E6 * 9100} = 4.9 \text{ pF}$$

In Fig.15 on next page we plotted for this case R_{1max} versus C_0 for R_0= 300 kOhm and infinity. For C_0= 40 pF and R_0= 300 kOhm we can read from the curve the value R_{1max}= 10.9 Ohm as calculated above.

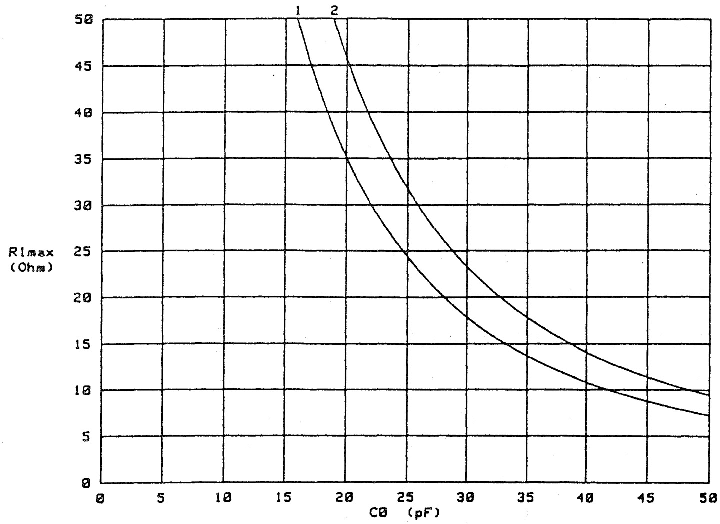


Fig.15 R_{1max} versus C_0 with R_0 as a parameter. Curve #1: R_0= 300 kOhm, #2: infinity. For conditions: See Example 1.

Example 2.

A PCF84C40 has a 6 MHz oscillator and operates at a supply voltage of 5 V minimum. Its quartz resonator requires a load capacitance of 20 pF, resonator shunt capacitance C_0= 4 pF, shunt resistance R_0 can be neglected.

We calculate first the required parallel (load) reactance:

$$|X_p| = \frac{1}{2 * \pi * 6E6 * 20E-12} = 1326 \text{ Ohm}$$

From Fig. 22 follows (intersection 6 MHz, 1326 Ohm): C_{ext}= 30 pF.

The Q factor is not given for V_{dd}= 5 V so we take the one for 4.5 V (Fig. 25). At 6 MHz and for C_{ext}= 30 pF we read Q_{min}= 25.

Now we continue as in Example 1.

$$R_p = 25 * 1326 = 33150 \text{ Ohm}$$

C_0= 4 pF so $$|X_0| = \frac{1}{2 * \pi * 6E6 * 4E-12} = 6631 \text{ Ohm}$$

R_0= negligible so $R_{pc} = R_p = 33150$ Ohm

$$|X_{pc}| = 1/(1/1326+1/6631) = 1105 \text{ Ohm}$$

$$Q_{pc} = 33150/1105 = 30$$

$$R_{1max} = \frac{33150}{1 + 30^2} = 36.8 \text{ Ohm}$$

This resistance value is rather low and it will not be easy to find a quartz resonator with such a low guaranteed value. The problem can be solved by leaving out the external capacitors C_{ext} at least as there is not a very strong requirement on frequency accuracy. When the components of the resonators equivalent circuit are known you can calculate the frequency shift by using the formulas in Chapter 2.2. In the next example we show the influence of leaving out C_{ext} on R_{1max}.

Example 3.

C_{ext} is zero now so we find from Figs. 22 and 25: $|X_p|$ = 6000 Ohm and Q_{min}= 6.6.

$$R_p = 6.6 * 6000 = 39600 \text{ Ohm}$$

C_0= 4 pF so $$|X_0| = \frac{1}{2 * \pi * 6E6 * 4E-12} = 6631 \text{ Ohm}$$

R_0= negligible so $R_{pc} = R_p = 39600$ Ohm

$$|X_{pc}| = 1/(1/6000+1/6631) = 3150 \text{ Ohm}$$

$$Q_{pc} = 39600/3150 = 12.6$$

$$R_{1max} = \frac{39600}{1 + 12.6^2} = 248 \text{ Ohm}$$

The load capacitance for the resonator is derived from $|X_p|$:

$$C_{load} = \frac{1}{2 * \pi * 6E6 * 60000} = 4.4 \text{ pF}$$

3.4 Oscillator start-up.

Start-up of an oscillator is only possible if two conditions are met:

1) there must be a start signal at the oscillation frequency.

2) if, after applying the startsignal, the level of the output signal is too small then the oscillator loop gain must be >1.

Normally the first condition is always met because of the presence of thermal noise or switch-on spikes. With a loop gain >1 the component at the oscillation frequency will be amplified and will grow exponentially to a suitable level.

The value of R_{1max}, as calculated in the preceeding chapter, guarantees a loop gain of 1 so we must take a lower value to ensure reliable start-up.

However, simply taking a certain safety factor solves the start-up problem only partially. A start-up which is reliable but takes too much time is not acceptable either.

The amplitude of a starting oscillator grows exponentially with a time constant:

$$\tau = -\frac{2\,L_1}{R_t}$$

with L_1: series inductance of resonator

$R_t = R_1 + R_{osc}$

R_1: series resistance of resonator

R_{osc}: (negative) resistance presented by the oscillator to the series circuit $L_1 - C_1 - R_1$.

The calculation for R_{1max} in Chapter 3.3 gave us in fact the value of R_{osc} for a loop gain of 1: $R_{osc} = -R_{1max}$. In this case $R_t = 0$ and τ becomes infinite and the oscillator voltage does neither increase nor decrease. For $R_t < 0$ we will find a positive value for τ and the oscillator voltage will grow exponentially. Taking for example the ceramic 3.58MHz resonator from Table 1 with $L_1 = 440\,\mu H$ and $R_1 = 8$ Ohm and the oscillator circuit from Example 1 with $R_{osc} = -R_{1max} = -10.9$ Ohm gives:

$$\tau = -\frac{2 * 440E-6}{8 - 10.9} = .30 \text{ ms}$$

A practical quartz resonator for the circuit of Example 3 has: $L_1 =$ 41 mH and $R_1 < 60$ Ohm, so with $R_{osc} = - R_{1max} = -248$ Ohm we find:

$$\tau = -\frac{2 * 41E-3}{60 - 248} = .44 \text{ ms}$$

At switch-on the oscillator voltage will have a certain start value, growing exponentially to a level suitable to drive the digital circuitry. For each period τ the voltage will increase by a factor e (= 2.71). The number of periods (N) can be approximated by the following formula:

$$N \approx \ln \left(16 f_{osc}^2 L_1 \frac{C_0 + C_{it}}{C_{it}} \left(C_0 + \frac{C_{it} \ C_{ot}}{C_{it} + C_{ot}}\right)\right)$$

with f_{osc}: oscillation frequency

L_1: resonator series inductance

C_0: resonator parallel capacitance

C_{it}: total input capacitance = C_{ext} + 7.0 pF

C_{ot}: total output capacitance = C_{ext} + 7.8 pF

The start-up time is:

$$t_{start-up} = N \tau$$

We calculate the start-up time for the ceramic and quartz resonator. For both it holds that $C_{ext} = 0$ pF.

For the 3.58 MHz ceramic resonator with $C_0 = 40$ pF and $L_1 = 440$ µH we find:

$$N \approx \ln \left(16 * 3.58E6^2 * 440E-6 * \frac{40 + 7.0}{7.0} * \left(40 + \frac{7.0 * 7.8}{7.0 + 7.8}\right) E-12\right) =$$

$$\ln (26.5) = 3.3$$

So: $t_{start-up} = 3.3 * .30E-3 = 1.0$ ms

For the 6 MHz quartz resonator with $C_0 = 4$ pF and $L_1 = 41$ mH:

$$N \approx \ln\left(16 * 6E6^2 * 41E{-}3 * \frac{4 + 7.0}{7.0} * \left(4 + \frac{7.0 * 7.8}{7.0 + 7.8}\right)E{-}12\right) =$$

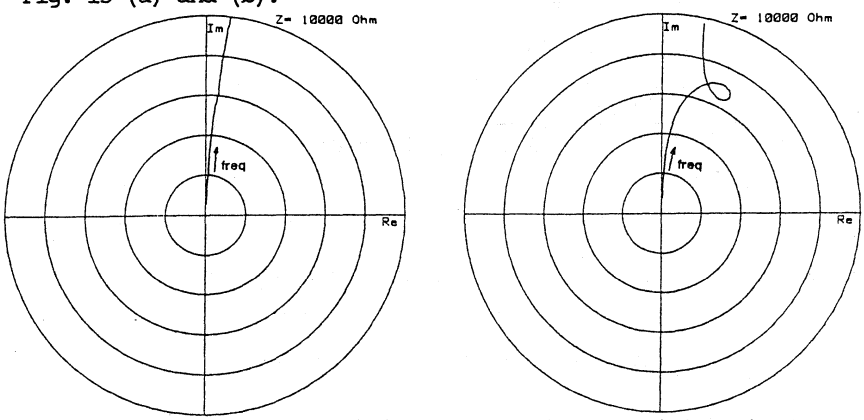
$$\ln(285) = 5.7$$

$$t_{start\text{-}up} = 5.7 * .44E{-}3 = 2.5 \text{ ms}$$

The value of the calculation method for the start-up time must not be overestimated. It gives a good indication how it is related to circuit- and resonator-parameters. Its value can best be demonstrated with two examples. A typical oscillator-quartz resonator combination had a measured start-up time of 2.8 ms and a calculated time of 2.2 ms. Replacing the quartz resonator by a ceramic one gave a measured start-up time of appr. .2 ms and a calculated time of .10 ms.

3.5 Parasitic effects in resonators.

A parasitic effect observed in ceramic resonators is illustrated by Fig. 15 (a) and (b).



(a) Normal (b) With parasitic swing mode

Fig.15 Real and imaginary components of ceramic resonator impedance versus frequency.

The real and imaginary series components of the resonator impedance are plotted as a function of frequency in a complex plane. The curve in (a) has the normal smooth character, the one in (b) has a pronounced "kink". This is probably because of a parasitic swing mode. The Q factor of the device is the ratio between imaginary and real part, Im/Re. It is quite obvious that the "kink" has a considerable influence on this factor. Preventing these effects is a task of the resonator manufacturer.

In quartz resonators the parasitic "drive level dependency" effect is well-known. Because piezo-electrical devices vibrate in a mechanical manner they are sensitive to impurities like dust for example. This makes the loss resistance R_1 dependent on the vibration amplitude or drive level. Fig. 17 gives curves for R_1 of a "normal" and of a "poor" quartz resonator. If the oscillator circuit presents

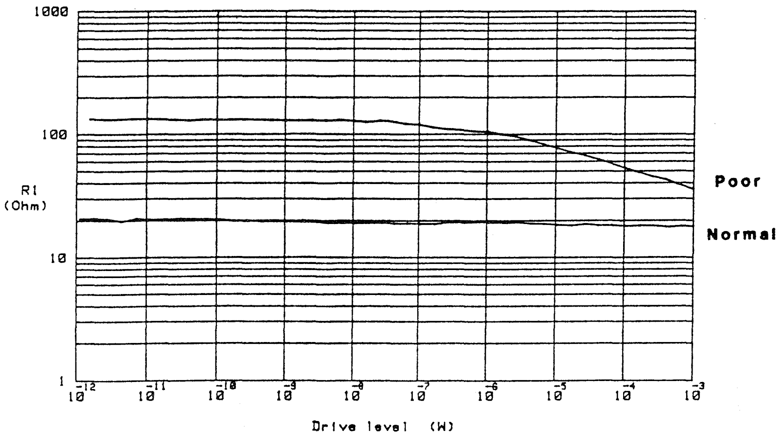


Fig.17 Series loss resistance of quartz resonator versus drive level.

a resistance of e.g. -100 Ohm to the resonator series circuit and the resonator is tested at a drive level of 1 mW then also the "poor" device seems to be good enough. However, with this device the oscillator will only start if the start drive level exceeds 1 uW because only under this condition R_1 drops below 100 Ohm. As practical start drive levels are in the pico and nano Watt range oscillation will not occur.

4. Preferred types.

f_{osc} = 3.58 MHz. (PCD33XX and PCF84CXX)

Quartz: Philips 4322 143 04401

Ceramic: Murata CSA3.58MG310VA

For the PCF84CXX family the minimum supply voltage is limited when using the ceramic resonator. For stand-alone applications the value is 2.8 V. When the output is coupled via 3.9 pF to the input of a PCD3311/12 oscillator the minimum voltage is 3.0 V. For the PCD33XX family there are no restrictions.

f_{osc} = 6 MHz. (PCF84CXX)

Quartz: Philips 4322 143 04710

5. Conclusions.

A method is given to specify resonators for the PCD33XX (3.58 MHz only) and PCF84CXX (1 to 10 MHz) oscillators over their specified supply voltage and temperature range. The value of the maximum acceptable resonator series loss resistance is found for given values of its shunt capacitance and resistance.
Examples showed that this loss resistance is strongly dependent on shunt capacitance and on external circuit capacitance. Leaving out the latter, if it is not strictly needed, is a better choice.
Start-up time calculations give an indication what might be expected but are not accurate especially for the fast starting devices.
Care must be taken that the manufacturer fulfils the resonator specification over the whole drive level and impedance range.

6. Literature.

1. Robert Meyer and David Soo, IEEE Journal of Solid-State Circuits, April 1980: "MOS Crystal Oscillator Design".
2. Mark Unkrich and Robert Meyer, IEEE Journal of Solid-State Circuits, February 1982: "Conditions for Start-Up in Crystal Oscillators".
3. Andreas Rusznyak, IEEE Transactions on Circuits and Systems, March 1987: "Start-Up Time of CMOS Oscillators.

Note 1: Rusznyak in [3] gives the complete mathematical derivation for the Pierce oscillator and shows that oscillation is also not possible above a certain (high) level of transconductance. As this level is 3 to 4 orders of magnitude higher than the minimum transconductance we are allowed to use the calculation method of this report.

Note 2: In several application reports the value of the the coupling capacitance is given as 27 pF. With the oscillator of the PCD3311/12 switched "on" this gives an acceptable load on the master oscillator. However, if the oscillator is switched "off" its input is connected to the positive supply voltage giving in a low input impedance. Under worst case conditions there is a risk that the master oscillator will not start-up or stops oscillating. Reducing the coupling capacitance to 3.9 pF prevents this effect. The resulting voltage attenuation between output of master and input of slave oscillator is allowed.

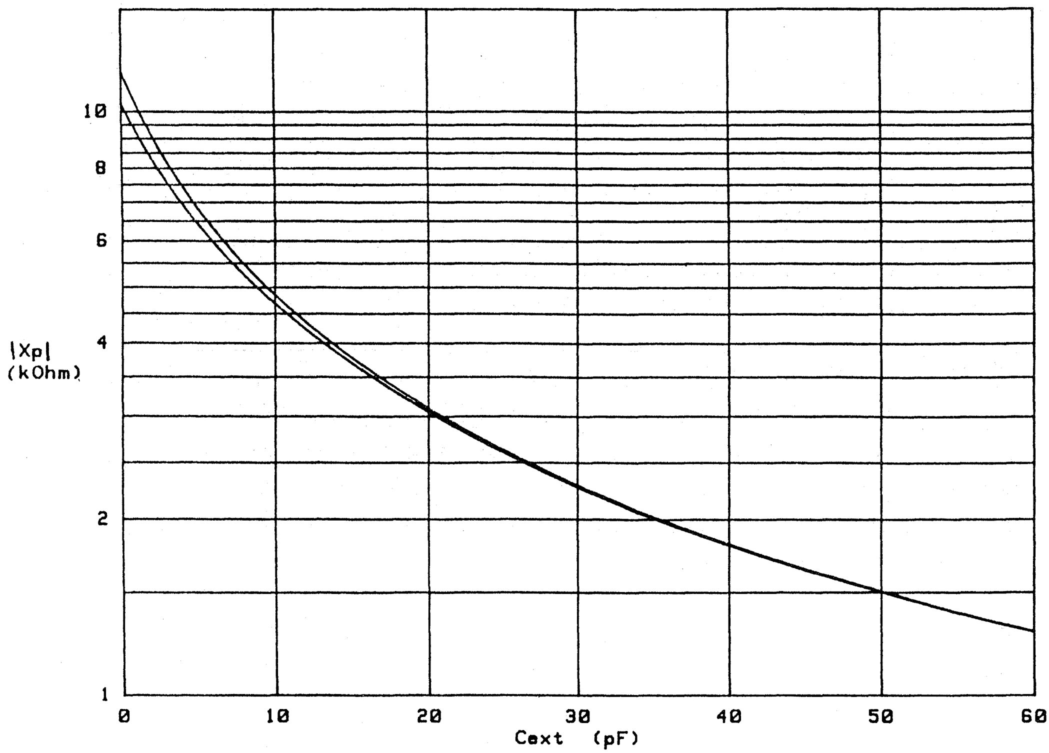


Fig.18 Stand-alone PCD33XX (f= 3.58 MHz):
Required parallel reactance versus external capacitance.
Upper curve: V_{dd}= 6.0 V, lower curve: V_{dd}= 2.5 V.

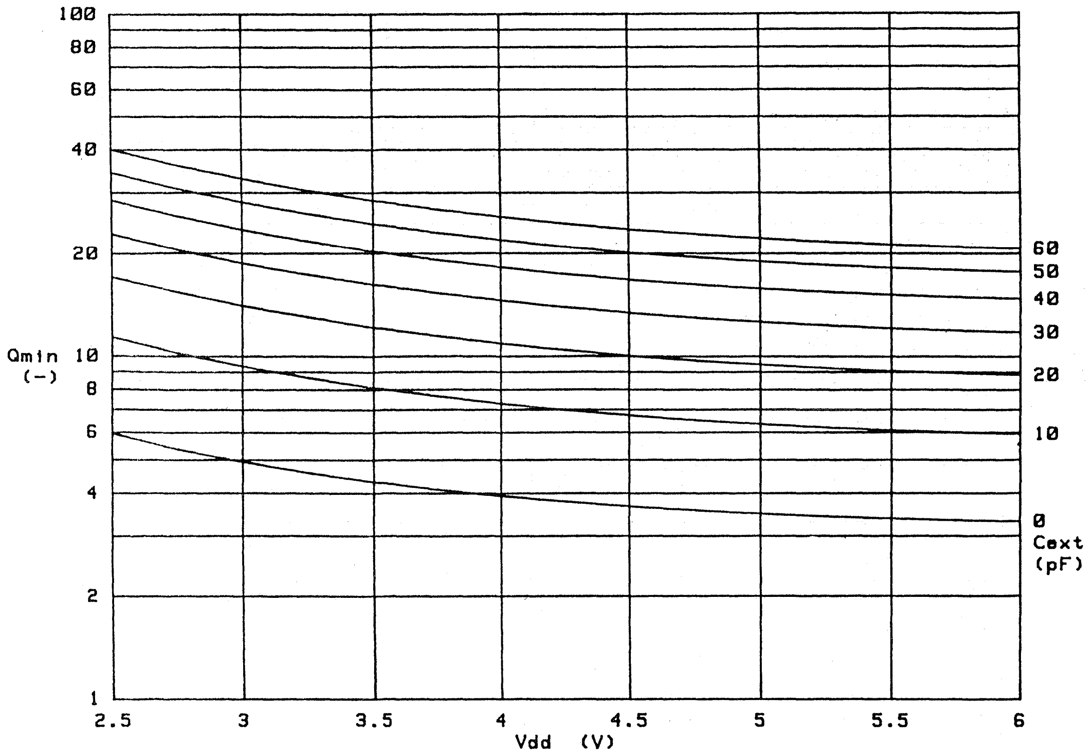


Fig.19 Stand-alone PCD33XX (f= 3.58 MHz):
Required minimum Q factor versus supply voltage with external capacitance as a parameter.

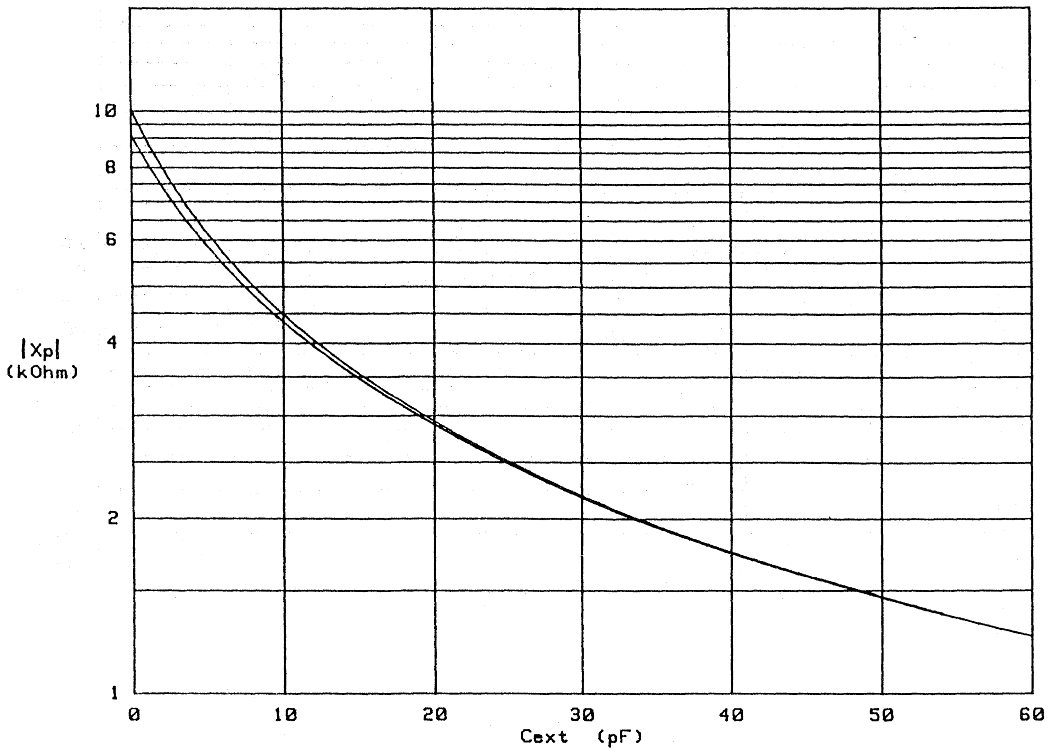


Fig.20 PCD33XX, output coupled via 3.9 pF to input of PCD3311/12, (f= 3.58 MHz):
Required parallel reactance versus external capacitance.
Upper curve: V_{dd}= 6.0 V, lower curve: V_{dd}= 2.5 V.

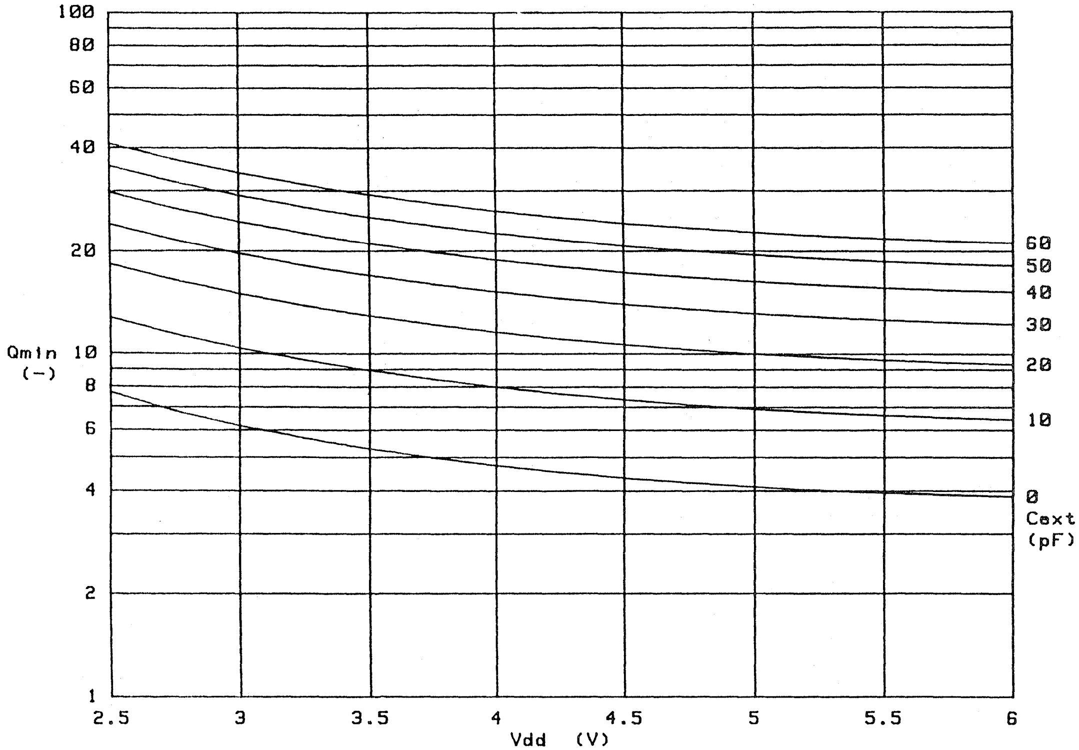


Fig.21 PCD33XX, output coupled via 3.9 pF to input of PCD3311/12,
(f= 3.58 MHz):
Required minimum Q factor versus supply voltage with external capacitance as a parameter.

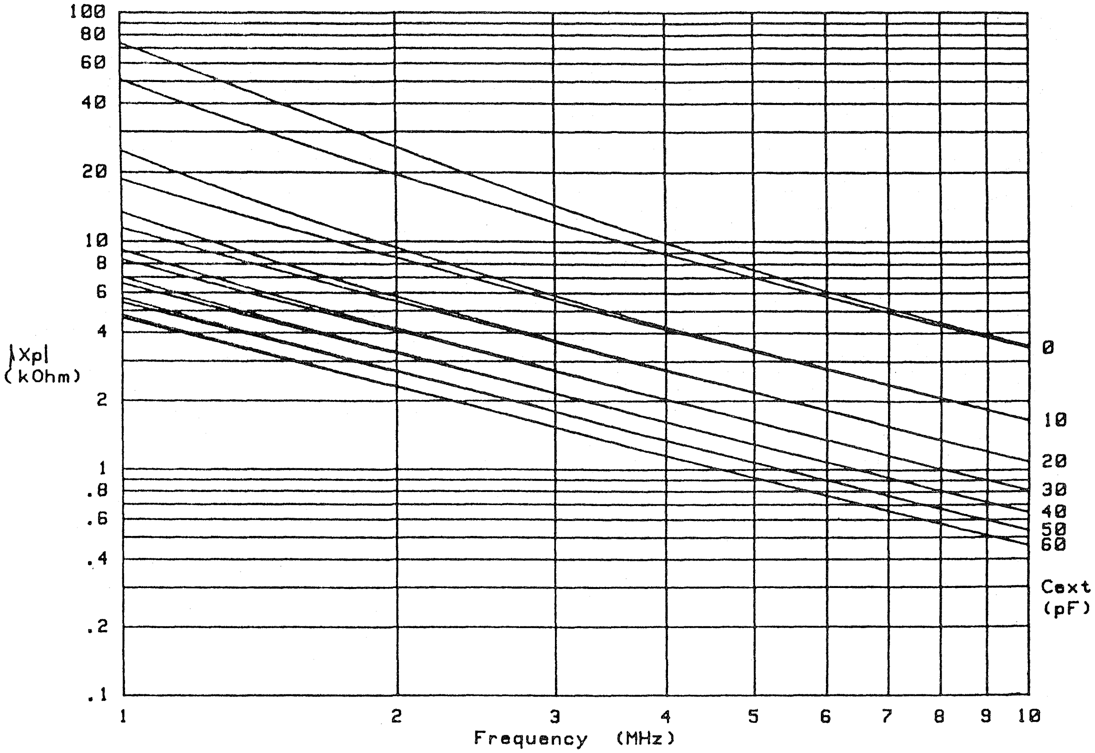


Fig.22 Stand-alone PCF84CXX:

Required parallel reactance versus frequency with external capacitance as a parameter. Upper curve: V_{dd}= 5.5 V, lower curve: V_{dd}= 2.5 V.

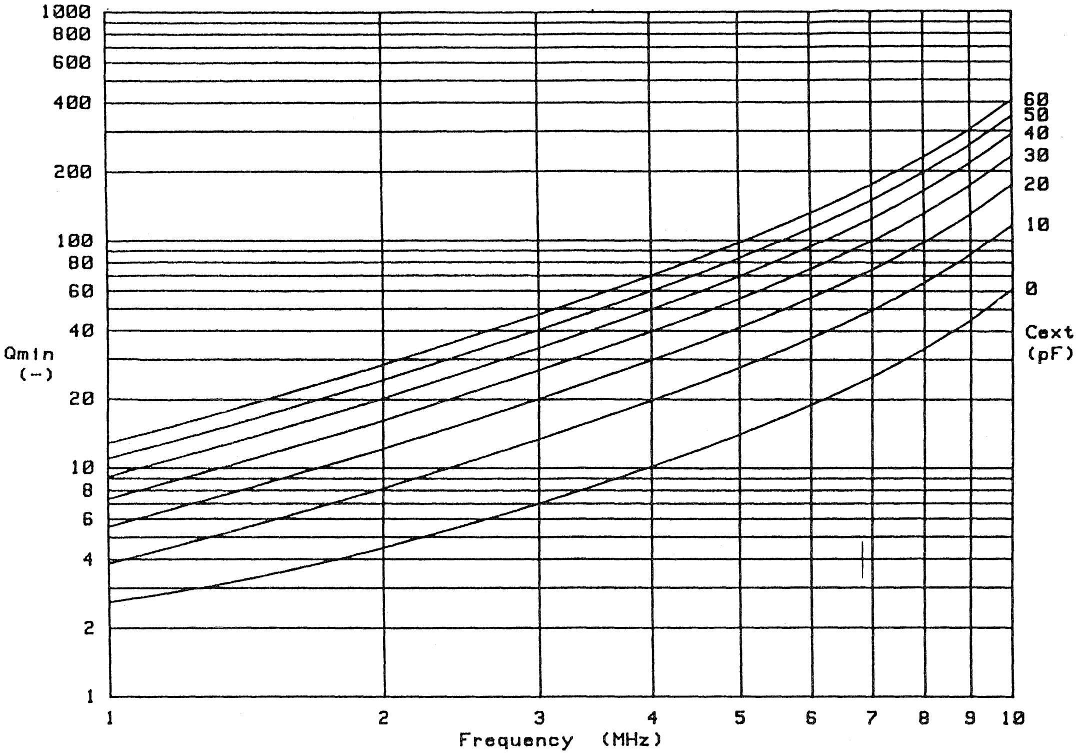


Fig.23 Stand-alone PCF84CXX (V_{dd}= 2.5 V):

Required minimum Q factor versus frequency with external capacitance as a parameter.

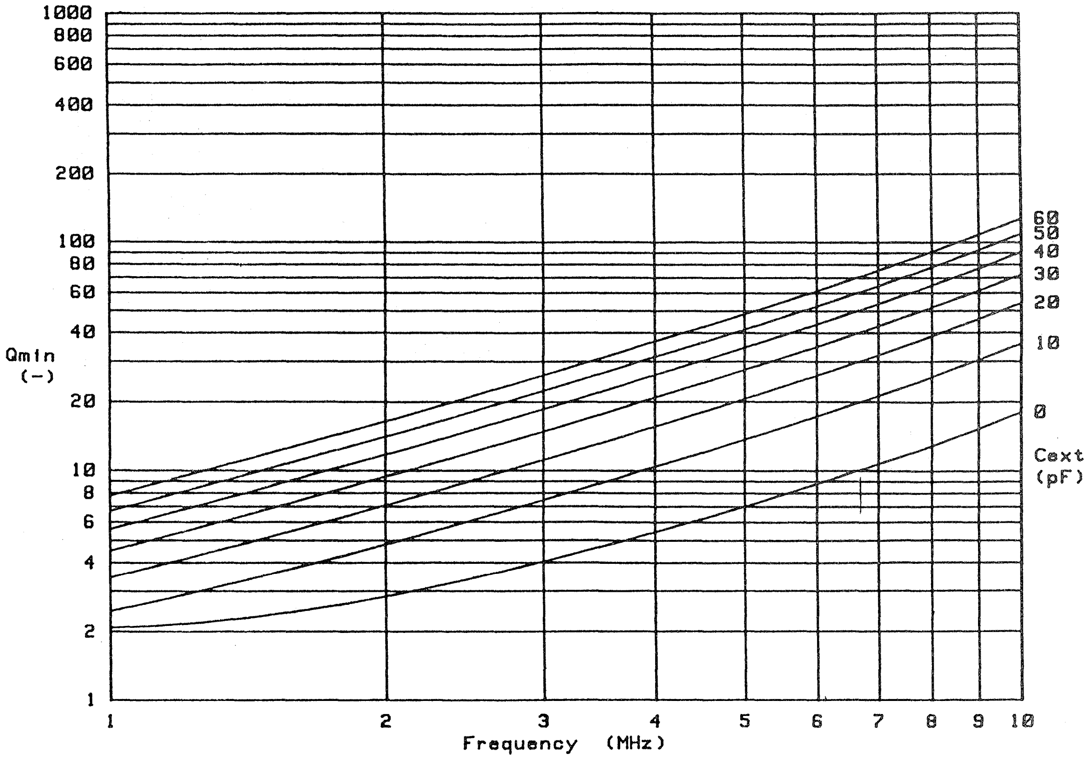


Fig.24 Stand-alone PCF84CXX (V_{dd}= 3.5 V):

Required minimum Q factor versus frequency with external capacitance as a parameter.

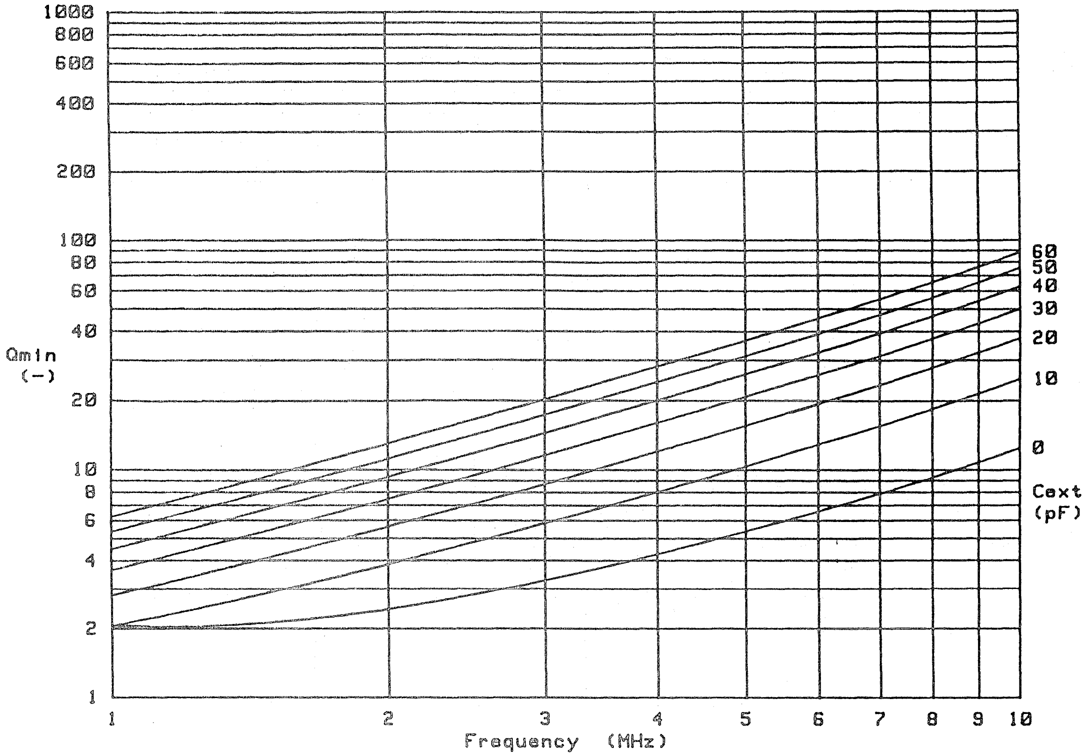


Fig.25 Stand-alone PCF84CXX (V_{dd}= 4.5 V):

Required minimum Q factor versus frequency with external capacitance as a parameter.

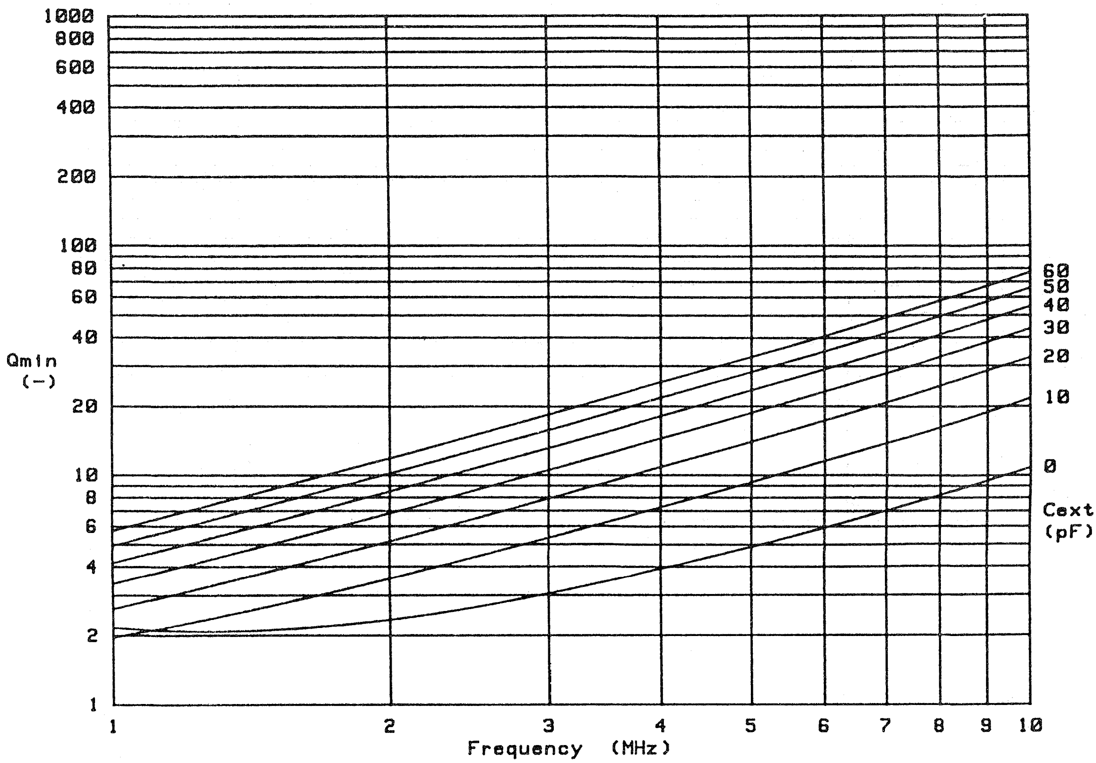


Fig.26 Stand-alone PCF84CXX (V_{dd}= 5.5 V):

Required minimum Q factor versus frequency with external capacitance as a parameter.

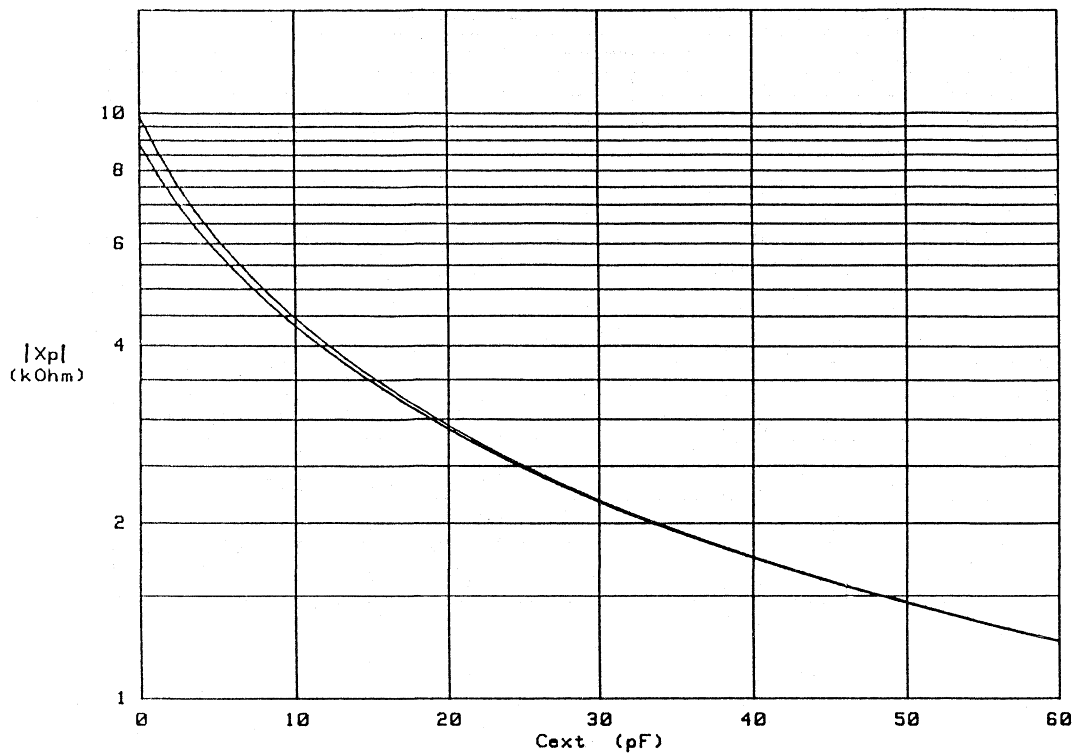


Fig.27 PCF84CXX, output coupled via 3.9 pF to input of PCD3311/12, (f= 3.58 MHz):
Required parallel reactance versus external capacitance.
Upper curve: V_{dd}= 6.0 V, lower curve: V_{dd}= 2.5 V.

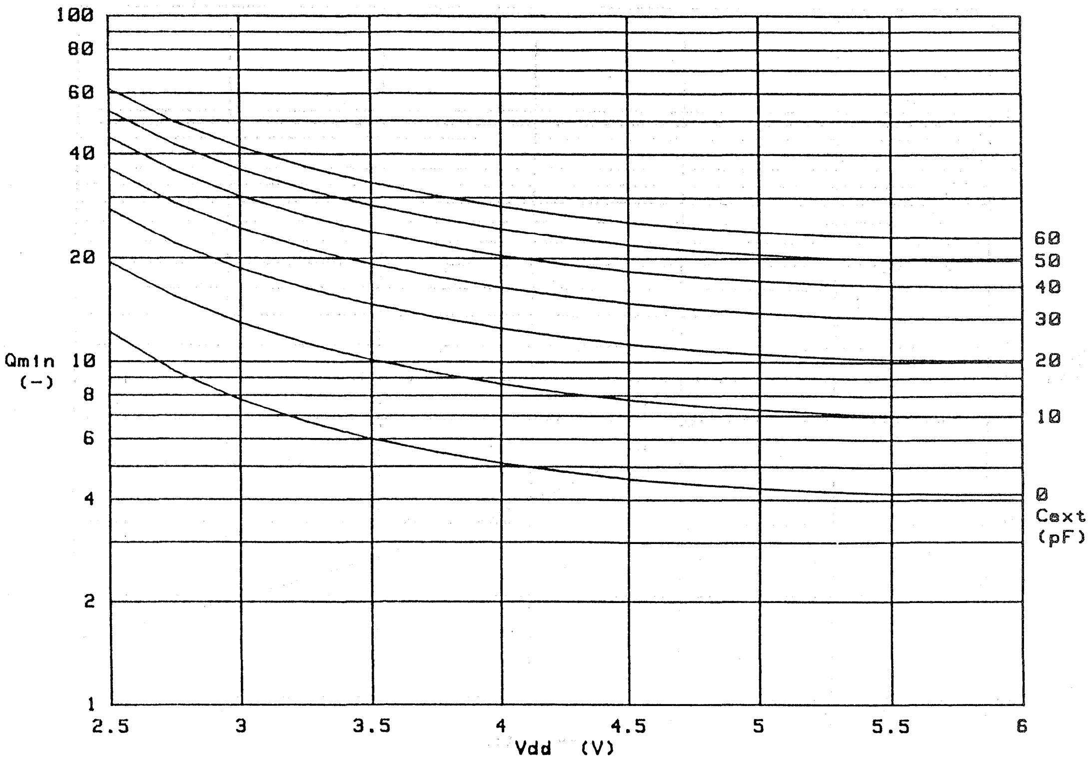


Fig.28 PCF84CXX, output coupled via 3.9 pF to input of PCD3311/12, (f= 3.58 MHz):
Required minimum Q factor versus supply voltage with external capacitance as a parameter.

APPLICATION NOTE Nr	ETT/AN8903
TITLE	Galvanic separation of the IIC-bus
AUTHOR	H. Leander
DATE	March 1989

Summary

In some TELEPHONE applications there is a need for galvanic separation of the IIC-bus, for example if IIC devices coupled to the bus are not fed from the same internal power supply.
In this note, a technical solution is given where the IIC-bus specification can be met.
As galvanic separation device, an optocoupler is used.

Table of Contents:

INTRODUCTION.

In feature phones, controlled by a microprocessor, communication with peripheral devices such as RAM, CLOCK-IC etc. is done via the IIC-bus. sometimes the devices connected to the IIC-bus are not fed from the telephone line only, for example some devices are fed from the mains supply. In the last case direct interconnection of those devices is not possible, but a galvanic separation is needed.

The IIC-bus is a two wire bidirectional bus system, thus we need 2 separation devices for each line.
To meet the IIC-bus specification the separation device must have the following characteristics:

. Low forward on current If (cmos driven).
. Low power supply (must operate at 3 Volts)
. Fast rise time Tr (max. 1 usec).
. Fast fall time Tf (max. 300 nsec)
. Low propagation delay time Tpdr (125 nsec)
. Low input capacitive load , max IIC-bus load is 400 pF.

The low tpdr is of importance if the minimum Hold and Setup time has to be realised at maximum transmission frequency.

Low If and power supply is due to the feature-phone specification.

THE SIGNAL NAMES.

For measurements,an input signal INP is generated and connected to INP1 or INP2 (see fig. 1 and fig. 7).

INP0 is available as a reference for measurements, because INP is influenced by the load of INP1 or INP2.

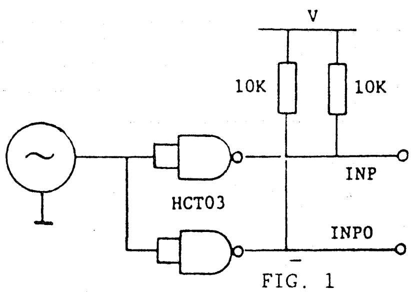


FIG. 1

THE FIRST SOLUTION.

In the schematic diagram of FIG. 2, a solution is given using the CNG36 as insulation device.
Due to the fact that this optocoupler has no internal pulse shaper, and has a TF/TR of 14μS, an external Schmitt trigger is used as pulse shaper to prevent oscillation during switching.

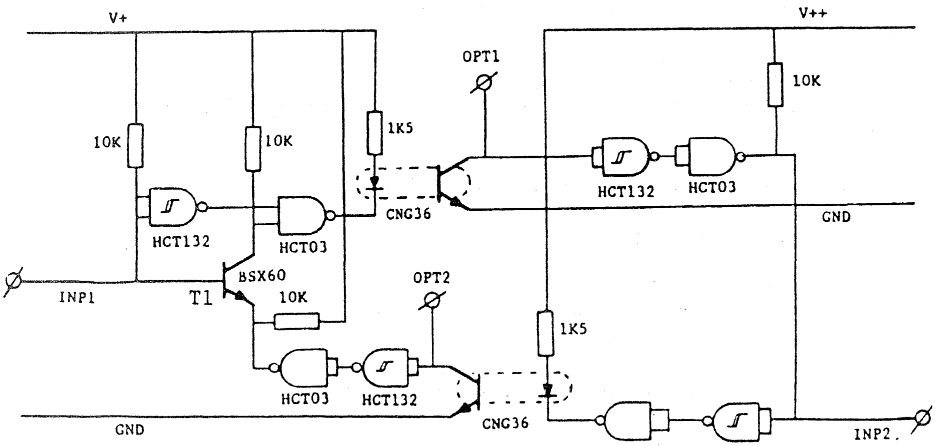


FIG. 2

THE PRINCIPLE OF OPERATION.

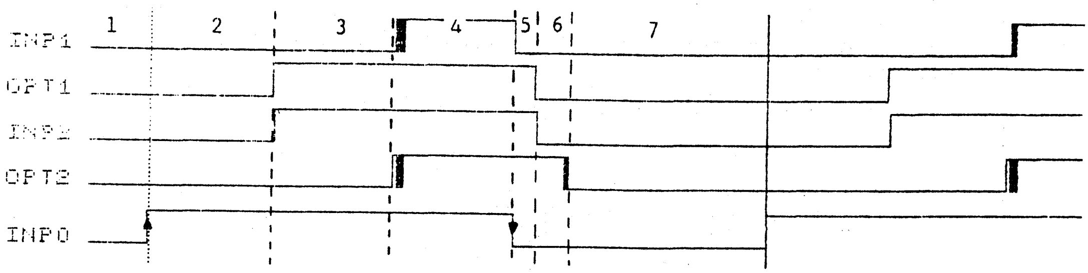


FIG. 3

INP1 is the driving input.

Time interval 1 : Start position.
  INP1 and INP2 : LOW.
  OPT1 and OPT2 : LOW.
  T1 : OFF.
Switch INP1 from LOW to HIGH, enter switch off delay of OPT1.

Time interval 2 : Switch off delay OPT1.
  INP2 : still LOW.
  OPT1 and OPT2 : still LOW.
  T1 : is switched ON.
  INP1 : is clamped at (VBE T1+VLOW HCT03).

Time interval 3 : Switch off delay OPT2.
  OPT1 : HIGH after switch off delay.
  INP2 : follows OPT1 to HIGH, enter switch off delay of OPT2.
  OPT2 : still LOW.
  T1 : still ON
  INP1 : still clamped at (VBE T1+VLOW HCT03).

Time interval 4 : Stable position.
  OPT2 : HIGH after switch off delay.
  T1 : is switched OFF.
  INP1 : follows OPT2 to HIGH.
  OPT1 : is HIGH.
  INP2 : is HIGH.
Switch INP1 from HIGH to LOW, enter switch on delay of OPT1.

Conclusion : The low to high propagation delay is, two times the switch off delay of the optocouplers.

Time interval 5	:	Switch on delay OPT1.
OPT2	:	is HIGH
OPT1	:	is HIGH.
INP2	:	is HIGH.
INP1	:	is LOW
T1	:	is switched OFF.

Time interval 6	:	Switch on delay OPT2.
INP1	:	LOW.
T1	:	OFF.
OPT1	:	LOW after switch on delay.
INP2	:	follows OPT1 to LOW, enter switch on delay of OPT2.
OPT2	:	still HIGH.

Time interval 7	:	Stable position.
INP1	:	LOW.
T1	:	OFF.
OPT1	:	LOW.
INP2	:	LOW.
OPT2	:	HIGH, after switch on delay

Conclusion : The high to low propagation delay is, two times the switch on delay of the optocouplers.

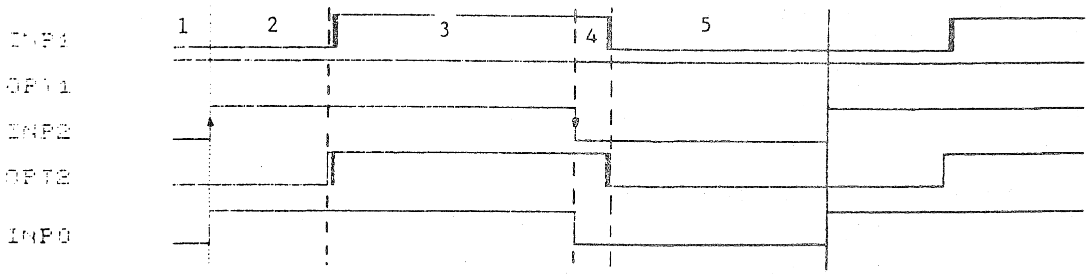


FIG. 4

INP2 is the driving input.

Time interval 1 : Start position.
INP2 : LOW.
OPT1 and OPT2 : LOW.
T1 : ON.
INP1 : is clamped at (VBE T1+VLOW HCT03).
Switch INP2 from LOW to HIGH, enter switch off delay of OPT2.

Time interval 2 : Switch off delay OPT2.
INP1 : still LOW.
OPT2 : still LOW.
OPT1 : HIGH.
T1 : ON.
INP1 : still clamped at (VBE T1+VLOW HCT03).

Time interval 3 : Stable position.
OPT2 : HIGH after switch off delay.
T1 : switched OFF.
INP1 : follows OPT2 to HIGH.
OPT1 : HIGH.
Switch INP2 from HIGH to LOW, enter switch on delay of OPT2.

Conclusion . The low to high propagation delay is equal to the switch off delay of one optocoupler (OPT2).

Time interval 4 : Switch on delay of OPT2.
OPT1 : HIGH.
OPT2 : still HIGH.
T1 : still OFF.
INP1 : still HIGH.
INP2 : LOW.

Time interval 5 : Stable position.
OPT2 : LOW after switch on delay.
T1 : switched ON.
INP1 : is clamped at (VBE T1+VLOW HCT03).
OPT1 : HIGH.

Conclusion : The high to low propagation delay is equal to the switch on delay of one optocoupler (OPT2).

THE MEASUREMENT RESULTS OF FIG. 2.

In fig. 5 the measurement results are given with input 1 as the driving input.

In fig. 6 input 2 is the driving input, the maximum frequency used in both cases is 10 kHz

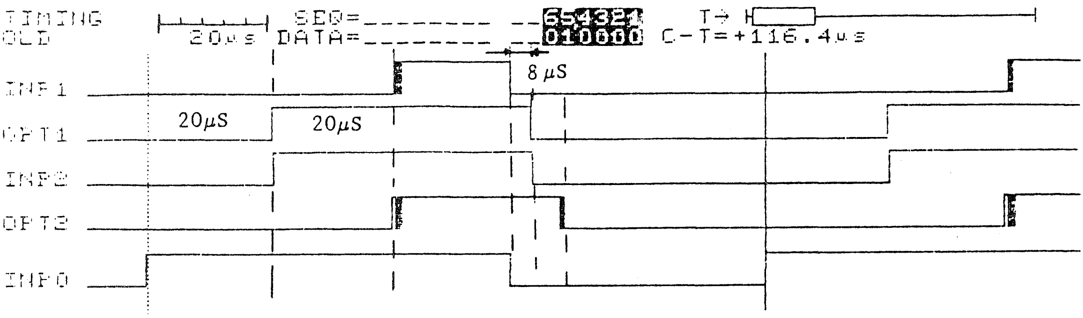


FIG. 5

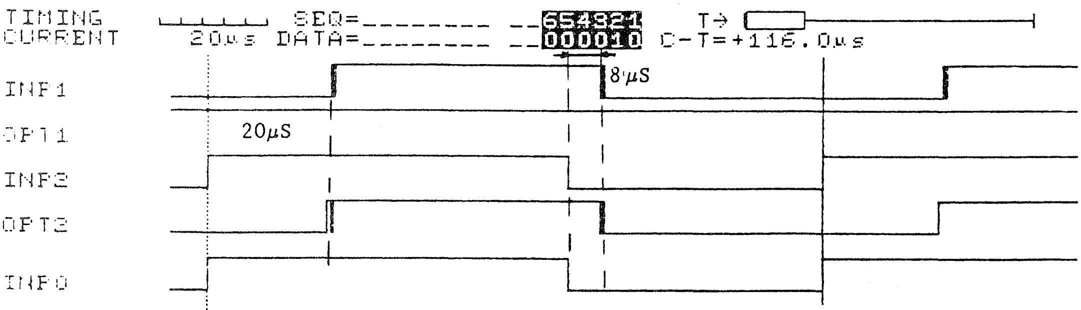


FIG. 6

From the figures 5 and 6 we can conclude that this solution can not meet the IIC specification .
However with software a Master - Slave configuration with a maximum transmission frequency of 10 kHz can be realised.
The software must then written such that the extra delay of 40 Usec is incorporated in the IIC protocol.

An other way is to decrease the Ton/Toff time by means of extra hardware, or use more complex optocouplers as in solution 2.

THE SECOND SOLUTION.

In this solution as can be seen in fig. 7, a HP2630 is used as insulating device. The advantage of this type is that it is internally gated ,and has a Ton/Toff time of 125 Nsec.

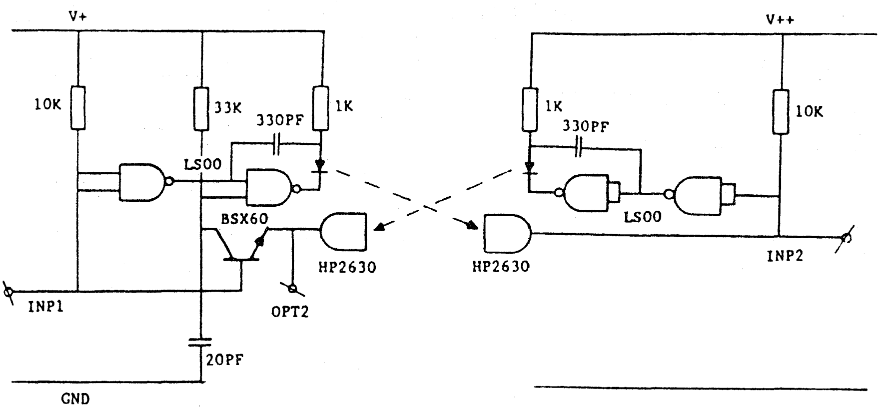


FIG. 7

Note:

. This circuit is used at our lab., as insulation interface between the MAB 8400 IIC lines and an external device.

. Newer types with lower If are available such as the HPCL2300, but at the time this circuit is build and tested there was no hardware available but the HP2630.

. The HPCL2300 has a Ton/Toff which is comparable to the HP2630, but has a lower If.

THE MEASUREMENT RESULTS OF FIG. 7.

In fig. 8 the measurement results are given with input 1 as driving input.

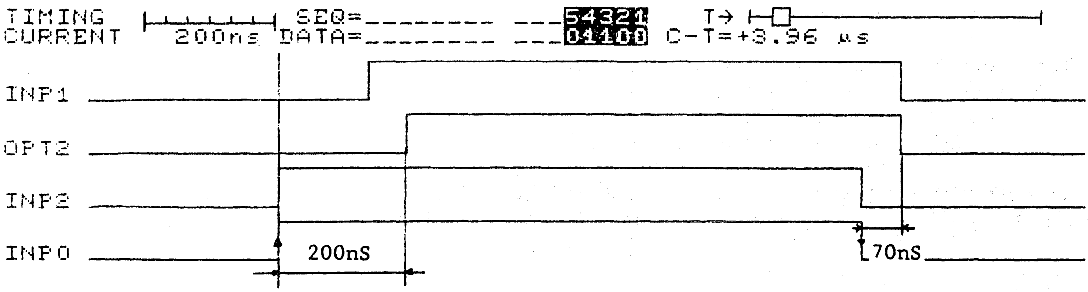


FIG. 8

In fig. 9 input 2 is the driving input, the frequency used was 500KHz which is well beyond the maximum transmission frequency of the IIC bus. The next advantage is that the hardware used is less than that of solution 1.

Technically solution 2 is promising though still work has to be done, such as testing its function under the IIC bus condition.

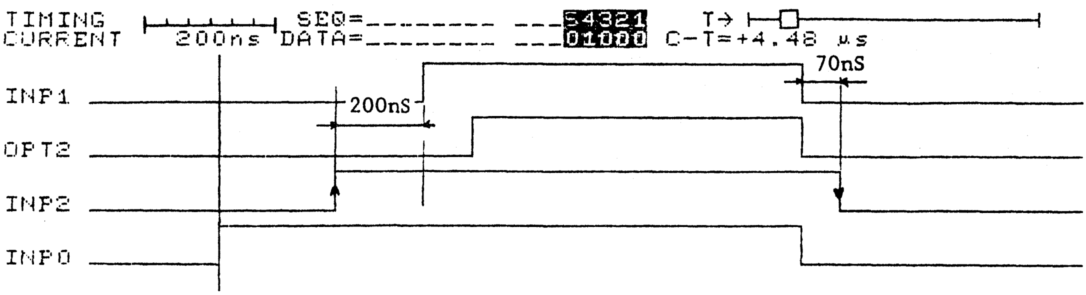


FIG. 9

CONCLUSION.

From a technical point of view solution 2 seems to be usable.
In general a gated insulation device with low If and fast Ton/Toff times is a must to fullfil the specification of the IIC bus .

Low If is needed if CMOS is used to deliver the If, or in applications where low stand by power is specified.

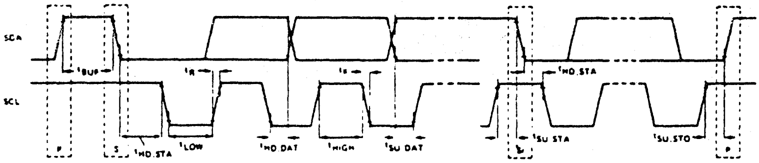
Fast Ton/Toff times are needed to fullfil the HOLD and SETUP times of the IIC bus and the maximum transmission frequency.

Solution 1 can be used under limited conditions i.e.:

maximum transmission frequency of 10 kHz.
master-slave configuration.
extra software, to control the delays.

The circuits are functionally tested , but NOT tested under IIC_bus_test_conditions, temperature and voltage variations.

THE IIC-BUS SPECIFICATION



parameter	symbol	min	max	units
SCL clock frequency	f_{SCL}	0	100	kHz
Time the bus must be free before a new transmission can start	t_{BUF}	4,7	-	µs
Hold time START condition. After this period the first clock pulse is generated	$t_{HD;STA}$	4,0	-	µs
The LOW period of the clock	t_{LOW}	4,7	-	µs
The HIGH period of the clock	t_{HIGH}	4,0	-	µs
Set-up time for START condition (only relevant for a repeated start condition).	$t_{SU;STA}$	4,7	-	µs
Hold time DATA for CBUS compatible masters	$t_{HD;DAT}$	5,0	-	µs
		0*	-	µs
Set-up time DATA	$t_{SU;DAT}$	250	-	ns
Rise time of both SDA and SCL lines.	t_R	-	1	µs
Fall time of both SDA and SCL lines.	t_F		300	ns
Set-up time for STOP condition	$t_{SU;DAT}$	250	-	ns

6 EMC

CONTENTS

APPLICATION NOTE Nr	ETT89016
TITLE	Measures to meet EMC requirements for TEA1060 - family speech transmission circuits
AUTHOR	M.Coenen, K. Wortel
DATE	October 1989

Table of Contents:

1. Introduction

Many of todays electronic apparatus have to be used in a RF-signal polluted environment. The number of wireless communication paths has increased quite rapidly for various applications e.g. : national broadcasting for radio and television, citizen band, short-wave communication, beacons, Radio amateurs, and cellular telephone. They cover the frequency spectrum from 10 kHz up to 1 GHz and above.

The transmitted energy of such devices may be quite large or the distance between the RF-source and other equipment may be quite small and may become a source of interference to other kinds of apparatus such as a common electronic telephone set.

Many Postoffices and/or National Authorities have defined requirements with respect to the functional specifications of the telephone sets. With respect to EMC (ElectroMagnetic Compatibility) no common European or International specification exists (yet) and the individual standards does not make it much easier.

The existing immunity test methods can be divided in three main groups:
- differential mode current injection,
- electromagnetic field generated by a strip-line (TEM-cell),
- common mode current injection,

and commonly cover the frequency range from 10 kHz up to 150 (200) MHz.

A test method is incomplete without defining the limits with respect to the demodulated signal towards the a,b-lines or the earpiece capsule to which the demodulation becomes discernable. Here again the methods can be divided in two main groups:
- acoustic measurements (measuring SPL or dBPa) or electrical measurements across the earpiece capsule,
- electrical measurements on the a,b-lines

and the required S/I-ratio (signal-to-interference ratio) may vary between 26 - 40 dB. The nominal signal (send and/or receive) level at 1 kHz is set at 100 mV across the telephone line impedance, $\simeq$ 600 Ω.

In order to satisfy the most stringent requirements, the German proposal was adopted to verify the immunity performance of our TEA 1060 applications.

2. Immunity test system

The printed circuit boards on which the speech transmission ICs are used in their applications are commonly small compared to the wavelengths of the applied interference signals (frequency < 300 MHz $\rightarrow$ λ > 1 meter). The traces on the printed circuit board should be kept small (< 1/10) compared to this wavelength to ensure that direct pick-up will be negligible.
In this situation the only part of the telephone set capable of acting as an antenna in order to receive a part of the interference signal will be the leads towards the public telephone network on one hand and the leads towards the handset on the other.

When the telephone set is placed in a strip-line (or TEM-cell), the RF-induced signal will be highly sensitive to the geometry of these leads, while during the current injection test these geometric constraints are excluded.

2.1. ElectroMagnetic field

With a parallel strip-line a plane-wave electromagnetic field is generated between the two (or three) flat conductors being the transmission line. The generated electric fieldstrength is equal to the voltage between the conductors divided by the distance between them.

The telephone set with its handset and its leads to the public telephone network receive this RF-field. The induced signal at low frequencies will be low due to the un-efficiency of the "antenna" (lead length) which is present between the two planes. At higher frequencies the value of the induced signal will show resonances due to the geometry of the set-up between the planes. Furthermore, the resonant frequencies will vary with the geometry and within a reasonable test sequence no "worst-case" response can be found.

2.2. Current injection

As explained above, the test set-up geometry may have a large impact on the results obtained and for reproduceability sake these kind of tests should be avoided.
With the current injection method the interfering signal is directly applied to the telephone set as a common-mode signal. The common-mode impedance of the source is chosen to be an "average" of 150 Ω [1].
Therefore, coupling networks (CI-networks) have to be placed between the 50 Ω RF-generator and the telephone set under test. The telephone set with the handset are placed on a non-metallic support of 100 mm height above a metallic reference plane on which the coupling networks are mounted. The lead length of the a,b-lines, between the public telephone coupling network, known as T-network, and the telephone set, should be limited to 300 mm. This to avoid resonances due to the geometry (lead length) of the test set-up. The handset should be wrapped in copper foil and this should be connected to reference by a series-impedance of 500 Ω + 200 pF, known as the artificial hand [2].

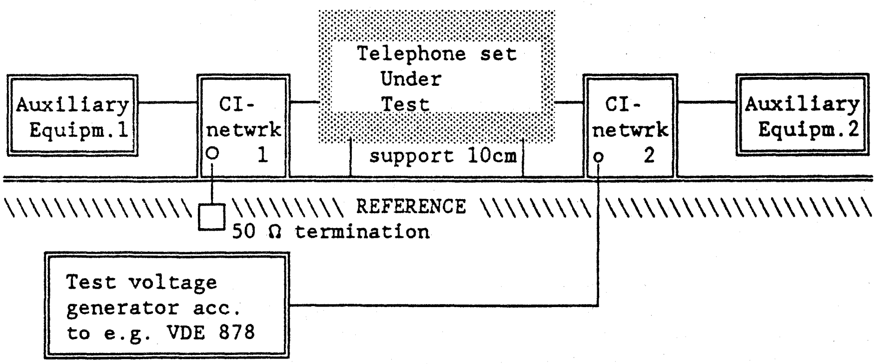


Figure 1. Schematic set-up for the current injection test.

The RF-signal is an amplitude modulated signal. The modulation depth is 80 % and the modulating frequency is 1 kHz.

By forcing the RF-current to flow through that telephone set under test, the direct immunity is tested as well as the immunity to RF-fields due to the re-radiation caused by that current.

Measurements have shown that sets complying with the current injection method at normal RF-signal levels indicated in table 1, column "normal" gave an immunity to EM-fields, under the same S/I-criteria, of 10 Volt/meter and higher.

Frequency range MHz	Applied voltage normal	enhanced
.01- .1	1.5	1.5
.1-30	3.0	6.0
30-150	.5	1.5 -

Table 1. The RF-signal levels that have to be applied to the telephone set under test.

3. Measures to improve the basic EMC performance (M1)

This section discusses the measures to improve the EMC performance of the speech and transmission ICs and the effect of the extra EMC components on the basic functioning of the TEA 1060-applications as described in report ETT 8613. The five capacitors at IR, MIC, LN and V_{CC} have improved the application quite a bit, but according to the more stringent requirements other necessary measures needed to be taken. The immunity performance of the TEA 1060 applications with respect to the basic EMC measures, see annex 1, with this current injection method and normal RF-levels which are used throughout this evaluation, is given in annex 2a); detection on the a,b-lines, and 2b); detection at the earpiece side.

First the measures at the microphone input, the earpiece output pins and the connection to the a,b-lines are discussed. The measures to improve immunity are given in annex 3, and the EMC-performance is given in annex 4a, detection on the a,b-lines, 4b, detection at the earpiece side.

3.1. Practical implementation

The RF-signals are mainly picked up by the leads connected to the telephone base set, see figure 2. The easiest way to overcome unwanted demodulation is to let the RF-induced current flow through an extra conductor instead of through the reference of the telephone circuitry.
Care must be taken that the re-radiation of the RF-currents through this extra conductor will not be picked up by the telephone circuitry again. Therefore an open space must be made between the circuitry and this conductor to reduce mutual coupling between the two. No long wires, which run over the PCB may be used for this connection !!

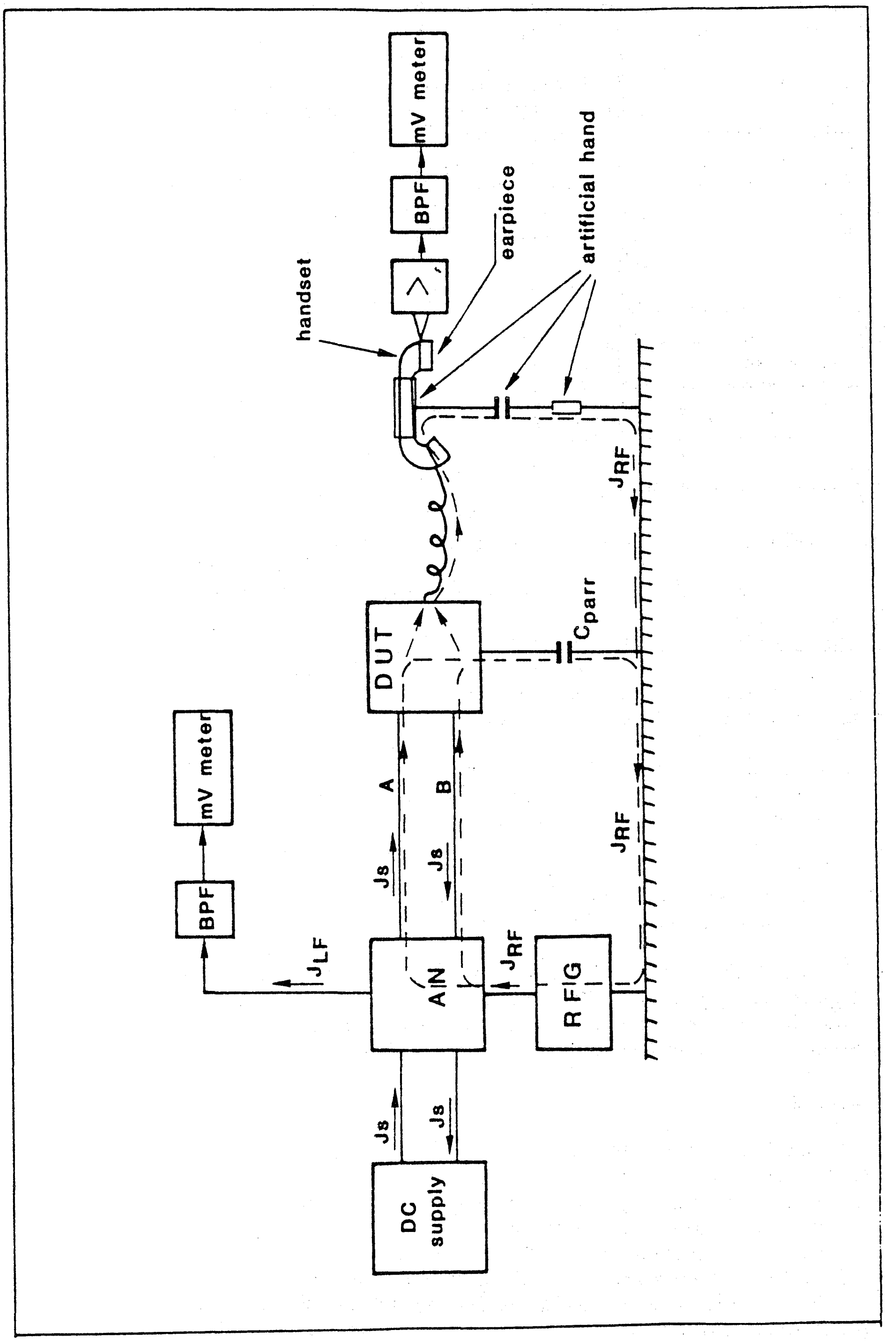
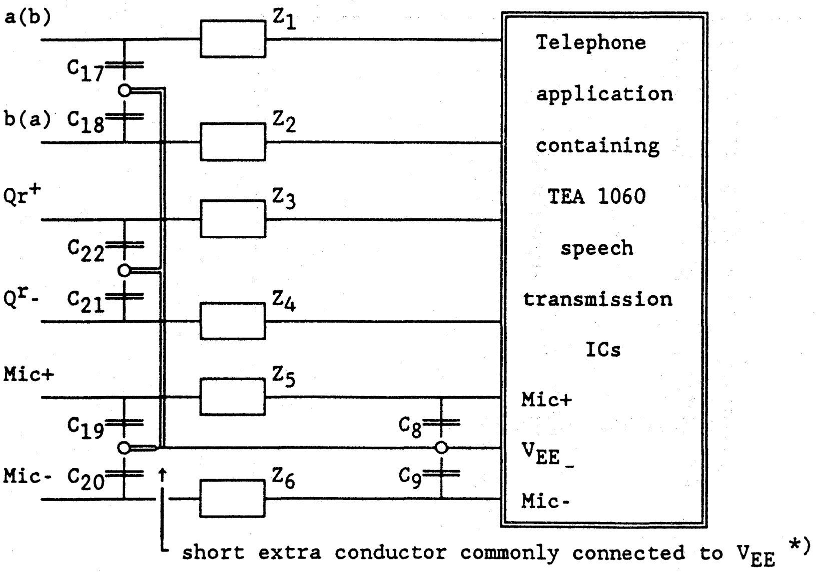


Figure 2. RF-current flow through the telephone application according to the chosen test method.



*) If other peripherals of the application are used, the same network must be continued.

Figure 3. The modified circuit to improve the immunity of Telephone sets containing TEA 1060-family speech transmission IC (see also annex 3 and 5).

The effect of these measures on functionality will be discussed in the next paragraphs.

3.1.1. Microphone inputs:

To prevent demodulation of the RF-injected signals in the microphone stage extra filters at the input must be applied:

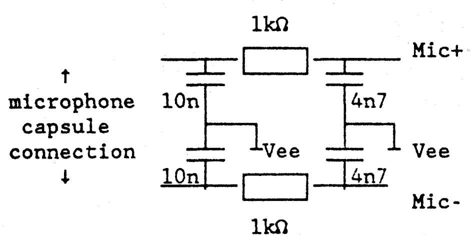


Figure 4. "Extra filtering at microphone inputs"

The 10nF capacitors decouple the RF-signal entering via the handset cord. The RF-signal which is then still present is filtered by means of the 1kΩ resistors and 4n7 capacitors.

cut-off frequency: 33.86kHz

As regards frequency dependence the extra filter does not have a noticeable influence on sending gain for frequencies between 300 and 3400Hz.

However, since the input impedance of the microphone amplifier is not infinite attenuation occurs. And due to the fact that the input impedance has a certain tolerance the attenuation shows a tolerance also. This means extra tolerance on the gain setting occurs!
In table 2, the attenuation, impedance and their tolerance are shown for several types of the TEA 1060 family:

	minimum	typical	maximum	
TEA 1060 input impedance:	3.3	4.1	4.9	kΩ
Attenuation with 1kΩ:	-2.30	-1.90	-1.61	dB
extra tolerance:	-0.29		+0.40	dB
TEA 1061 input impedance:	16.5	20.4	24.3	kΩ
Attenuation with 1kΩ:	0.51	0.42	0.35	dB
extra tolerance:	-0.10		+0.07	dB
TEA 1067/68 input impedance:	25.5	32.0	38.5	kΩ
Attenuation with 1kΩ:	0.33	0.27	0.22	dB
extra tolerance:	-0.05		+0.06	dB

Table 2. Gain and impedance tolerances due to the RF-immunity measures

The extra tolerance on the sending gain for TEA 1061/67/68 of less then 0.1 dB seems acceptable, but for the TEA 1060 the extra tolerance is rather high. With a 180Ω resistor (increase 4n7 to 27n for same cut off frequency) instead of the 1kΩ for the TEA 1060 the extra tolerance on sending gain is less then 0.1 dB which is in line with the rest of the family.

3.1.2. Earpiece outputs

For the earpiece outputs also filtering is applied to prevent RF-signals to enter the internal TEA 1060 circuitry:

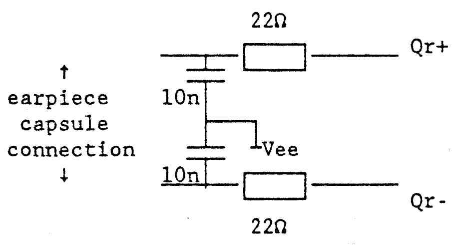


Figure 5. "Extra filtering at earpiece outputs"

The 10nF capacitors first decouple the handset cord. The 22Ω resistors attenuate the RF-signal which is still left. Since demodulation in a non linear device as a transistor has a quadratic relationship with the applied level, only little RF-attenuation will improve a lot.

The filter from Qr to the earpiece has no influence on speech signals between 300-3400Hz, cut-off frequency: 723kHz.

The disadvantage of the extra components for LF:

1. Receiving gain decreases
2. Maximum output level decreases

The actual influence is dependent on the impedance of the earpiece.

Example:
Influence gain + max output level: (ignore small change Qr+ level for max output)

150Ω single ended: -1.19 dB

450Ω bridge tied load: -0.81 dB

The effect on gain can easily be compensated (increase resistor between Qr+ and GAR in basic application) but the effect on maximum output level will remain.

3.1.3. Telephone line:

In front of the diode bridge 2 extra decoupling capacitors are desired to prevent RF-signals to enter the TEA 1060 (see figure).
For the application this means care has to be taken with Balance return loss. With a 600Ω set impedance still more than 20dB of BRL (balance return loss) can be reached if 10nF is present at the line.
Since the two 4n7 capacitors form a load at the line of only 2n35 still 1-6n8 can be connected to pin LN of the TEA 1060 after the diode bridge.

The coil after the diode bridge will form a high impedance at high frequencies forcing decoupling via the line capacitors rather than through the TEA 1060.
For audio frequencies the influence of the coil can be neglected, but its low frequency series resistance should be as low as possible!

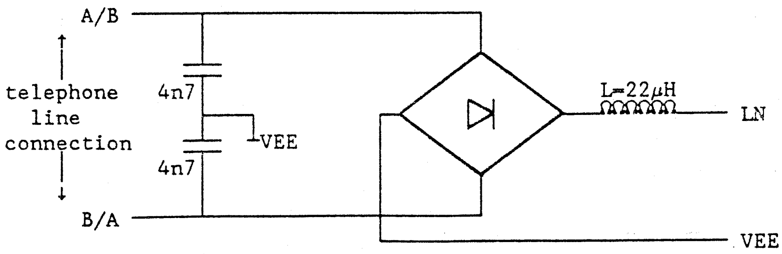


Figure 6. "Extra filtering at a,b-lines"

3.2. Conclusions

The following measures to improve the RF-immunity can be taken without influencing the TEA 1060-family functioning too muchon the common application board CAB 3391, see annex 7:

Microphone inputs:

- Microphone cord decoupling: 2 x 10nF
- Series resistors : 2 x 180Ω (TEA 1060 only), 2 x 1kΩ (rest)
- Mic. inputs decoupling : 2 x 27 nF (TEA 1060), 2 x 4.7nF (rest)

Receiver outputs:

- Earpiece series resistors: 2 x 22Ω
- Earpiece cord decoupling : 2 x 10nF

Telephone line:

- Telephone line decoupling in front of diode bridge: 2 x 4n7
- Series inductor after diode bridge : 1 x 22μH
- reduction of C_{LN} from 10 nF to : 1 x 5n6

4. Measures to reach enhanced immunity (M2)

Some problems remain with respect to the internal reference (SLPE) which is used inside the speech transmission IC. The transmission stage of the TEA 1060 may demodulate the RF-signal into a 1 kHz base band signal which can be measured on the a,b-lines. The side-tone network couples this signal to the input of the receiving amplifier. This signal then, becomes audible at the earpiece side.

4.1. Output stage:

The EMC measures which have to be taken at the output stage need to be chosen properly because the output stage can become instable.

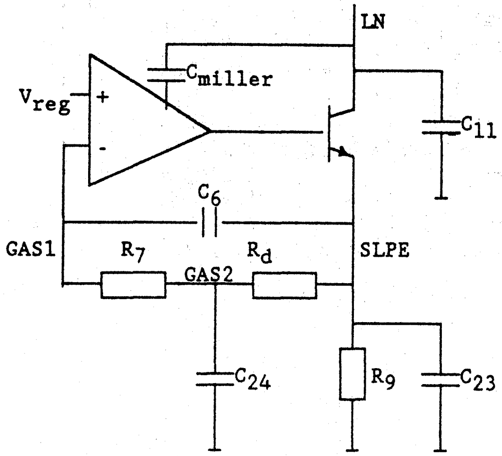


Figure 7. Transmission stage of the TEA 1060

From this network the ranges for the values of the capacitances at the pins: SLPE, GAS1, GAS2 and LN have been optimised to improve the RF-immunity without affecting the stability of the circuit.

Optimal selection of the capacitors:

* C_6 - 100pF minimum, larger values can be used for High Frequency roll-off of the transmission amplifier.

* C_{24} - 100pF maximum for all possible applications.

$$\text{typ. } C_{24} << \frac{(R_7 + R_d).C_6}{R_d}$$

* C_{11} - 1 to 10nF, larger values are not practical with respect to the BRL!

* C_{23} - 220pF maximum.

4.2. Conclusion

The immunity of the TEA 1060-family can be further improved by adding 2 capacitors; C_{24} = 100 pF, C_{23} = 220 pF, see annex 5. The results on the immunity can be found in annex 6a, for the detection on the a,b-lines and 6b, for the detection at the earpiece side.

5. Other measures

5.1. Do:

* Apply series resistors (> 1 kΩ) with the DTMF, PD, MUTE lines between the dialer or µP and the speech transmission IC.

* Place R_{SLPE}, R_{STAB} as close as possible to the TEA 1060

* Place C_{REG}, C_{IR}, C_{MIC}s and C_{LN} as close as possible to the TEA 1060

* Make a ground loop around the TEA 1060 but_avoid other then functional currents to run through this loop!

5.2. Don't:

* Connect large circuit parts to SLPE without decoupling capacitor C_9

* Decouple e.g. LN or IR at several reference points on your PCB without any series impedances. It will cause resonances with the copper trace inductances with high Q-factors.

* Apply a capacitor in parallel to R_{STAB}. It will lower the phase margin at lower frequencies (1-5 MHz).

6. Conclusion

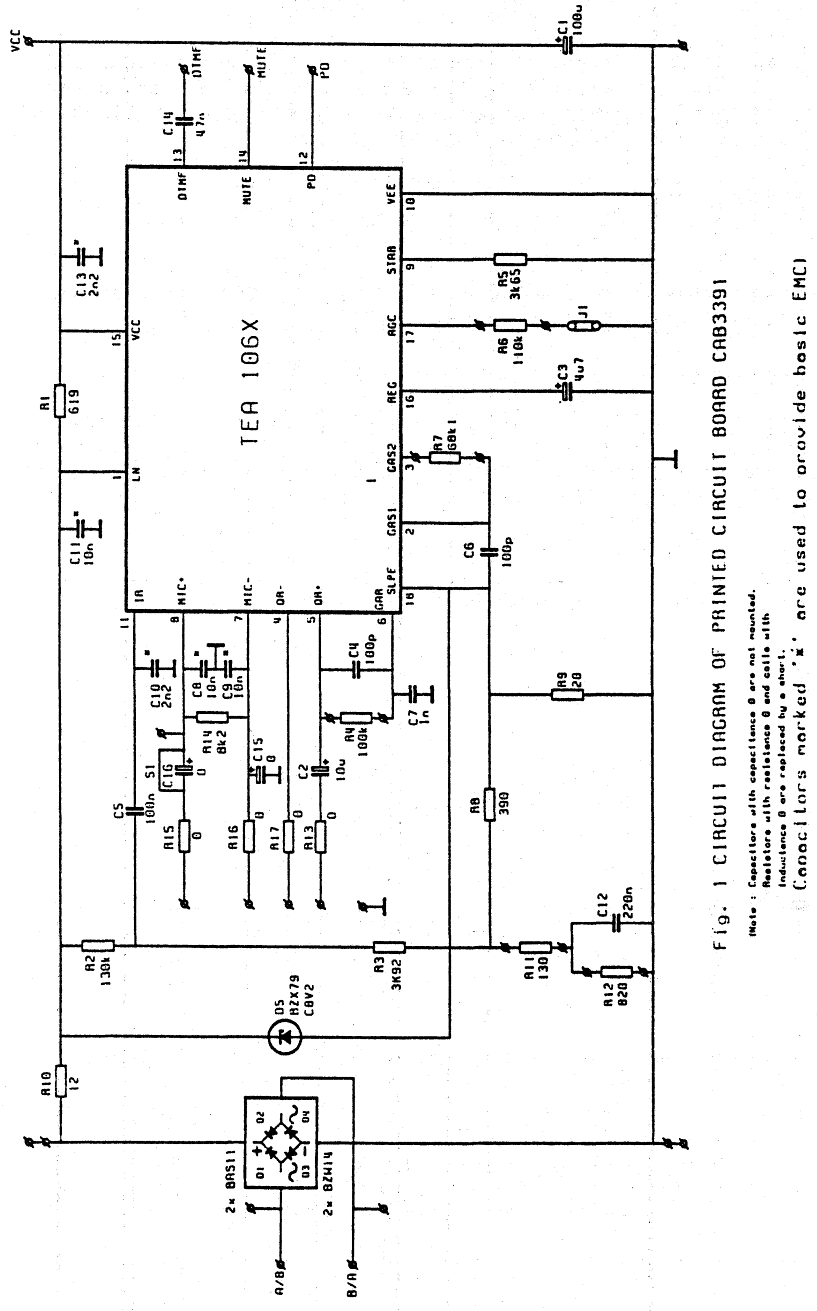
The final result of the measures partly depends on the remaining part of the circuitry connected to the TEA 1060. This because the performance of the IC has been tested mainly on its application board, while commonly other circuit parts will highly influence the RF-current distribution throughout the telephone set.

With the above described measures it is possible to create a telephone set application which will give at least a S/I-ratio of 26 to 40 dB both at the a,b-lines and at the earpiece-side in the entire frequency range 100 kHz up to 150 MHz under "normal" RF-severity levels.

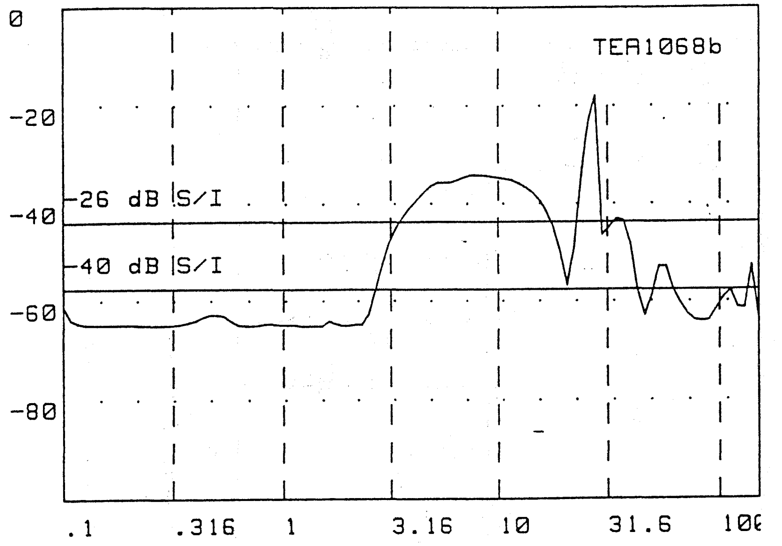
Reproduceability of the results obtained from the test set-up which is further explained in VDE 0878 [3] has shown to be within a few dBs (≤ 3).

7. References

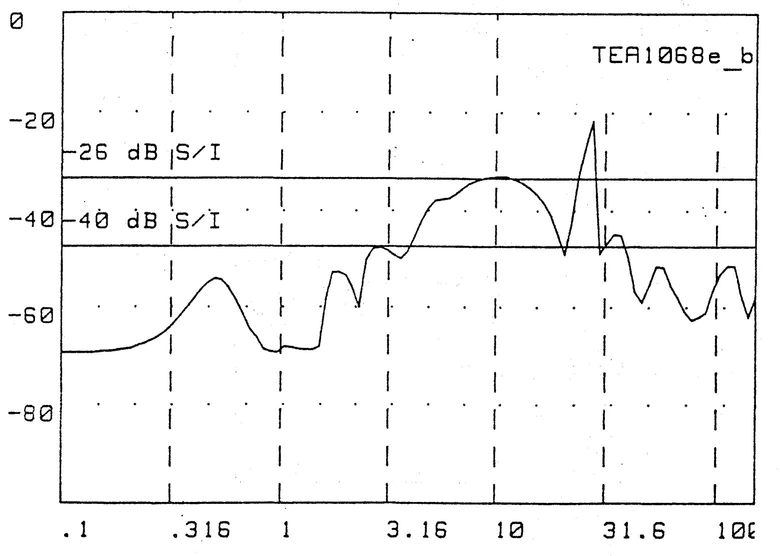
[1.] IEC CISPR publication 20,
[2.] IEC CISPR publication 16,
[3.] VDE 0878 part 1 and 200, 1987.



Annex 1 TEA 1060 application with 'basic' EMC performance
(add 5 capacitors)



a) detection on a,b-lines



b) detection at earpiece side.

Annex 2 EMC performance of basic application.

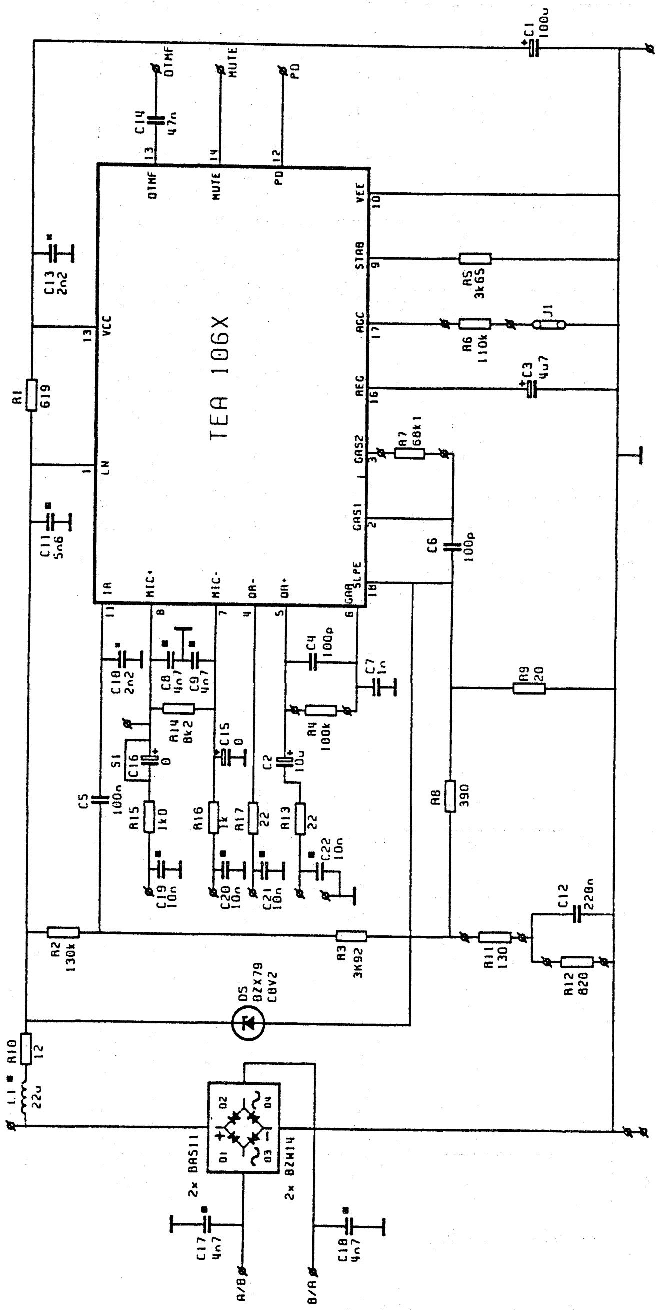
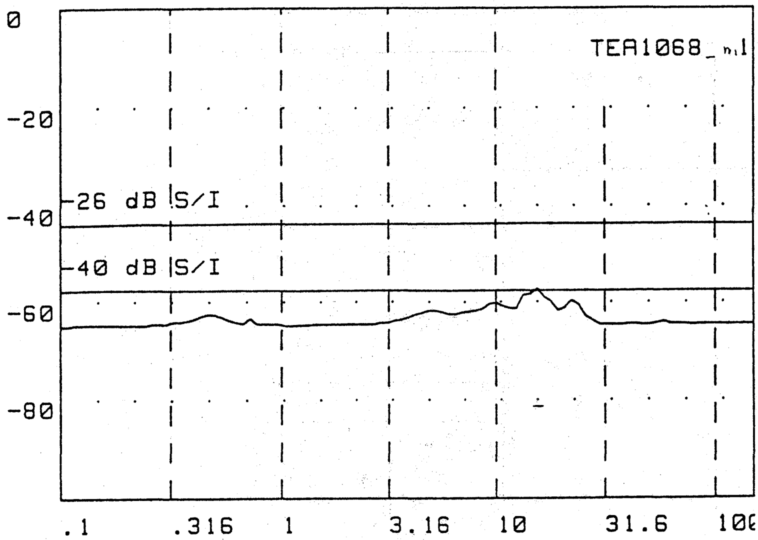


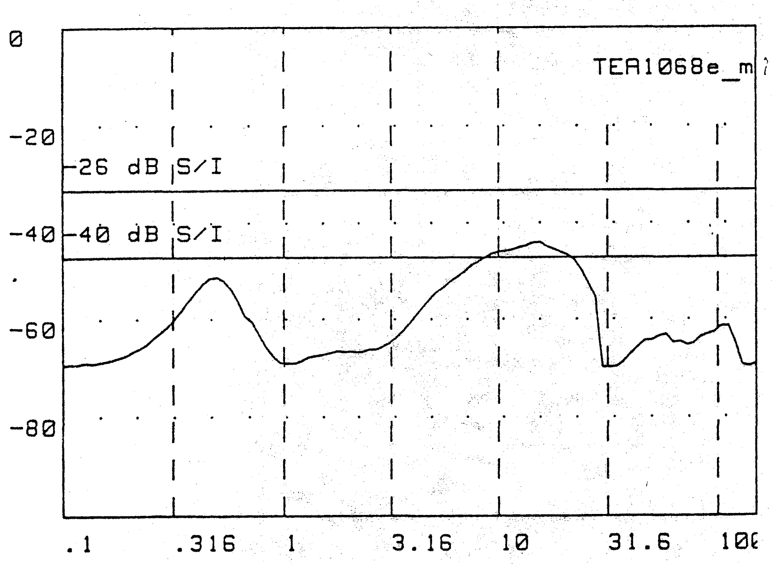
Fig. 2 CIRCUIT DIAGRAM OF PRINTED CIRCUIT BOARD CAB3391 (Mod.1)

(Note : Capacitors with capacitance 0 are not mounted.
Resistors with resistance 0 and coils with
Inductance 0 are replaced by a short.
Components marked '■' are used to improve basic EMC)

Annex 3 TEA 1060 application with 'improved' EMC (M1) performance
(add 11 components)



a) detection on a,b-lines



b) detection at earpiece side.

Annex 4 EMC performance of improved application.

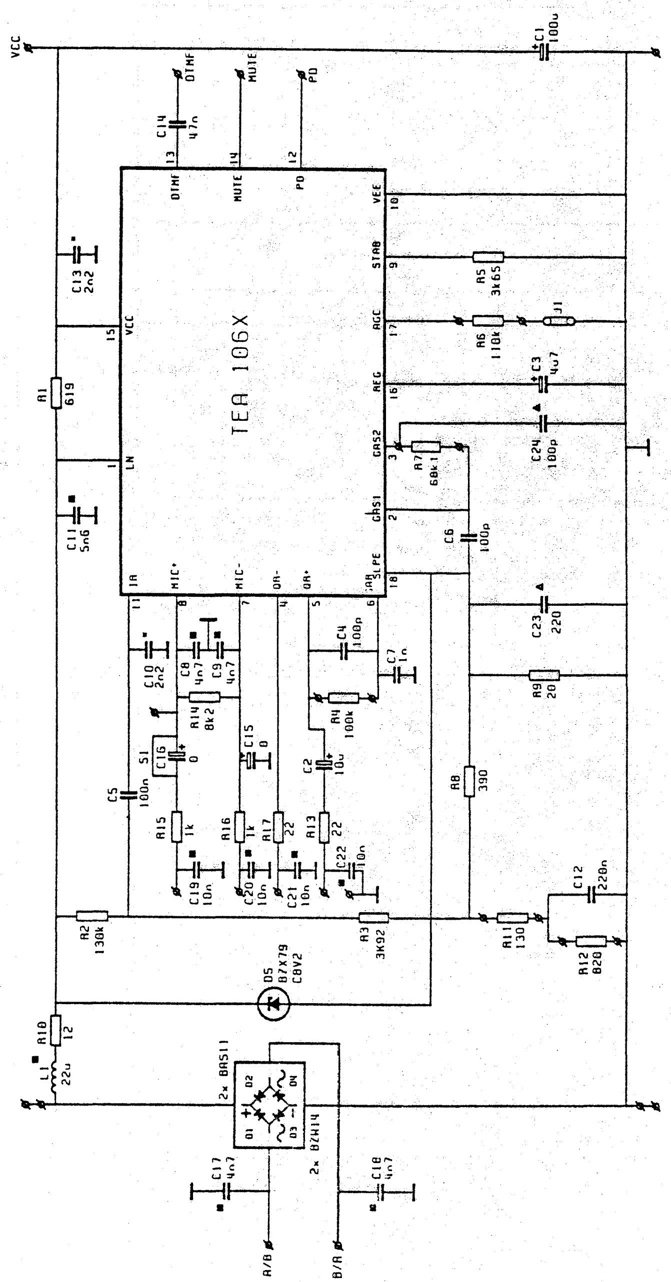
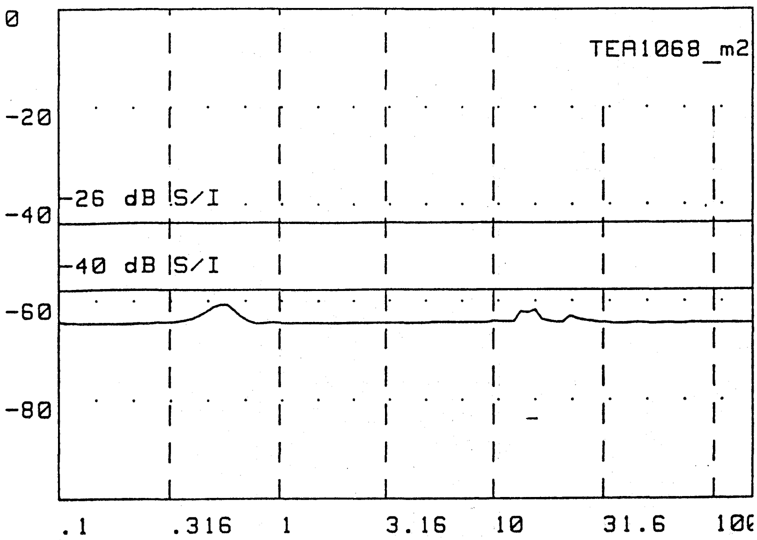


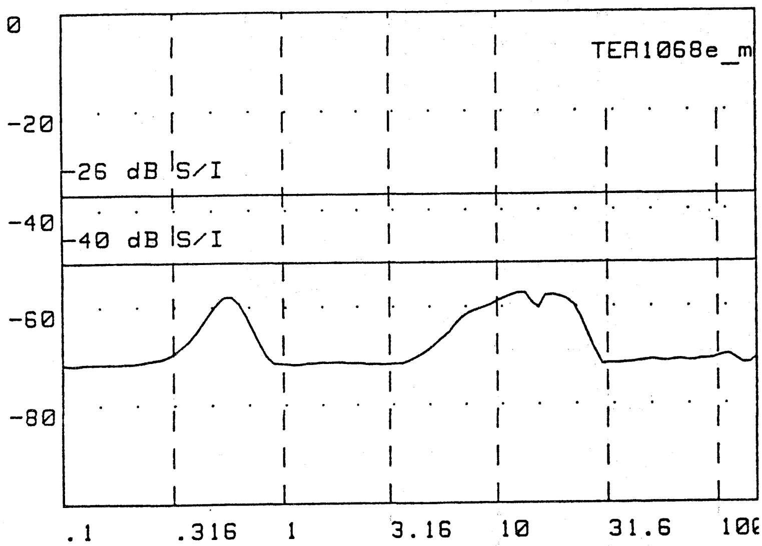
Fig. 3 CIRCUIT DIAGRAM OF PRINTED CIRCUIT BOARD CAB3391 (Mod.2)

(Note: Capacitors with capacitance 0 are not mounted. Resistors with resistance 0 and coils with inductance 0 are replaced by a short.
Capacitors marked 'Δ' are used to provide enhanced EMC)

Annex 5 TEA 1060 application with 'enhanced' EMC (M2) performance (add 2 capacitors)

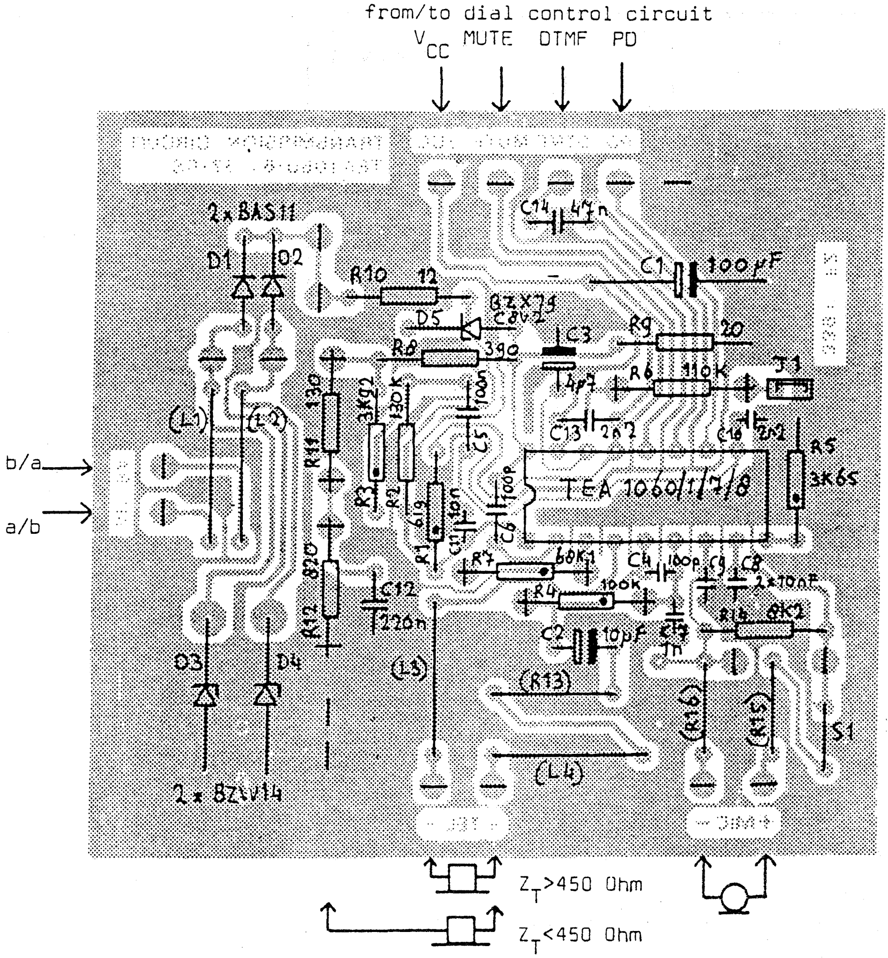


a) detection on a,b-lines



b) detection at earpiece side.

Annex 6 EMC performance of enhanced application.



Annex 7 Layout of the TEA 1060 application board CAB 3391.

7 DESIGN - EXAMPLES

CONTENTS

APPLICATION NOTE Nr	ETT/AN91010
TITLE	User's Manual PR4535X DEMO board with TEA1083A -TEA1064A call progress monitoring application
AUTHOR	F. van Dongen
DATE	June 1991

Table of Contents:

1. INTRODUCTION.

The printed circuit board (PCB), with identification number PR4535X, contains the application of the call progress monitoring circuit TEA1083A combined with the transmission circuit TEA1064A. It can be applied also with the TEA1083 instead of the TEA1083A.

The TEA1083A and the TEA1083 are members of the TEA108X "line monitoring IC-family" as indicated in the overview of table 1.
The TEA1083A is a 16-pins IC, containing a supply and start circuit, loudspeaker amplifier, loudspeaker enable (LSE) function and a power down (PD) circuit. The TEA1083 is an 8-pins IC with the same provisions as the TEA1083A, except for the power down facility. The TEA1083 has no PD input.

This documentation describes the transmission and the monitoring part of the TEA1083A / TEA1064A application. The demonstration board (PR4535X) contains the TEA1083A and the TEA1064A, however, the TEA1083A (16-pins) may be replaced by the TEA1083 (8-pins). The PCB is prepared for use with TEA1083.
On the other hand, the TEA1064A (20-pins) may be replaced by the TEA1060, TEA1061, TEA1067 or TEA1068 (TEA106X) for which an 18-pins shadow IC socket is mounted on the board. Modification instructions of the PCB are summarized in appendix 1.

Table 1. Overview of TEA108X line monitoring ICs and there features.

Product:	TEA1082	**TEA1083**	**TEA1083A**	TEA1085/TEA1085A
Application area 1)	call progress monitoring			listening-in
PD facility	x	.	x	x
MUTE or LSE facility	.	x 2)	x 2)	x 3)
Dynamic limiter	.	.	.	x
Howling limiter	.	.	.	x
VBB setting	.	.	.	x
SEL 4)	x	x	x	x
BTL 4)	.	.	.	x
Number of pins 5)	8	8	16	24

Notes:

1) A call progress monitor, in order to enable audible (or visible) monitoring of the progress of the call attempt, is recommended by the ETSI (European Telecommunications Standards Institute) for telephone sets with automatic on-hook dialling facilities. ETSI: a minimum level of 50dBA shall be guaranteed at 50cm distance from the set, at 440Hz and a line level of -20dBm (600Ω). This corresponds with a minimum level of about $100mV_{rms}$ (Pout ≥ 0.2mW) across a 50Ω loudspeaker, Philips type AD2071/Z50.
A listening-in set has to offer the user more facilities. Howling limiting, to reduce annoying loudspeaker and line signals, and dynamic limiting of the loudspeaker signal with respect to supply conditions can be required. Acoustic output levels for listening-in are in the

order of 70-75dBA, corresponding with a loudspeaker level of $\approx 1V_{rms}$ (Pout≈20mW).

2) The TEA1083 and TEA1083A have a LSE (LoudSpeaker Enable) facility.

3) The MUTE of the TEA1085A has a logic input; the MUTE of the TEA1085 a toggle input.

4) SEL: loudspeaker connected as single ended load.
BTL: loudspeaker connected as bridge tied load.

5) Consult product specification concerning package outlines.

The TEA1083A has to be applied in conjunction with a transmission circuit of the TEA1060-family which provides the common line interface for speech as well as for dialling. The TEA1083A offers a monitoring facility of the received line signal via the connected loudspeaker, during on-hook dialling for instance. Listening-in during speech mode is possible, however with reduced performance.

Chapter 2 describes the use of the PCB and the application of the TEA1083A combined with the TEA1064A with respect to line current and voltage, while chapters 3 and 4 describe the two ICs separately.

Detailed information about the TEA1083A and the TEA1064A can be found in the documentation as given in the references of chapter 7. For the TEA1064A is referred to the application report of the TEA1064 which for most of the information is also valid for the TEA1064A. However, differences are present with respect to AGC characteristics, AC start up time, and the presence of a microphone mute possibility.

2. MONITORING COMBINATION OF THE TEA1083A AND TEA1064A.

Fig.1 shows the block diagram of the monitoring combination on the PCB.

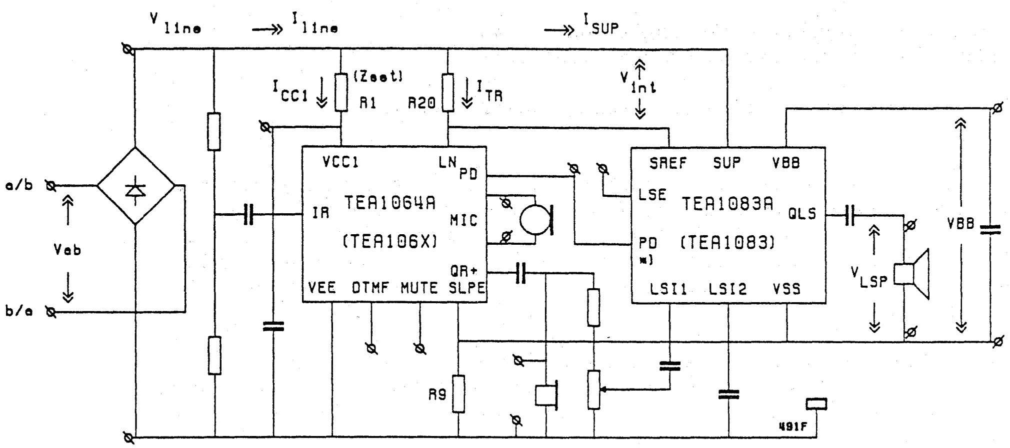


Fig.1 Block diagram of the monitoring application on PCB PR4535X.
*) PD input: TEA1083A only !

The TEA1083A is connected between the positive line terminal and pin SLPE of the TEA1064A. The transmission characteristics such as set impedance, gain settings, gain control etc. are not affected.
The TEA1064A application on the PCB is according to the regulated line voltage mode.

The complete circuit diagram of the application is given in fig.2 while the print layout with components is shown in fig.3.

2.1. How to use the demonstration board.

The terminals to connect the line, transducers and dialling signals are illustrated in fig.3.

The demonstration board can be used after connection of the line supply to the a/b-b/a terminals, and a loudspeaker to the LSP-SLPE terminals. An audio signal on the line can be heard through the loudspeaker. The level of the loudspeaker signal can be controlled by means of the potentiometer.
LSE is interconnected with the VBB voltage level, by means of jumper J2, to enable the loudspeaker. The monitoring function will be disabled when J2 is removed.

A complete telephone set can be build by connecting the handset, dialler and a ringer. The board is assembled for use with a dynamic microphone, which has to be connected between the MIC- and MIC+ terminals. It has to be modified for applications with an asymmetrical microphone. A symmetrical or asymmetrical earpiece can be connected between the TEL- and TEL+ terminals or the TEL+ and VEE terminals.

The DTMF terminal can be used to transmit a DTMF signal to the line by applying a 'HIGH' level to the MUTE terminal. The DTMF dialler can be supplied via the terminals VP-SLPE (see chapter 4.1).
A DP terminal is available for pulse dialling applications. A 'HIGH' level at DP reduces the current consumption of the TEA1064A and the TEA1083A.
The TEA1083 has no Power Down facility !

The PCB is designed in such a way that several parameters can be easily adapted. A brief description of how to change them is given in the following paragraphs.

Notes:
- The TEA1083A contains no dynamic limiter; overload of the loudspeaker amplifier results in signal distortion and reduces the VBB voltage.
- A relative high signal gain from line to loudspeaker (depends on the position of the potentiometer tap) combined with a short distance between microphone and loudspeaker can result in annoying loudspeaker and line signals (howling effect).

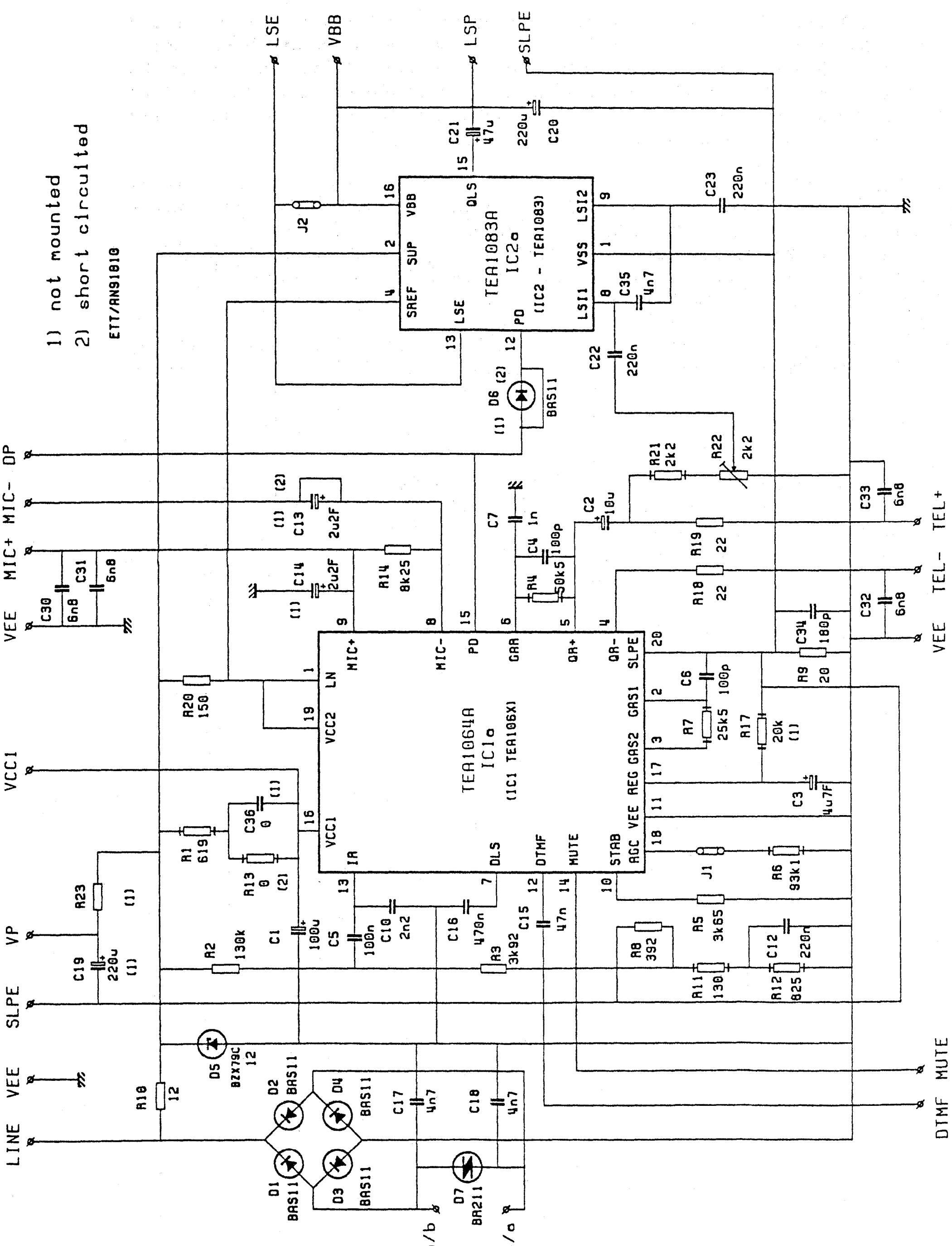


Fig.2 Application diagram of the TEA1083A / TEA1064A combination.

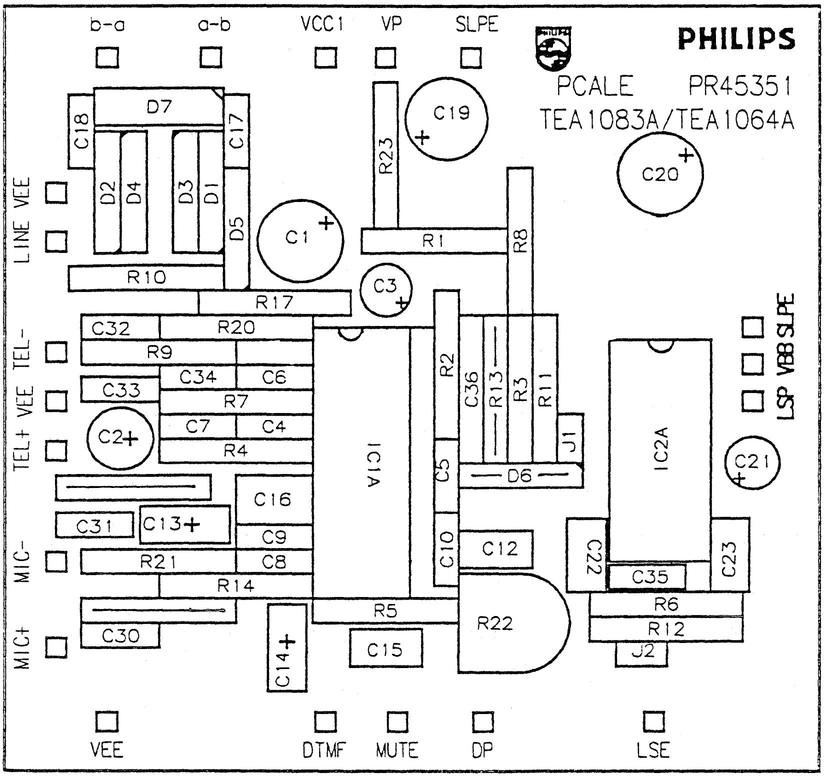


Fig.3 View of the components side of demonstration board -PR4535X-.

Remarks: C19, R23, R13, C36, C14, C13, R17 and D6 are not mounted.
R13, C13 and D6 are short circuited.
R18 and R19 are mounted at the soldering side.

2.2. Peripheral supply / logic inputs.

The TEA1083A can supply peripheral circuits between the stabilised supply voltage VBB (typ. 2.95V) and VSS. The VBB voltage can not be adjusted.
The ground reference for peripherals is terminal SLPE of the TEA1064A (VSS of the TEA1083A). The PD input of the TEA1083A (TEA1083 has no PD input) and the inputs of the TEA1064A (Power Down, Mute and DTMF) are all referred to SLPE. The Power Down inputs of the TEA1083A and TEA1064A have the same structure and can be controlled in parallel from the same source, via terminal DP.

Terminal VEE is the reference of the Mute, Power Down and DTMF inputs of the other members of the TEA1060-family (= TEA106X). Interface circuitry can be required between TEA1083A, TEA106X and peripherals depending on the supply of the peripherals. See appendix 1.

Circuits without connections to either PD, MUTE or DTMF and with a relatively low current consumption (< 200µA), like electret microphones, can be powered between the terminals VCC1 and VEE.

2.3. Start-up / line current split-up.

The internal start circuit of the TEA1083A ensures normal start-up of the TEA1064A whereby the capacitor on pin VBB is charged quickly during this start up phase.

At 20mA line current the TEA1064A is available for speech functions within 60ms, while the VBB capacitor (C20 = 220µF) reaches its typical voltage of 2.95V within 300ms when no extra peripheral current is consumed from VBB. The monitoring function is available within 350ms.

As shown in fig.1, the line current is split up into the currents I_{TR} and I_{CC1} (for the TEA1064A) and I_{SUP} which is available for the monitoring function of the TEA1083A. These currents are:

$$I_{TR} = \frac{V_{int}}{R20} = \frac{0.5}{150} = 3.3\text{mA}, \quad I_{SUP} = I_{line} - I_{CC1} - 3.3\text{mA}$$

where:
- I_{TR} is the current through R20 to bias the output stage / voltage stabilizer of the TEA1064A.
- I_{SUP} is the current into SUP of the TEA1083A.
- I_{CC1} is the current consumed from V_{CC1} to supply the internal circuitry of the TEA1064A (≤ 1.5mA) and the peripheral connected to V_{CC1}.
- V_{int} is an internal bias voltage of the TEA1083A between SUP and SREF (500mV typical)
- R20 on the PCB is 150Ω.

Only a part of I_{SUP} is available to power the loudspeaker. In the TEA1083A I_{SUP} is used to supply the internal circuitry (I_{SUP0} = 2.5mA), while the remaining part of I_{SUP} is available for powering the loudspeaker.
However, the current which can be applied to power the loudspeaker is in practice less than $I_{SUP}-I_{SUP0}$ due to the reduced efficiency of the supply of

the TEA1083A at large line signals. This will be described in the next paragraph.

2.4. Line voltage and loudspeaker level versus line current.

The line voltage Vab between the a/b-b/a terminals and Vline (LINE-VEE) are shown in fig.4. Vab can be calculated by:

$$Vab \simeq 2.Vd + V_{LN-SLPE} + V_{int} + I_{line} * R10 + (I_{line} - I_{CC1}) * R9$$

in which:
- 2.Vd is the voltage drop across the diode bridge.
- R9 = 20Ω, R10 = 12Ω (on the PCB).
- $V_{LN-SLPE}$ (3.3V typically) is the reference voltage of the TEA1064A.

$V_{LN-SLPE}$ can be increased by means of R17 which is not placed on the PCB; R17 = 20kΩ results in $V_{LN-SLPE}$ = 4.3V.

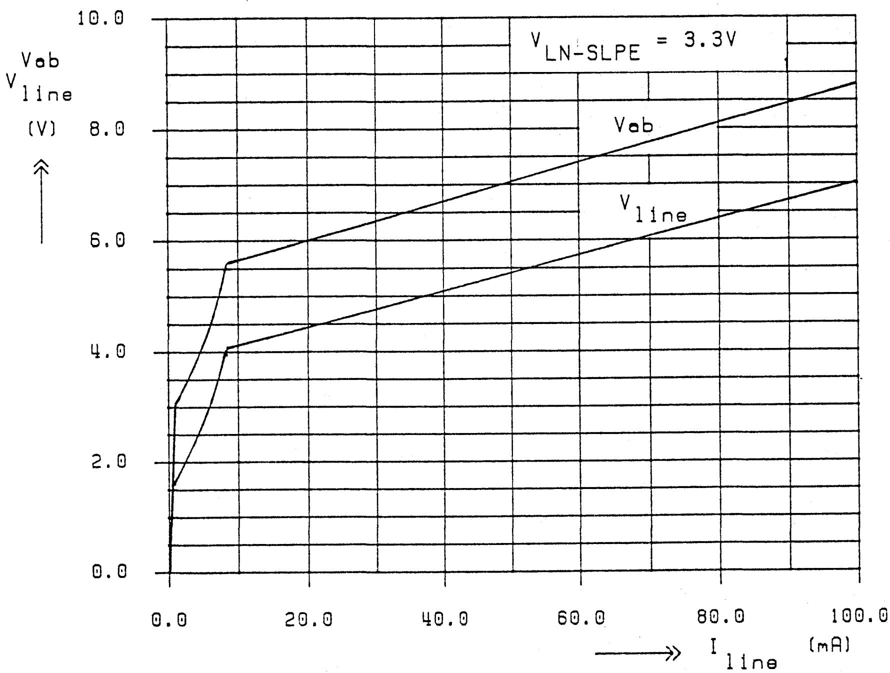


Fig.4 Line voltage as a function of the line current.

Fig.5 shows the maximum output voltage (v_{LSP} in V_{rms}) of the loudspeaker amplifier of the TEA1083A, at 5% harmonic distortion, as well as the generated power PO into the loudspeaker, both as a function of the line current.
The output voltage v_{LSP}, and the output power PO, are given with a load (R_{LSP}) of 50Ω between the PCB terminals LSP and SLPE. For other loudspeaker impedances (than 50Ω), see paragraph 3.2.2.

The output voltage is limited by the supply current up to $I_{line} \approx 17mA$, as shown in fig.5, while the signal is limited by VBB at $I_{line} > 17mA$. An output power of 10mW can be obtained at about 13mA line current.

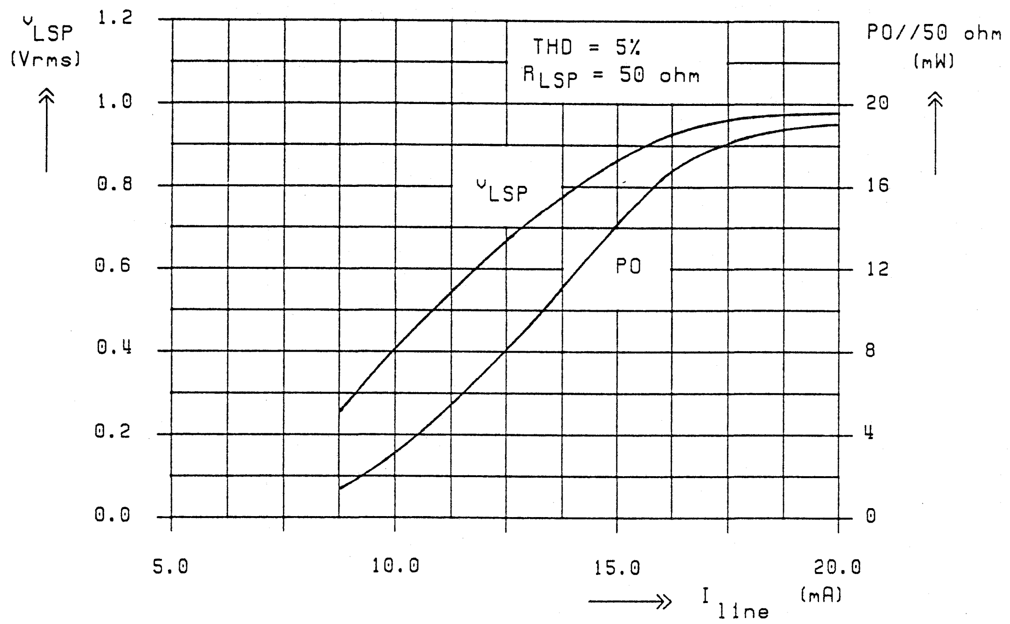


Fig.5. Maximum output voltage and output power versus I_{line}.

The influence of the efficiency of the supply of the TEA1083A is indicated in fig.6. The maximum output voltage of the loudspeaker amplifier is here shown as a function of the line voltage (continuous signal) at line currents from 10mA to 25mA and a voltage space of 850mV (typical) between SUP and VBB. This voltage difference is a result of the typical reference voltage of the TEA1064A, V_{int} and the supply voltage VBB of the TEA1083A.

The efficiency of the supply will decrease as soon as the momentary line voltage becomes less than the VBB + 250mV potential level. The efficiency will be maximum if:

$$v_{line-rms} \leq (V_{SUP-VBB} - 250mV) / \sqrt{2}.$$

The conditions of fig.6 are: THD $\leq$ 5%, $V_{LN-SLPE}$ = 3.3V and VBB = 2.95V. Larger signal levels at the LSI inputs, in order to increase v_{LSP}, result in higher distortion figures and/or a decrease of the VBB voltage; see chapter 3.1.

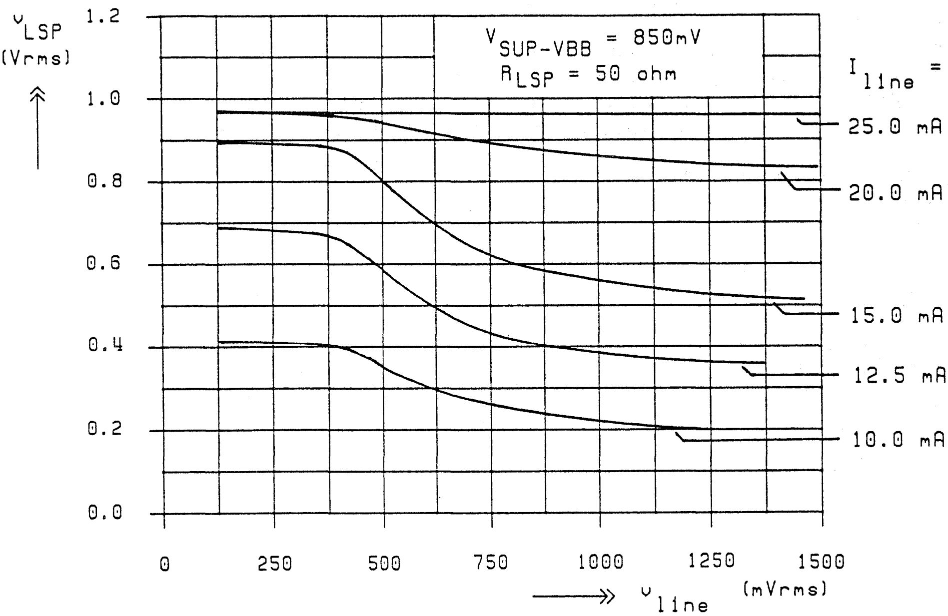


Fig.6. Maximum output voltage as a function of the line signal level.

2.5. Parallel operation / low line current behaviour.

At line currents below the internal threshold current (8mA typically) of the TEA1064A, the internal reference voltage $V_{LN-SLPE}$ is automatically adjusted to lower values, to facilitate operation of the application with more telephone sets in parallel (ref.3).
The speech part will not be disturbed by the TEA1083A application in that low voltage region, but sending gain, receiving gain and output swing capabilities of the TEA1064A (and TEA1067) are reduced according to the low voltage behaviour of these transmission ICs.
The line monitoring possibilities are rather limited in this region. The current to power the loudspeaker is reduced and the supply voltage VBB of the TEA1083A is decreased corresponding to the reduced voltage at the line and pin SUP of the TEA1083A.

For the TEA1060, TEA1061 and TEA1068, which are not equipped with a low voltage function, the line voltage of this monitoring application is not well defined for line currents less than about 8mA. The line voltage can show LF relaxations caused by the start circuit of the TEA1083A.

2.6. PD function / current reduction during supply breaks.

The TEA1083A as well as the TEA1064A are provided with a PD function for use in pulse dialling and register recall (timed loop break) applications.
The PD inputs of both IC are wired to the DP terminal which can be controlled by the dialler.
During line breaks the ICs are without continuous power and have to be supplied by the charge of buffer capacitors C1 and C20.
A 'HIGH' level applied at the DP terminal forces both ICs into the PD mode, thereby reducing the current consumption from the VCC1 and VBB supply points.
The current consumption during PD is specified in the data sheets of the TEA1083A and TEA1064A.

3. TEA1083A CALL PROGRESS MONITORING CIRCUIT.

3.1. Stabilised supply voltage VBB.

The internal circuitry of the TEA1083A is supplied by VBB, which is stabilised by an internal shunt stabilizer at a fixed level of 2.95V typically. This supply voltage, decoupled by C20 (220μF), can be used also for peripheral supply.

Note:
The loudspeaker amplifier of the TEA1083A contains no dynamic limiter. The gain is not reduced and the current consumption is not limited at low supply currents and overload conditions.
These conditions result in a decrease of the VBB voltage.

3.2. Loudspeaker amplifier.

The loudspeaker amplifier of the TEA1083A has symmetrical inputs LSI1 and LSI2. One of them is connected with the receiver output (QR+) of the transmission IC via an attenuator, while the other is connected with VEE (via C23); VEE is the reference of the receiver outputs QR+ and QR-.
The input impedance measured between the two inputs LSI1 and LSI2 is 19kΩ typical.

The loudspeaker has to be connected between the terminals LSP and SLPE.

3.2.1. Amplifier gain / volume control.

The gain between inputs and output is fixed to 35dB. Volume control of the total monitoring receive gain is obtained by attenuation of the input signal via R21 and potentiometer R22 (fig.2).

The value of R21 and R22 on the PCB are both 2.2kΩ. R21 has to be chosen in accordance with the earpiece sensitivity and the required maximum loudspeaker signal and maximum signal distortion.

The overall receive gain from line to loudspeaker output, at potentiometer R22 in maximum position, measures +22dB.
The adjusted gain on the PCB from line to QR+ (or QR-) output measures −7dB (R4 = 50.5kΩ), −13dB from line to LSI1 input (R22-max. = R21) and +22dB from line to the PCB terminal LSP.

3.2.2. Output voltage / signal distortion.

The output voltage (v_{LSP} in V_{rms}) and the signal distortion (THD in %) both as a function of the input level (at 1kHz) between LSI1 and LSI2, are shown in fig.7.
The output stage of the TEA1083A is designed for application with a 50Ω loudspeaker. The specified maximum output power is 16mW typically with a 50Ω loudspeaker (for example Philips type AD2071/Z50) at $I_{SUP} \geq$ 12mA.

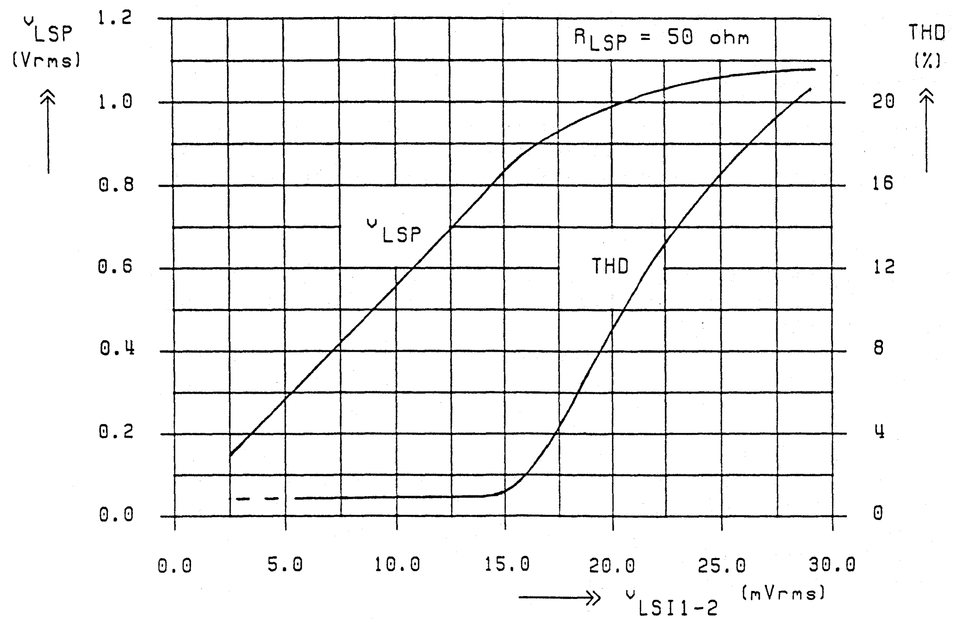


Fig.7 Output level and harmonic distortion versus input level.

The maximum signal level across the loudspeaker, and maximum power into the loudspeaker, depends on the maximum output current of the amplifier, VBB level, available supply current and loudspeaker impedance.
The limitations of the loudspeaker amplifier of the TEA1083A (and TEA1083) are:

- Maximum output current, from pin QLS, of about 70mA-peak.
- Maximum output swing at pin QLS, which is limited by the VBB voltage, measures approximately 1.4V-peak.
- Output level, depending on I_{SUP} and loudspeaker impedance R_{LSP}, of:

$$v_{LSP\text{-}rms} \approx (I_{SUP} - I_{SUPO}) * R_{LSP} * \pi / \sqrt{2}$$

The value of C21 has to be chosen in accordance with the loudspeaker impedance in order to keep the voltage drop across C21 low at low audio frequencies. The output impedance of the loudspeaker amplifier is < 2.5Ω.

3.3 Loudspeaker enable function.

The TEA1083A is provided with an enable input to enable or disable the loudspeaker amplifier. The LSE terminal is connected to a 'HIGH' level (VBB voltage) by means of jumper J2 to enable the amplifier. The loudspeaker amplifier is disabled when J2 is removed. LSE has to be 'LOW' (input open or connected to VSS) to disable the loudspeaker amplifier.

4. TEA1064A TRANSMISSION CIRCUIT.

4.1. Regulated $V_{LN-SLPE}$ voltage.

Fig.2 shows the application of the TEA1064A as transmission IC in the call progress monitoring application. Pin VCC2 is connected to pin LN to obtain a stabilised voltage between LN and SLPE. The voltage at terminal LN is:

$$V_{LN} = V_{LN-SLPE} + (I_{line} - I_{CC1}) * R9$$

The reference voltage $V_{LN-SLPE}$ is typically 3.3V, which can be increased by means of R17.

The value of R9 on the PCB is 20Ω. Changing R9 influences the microphone gain (chapter 4.2), DTMF gain (chapter 4.3), gain control characteristics (chapter 4.5), side tone level (chapter 4.6), threshold current of the low voltage function (chapter 2.5) and the maximum output swing on the line. If R9 has to be changed follow the adjustment procedure of the TEA1064A as given in ref.2 or 3.

An extra supply voltage VP (fig.2) for peripherals can be constructed between LINE and SLPE by means of an RC smoothing filter consisting of the components R23 and C19. Consult ref.3 to obtain the components values of C19 and R23 and for the application consequences. The PCB is prepared for this function, however R23 and C19 are not placed.

4.2. Microphone amplifier.

4.2.1. Amplifier inputs.

The printed circuit board is ready for use with microphone capsules which require a symmetrical input (dynamic, magnetic and piezo electric). The input impedance of the TEA1064A microphone amplifier is high (typ. 64kΩ); with R14 = 8.25kΩ (figure 2) a more low-ohmic termination is obtained. If an asymmetrical input is required (electret) the MIC- input is used as a signal input via C13. The MIC+ input should be connected to ground by C14.

The value of both capacitors is determined by the lower cut-off frequency required. For testing purposes only (low-ohmic drive) 2.2μF is a suitable value. The PCB is prepared to place these capacitors.

Since the TEA1064A has a microphone gain that can be set between 44 and 52 dB for R9 = 20Ω (see chapter 4.2.2.) an external attenuation network (or another type of transmission IC) is required if a sensitive type of microphone is used needing less than 44dB of gain.

4.2.2. Gain adjustment.

The gain of the microphone amplifier can be calculated using the following formula:

$$A_m = 1.356 * \left(\frac{R7 + 3.47 * 10^3}{R5 * R9}\right) * \frac{Ri * R_{line}}{Ri + R_{line}}$$

in which:
$Ri = R1 // R_p$; $R_p = 15.5k\Omega$
R_{line} is the load resistance at the line during the measurements (normally 600Ω).

Figure 8 shows the gain of the microphone amplifier as a function of R7 for different values of R9.

On the board the following resistors are mounted:
R1 = 619Ω; R5 = 3.65kΩ; R7 = 25.5kΩ; R9 = 20Ω. This results in a gain of 44dB (assuming R_{line} = 600Ω).

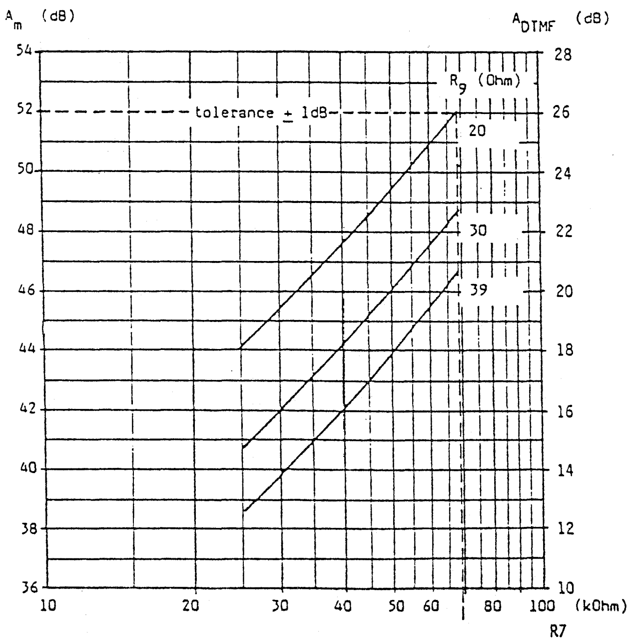


Fig.8. Microphone gain as a function of R7.

To ensure stability of the microphone amplifier capacitor C6 is connected between GAS1 and SLPE. C6 can be adjusted to obtain a first-order low-pass filter with time constant R7C6. On the board C6 = 100pF (f_{3dB} = 62kHz).

4.2.3. Microphone mute / dynamic limiter.

A logic low level at the DLS/MMUTE pin (e.g. via a 3.3kΩ resistor to V_{EE}) inhibits the microphone amplifier and has no influence on the receiver and DTMF amplifiers.
Removing the low level at the DLS/MMUTE pin provides the normal function of the microphone amplifier after a time proportional to C16.

To prevent distortion of the transmitted microphone signal a dynamic limiter is incorporated in the microphone amplifier. The use of such a limiter also improves the sidetone performance considerably (less distortion and a limited sidetone level in overdrive conditions).

When peaks of the transmitted signal on the line exceed an internally determined threshold the gain of the transmitter amplifier is reduced rapidly. The time in which the gain reduction is realized (the attack time t_{att}) is very short. The circuit stays in the gain-reduced condition until the peaks of the transmitted signal remain below the threshold level. The gain then returns to its normal value after the release time t_{rel}.

Both the attack and the release time are proportional to the value of the capacitor C16 connected to pin DLS of the TEA1064A. The recommended value for C16 is 470nF (also mounted on the board).

4.3. DTMF amplifier.

By applying a HIGH voltage to the MUTE input of the TEA1064A, the device is switched over from the speech mode to the dialling mode. In this mode DTMF signals applied to the DTMF input are transmitted to the line. The microphone and receiver amplifiers are blocked now and the dynamic limiter will not be activated.
The gain of the DTMF amplifier is:

$$20 * {}^{10}\log A_{DTMF} = 20 * {}^{10}\log A_m - 26\text{dB} \qquad \text{(See fig.8)}$$

Once the microphone gain has been adjusted, the DTMF gain is fixed. This means that the DTMF input level must be adjusted to the correct value by means of an external attenuator at the DTMF input.

4.4. Receiver amplifier.

4.4.1. Amplifier outputs.

Earpieces with an impedance up to Z_T = 450Ω must be driven in single-ended mode. For impedances larger than 450Ω differential drive is possible.

The maximum output swing on the earpiece depends on the DC voltage at VCC1, and therefore also on the adjusted voltage drop (V_{LN-VEE}) over the circuit.

4.4.2. Gain adjustment.

The gain of the receiver amplifier (from IR to QR) can be calculated by means of:

$$\text{single-ended drive:} \quad A_{TA} = 1.314 * \frac{R4}{R5} * \frac{Z_T}{Z_T + 4},$$

$$\text{differential drive:} \quad A_{TS} = 2.628 * \frac{R4}{R5} * \frac{Z_T}{Z_T + 8},$$

in which: Z_T = earpiece impedance (in Ω).

Figure 9 shows the receiver gain as a function of R4.

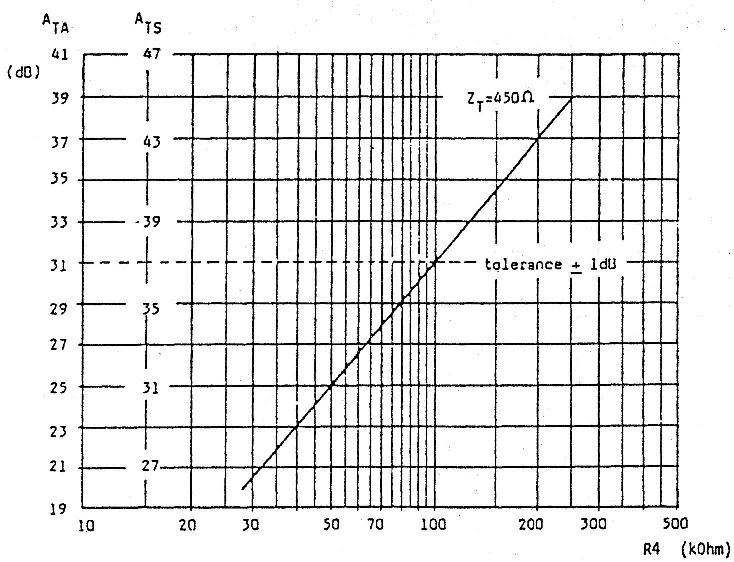


Fig.9. Gain of the receiver amplifier as a function of R4.

On the board R4 = 50.5kΩ resulting in a gain A_{TA} of 25dB. The total receive gain from line to earpiece output QR+ can be obtained by subtracting the attenuation of the anti-sidetone network (32dB with the component values of fig.2) from A_{TA}, resulting in an overall gain of −7dB.

C4 and C7 (fig.2) prevent parasitic oscillations of the receiver amplifier while C4 can also be adjusted to obtain a first-order low-pass filter with time constant R4C4. The value of C7 should be 10 times C4.

A high-pass filter is formed by C5 (together with the input impedance of pin IR) and by C2 (together with the earpiece impedance).

On the board the following components are mounted: C2 = 10μF, C4 = 100pF, C5 = 100nF and C7 = 1nF. The resistors R18 and R19 (both 22Ω) are mounted on the PCB to restore the decrease of phase margin of the receiver amplifiers caused by the capacitive load of C32 and C33.

4.4.3. Confidence tone.

During DTMF dialling the dialling tones can be heard at a low level in the earpiece. For both single-ended and differential drive the gain from the DTMF input to the receiver output, A_{CT}, is given by (fig.9):

$$20 * {}^{10}\log A_{CT} = 20 * {}^{10}\log A_T - 49dB.$$

4.5. Automatic gain control.

The line-loss compensation circuit of the TEA1064A regulates the microphone and receiver gain (ΔA_{vd}) with the line current (line length). The DTMF amplifier is not affected by the circuit.
The gain control range is 6dB corresponding with a 5km line of 0.5mm diameter copper twisted pair (DC resistance 176Ω/km, average AC attenuation 1.2dB/km).

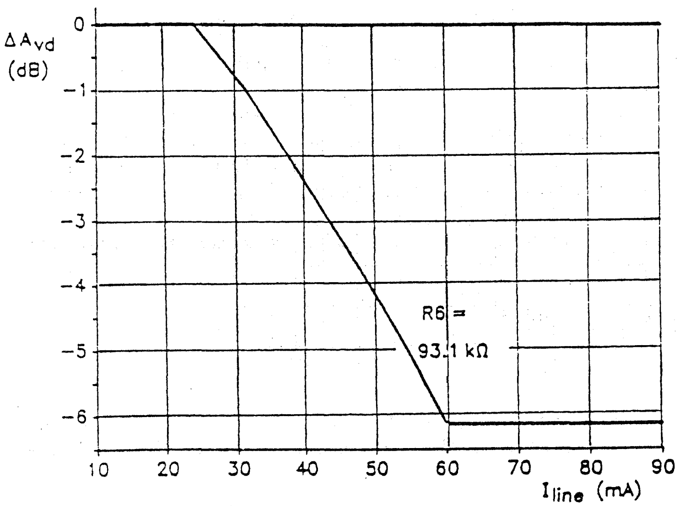


Fig.10. Control of the microphone and receiver gain versus I_{line}.

If no automatic gain control is required jumper J1 on the printed circuit board has to be removed. Resistor R6 must be chosen in accordance with the exchange supply voltage and its feeding bridge resistor.
Figure 10 shows the gain-control characteristics with R6 = 93.1kΩ and R9 =

20Ω optimised for a 600Ω feeding bridge and an exchange supply voltage of 48V; ΔA_{vd} is the reduction of the microphone and the receiver gain.

4.6. Anti-sidetone circuit.

On the PCB the TEA1060-family bridge is used to suppress the sidetone. For details of this bridge refer to ref.3.
Maximum suppression is obtained when the following two conditions are fulfilled:

a) $R9 * R2 = R_i * (R3 + R8)$ where $R_i = R_1//R_p$ and $R_p = 15.5k\Omega$

b) $$Z_{bal} = \frac{R3 * (R8 + R9)}{R2 * R9} * Z_{line} = k * Z_{line}$$

Normally R1 and R9 are fixed and R2, R3, and R8 can be chosen by the designer. These resistors have to be chosen to meet the following criteria:

- compatibility with a standard capacitor from the E6 or E12 range for the capacitor used in Z_{bal}
- $|Z_{bal}//R8| << R3$ to avoid influence of Z_{bal} on the receiver gain
- $|Z_{bal} + R8| >> R9$ to avoid influence on the microphone gain
- $(R3 + |R8//Z_{bal}|) << 20k\Omega$ to avoid the influence of the input impedance at IR (21 kΩ ± 4kΩ) on the bridge attenuation.

In practice Z_{line} varies strongly with line length and type of cable. Therefore an average value has to be chosen for Z_{bal}.
On the printed circuit board Z_{bal} is optimised for a cable length of 5km with a diameter of 0.5mm copper with a DC resistance of 176Ω/km and a capacitance of 38nF/km resulting in an average AC attenuation of 1.2dB/km.
The anti-sidetone network attenuates the signal from the line to the input of the receiver amplifier with 32dB.
For compatibility of the capacitor value in Z_{bal} with a standard capacitor from the E6 range (220nF):

$$k = \frac{140nF}{220nF} = 0.64$$

For R3 a value of 3.92kΩ has been chosen. So Z_{bal}, R8, and R2 can be calculated resulting in the practical values R11 = 130Ω, R12 = 825Ω, C12 = 220nF, R8 = 392Ω, and R2 = 130kΩ.

In case of complex set impedance (chapter 4.7) the bridge can be rebalanced by readjusting the values of R8 and Z_{bal}, and either R2 and R9 (see ref.4). Changing R2 and R9 has also consequences on other set parameters.
R2 determines the attenuation from line to receiver amplifier input IR, while changing R9 influences microphone gain, dtmf gain, etc. as indicated in chapter 4.1.

4.7. Set impedance.

The set impedance of the PCB application is provided by R1 (619Ω). Complex set impedance, consisting of Ra + Rb//Cp, can be obtained by means of R1, R13 and C36 (on the PCB: R13 = 0Ω; C36 is not mounted). Changing R1 has consequences for sidetone behaviour (chapter 4.6), while the value of the DC resistance of the complex impedance influences the voltage at terminal VCC1 which can influences the transmission properties (ref.3).
Complex set impedance results in a frequency dependent transmission gain.

5. PROTECTION.

Protection is achieved by means of a break over diode (BR211/240V) between the a/b lines in combination with R10 (12Ω) and zener diode D5 (12V) between LINE and VEE. The protection is meant for application with pulse dialling and DTMF dialling.

6. ELECTROMAGNETIC COMPATIBILITY.

EMC components C17, C18, C30, C31, C32, C33, C34 and C35 are mounted on the PCB to improve the EMC behaviour (reduction AM detection sensitivity) of the speech as well as of the call monitoring circuit part. The EMC behaviour is shown in fig.11, 12 and 13.

Measured are the demodulated signal levels at the a/b terminals (fig.11), across the earpiece (fig.12) and across the loudspeaker (fig.13). The -26dB and -40dB S/I (signal interference) lines are referred to a mean speech signal level of 100mV at the telephone line.
The EMC requirements are -26dB or -40dB, depending on the national PTT organization.

The measurement conditions are as follows:

- The demonstration board (PR4535X) is carried out with double sided wiring and two ground reference layers both connected to VEE of the TEA1064A.
- I_{line} = 27mA, microphone (200Ω) and earpiece (200Ω) both symmetrically connected.
- Position of R22: gain from line to LSP terminal is 10dB.
- Current injection method (common mode on a/b-b/a terminals).
- RF levels: 3Vrms at 100kHz-30MHz and 0.5Vrms at 30MHz-150MHz.
- Amplitude modulation: 80%, 1kHz.
- Test set up as explained in German requirements VDE 0878 (ref.5).

Remark: Avoid long connection wires and test equipment connected to the SLPE terminal(s). They worsen the EMC behaviour and can give instability of the loudspeaker amplifier.

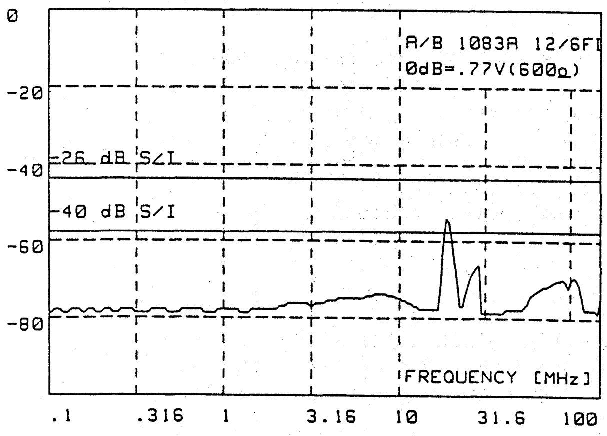


Fig.11 Demodulated signal level across the a/b-b/a terminals versus frequency.

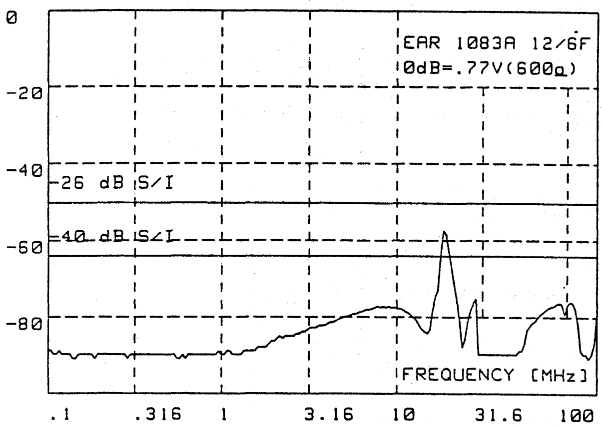


Fig.12 Demodulated signal level across the earpiece versus frequency.

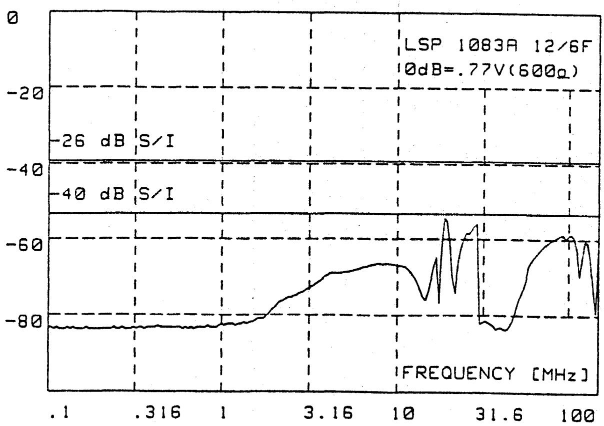


Fig.13 Demodulated signal level across the loudspeaker versus frequency.

7. REFERENCES.

1) - Preliminary data sheet TEA1083/TEA1083A "Call progress monitoring circuit for line powered telephone sets". F.v.Dongen.
Report number ETT/AN91008 (June 1991).

2) - Documentation for the TEA1064A printed circuit board CAB3458.
P.Biermans / P.J.M.Sijbers / K.Wortel.
Report number ETT/AN90001.

3) - PCALE laboratory report, "Application of the versatile speech transmission circuit TEA1064 in full electronic telephone sets", F.v.Dongen / P.J.M.Sijbers.
Report number ETT89009, (12nc. 9398 068 80011).
(Application report TEA1064A; to be published).

4) - Designers' Guide: "TEA1060-family - Versatile Speech/Transmission ICs for Electronic Telephone Sets", P.J.M.Sijbers.
(12nc. 9398 341 10011).

5) - VDE 0878, part 1 and part 200; 1987.

6) - Test procedure of the PR4535X: ETT/IR91042.

APPENDIX 1: PCB MODIFICATION FOR TEA1083 AND/OR TEA106X.

This appendix gives the modifications of the PCB when the TEA1083A has to be replaced by the TEA1083 and/or the TEA1064A has to be replaced by the TEA106X (TEA1060, TEA1061, TEA1067 or TEA1068).

TEA1083:

The PCB is prepared for use with the TEA1083. It can be mounted next to the 16-pins IC-socket of the TEA1083A as shown in fig.A1, if necessary, via an 8-pins IC-socket.
The DP terminal is connected to the TEA1064A only in this case, because the TEA1083 has no PD input.
No further modifications are required.

TEA1060, TEA1061, TEA1067 or TEA1068:

The TEA106X can be placed in the 18-pins shadow IC socket, which is mounted next to the 20 pins socket of the TEA1064A, as shown in fig.A1.

The following PCB modifications have to be carried out to obtain the same gain and AGC settings as with the TEA1064A (see also ref.4):

TEA1060:	receive gain	adapt R4 = 100kΩ.
	automatic gain control . . .	adapt R6 = 100kΩ.
TEA1061:	microphone gain	adapt R7 = 140kΩ.
	receive gain	adapt R4 = 100kΩ.
	automatic gain control . . .	adapt R6 = 100kΩ.
TEA1067:	automatic gain control . . .	adapt R6 = 100kΩ.
TEA1068:	receive gain	adapt R4 = 100kΩ.
	automatic gain control . . .	adapt R6 = 100kΩ.

TEA1060/1061/1067/1068 (TEA106X):

For adaption of the line voltage is referred to the documentation of the concerned devices.

Diode D6 is not mounted and bridged by a short connection, as shown in fig.2, because of the common reference (SLPE) of the PD inputs of the TEA1064A and the TEA1083A. This short connection has to be removed, and D6 (BAS11) has to be placed in case the PD inputs are controlled by a peripheral which is referred to VEE.
This modification is not useful in case the TEA1083 is applied.

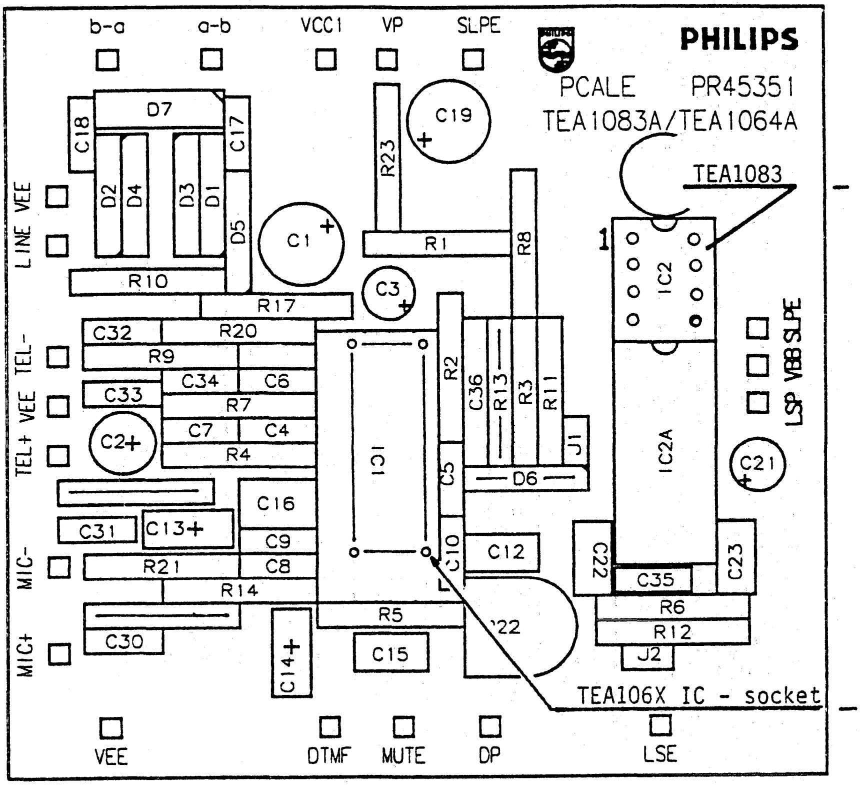


Fig.A1 View of the components side of the PCB showing the mounting place of the TEA1083 and the shadow socket of the TEA106X.

APPLICATION NOTE Nr	ETT/AN91012
TITLE	User's Manual PR4516X DEMO board with TEA1085A -TEA1064A listening in application
AUTHOR	F. van Dongen
DATE	July 1991

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1. INTRODUCTION.

The printed circuit board (PCB), with identification number PR4516X, contains the application of the listening-in circuit TEA1085A combined with the transmission circuit TEA1064A.

The TEA1085A offers a listening-in facility of the received line signal via the connected loudspeaker. It has to be applied in conjunction with a transmission circuit of the TEA1060-family which serves for the common line interface for speech as well as for dialling.

The TEA1085A incorporates a start circuit, supply circuit, loudspeaker amplifier, logic inputs for gain setting, dynamic limiter, power down circuit, mute facility and moreover a "Larsen Level Limiter" to prevent annoying howling effects.

This PCB is described for and delivered with the TEA1085A and TEA1064A. The TEA1085A may be replaced by the TEA1085; see the notes in this chapter concerning the differences between TEA1085A and TEA1085.
The TEA1064A (20-pins) may be replaced by the TEA1060, TEA1061, TEA1067 or TEA1068 for which a 18 pins shadow IC socket is mounted on the board.
Modification instructions are summarized in appendix 2.

Chapter 2 describes the application of the TEA1085A combined with the TEA1064A with respect to line current and voltage, while chapters 3 and 4 describe the two ICs separately.

Detailed information about the TEA1085A and the TEA1064A can be found in the documentation as given in the references of chapter 7. For the TEA1064A is referred to the application report of the TEA1064 which for most of the information is also valid for the TEA1064A. However, differences are present with respect to AGC characteristics, AC start up time, and the presence of a microphone mute possibility.

Notes:
- The difference between the TEA1085A and TEA1085 concerns the MUTE facility.
  - TEA1085A: the MUTE is provided with a logic input. A LOW level at the MUTE input (or open input) corresponds to the standby condition of the loudspeaker amplifier. The amplifier is enabled (listening-in condition) at MUTE = HIGH.
  - TEA1085: the MUTE is provided with a toggle input. Switching over, between standby and listening-in condition of the loudspeaker amplifier, occurs at the positive edges of the MUTE input signal.
- The MUTE-pin of the IC is wired to the PCB terminal LSE (loudspeaker enable).

2. LISTENING-IN COMBINATION OF THE TEA1085A AND TEA1064A.

Fig.1 shows the block diagram of the listening-in combination on the PCB. As shown, the TEA1085A is connected between the positive line terminal and pin SLPE of the TEA1064A. The transmission characteristics such as set impedance, gain settings, gain control etc. are not affected.
The TEA1064A application on the PCB is according to the regulated line voltage mode. The regulated supply voltage mode of the TEA1064A is not taken into account due to the presence of a stabilised voltage VBB of the TEA1085A.

The complete circuit diagram of the application is given in fig.2 while the components side of the demonstration board is shown in fig.3.
The interfaces between PCB application, transducers, dialler and a/b lines are indicated.

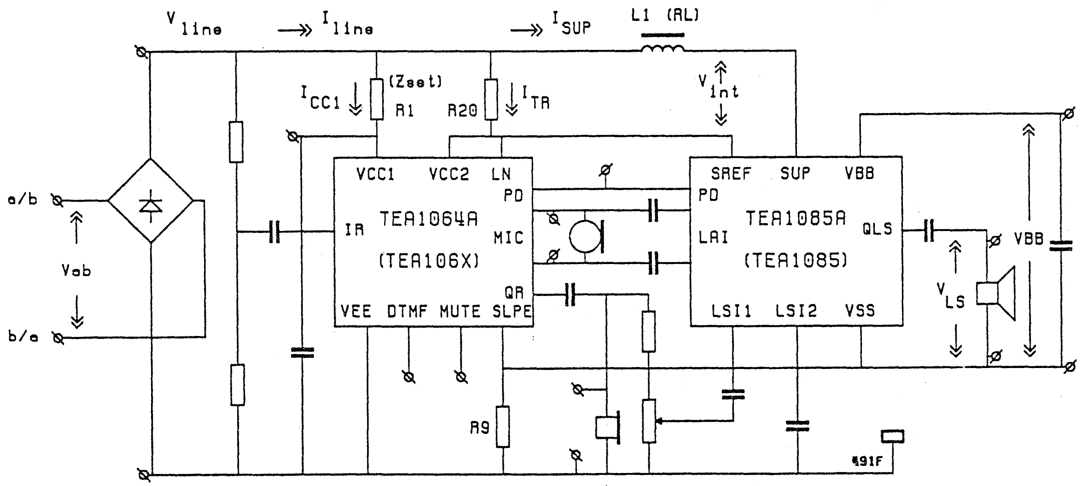

Fig.1 Block diagram of the listening-in application on PCB PR4516X.

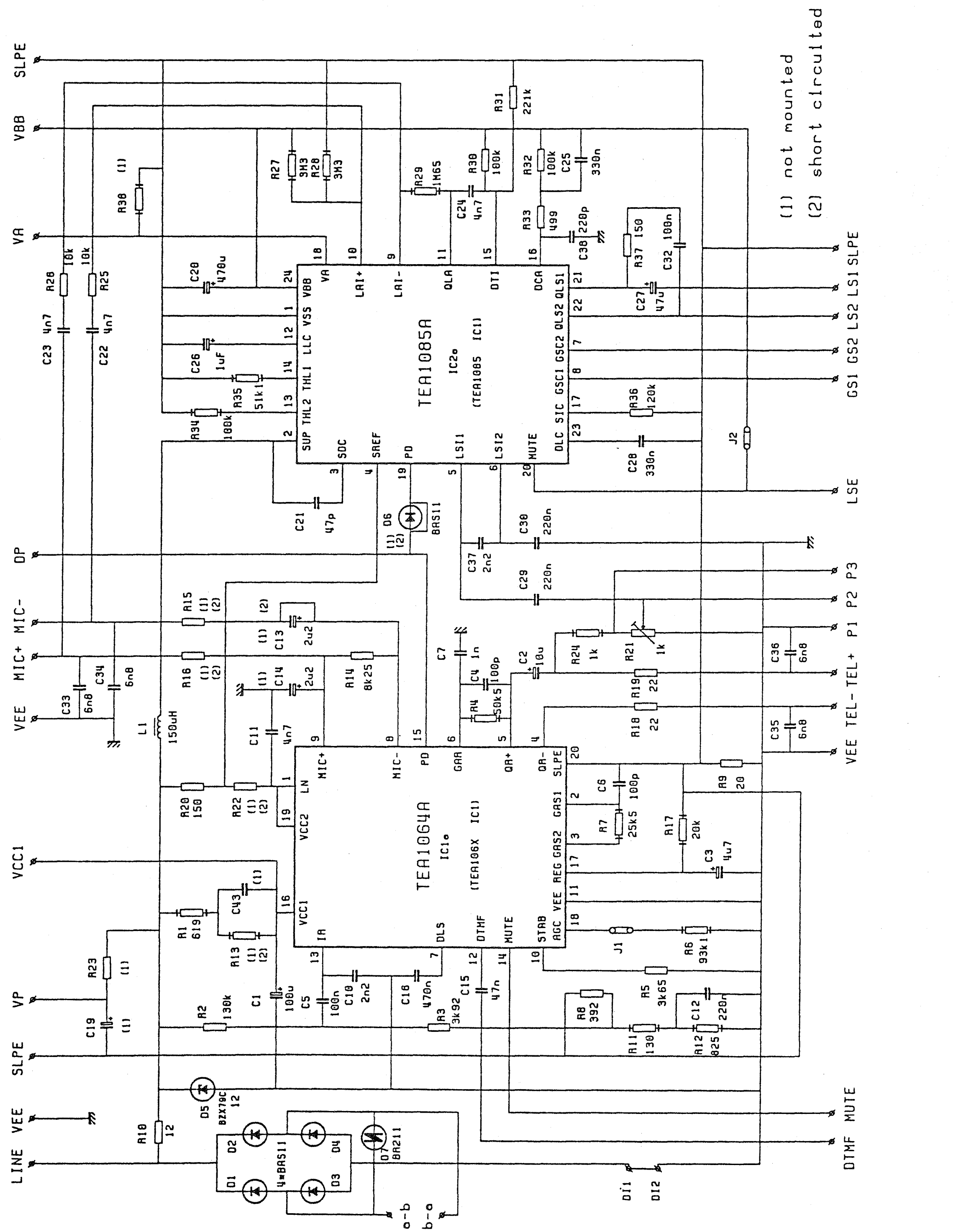

Fig.2 Application diagram of the TEA1085A / TEA1064A combination.

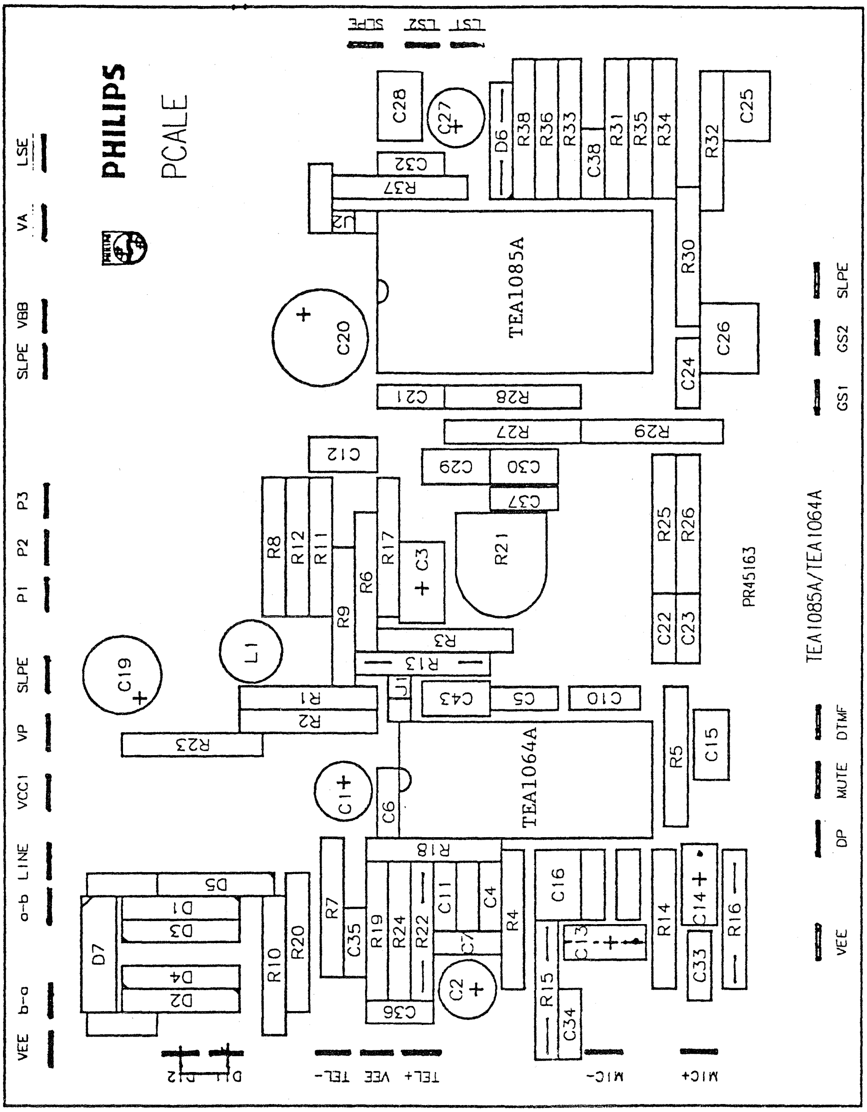

Fig.3 View of the components side of the demonstration board.

Remarks:
- R15, R16, C13, C14, C19, R23, C43, R13, R22, R38 and D6 are not mounted.
- R15, R16, C13, R13, R22 and D6 are short circuited.

2.1. How to use this demonstration board.

The PCB terminals to connect the line, transducers, DTMF and logic signals and the terminal to adjust the VBB voltage are drawn in fig.3.

The demonstration board can be used after connection of the line supply to the a/b-b/a terminals and a loudspeaker to the LS1-SLPE or LS1-LS2 terminals. An audio signal on the line can be heard through the loudspeaker. The level of the loudspeaker signal can be controlled by means of the potentiometer, or if necessary, in combination with the logic levels at the GS1 and GS2 terminals.
Terminal LSE (loudspeaker enable) is interconnected with the VBB voltage level, by means of jumper J2. The listening-in function is disabled when J2 is removed.

A complete telephone set can be build by connecting a handset, ringer and dialler. The board is assembled for use with a dynamic microphone; the connection terminals are MIC- and MIC+. It has to be modified for applications with an asymmetrical microphone. A symmetrical or an asymmetrical earpiece can be connected between the TEL- and TEL+ terminals or between the TEL+ and VEE terminals.

The DTMF terminal can be used to transmit a DTMF signal to the line by applying a 'HIGH' level to the MUTE terminal. The DTMF dialler can be supplied via the terminals VP-SLPE (see chapter 4.1).

Terminal DP is available for pulse dial applications. A 'HIGH' level at the DP terminal reduces the current consumption of the TEA1064A and the TEA1085A. An interrupter can be connected between DI1 and DI2.
DI1 and DI2 are interconnected on the PCB.

The PCB is designed in such a way that several parameters can easily be adapted. A brief description of how to change them is given in the following paragraphs. An adjustment procedure for the application is given in appendix 1.

2.2. Peripheral supply / logic inputs.

The TEA1085A can supply peripheral circuits between the stabilised supply voltage VBB (typ. 3.6 V) and VSS. The ground reference for peripherals is terminal SLPE; SLPE of the TEA1064A is interconnected with VSS of the TEA1085A. The logic inputs of the TEA1085A (GSC1, GSC2, PD and MUTE) and the inputs of the TEA1064A (PD, MUTE and DTMF) are all referred to SLPE. The Power Down inputs of both ICs have the same structure and can be controlled in parallel from the same source, via terminal DP.

The MUTE input of the TEA1085A (LSE terminal on the PCB !) and TEA1064A have the same pin-name, but have a different function. They should not be controlled from the same source.

For the other members of the TEA1060-family, terminal VEE is the reference for the MUTE, PD and DTMF inputs. Interface circuitry can be required between TEA1085A, TEA106X and peripherals depending on the supply of the peripherals. See appendix 2 for application adaptations when the TEA1064A is replaced by another IC of the TEA1060-family.

Circuits without connections to either PD, MUTE or DTMF and with a relatively low current consumption like electret microphones can be powered between VCC1 and VEE of the TEA1064A.

2.3. Start-up / line current split-up.

After going off-hook, the internal start circuit of the TEA1085A ensures normal start-up of the TEA1064A whereby the capacitors on pins DLC, LLC and VBB are charged quickly during this start up phase.
At 20mA line current the TEA1064A is available for speech functions within 60ms, while the VBB capacitor (C20 = 470µF) reaches its typical voltage of 3.6V within 300ms when no extra peripheral current is consumed from VBB. The listening-in function can be used within 400ms; the LSE terminal has to be 'HIGH' in this case.

As shown in fig.1 the line current is split up into the currents I_{TR} and I_{CC1} (for the TEA1064A) and I_{SUP} which is available for the listening-in function of the TEA1085A. These currents are, taking into account the DC resistance of coil L1 (RL):

$$I_{TR} = (I_{line} - I_{CC1}) * \frac{RL}{R20 + RL} + \frac{V_{int}}{R20 + RL}$$

$$I_{SUP} = (I_{line} - I_{CC1}) * \frac{R20}{R20 + RL} - \frac{V_{int}}{R20 + RL}$$

where:
- I_{TR} is the current through R20 to bias the output stage/voltage stabilizer of the TEA1064A.
- I_{SUP} is the current into SUP available for the listening-in part.
- I_{CC1} is the current consumed from V_{CC1} to supply the internal circuitry of the TEA1064A (1.5 mA typically) and the peripheral connected to V_{CC1}.
- V_{int} is an internal bias voltage of the TEA1085A between SUP and SREF (315mV typical)
- R20 on the PCB is 150Ω.

The DC resistance of the applied coil on the PCB is 1.6Ω typically. In this case RL << R20 which results in:

$$I_{TR} \approx 2mA \quad \text{and} \quad I_{SUP} \approx I_{line} - I_{CC1} - 2mA \ .$$

Only a part of I_{SUP} is available to power the loudspeaker. In the TEA1085A I_{SUP} is used to supply the internal circuitry (I_{SUPO} = 4.2mA), while the remaining current $I_{LI\text{-}MAX} = I_{SUP} - I_{SUPO}$ is available for powering the loudspeaker.
However, the current which can be applied to power the loudspeaker is in practice less than $I_{LI\text{-}MAX}$ due to the limited efficiency of the supply of the TEA1085A and the sending modulation current which flows from SUP to

SLPE, through the TEA1085A. This will be explained in more detail in ref.6. See also the next paragraph and fig.6.

2.4. Line voltage / maximum output power.

The voltage drop over the application, without diode bridge, is given by:

$$V_{line} = V_{LN\text{-}SLPE} + V_{int} + I_{line} * R10 + (I_{line} - I_{CC1}) * R9$$

in which:
- R9 = 20Ω, R10 = 12Ω (on the PCB).
- $V_{LN\text{-}SLPE}$ is the reference voltage of the TEA1064A.

The reference voltage $V_{LN\text{-}SPLE}$ of the TEA1064A, which is typically 3.3V, has to be increased by means of R17.
Fig.4 shows the line voltage V_{line} as function of I_{line}: the reference voltage ($V_{LN\text{-}SLPE}$) of the TEA1064A has been set at respectively 4.3V, 4.0V and 3.7V (R17 is respectively 18kΩ, 24kΩ and 39kΩ). The value of R17 on the PCB is 20kΩ ($V_{LN\text{-}SLPE}$ = 4.2V typically).

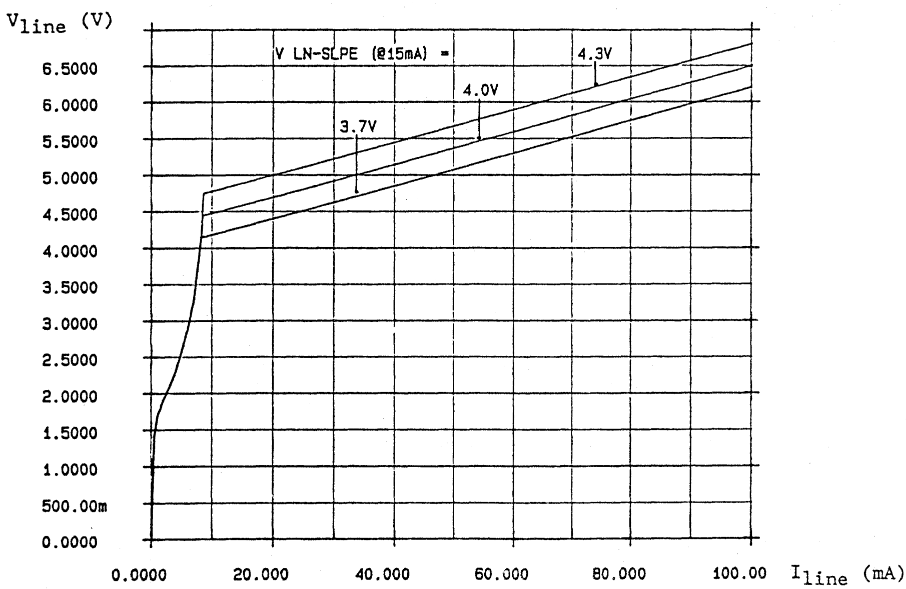

Fig.4 Line voltage V_{line} as a function of line current I_{line}.

Fig.5 shows the maximum output voltage (peak-peak) of the loudspeaker amplifier of the TEA1085A as a function of I_{line} at VBB = 3.6V and $V_{LN\text{-}SLPE}$ ≥ 3.7V.
The loudspeaker (50Ω) is connected as a single ended load (SEL) or as a bridge tied load (BTL) between PCB terminals LS1-SLPE, respectively LS1-LS2.

For SEL drive, the output voltage is limited by the current limiter up to $I_{line} \approx 20mA$, as shown in fig.5, while the peak limiter will be activated at $I_{line} > 20mA$. An output power of 20mW can be obtained at about 19mA line current for SEL loudspeaker drive.
For BTL loudspeaker drive, the current limiter controls the output voltage up to 45mA line current. The required minimum line current in this case is about 28mA to obtain the same 20mW output power.

The electrical maximum output power in SEL mode is about 20mW and about 65mW in BTL mode, both at VBB = 3.6V and 50Ω load.

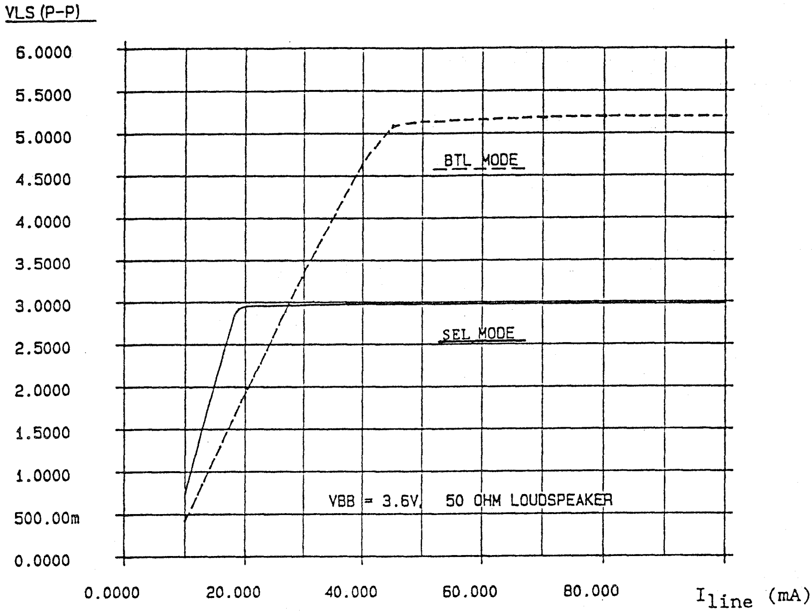

Fig.5. Maximum output voltage (peak-peak) of the loudspeaker amplifier as a function of I_{line}.

The influence of the efficiency of the supply of the TEA1085A is indicated in fig.6. The max. output voltage of the loudspeaker amplifier is here shown as a function of the DC voltage margin between SUP and VBB ($V_{SUP-VBB}$) for SEL and BTL mode at 20mA respectively 50mA line current. The signal levels at the line are $100mV_{rms}$ or $300mV_{rms}$ continuously, while the output load for both modes, SEL as well as BTL, is 50Ω.
The efficiency of the supply will decrease as soon as the momentary line voltage becomes less than the VBB + 0.25V potential level. The efficiency will be maximum if:

$$v_{line-peak} \leq (V_{SUP-VBB} - 0.25V)$$

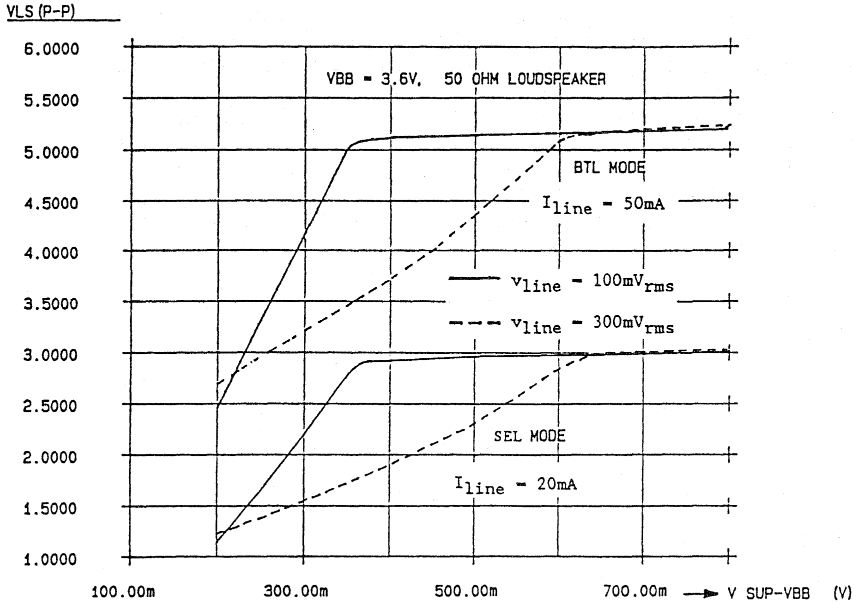

Fig.6. Maximum output voltage (peak-peak) as a function of $V_{SUP-VBB}$.

2.5. Parallel operation / low line current behaviour.

At line currents below the internal threshold current (8mA typically) of the TEA1064A, the internal reference voltage $V_{LN-SLPE}$ is automatically adjusted to lower values, to facilitate operation of the application with more telephone sets in parallel (ref.3).
The speech part will not be disturbed by the TEA1085A application in that low voltage region, but sending gain, receiving gain and output swing capabilities of the TEA1064A (and TEA1067) are reduced according to the low voltage behaviour of these transmission ICs.
The listening-in possibilities are rather limited in this region. The current to power the loudspeaker is reduced and the supply voltage VBB of the TEA1085A is decreased corresponding to the reduced voltage at the LINE terminal and pin SUP of the TEA1085A.

For the TEA1060, TEA1061 and TEA1068, which are not equipped with a low voltage function, the line voltage of this listening-in application is not well defined for line currents less than about 8mA. The line voltage can show LF relaxations caused by the start circuit of the TEA1085A.

2.6. PD function / current reduction during supply breaks.

The TEA1085A as well as the TEA1064A are provided with a PD function for use in pulse dialling and register recall (timed loop break) applications.
The PD inputs of both IC are wired to the DP terminal which can be controlled by the dialler.
During line breaks the ICs are without continuous power and have to be supplied by the charge of buffer capacitors C1 and C20.
A 'HIGH' level applied at the DP terminal forces both ICs into the PD mode, thereby reducing the current consumption from the VCC1 and VBB supply points.
The current consumption during PD is specified in the data sheets of the TEA1085A and TEA1064A.

3. TEA1085A LISTENING-IN CIRCUIT.

3.1. Stabilised supply voltage VBB.

The internal circuitry of the TEA1085A is supplied by VBB, which is stabilised by an internal shunt stabilizer at 3.6V typically. VBB, which is decoupled by C20 (470μF), can be used also for peripheral supply.

The stabilised VBB voltage can be increased by means of R38 (not mounted). The maximum voltage is restricted to 6.0V. Fig.7 gives the typical VBB voltage as a function of R38. Increase of VBB means a decrease of the voltage difference between SUP and VBB which can decrease the efficiency of the supply (as explained in chapter 2.4), depending on the mean level of the speech signal on the line.

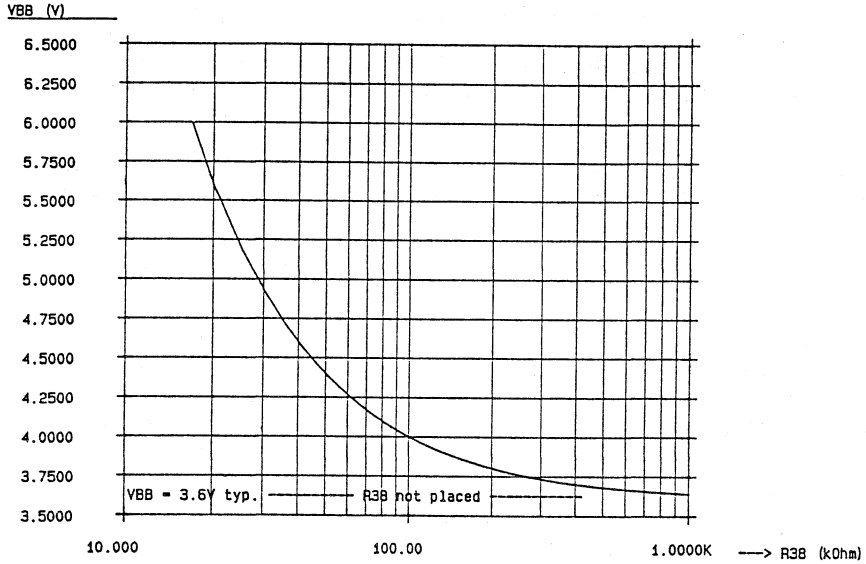

Fig.7. VBB voltage setting by means of R38.

A reduction of the VBB voltage is possible by means of a resistance between the terminals VA and VBB; the minimum level is restricted to 3.0V.

A modification of the VBB voltage means a change in the DC setting of the detector of the Larsen Level Limiter. R32 has to be adapted according chapter 3.4.2.

3.2. Loudspeaker amplifier.

The loudspeaker amplifier of the TEA1085A has symmetrical inputs LSI1 and LSI2. One of them is connected with the receiver output (QR+) of the transmission IC via an attenuator, while the other is connected with VEE (via C30) because VEE is the reference of the receiver outputs.

The input impedance measured between the two inputs LSI1 and LSI2 is 19kΩ typical.
The PCB terminals LS1, SLPE and LS2 can be used for connection of the loudspeaker as a SEL (single ended load) or as a BTL (bridge tied load).

3.2.1. Amplifier gain / gain select.

The gain between inputs and output is fixed to 35dB for SEL mode and 41dB for BTL mode. Volume control of the total listening-in receive gain is obtained by attenuation of the input signal via R24 and potentiometer R21 (fig.2) or by three fixed intervals of 6dB via the gain select PCB terminals GS1 and GS2 according table 1.

The value of R21 and R24 has to be in accordance with the receive gain of the transmission IC, which depends on the sensitivity of the earpiece.
The value of R21 and R24 on the PCB are both 1kΩ. R24 has to be chosen in such a way that for a mean speech level at the line, the signal across R21 is $18mV_{rms}$ which corresponds with a loudspeaker signal of $1V_{rms}$ at SEL loudspeaker drive.

The overall receive gain from line to loudspeaker output (without gain select and potentiometer R21 in maximum position) is +22dB for SEL and +28dB for BTL mode.
The adjusted gain on the PCB from line to QR+ (or QR-) output measures −7dB (R4 = 50.5kΩ), −13dB from line to LSI1 input (R24 = R21-max.) and +22dB from line to PCB terminal LS1 or +28dB to terminals LS1 and LS2.

An external potentiometer can be connected on the PCB terminals P1 up to P3. The preset potentiometer R21 on the PCB has to be removed in this case.

A combined control via potentiometer and GS1/GS2 terminals is also possible. The GS terminals can be left open or connected to SLPE when gain select is not used.

GS2	GS1	gain(dB)	gain reduction(dB)
0	0	35	0
0	1	29	6
1	0	23	12
1	1	17	18

Table 1.

Gain reduction via the GS1/GS2 terminals.

"0" = SLPE level
"1" = VBB level

3.2.2. Output power.

The output stages of the TEA1085A are optimised for application with a 50Ω loudspeaker. In SEL mode the output power is 20mW typically for a 50Ω loudspeaker (for example Philips type AD2071/Z50) at VBB = 3.6V nominally and at I_{line} > 19mA; see also fig.5.
The PCB is provided with series capacitor C27 between IC output QLS1 and PCB terminal LS1. The loudspeaker has to be connected between the PCB terminals LS1 and SLPE for SEL mode.

An output power of 40mW can be obtained at I_{line} > 35mA with the loudspeaker (50Ω) connected between the two PCB terminals LS1 and LS2 (BTL-mode). The series capacitor C27 is recommended for BTL applications in case of low line currents.
The available supply current for the power amplifiers can be reduced because of the offset current flowing between LS1 and LS2 (without C27), which can reduce the listening-in performances.

3.2.3. Dynamic limiter.

The TEA1085A uses an internal dynamic limiter to prevent distortion of the loudspeaker signal by reducing the amplifier gain under control of the peak limiter, VBB limiter or current limiter.

The peak limiter reacts when peaks of the output signal exceed internally determined thresholds and stays in the gain reduced condition until the peaks of the loudspeaker signal remain below the threshold levels.
The VBB limiter reduces the gain when VBB decreases below a threshold of typically 2.9V at VBB = 3.6V typically. The VBB limiter threshold is internally coupled with the VBB level.
The current limiter reduces the amplifier gain when the supply current I_{SUP} is too low to power the loudspeaker at the desired level .

The attack time at which the gain is reduced is very short for the peak and VBB limiter and relatively long for the current limiter.
Both attack time and release time, at which the gain returns to the nominal value, are proportional to the value of C28 connected to DLC.
The recommended value for C28 is 330nF.

3.3 MUTE or loudspeaker enable (LSE) function.

A 'HIGH' level at the LSE terminal (with respect to SLPE) enables the loudspeaker amplifier.
LSE 'open' or connected to SLPE disables the amplifier; the TEA1085A is in the standby mode.

On the PCB, the LSE terminal is interconnected with the MUTE input of the TEA1085A. Jumper J2 connect the LSE terminal to the VBB voltage.
Removing J2 disables the amplifier.

3.4. Larsen Level Limiter (LLL).

The Larsen Level Limiter (LLL) reduces annoying howling signals from the loudspeaker and on the telephone line.
A larsen signal will occur when the overall gain, from microphone via loudspeaker back to microphone, is more than one. This overall gain depends on the side tone transfer from microphone input to earpiece output QR+ (or QR-), the attenuation from QR+ (or QR-) to LSI1 (or LSI2) input, the loudspeaker amplifier gain and on the acoustic feedback from loudspeaker to microphone. The overall gain also depends on the sensitivity of the applied capsules and the distance between loudspeaker and microphone.

A possible howling is received by the microphone, is amplified and filtered by an active 3rd order high pass filter and is compared with a voltage level and checked for its time duration. This high pass filtering and time duration measurement are done to obtain a low sensitivity for own speech. When the larsen signal exceeds this first voltage threshold, capacitor C26 will be discharged. For normal speech this discharge is relatively small.

As soon as the first threshold is passed and the Larsen signal at DTI remains for more than 120ms (attack delay time), the LLL is switched over from listening-in to larsen mode. The gain of the loudspeaker amplifier is than reduced quickly. The remaining signal at DTI in this Larsen mode is determined by a second threshold. When the Larsen effect stops, by reducing the acoustic feedback from loudspeaker to microphone, the gain of the loudspeaker amplifier returns to the normal gain in 250ms (larsen release time).

A reference current of 10.4μA (set by R36) is used to determine the timing of the LLL system together with the value of the Larsen Limiter capacitor C26. The recommended value of R36 and C26 is 120kΩ respectively 1μF.

The PCB offers the possibility to adapt the gain of the preamplifier and the two voltage thresholds as explained in the following sections.

3.4.1. Microphone preamplifier gain.

The gain between the two microphone inputs LAI+/LAI- and output QLA has to be adjusted to the same value as the microphone gain of the applied transmission IC. The attenuation of the three cascaded high pass filters (see next chapter) must not be taken into account.

With a fixed value of R25 = R26 = 10kΩ (on the PCB), the gain can be adjusted according:

$$A_{pre} = \frac{v_{QLA}}{v_{mic}} = \frac{R29}{R26} .$$

The gain on the PCB is set to 44dB (R29 on the PCB is 1.65MΩ), which corresponds to the microphone gain of the TEA1064A (chapter 4.2.2). The +input is biased at VBB/2 by means of R27 and R28. The preferred value of R27 and R28 is 2 * R29 (R27 = R28 = 3.3MΩ on the PCB).

3.4.2. High pass filter / detector current.

The microphone input voltage is transferred into a current for the detector flowing into terminal DCA via three cascaded first order high pass filters with three separated cut off frequencies (ref.1) at:

$f1 = 1/(2 * \pi * C24 * R30//R31)$, $f2 = 1/(2 * \pi * C25 * R33)$ and
$f3 = 1/(2 * \pi * C23 * R26) = 1/(2 * \pi * C22 * R25)$.

With the components on the PCB: C24 = 4.7nF, R30 = 100kΩ, R31 = 220kΩ, C25 = 330nF, R33 = 510Ω, C23 = C22 = 4.7nF and R26 = R25 = 10kΩ the three cut off frequencies are approximately f1 = 500Hz, f2 = 1kHz and f3 = 3kHz.

The detector current into pin DCA consists of a DC current of I_{DCA} = 11μA, at VBB = 3.6V typical, and an AC component i_{DCA} = 2 * v_{DTI} (mA) at f > 1kHz.

Adjusting the VBB voltage, as described in chapter 3.1, has the consequence that R32 (100kΩ on the PCB) has to be modified to keep I_{DCA} = 11μA.
The new value of R32 is given by:

$$R32 = \frac{R30}{R30 + R31} * VBB * 90*10^3 - R33 \quad (\Omega)$$

3.4.3. Threshold levels.

The Larsen system switches from listening-in to larsen mode as the peak level of the signal at DTI exceeds the first threshold (after the attack delay time). This first threshold v_{DTI1} can be adjusted by R35 (51.1kΩ on the PCB), connected between pin THL1 and VSS. The adjusted level of v_{DTI1} on the PCB is 18.8mV (if f > 3kHz), but can be modified by changing R35:

$$R35 = \frac{R33 * 2.5}{v_{DTI1} + R33 * 11*10^{-6}} \quad (\Omega) \; .$$

After the attack delay time the gain of the loudspeaker amplifier is reduced to a level at which the residual Larsen signal will be determined by a second threshold. This threshold v_{DTI2} (6.9mV on the PCB) can be set by means of R34 (100kΩ on the PCB), connected between pin THL2 and VSS. R34 can be calculated by:

$$R34 = \frac{R33 * 2.5}{v_{DTI2} + R33 * 11*10^{-6}} \quad (\Omega) \; .$$

As soon as the overall gain is < 1, for instance by enlarging the distance between microphone and loudspeaker, the gain of the loudspeaker returns to its normal value within 250ms.

4. TEA1064A TRANSMISSION CIRCUIT.

4.1. Regulated $V_{LN-SLPE}$ voltage.

Fig.2 shows the application of the TEA1064A as transmission IC in the listening-in application. Pin VCC2 is connected to pin LN at which a stabilised voltage between LN and SLPE will be obtained. The voltage at terminal LN is:

$$V_{LN} = V_{LN-SLPE} + (I_{line} - I_{CC1}) * R9$$

The reference voltage $V_{LN-SLPE}$ is typically 3.3V, but has been increased for the application with the TEA1085A by means of R17 (see chapter 2.4). The total line voltage (without diode bridge) is obtained by this reference voltage increased with the voltage drops across R20, R9 and R10.

The value of R9 on the PCB is 20Ω. Changing R9 influences the microphone gain (chapter 4.2), DTMF gain (chapter 4.3), gain control characteristics (chapter 4.5), side tone level (chapter 4.6), threshold current of the low voltage function (chapter 2.5) and the maximum output swing on the line. If R9 has to be changed follow the adjustment procedure of the TEA1064A as given in ref.2 or 3.

An extra supply voltage VP for peripherals can be constructed between LINE and SLPE by means of an RC smoothing filter consisting of the components R23 and C19, see fig.2. Consult ref.3 for the values of C19 and R23 and for the application consequences. The PCB is prepared for this function, however R23 and C19 are not placed.

4.2. Microphone amplifier.

4.2.1. Amplifier inputs.

The printed circuit board is ready for use with microphone capsules which require a symmetrical input (dynamic, magnetic and piezo electric). The input impedance of the TEA1064A microphone amplifier is high (typ. 64kΩ); with R14 = 8.25kΩ (figure 2) a more low-ohmic termination is obtained. If an asymmetrical input is required (electret) the MIC- input is used as a signal input via C13. The MIC+ input should be connected to ground by C14.
The value of both capacitors is determined by the lower cut-off frequency required. For testing purposes only (low-ohmic drive) 2.2μF is a suitable value. The PCB is prepared to place these capacitors.

Since the TEA1064A has a microphone gain that can be set between 44 and 52 dB (with R9 = 20Ω, see also chapter 4.2.2.) an external attenuation network (e.g. R15, R16) is required if a sensitive type of microphone is used needing less than 44dB of gain.

4.2.2. Gain adjustment.

The gain of the microphone amplifier can be calculated using the following formula:

$$A_m = 1.356 * \left(\frac{R7 + 3.47 * 10^3}{R5 * R9}\right) * \frac{Ri * R_{line}}{Ri + R_{line}}$$

in which:
$Ri = R1 // R_p$; $R_p = 15.5k\Omega$
R_{line} is the load resistance at the line during the measurement (normally 600Ω).

Figure 8 shows the gain of the microphone amplifier as a function of R7 for different values of R9.

On the board the following resistors are mounted:
R1 = 619Ω; R5 = 3.65kΩ; R7 = 25.5kΩ; R9 = 20Ω. This results in a gain of 44dB (assuming R_{line} = 600Ω).

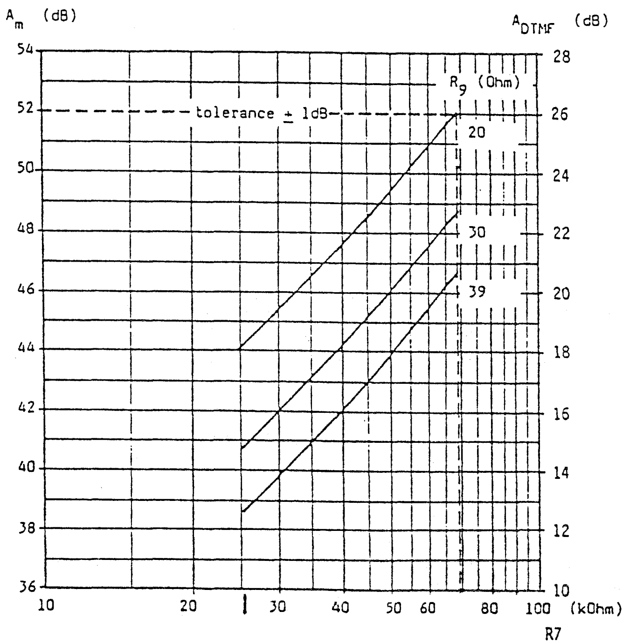

Fig.8. Microphone gain as a function of R7.

To ensure stability of the microphone amplifier capacitor C6 is connected between GAS1 and SLPE. C6 can be adjusted to obtain a first-order low-pass filter with time constant R7C6. On the board C6 = 100pF (f_{3dB} = 62kHz).

4.2.3. Microphone mute / dynamic limiter.

A low level at the DLS/MMUTE pin (e.g. via a 3.3kΩ resistor to V_{EE}) inhibits the microphone amplifier and has no influence on the receiver and DTMF amplifiers.
Removing the low level at the DLS/MMUTE pin provides the normal function of the microphone amplifier after a time proportional to C16.
To prevent distortion of the transmitted microphone signal a dynamic limiter is incorporated in the microphone amplifier. The use of such a limiter also improves the sidetone performance considerably (less distortion and a limited sidetone level in overdrive conditions).
When peaks of the transmitted signal on the line exceed an internally determined threshold the gain of the transmitter amplifier is reduced rapidly. The time in which the gain reduction is realized (the attack time t_{att}) is very short. The circuit stays in the gain-reduced condition until the peaks of the transmitted signal remain below the threshold level. The gain then returns to its normal value after the release time t_{rel}.
Both the attack and the release time are proportional to the value of the capacitor C16 connected to pin DLS of the TEA1064A. The recommended value for C16 is 470nF (also mounted on the board).

4.3. DTMF amplifier.

By applying a HIGH voltage to the MUTE input of the TEA1064A, the device is switched over from the speech mode to the dialling mode. In this mode DTMF signals applied to the DTMF input are transmitted to the line. The microphone and receiver amplifiers are blocked now and the dynamic limiter will not be activated.
The gain of the DTMF amplifier is:

$$20 * {}^{10}\log A_{DTMF} = 20 * {}^{10}\log A_m - 26\text{dB} \qquad \text{(See fig.8)}$$

Once the microphone gain has been adjusted, the DTMF gain is fixed. This means that the DTMF input level must be adjusted to the correct value by means of an external attenuator at the DTMF input.

4.4. Receiver amplifier.

4.4.1. Amplifier outputs.

Earpieces with an impedance up to $Z_T = 450\Omega$ must be driven in single ended mode. For impedances larger than 450Ω differential drive is possible.
A resistor in series with the earpiece can be mounted on the board if required (see ref.3 for details).

The maximum output swing on the earpiece depends on the DC voltage at VCC1, and therefore also on the adjusted voltage drop (V_{LN-VEE}) over the circuit.

4.4.2. Gain adjustment.

The gain of the receiver amplifier (from IR to QR) can be calculated by means of:

single ended drive: $$A_{TA} = 1.314 * \frac{R4}{R5} * \frac{Z_T}{Z_T + 4},$$

differential drive: $$A_{TS} = 2.628 * \frac{R4}{R5} * \frac{Z_T}{Z_T + 8},$$

in which: Z_T = earpiece impedance (in Ω).

Figure 9 shows the receiver gain as a function of R4.

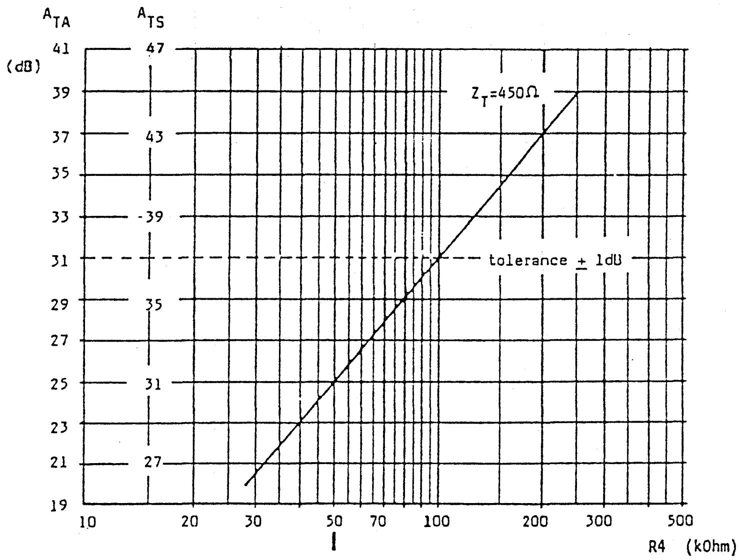

Fig.9. Gain of the receiver amplifier as a function of R4.

On the board R4 = 50.5kΩ resulting in a gain A_{TA} of 25dB. The total receive gain from line to earpiece output QR+ can be obtained by subtracting the attenuation of the anti-sidetone network (32dB with the component values of fig.2) from A_{TA}, resulting in an overall gain of -7dB.

C4 and C7 (fig.2) prevent parasitic oscillations of the receiver amplifier while C4 can also be adjusted to obtain a first-order low-pass filter with time constant R4C4. The value of C7 should be 10 times C4.

A high-pass filter is formed by C5 (together with the input impedance of pin IR) and by C2 (together with the earpiece impedance).

On the board the following components are mounted: C2 = 10μF, C4 = 100pF, C5 = 100nF, C7 = 1nF.

4.4.3. Confidence tone.

During DTMF dialling the dialling tones can be heard at a low level in the earpiece. For both single ended and differential drive the gain from the DTMF input to the receiver output, A_{CT}, is given by:

$$20 * {}^{10}\log A_{CT} = 20 * {}^{10}\log A_T - 49\text{dB} .$$

4.5. Automatic gain control.

The line-loss compensation circuit of the TEA1064A regulates the microphone and receiver gain with the line current (line length). The DTMF amplifier is not affected by the circuit.
The gain control range is 6dB corresponding with a 5km line of 0.5mm diameter copper twisted pair (DC resistance 176Ω/km, average AC attenuation 1.2dB/km).

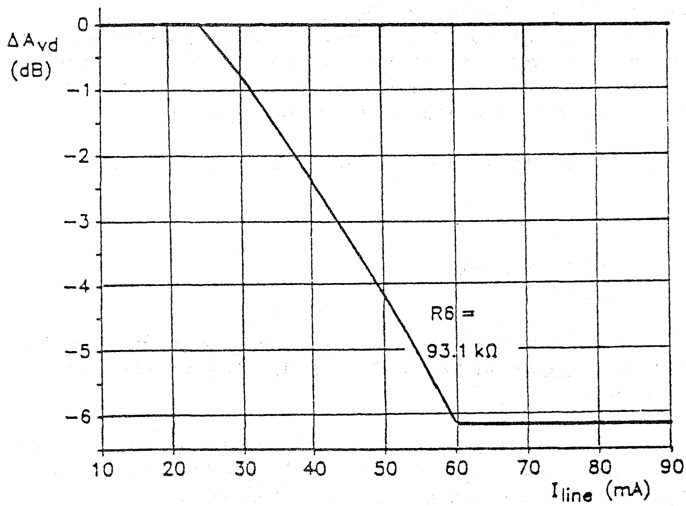

Fig.10. Automatic microphone and receiver gain control versus I_{line}.

If no automatic gain control is required jumper J1 on the printed circuit board has to be removed. Resistor R6 must be chosen in accordance with the exchange supply voltage and its feeding bridge resistor.
Figure 10 shows the gain-control characteristics with R6 = 93.1kΩ and R9 = 20Ω optimised for a 600Ω feeding bridge and an exchange supply voltage of 48V; ΔA_{vd} is the reduction of the microphone and receiver gain.

4.6. Anti-sidetone circuit.

On the PCB the TEA1060-family bridge is used to suppress the sidetone. For details of this bridge please refer to ref.3.
Maximum suppression is obtained when the following two conditions are fulfilled:

a) $R9 * R2 = R_i * (R3 + R8)$ where $R_i = R_1 // R_p$ and $R_p = 15.5k\Omega$

b) $$Z_{bal} = \frac{R3 * (R8 + R9)}{R2 * R9} * Z_{line} = k * Z_{line}$$

Normally R1 and R9 are fixed and R2, R3, and R8 can be chosen by the designer. These resistors have to be chosen to meet the following criteria:
- compatibility with a standard capacitor from the E6 or E12 range for the capacitor used in Z_{bal}
- $|Z_{bal}//R8| \ll R3$ to avoid influence of Z_{bal} on the receiver gain
- $|Z_{bal} + R8| \gg R9$ to avoid influence on the microphone gain
- $(R3 + |R8//Z_{bal}|) \ll 20k\Omega$ to avoid the influence of the input impedance at IR ($21\ k\Omega \pm 4k\Omega$) on the bridge attenuation.

In practice Z_{line} varies strongly with line length and type of cable. Therefore an average value has to be chosen for Z_{bal}.
On the printed circuit board Z_{bal} is optimised for a cable length of 5km with a diameter of 0.5mm copper with a DC resistance of 176Ω/km and a capacitance of 38nF/km resulting in an average AC attenuation of 1.2dB/km.
The anti-sidetone network attenuates the signal from the line to the input of the receiver amplifier with 32dB.
For compatibility of the capacitor value in Z_{bal} with a standard capacitor from the E6 range (220nF):

$$k = \frac{140nF}{220nF} = 0.64\ .$$

For R3 a value of 3.92kΩ has been chosen. So Z_{bal}, R8, and R2 can be calculated resulting in the practical values R11 = 130Ω, R12 = 825Ω, C12 = 220nF, R8 = 392Ω, and R2 = 130kΩ.

In case of complex set impedance (chapter 4.7) the bridge can be rebalanced by readjusting the values of R8 and Z_{bal}, and either R2 and R9 (see ref.4). Changing R2 and R9 has also consequences on other set parameters.
R2 determines the attenuation from line to receiver amplifier input IR, while changing R9 influences microphone gain, dtmf gain, etc. as indicated in chapter 4.1.

4.7. Set impedance.

The set impedance of the PCB application is provided by R1 (619Ω). Complex set impedance, consisting of Ra + Rb//Cp, can be obtained by means of R1, R13 and C43 (on the PCB: R13 = 0Ω; C43 is not mounted). Changing R1 has

consequences for sidetone behaviour (chapter 4.6), while the value of the DC resistance of the complex impedance influences the voltage at terminal VCC1 which can influences the transmission properties (ref.3).
Complex set impedance results in a frequency dependent transmission gain.

5. PROTECTION.

Protection is achieved by means of a break over diode (BR211/240V) between the a/b lines in combination with R10 (12Ω) and zener diode D5 (12V) between LINE and VEE. The protection is meant for application with pulse dialling and DTMF dialling.

6. ELECTROMAGNETIC COMPATIBILITY / STABILITY.

The EMC behaviour of the TEA1085A / TEA1064A application is different from the application of the TEA1064A alone and requires different measures. Necessary for this application is coil L1 (150μH), connected in the SUP supply wire of the TEA1085A, in combination with C11 (4n7) connected from LN of the TEA1064A to VEE. Both components, including C21 (47pF) connected from SUP to SREF of the TEA1085A, are also required for stability of this structure with the TEA1085A and TEA1064A or another transmission IC from the TEA1060-family.

Coil L1 (type LHL06-151K from TAIYO) has to remain a coil, without saturation effects, for the applied line current range of this application.

EMC components C33, C34, C35, C36, C37 and C38 are mounted on the PCB to improve the EMC behaviour of the speech and listening-in circuit part. The EMC behaviour is shown in fig.11, 12, 13 and 14.
Measured are the demodulated signal levels at the a/b terminals (fig.11), across the earpiece (fig.12), across the loudspeaker (fig.13) and at the detector current input DCA of the Larsen Level Limiter (fig.14).

The -26dB and -40dB S/I (signal interference) lines are referred to a mean speech signal level of 100mV at the telephone line.
The EMC requirements are -26dB or -40dB, depending on the national PTT organization.

The measurement conditions are as follows:
- The demonstration board (PR4534X) is carried out with double sided wiring and two ground reference layers both connected to VEE of the TEA1064A.
  I_{line} = 27mA, microphone (200Ω) symmetrically connected, earpiece (200Ω or 47nF) symmetrically connected.
- Current injection method (common mode on a/b-b/a terminals).
- RF levels: 3Vrms 100kHz-30MHz, 0.5Vrms 30MHz-150MHz
- Amplitude modulation: 80%, 1kHz.
- Test set up as explained in German requirements VDE 0878 (ref.5).

Remark:
Avoid long connection wires and test equipment connected to the SLPE terminal(s). They worsen the EMC behaviour and can give instabilities in the loudspeaker amplifier.

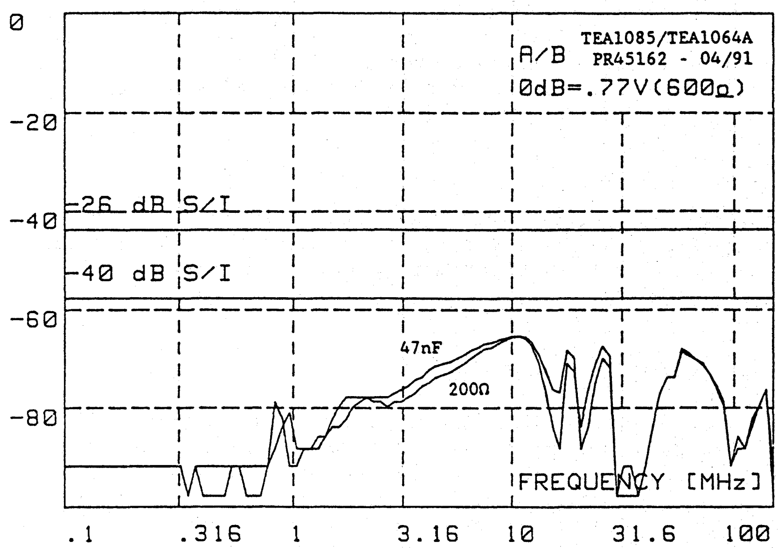

Fig.11 Demodulated signal level across the a/b terminals versus frequency.

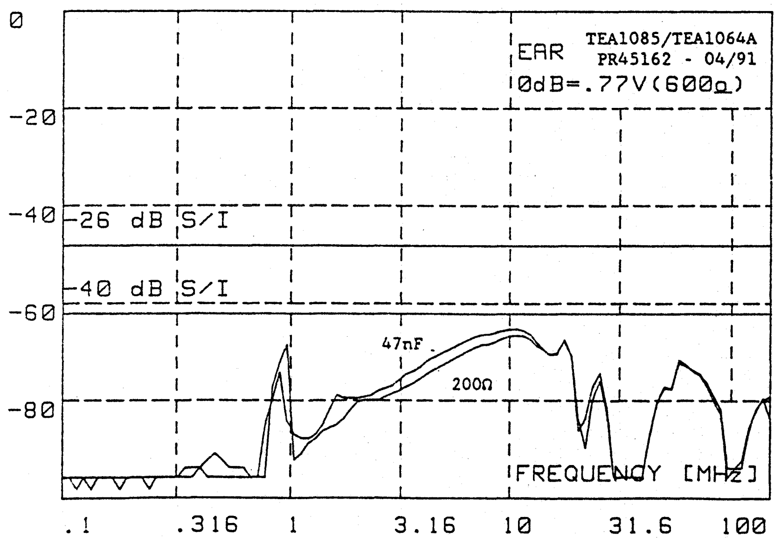

Fig.12 Demodulated signal level across the earpieces versus frequency.

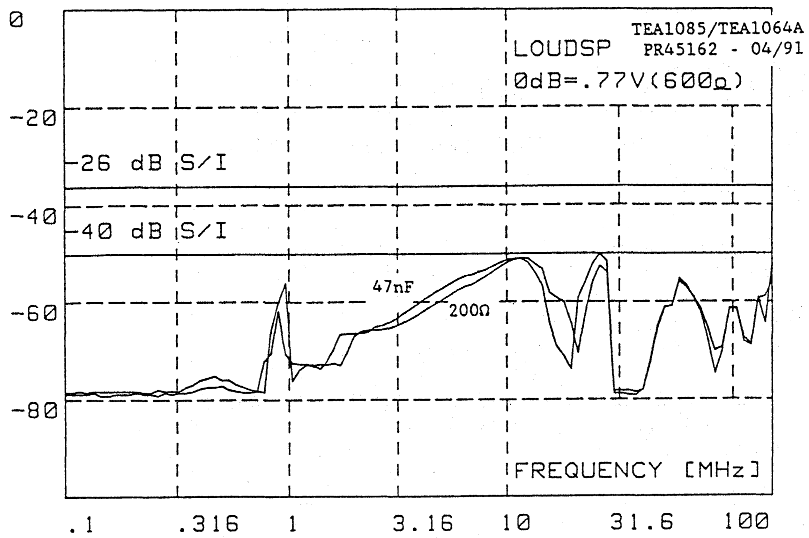

Fig.13 Demodulated signal level across the loudspeaker versus frequency.

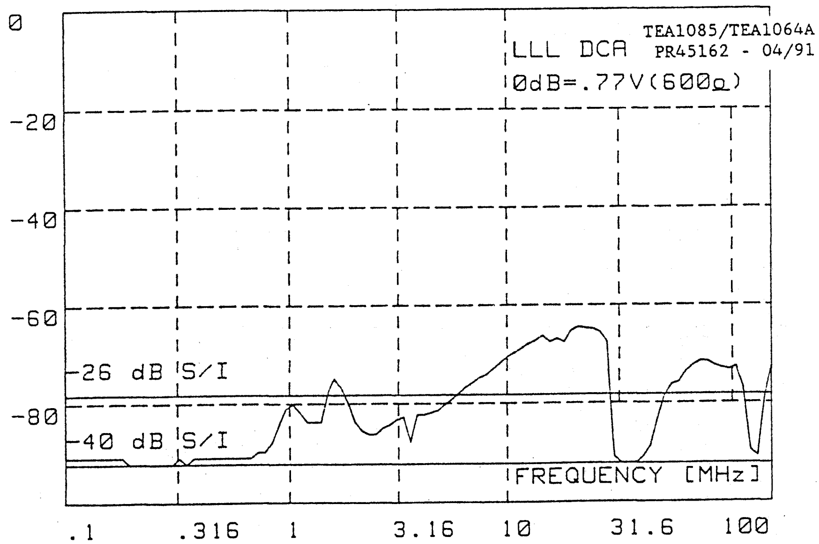

Fig.14 Demodulated signal level at the detector input DCA versus frequency.

7. REFERENCES.

1) - Data sheet TEA1085; November 1990
(data sheet TEA1085A/TEA1085; to be made).

2) - Documentation for the TEA1064A printed circuit board CAB3458.
P.Biermans / P.J.M.Sijbers / K.Wortel.
Report number ETT/AN90001.

3) - PCALE Laboratory Report, "Application of the versatile speech / transmission circuit TEA1064 in full electronic telephone sets",
F.v.Dongen / P.J.M.Sijbers.
Report number ETT89009, (12nc. 9398 068 80011).
(Application report TEA1064A; to be published).

4) - Designers' Guide: "TEA1060-family - Versatile Speech/Transmission ICs for Electronic Telephone Sets", P.J.M.Sijbers.
(12nc. 9398 341 10011).

5) - VDE 0878, part 1 and part200; 1987.

6) - PCALE Laboratory Report, "TEA1085A / TEA1085 - A Listening-in facility for electronic telephone sets -".
Report number ETT/AN91016.

7) - Test procedure of the PR45163: ETT/IR91059

APPENDIX 1: Adjustment procedure of the TEA1085A/TEA1064A application.

	See chapter - figure:
- Adjust VBB (if necessary) by R38	3.1. - 7.
- Adjust $V_{LN-SLPE}$ of the TEA1064A by R17 to obtain $V_{SUP-VBB}$ > 0.4V, taking into account VBB-max.	2.4.
- Adjust microphone gain by R7	4.2.2. - 8.
- Adjust receive gain by R4	4.4.2. - 9.
- Adjust microphone preamplifier gain of LLL, by R29, adapt R27 and R28 according to 2 * R29.	3.4.1.
- Choose loudspeaker drive mode, SEL or BTL, depending on required output power and available minimum line current	2.4. - 5 and 6.
- Adjust attenuation of network R24-R21	3.2.1.
- Modify R32, depending on VBB setting	3.4.2.
- Follow instructions in reference 2 or 3 to obtain the specific settings of the transmission IC (do not use the stabilized supply option).	

Notes:

- The gain of the loudspeaker amplifier can completely be reduced (by the Larsen Level Limiter) and remains reduced, during tests of the listening-in facility with continuously microphone signals. An input signal of more than 1mVrms (1kHz; continuous) between the PCB terminals MIC-/MIC+ activates the Larsen Level Limiter. It will be switched into the "Larsen mode" and remains in this mode as long as this input signal is present.

  The influence of the Larsen Level Limiter on the loudspeaker amplifier gain can be cancelled (for the time being) by means of a short circuited resistor R30.

APPENDIX 2: PCB modification for TEA1085 and/or TEA106X.

This appendix gives the modifications of the PCB when the TEA1085A has to be replaced by the TEA1085 and/or when the TEA1064A has to be replaced by the TEA1060, TEA1061, TEA1067 or TEA1068.

TEA1085:

The TEA1085 can be placed in the socket of the TEA1085A.

The modifications of the PCB (with TEA1085) concerns the loudspeaker enable function which has to be operated by a switch to control the MUTE 'toggle' circuitry of the TEA1085. The switch may be a push button switch with a 'break' or 'make' contact.

- Remove jumper J2.
- Connect the switch between the terminals LSE and SLPE. Switching the listening-in circuit over between standby and listening-in occurs at the positive going edges of the MUTE signal.
- Connect a debounce capacitor of 10nF across the switch contacts.

TEA1060, TEA1061, TEA1067 or TEA1068:

The TEA1060, TEA1061, TEA1067 or TEA1068 can be placed in the 18 pins shadow IC socket, which is mounted beside the 20 pins socket of the TEA1064A.

Fig.A2 shows the components side of the PCB on which both IC sockets are indicated. The following modifications of the PCB are given to obtain the same settings as with the TEA1064A, if the TEA1064A is replaced by the indicated transmission IC (see also ref.4):

TEA1060:	line voltage	remove R17.
	receive gain	adapt R4 = 100kΩ.
	automatic gain control	adapt R6 = 100kΩ.
TEA1061:	line voltage	remove R17.
	microphone gain	adapt R7 = 140kΩ.
	receive gain	adapt R4 = 100kΩ.
	automatic gain control	adapt R6 = 100kΩ.
TEA1067:	line voltage	adapt R17 = 33kΩ.
	automatic gain control	adapt R6 = 100kΩ.
TEA1068:	line voltage	remove R17
	receive gain	adapt R4 = 100kΩ.
	automatic gain control	adapt R6 = 100kΩ.

TEA1060/1061/1067/1068 (TEA106X):

Interface circuitry between the logic inputs of the TEA1085A, TEA106X and peripheral circuits depending on the ground reference of the peripherals (VEE or SLPE). Diode D6 has to be placed on the PCB to interconnect both PD inputs of the TEA1085A and TEA106X. The short connection, placed on the PCB instead of D6, has to be removed.

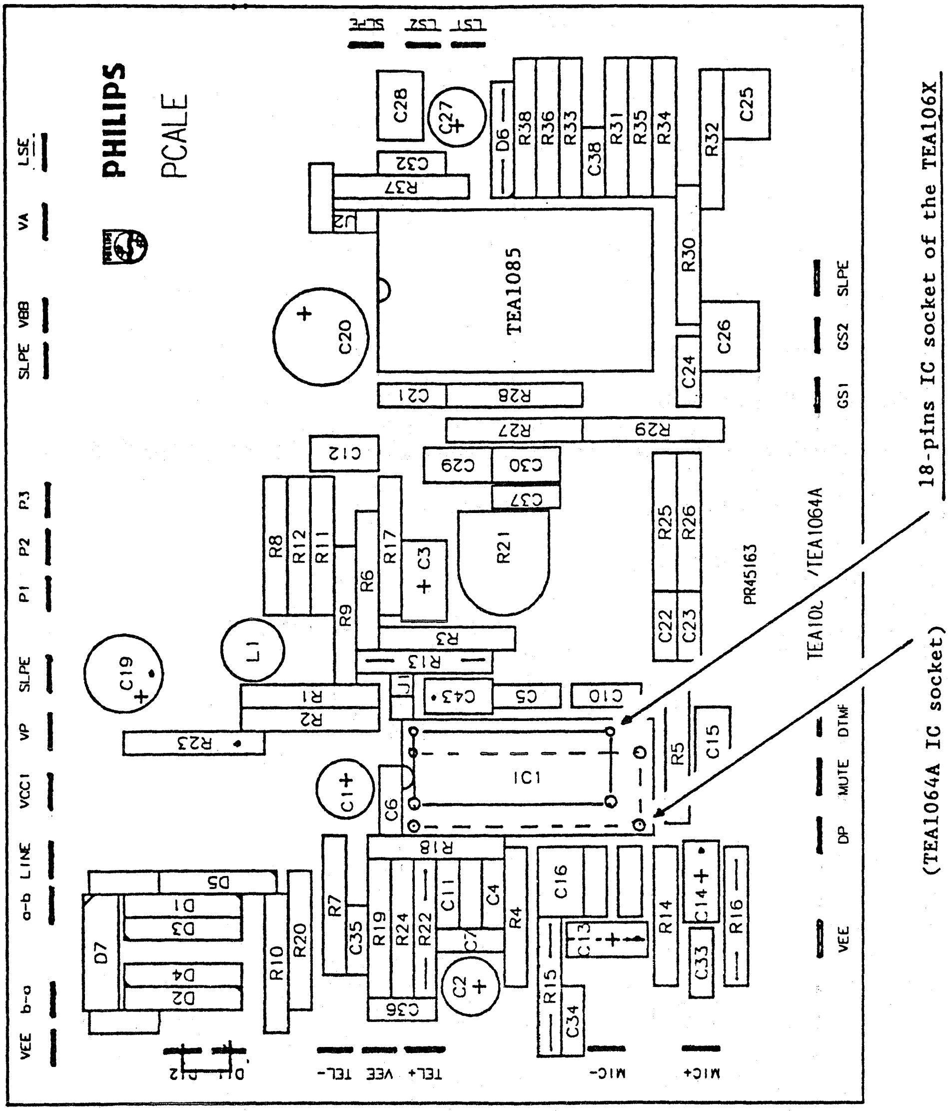

Fig.A2 View of the components side of the PCB showing the shadow socket of the TEA106X.

APPLICATION NOTE Nr	ETT/AN94001
TITLE	User manual for OM4750: Demonstration board TEA1093 and TEA1094
AUTHOR	R. van Leeuwen, C. Voorwinden
DATE	April 1994

Table of Contents:

1 Introduction

This report is meant as a guideline for the demonstration board number OM4750 with the TEA1093/TEA1094 handsfree chip and the TEA1062/TEA1067 transmission circuit. The board is meant to demonstrate the performance of the TEA1093/TEA1094.

In this report only a brief explanation is given of the TEA106x and TEA1093/TEA1094 parameter setting. More technical details and application examples of the TEA1093/TEA1094 can be found in the appropriate application reports and specifications (see references 1 up to 4). More details of the TEA106x can be found in the data handbook (reference 5) and the TEA1060 designers guide (reference 6).

This introduction continues with an explanation about how to use the TEA1093/TEA1094 demonstration board. In chapter 2 and 3 of this report the most important parameter settings of the board are explained.

The board can be connected to the telephone line by the a/b connectors, or by a 6-pins socket which is connected via a 6x2 jumper matrix to the a/b connectors. On the bottom left part of the board three control switches are applied, which have the following functions:

S1: switching the loudspeaker on or off.
S2: selection of handsfree or handset operation.
S3: mute the microphone channel.

The volume of the speaker can be adjusted by the potentiometer.

With the aid of another 4x6 jumper matrix on the board the microphone and earpiece signals can be connected to the appropriate pins of a 4-pins socket. The jumper setting depends on the connection in the handset cord.

The TEA1062 and TEA1067 can be simply plugged in. The state of the jumpers J4 and J5 depends on whether a TEA1093 or a TEA1094 is placed on the board.

Basically the three switches S1,S2 and S3 provide three possible operation modes:

Handset operation

During handset operation, selected by S2, the handset microphone can be muted by S3, for privacy reasons. The loudspeaker should be switched off by S1.

Listening-in operation

If the loudspeaker is switched on by S1 during handset operation, the listening-in function is enabled and the TEA1093 prevents the set from howling. With switch S3, the microphone can be muted.

Handsfree operation

Handsfree operation can be selected by S2. In handsfree mode, both the handsfree microphone and the loudspeaker can be muted separately by means of the switches S1 and S3.

The switches S1, S2 and S3 can be overruled by a microcontroller which can be connected to the microcontroller interface signals of the board. Jumpers J1, J2, and J3 need to be removed in such a case.

2 TEA106x application

This chapter describes the telephone line connection, the TEA106x parameter settings and the AC signal interfacing to the TEA1093/TEA1094 (see the circuit diagram in figure 3).

Two sockets are mounted on the printed circuit board: a 16 pins socket for the TEA1062, and an 18 pins socket for the TEA1067. Major difference between the TEA1067 and the TEA1062 is the fact that the TEA1062 has no power down function and no inverting receiver output stage.

2.1 Telephone line connection

The telephone line can be connected to the a/b connectors. The polarity guard consists of 4 diodes. The BR211-240 is a breakover diode with a breakover voltage between 211 and 269V and protects together with R_{10} and D_5 the application against over voltage. There is also a kind of ringer implemented: diode D_{11} lights up if a ringer signal is present on a/b.

2.2 DC settings

In case of a TEA1093, the TEA106x is supplied at pin LN with a constant current determined by the voltage V_{sup} - V_{sref} = 315 mV and the resistor R_{sref}. If R_{sref1} = 100Ω this results in a current of about 3 mA flowing into the TEA106x output stage. R_{ref2} is placed between SREF and LN to avoid strong clipping at the bottom of the line signal.

At a line current of 15 mA, the voltage at pin LN is 4.0 V for the TEA1062 and 3.9V for the TEA1067. This voltage can be increased by adding a resistor R_{17}. On the board R_{17}=39kΩ resulting in a voltage at pin LN of 4.5V for both the TEA1062 and TEA1067. Adding the voltage drop over R_{sref1} (315mV), R_{sref2} (315mV), R_{10} (182mV at 15 mA) and the polarity guard (≈1.5V), gives 6.8V between the a and b connectors at 15 mA.

In case of TEA1094 pin LN is connected to the line by jumper J5. The line voltage is determined in the TEA1067/TEA1062 specifications. The pins 7 and 9 of the TEA1094 are internally not connected. The TEA1094 need not to be referenced to the voltage on SLPE, but its ground can be connected to VEE, which makes a simpler interface with the microcontroller possible: D_6 and R_{mutet} are not necessary.

2.3 Set impedance

The set impedance is mainly determined by resistor R_1 and the internal resistance of the TEA106x:

$$Z_{set} = \frac{R_1 R_i}{R_1 + R_i} \tag{1}$$

in which $R_i = 16.2\ k\Omega$.

On the board resistor R_1 is split into R_{1a} and $R_{1b}//C_{1b}$ to have the possibility to create a complex impedance. R_{1b} is chosen 0Ω and R_{1a} is 619 Ω, providing a Z_{set} of approximately 600 Ω in the frequency range of 300Hz up to 3400Hz.

2.4 Side tone

In order to obtain optimum side tone suppression, the following conditions should be met:

$$R_9 R_2 = R_1 (R_3 + R_8 // Z_{bal}) \tag{2}$$

$$\frac{Z_{bal}}{Z_{bal} + R_8} = \frac{Z_{line}}{Z_{line} + R_1} \tag{3}$$

where:

where:	Z_{bal}	=	the network consisting of R_{11}, R_{12} and C_{12}
	Z_{line}	=	the load at the line
	R_1	=	$R_{1a} + R_{1b}//C_{1b}$

Equation (2) can be met if $R_8//Z_{bal} << R_3$ and if R_9, R_1 and R_3 are real (not complex).

Equation (3) can be met if:

$$Z_{bal} = \frac{R_8}{R_1} Z_{line} = k\ Z_{line} \tag{4}$$

In practice, Z_{line} will vary strongly with line length and cable type. Consequently, an average value for Z_{bal} has to be chosen. On the board Z_{bal} is optimised for a 0.5 mm copper cable of 5 km with an attenuation of 1.2 dB/km, a DC resistance of 176Ω/km and a capacitance of

38 nF/km. The anti side tone circuit attenuates the received signal from the line with approximately 32 dB.

2.5 Microphone amplifier

Microphone input MIC- of the TEA106x is used as a signal input. MIC+ is referenced to SLPE since the microphone output of the TEA1093 is also referenced to SLPE. In case of TEA1094, GNDMIC and MIC+ are referenced to VEE. The resistor network R_{20}, R_{21} and R_{22} attenuates the signal coming from the TEA1093 with 20 dB. This is done to improve the noise behaviour in Rx mode (for details see reference 1 and 3). Capacitors C_{21} and C_{22} block DC and form a high pass filter with R_{20}, R_{21} and R_{22}. The cut-off frequency is set at about 270 Hz.

The gain from the MIC+/- inputs to the line depends on the value of resistor R7 between pins GAS1 and GAS2 in the following way:

$$A_{MIC+,MIC-\ to\ LN} = 1.356\ x\ \frac{R_7 + 3470\Omega}{R_5\ R_9}\ x\ \frac{Z_{set}\ Z_{line}}{Z_{set} + Z_{line}} \qquad (5)$$

On the board, resistor R_7 is set to 27.4 kΩ, giving a microphone gain of approximately 44.5 dB (Z_{line} = 600Ω) from MIC+, MIC- to the line. If another gain setting is required, the value of R_7 may be varied between 27.4 kΩ and 68kΩ (gains setting 44 dB up to 52 dB).

Capacitor C_6 between GAS1 and SLPE should be at least 100 pF for stability reasons. It is used to create a low pass first order filter and can be adjusted to suit user requirements. The mounted value is 1.2nF giving a cut-off frequency of about 4.3 kHz. Capacitor C_{20} should have a value 10 times larger than C_6. The mounted value is therefore 12nF.

2.6 DTMF amplifier

The DTMF amplifier of the TEA106x is enabled if a high level (>1.5V) at microcontroller interface pin MUTE is applied. In mute condition, the microphone amplifier and the earpiece amplifier are muted and only a confidence tone can be noticed at the earpiece output.

The gain from the DTMF input to the line is 26.5dB lower than the microphone gain. Thus, with R_7=27.4kΩ the DTMF gain equals: 44.5dB - 26.5dB = 18dB.

The gain from the DTMF input to QR+ (confidence tone) is 50 dB lower than the earpiece gain from IR to QR+ (see paragraph 2.7). With R_4 = 68.1kΩ the gain from DTMF input to QR+ output equals: 27.8dB - 50dB = -22.2dB.

2.7 Earpiece amplifier

The receiving signal coming from the line is attenuated approximately 32 dB in the sidetone bridge before it enters pin IR. The gain from IR to QR+ (QR+ = QR for the TEA1062) is dependent on resistor R_4 between QR+ and GAR in the following way:

$$A_{IR\ to\ QR} = 1.314\ x\ \frac{R_4}{R_5}\ x\ \frac{Z_{ear}}{Z_{ear} + 4\Omega} \qquad (6)$$

where: Z_{ear} = earpiece impedance
R_5 = resistor at pin STAB which is $3.65k\Omega$

On the board, resistor $R_4 = 68.1k\Omega$ giving a receiver gain of 27.8dB from IR to QR+. The overall gain from the line to QR+ will therefore equal -4.2dB. If another gain setting is required, the value of R_4 may be varied between $28k\Omega$ and $100k\Omega$ for the TEA1062 corresponding to a gain setting from line to QR+ from -12dB up to -1dB. For the TEA1067, R_4 may be varied between $28k\Omega$ and $250k\Omega$ corresponding to gain settings of -12dB up to 7dB. Besides, the TEA1067 has an inverting output (QR-) for BTL drive (gain increases 6dB).

Capacitor C_4 should be at least 100pF for stability and is also used to form a low pass filter. The mounted value is 560pF providing a cut-off frequency in conjunction with R_4 of about 4.2kHz. Capacitor C_7 should be 10 times larger than C_4. Therefore 5.6 nF is mounted.

Pin QR+ interfaces to pin RIN1 of the TEA1093/TEA1094 via capacitor C_{rin1}. Since output QR+ is referenced to pin VEE of the TEA106x, the second input of the TEA1093/TEA1094 pin RIN2 is also connected to this reference via a DC blocking capacitor (C_2) and T1. T1 is controlled by switch S2 and will conduct if the handset mode is selected.

2.8 Automatic gain control

The built-in automatic gain control (AGC) circuit of the TEA106x attenuates the gain of the microphone and earpiece amplifiers at high line currents. For handsfree operation, this is a nice feature since sidetone is optimised for 5km, and thus will be worse for shorter lines and for longer lines. With AGC active the loop gain from MIC+,MIC- to QR+ is kept more or less constant over the range from 0 up to 5km. For longer lines, this loop gain will increase.

With $R_6 = 110k\Omega$ the gain of the microphone amplifier and the earpiece amplifier is reduced at line currents larger than 23mA. At 61mA, the maximum attenuation occurs (6dB) and above this value the gains are kept constant.

3 TEA1093/TEA1094 application

This chapter describes the parameter setting of the TEA1093/TEA1094. The circuit diagram is shown in figure 3.

3.1 DC setting

The TEA1093 splits up the line current into a constant current for the TEA106x (approximately 3 mA) and handles the excess current for generating its own stabilized supply voltage VBB (3.6V nominal). In case, due to AC line signals, the line voltage drops below the VBB voltage plus 0.4V, the excess current is sunk to GND to prevent distortion.

For stability reasons, a capacitor C_{sref} of 4n7 is needed between pins SREF and VEE. For correct functioning of the TEA1093/TEA1094 a capacitor of 470 µF is mounted at VBB (C_{vbb}).

The TEA1094 has to be supplied on pin Vbat, via R_{32}. R_{32} is added on the demoboard to protect, with the aid of zener diode D_{12}, the V_{BB} input for voltages larger than 12V. The supply has to be stabilized, V_{BB} is not stabilized in the TEA1094. Therefore pin 15 is not connected. Pin 17 is also not connected. Pin 7 and 9 are also not connected. For avoiding ground bouncing the supply has to be referenced to the 'ground starpoint' via GND.

3.2 Microphone amplifier

To be able to connect a handsfree as well as a handset microphone, an analogue switch (IC4) is applied at the microphone input MIC of the TEA1093/TEA1094 see figure 3. The analogue switch transfers either the signals coming from the handset or the handsfree microphone to input MIC. It is controlled by S2.
Both microphones are supplied from VBB via the RC smoothing filter formed by R_{vbbm} and C_{vbbm} Their sensitivities are set with R_{mic1} for the handsfree microphone and with R_{mic2} for the handset microphone.

The overall gain needed for the handsfree microphone can be adjusted by varying R_{gat}. An appropriate line level has been achieved with R_{gat} set to 95.3kΩ in combination with R_{mic1} set to 2.2kΩ. Capacitor C_{gat} of 390 pF forms a low pass filter together with R_{gat} (4.3kHz). Capacitor C_{mic1} of 18nF in parallel with R_{mic1} also forms a low pass filter (4kHz). The gain from MIC to MOUT is proportional to the resistor between GAT and MOUT. R_{gat} may be varied between 9.53kΩ up to 95.3kΩ providing a gain setting range of 5 dB up to 25 dB.

The handset microphone gain is determined by varying the sensitivity resistor R_{mic2} up to the required level. The handset microphone signal has to be attenuated with respect to the handsfree microphone signal, because the distance between the speaking person and microphone in case of the handsfree microphone is larger than in case of the handset microphone. The electrical attenuation equals: R_{mic2}/R_{mic1}. On the board R_{mic2} is set to 100Ω and series resistors R_{25} and R_{26} are used to adjust DC current through the microphone, and for EMC. Capacitor C_{mic2} in parallel with R_{mic2} forms a low pass filter (4.1kHz).

After the analogue switch, the microphone signal is coupled to MIC via capacitor C_{mic} of 33nF. With this capacitor and the input impedance at MIC (20kΩ) a high pass filter is realized (240Hz).

The microphone amplifier of the TEA1093/TEA1094 can be muted by means of switch S_3 or by applying a voltage >1.5V between connectors MUTET and GND (remove J3 in the latter case).

3.3 Loudspeaker amplifier

The input RIN1 of the loudspeaker amplifier is connected to QR+ via C_{rin1} of 33 nF. Input RIN2 is connected to VEE of the TEA106x via C_{rin2} of 33nF. With these DC blocking capacitors and the input impedance at RIN1 and RIN2 (2 *x* 20 kΩ) a high pass filter is realized (240Hz).
The gain of the loudspeaker amplifier from the inputs RIN1 and RIN2 to LSP1 is proportional to the resistor between GAR and LSP1. On the board R_{gar} is set to 221kΩ resulting in a loudspeaker gain of 28.4dB. With the TEA106x receiver gain being set at -4.2dB, the overall gain from telephone line to LSP1 is 24.2dB. If another gain setting is required for the TEA1093/TEA1094 loudspeaker amplifier, resistor R_{gar} may be varied between 11.5kΩ and 365kΩ providing a gain setting range of 3dB up to 33dB.
At pin LSP2, in case of the TEA1093, the inverted signal from LSP1 is present and may be used to obtain a bridge tied load drive configuration. In this configuration, the gain is increased 6dB. On the board, the loudspeaker is connected as a single ended load. In case of the TEA1094 pin 4 (LSP2 in case of TEA1094) is internally not connected, because the voltage of the external supply can be assumed sufficient, and BTL has a lower efficiency than SEL.

Capacitor C_{gar} of 180pF is mounted to form a low pass filter with resistor R_{gar} (4kHz).

The loudspeaker amplifier can be muted by making pin *DLC/$\overline{MUTER}$* lower than 0.2V. On the board this is realized with an analogue switch (IC4). The switch is controlled by S_1 or by a microcontroller. If the loudspeaker is muted, the analogue switch will close and C_{dlc} is discharged via current limiting resistor R_{muter}.

3.4 Volume control

Every increase of R_{vol} (connected to pin VOL) with 950Ω results in a 3dB lower loudspeaker gain. On the board a potentiometer is mounted of 10kΩ providing a volume range of about 32dB.

3.5 Dynamic limiter

The dynamic limiter will reduce the gain of the loudspeaker amplifier in the following cases:

a) VBB drops below 2.8V.
b) VBB starts dropping because supply current is insufficient (not for TEA1094).
c) The loudspeaker signal starts to cause saturation.

The attack and release times of these limiters are all proportional to the value of capacitor C_{dlc} connected to DLC. With C_{dlc} = 470nF, attack times are in the order of a few milliseconds for limiters a) and c). Limiter b) is much slower and has an attack time in the order of seconds.

3.6 Switching range

The switching range is proportional to the ratio between R_{swr} and R_{stab}. The resistor R_{stab} is fixed to 3.65kΩ. The resistor R_{swr} can be varied between 3.65kΩ and 1.45MΩ resulting in a switching range of 0dB up to 52dB.

The volume setting affects the switching range such that the sum of the microphone and loudspeaker amplifier gain is kept constant. Therefore, volume control has a maximum range equal to the switching range.

As explained in more detail in the application report (see reference 3) the following formula applies for calculation of the loop gain:

$$A_{loop} = A_{tx1093/4} + A_{tx106x} + A_{st} + A_{rx106x} + A_{rx1093/4} + A_{ac} - A_{swr} \tag{8}$$

where:
$A_{tx1093/4}$ = microphone amplifier gain of TEA1093/TEA1094 (set at 5dB)
A_{tx106x} = microphone amplifier gain of TEA106x (set at 44.5dB)
A_{st} = sidetone coupling (about -7dB worst case)
A_{rx106x} = earpiece amplifier gain of TEA106x (set at -4.2dB)
$A_{rx1093/4}$ = loudspeaker amplifier gain of TEA1093/TEA1094 set at 28.4dB)
A_{ac} = acoustic coupling (-37dB measured)
A_{swr} = switching range (to be set)

Filling in the known figures gives:

$$A_{loop} = 29.7\,dB - A_{swr} \approx 30\,dB - A_{swr} \tag{9}$$

In order to prevent howling, the loopgain has to be below 0dB with a margin of 10dB up to 20dB (A_{margin}). This guarantees good balance return loss figures and stable operation.

On the board the switching range is set to 40dB via the 365kΩ resistor R_{swr}, providing a howling margin A_{margin} of 10 dB.
If AGC is used, the gain from MIC+/- to QR of the TEA106x is lowered with approximately 12dB for line lengths below 5km. In that case the howling margin is in the order of 22dB.

3.7 Envelope detectors

The sensitivity of the detectors can be adjusted with R_{tsen} and R_{rsen}. They can best be adjusted such that currents in the order of $10\mu A_{rms}$ are flowing through them at nominal signals. The DC blocking capacitors C_{tsen} and C_{rsen} form a high pass filter together with R_{tsen} and R_{rsen}.

The dial tone detector threshold is proportional to the value of R_{rsen}. In formula:

$$R_{rsen} = V_{dial} \times 78.7\ k\Omega \tag{10}$$

On the board R_{rsen} is set to 4.75kΩ giving a dial tone detector threshold of about 60mV between RIN1 and RIN2. At the telephone line this means the dial tone detector level is set to 100mV (receive gain TEA106x is -4.2dB).
Capacitor C_{tsen} in series with R_{tsen} blocks DC and forms a high pass filter (225Hz).

As explained in more detail in the appropriate application reports, the value of R_{tsen} can be determined with the following formula:

$$20\log(R_{tsen}) = 20\log(R_{rsen}) - A_{tx1093/4} - A_{tx106x} - A_{st} - A_{rx106x} + A_{tsen} - \frac{1}{2}A_{margin} \tag{11}$$

In which A_{tsen} is the gain from MIC to TSEN of the TEA1093/TEA1094 = 40dB

Filling in the known figures gives:

$$20\log(R_{tsen}) = 20\log(R_{rsen}) - 3dB \tag{12}$$

Thus Rtsen = 3.3kΩ.

With AGC active (affects A_{tx106x} and A_{rx106x} and A_{margin}):

$$20\log(R_{tsen}) = 20\log(R_{rsen}) + 3dB \qquad \textbf{(13)}$$

thus $R_{tsen} = 6.7k\Omega$

As a compromise between the two situations, R_{tsen} is set to $3.92k\Omega$ on the board. The 150nF capacitor in series with R_{tsen} blocks DC and forms a high pass filter (220Hz).

The charge and discharge times of both signal envelope detectors are proportional to the value of the capacitors C_{tenv} and C_{renv} giving a maximum rise slope in the order of 85dB/ms maximum, and a maximum fall slope of 0.7dB.ms.
The noise envelope timing is proportional to the value of the capacitors C_{tnoi} and C_{rnoi}. On the board 4.7µF is mounted at TNOI providing a maximum rise slope of 0.07dB/ms maximum, and a maximum fall slope in the order of 0.7dB/ms.

3.8 Switch over timing

The switch over time from Tx to Rx or vice versa is proportional to the value of capacitor C_{swt} connected to pin SWT. On the board 220nF is mounted for C_{swt}, resulting in switch over times of around 13ms.

The switch over time from Idle mode to Tx or Rx is also proportional to the value of C_{swt}. With the value chosen, switch over will take 4ms.

The timing from Tx or Rx to idle mode is determined by resistor R_{idt} in conjunction with C_{swt}. On the board 2.2MΩ is mounted for R_{idt} resulting in an idle mode timing of about 2s.

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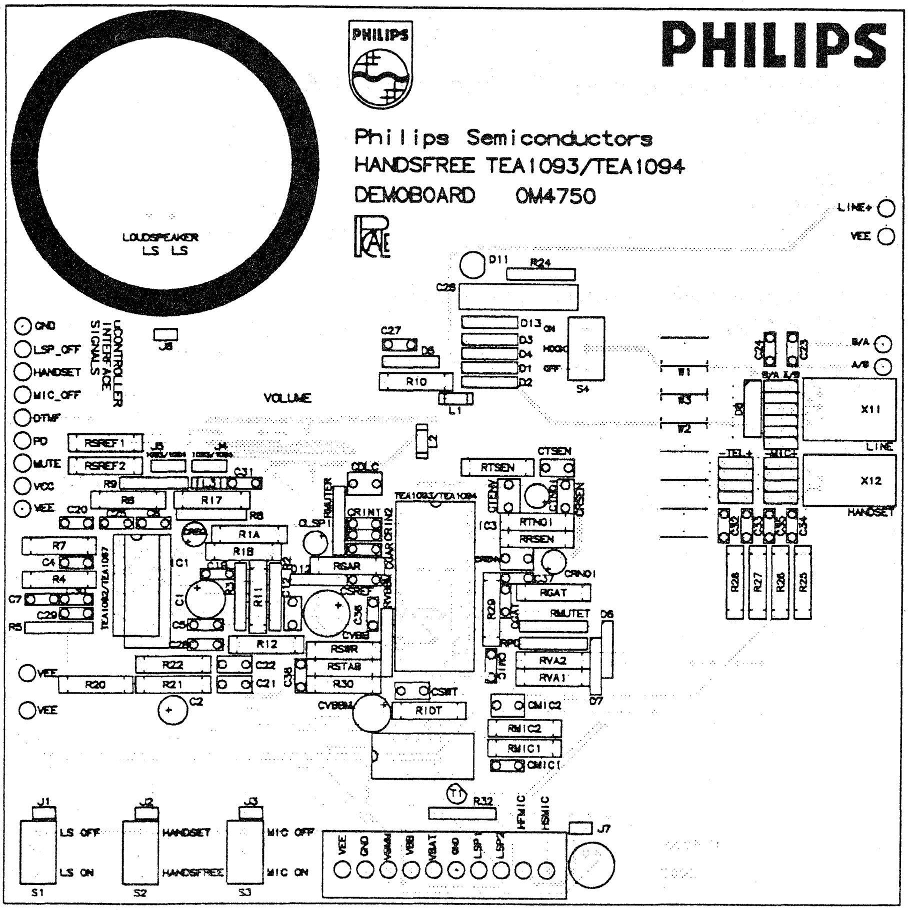


Figure 1 Position diagram of the demonstration board OM4750.

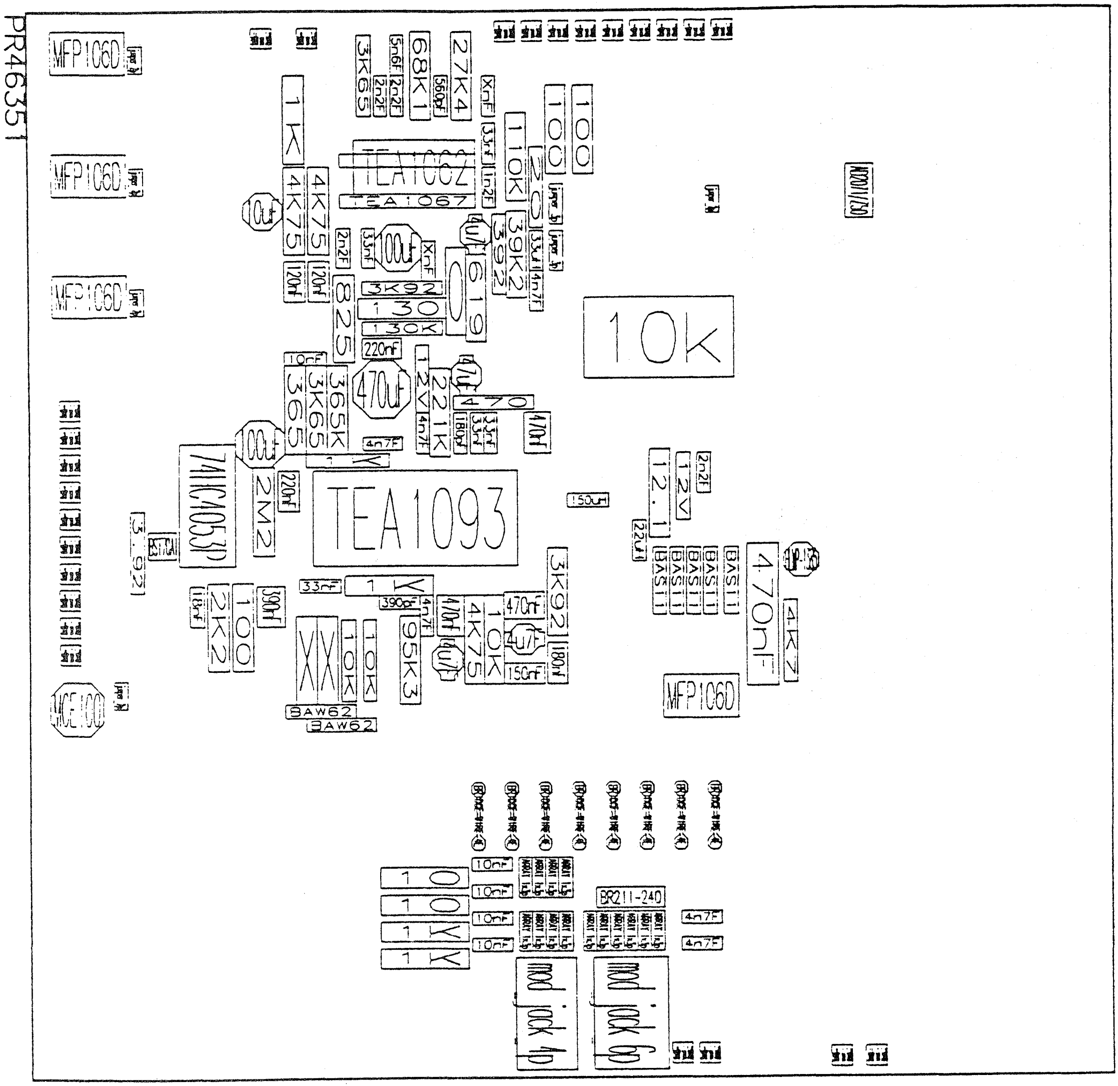
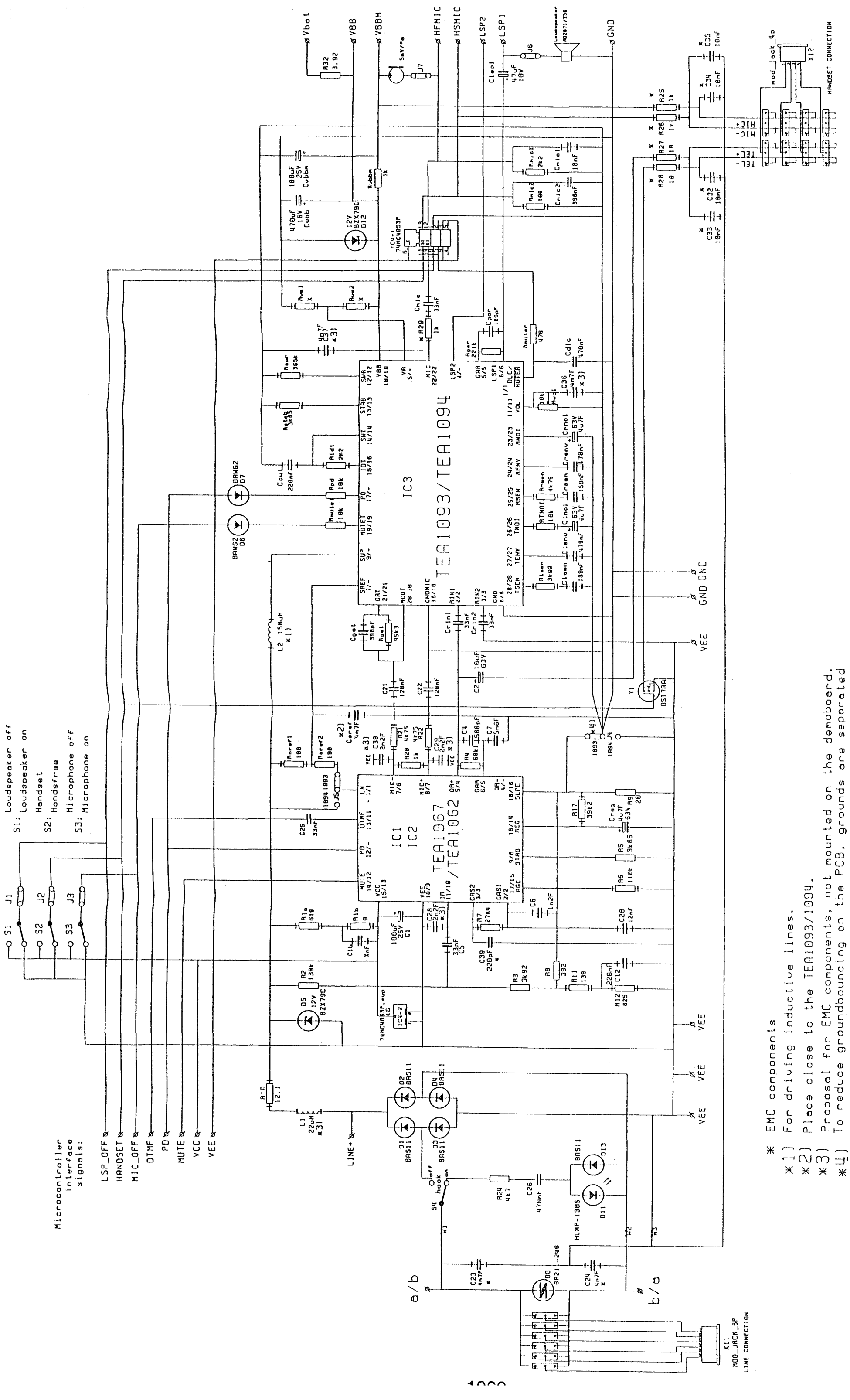


Figure 2 Position diagram with component types and values of the demonstration board OM4750.



* EMC components
*1) For driving inductive lines.
*2) Place close to the TEA1093/1094.
*3) Proposal for EMC components, not mounted on the demoboard.
*4) To reduce groundbouncing on the PCB, grounds are separated as indicated, and connected to a star point at J4.

Figure 3 Circuit diagram of the demonstration board OM4750.

APPLICATION NOTE Nr	ETT/AN90017
TITLE	Software controlled ringer for German market
AUTHOR	C. H. Voorwinden, K. Wortel
DATE	July 1990

1. Introduction and description of the concept

In telephone sets with listening-in it is possible to incorporate a software controlled ringing function. The application described is a proposal for a concept that makes use of this possibility and also fulfills the German TEL02 requirements. Special requirements like earth switch, long line breaks and redial are not taken into account.

The base of the set is formed by the speech-transmission circuit TEA1064A, the listening-in circuit TEA1085, the microcontroller with EEPROM PCD3346 and the DTMF-generator PCD3312.
The functional requirements of the TEL02 ringer specification concern the detection of the ring signal on the line and production of a ringing melody via the loudspeaker. For the detection of the ring signal its frequency and magnitude are used as parameters. The production of the ring melody must be controllable via keyboard entries. Parameters to be controlled are the melody, its frequency and its volume. The concept of figure 1.1 provides such.

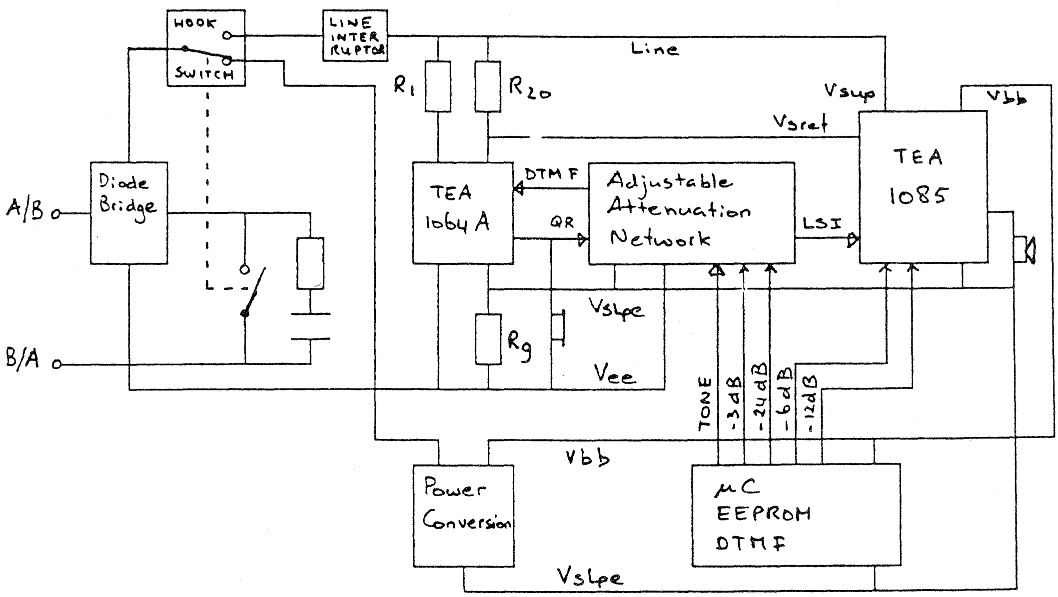


Figure 1.1: concept of the software controlled ringer

First the concept is presented and in the next chapters each block of the concept will be discussed in closer detail.
The set is connected to the line via a bridge and hook switches. When the handset is off-hook the line is connected to the speech-transmission circuit TEA1064A via a line interrupter. This is the normal speech-dialling condition. The line interrupter is used for pulse dialling and the TEA1085 can be used for listening-in.

When the handset is on-hook the line is connected to a power converter that converts the high voltage low current ring signal into a low voltage high current signal. The converted power is used to drive a low ohmic loudspeaker via the listening-in circuit TEA1085.
During the speech-dialling condition the biassing of the TEA1064A is controlled by the TEA1085. A reference voltage of 0.315 Volts is applied over R20=150Ω and as a result the TEA1064A is supplied with a constant current of 2.1mA. The rest of the line current is available for listening-in and the peripherals. The ground reference of the TEA1085 and the peripherals (microcontroller, DTMF-generator) is shifted with respect to the ground reference of the TEA1064A, ie. Vslpe in stead of Vee.
The output of the power converter is connected to the output of the TEA1085. As a result, during the ringing condition only the TEA1085 and the peripherals are supplied. The TEA1064A is not supplied. This is to ensure a minimum current consumption of the set.
In the block diagram of figure 1.1 the peripherals consists of a DTMF-generator that has to generate the ringing melody and a microcontroller that controls it. Both of course have their usual function during speech and dialling.
The AC outputs of the blocks are tied together via the 'adjustable attenuation network'. This network adapts, in some cases software controlled, the incoming signals like DTMF or earpiece signals to the correct values for further processing like listening-in or dialling.
One of the essential features of the software controlled ringer is volume control via keyboard entries. For a smooth volume control, steps of 3dB are needed. The listening-in circuit TEA1085 already has an attenuator of 6dB and an attenuator of 12dB integrated. An extra 3dB is included in the attenuation network. As a result steps of 3dB are provided. Because for the German market at least 10 steps are required also a 24dB attenuator is included in the attenuation network.

In chapter 2 the block noted as 'bridge, hook switches, line interrupter, power conversion' will be discussed. In chapter 3 an application of the speech-transmission circuit TEA1064A will be discussed that fulfills the German requirements, see appendix B. This German application is optimised for use with the listening-in circuit TEA1085. The application of the TEA1085 will be discussed in chapter 4. In chapter 5 the microcontroller and the DTMF-generator are discussed. The adjustable attenuation network is the subject of chapter 6. Then the complete concept is worked out and in chapter 7 the measurement results are shown. The electrical diagrams are placed in appendix A.

2. The line interface and power conversion

This part of the set is depicted in figure A.1 and consists of:

- a diode bridge
- two hook switches (drawn in on-hook position)
- a positive rail line interrupter
- set protection
- a power converter

The bridge is for double rectifying the line signal to ensure that only one polarity is applied to the set. To prevent breakdown of the circuits used, after the interrupter a 12Volts zener protection is placed. To protect the whole set, a 140Volts break over diode is placed in front of the bridge. The zener in the interrupter is present for protecting the gate of the PFET. The interrupter itself is needed for pulse dialling.
When the handset is on hook the telephone set is in the ringing condition. The switches are in that position that the power converter is connected to the line via the bridge and an RC-network. The capacitor is present for DC-blocking and according to TEL02 it is 0.8μF. The resistor is present for protecting the power converter from too large currents. In the presented application it is chosen to be 500Ω.
Power conversion during ringing is necessary. For generating the ringing melody a low ohmic loudspeaker is used (50Ω). To produce a sound pressure level high enough to fulfill the German requirements more current must be supplied to the loudspeaker than available from the line. The ring signal on the line has a fairly high voltage, about 50Vrms. Because for generating the ringing melody only a small voltage is needed, it is possible to convert the ring signal. The power converter which does such has to fulfill several requirements. It must sustain high voltages (140V), it must have a constant impedance ($\geq$8kΩ) and it has to produce information about the frequency of the ring-signal in the form of a waveform shaped signal. This signal can be used by the microcontroller to decide whether the ring frequency is within the TEL02 limits or not, see chapter 5. A type of converter suitable for this purpose is a switched mode power supply with additional features (eg. Motorola TCA3385 or discrete solution).
When the handset is lifted the switches will take the other position and the line is no longer connected to the power converter but to the line interrupter. Then the set is in the speech or dialling condition.

3. The speech-transmission part

Recently, transmission requirements in Germany have been changed and strengthened. Therefore the application proposal of [ref.1] does not fulfill the requirements for TEL02 of today. [Ref.1] was based on the requirements mentioned in 'FTZ pflichtenheft, FTZ 121 TR 8-8' of january 1988. Most important electrical differences are:

- Measurement of frequency curves performed with complex line termination (used to be 600Ω)
- Balance return loss: 300Hz - 500Hz: 14dB (was 12dB)
  500Hz -2500Hz: 18dB (was 15dB)
  2500Hz -3400Hz: 14dB (was 12dB),

with the complex reference impedance being:

Zref = (220Ω)+(820Ω//115nF).

Since for TEL02 also listening-in is required, the TEA1085 will be used in combination with the TEA1064A. This chapter describes an application of these circuits according the requirements mentioned.
Figure A.2 shows the circuit diagram used to fulfill requirements. Throughout this chapter, this figure is referred to.

Balance return loss setting:

As explained the balance return loss figures have been strengthened in Germany. The set impedance of a TEA1064A equipped telephone set is mainly determined by the impedance between the line and VCC1. In order to obtain sufficient margin the reference impedance is connected between LN and VCC1, see figure A.2.
Measurement results on balance return loss can be found in chapter 7.

Sidetone attenuation:

In Germany sidetone is measured acoustically. Therefore it is extremely difficult to obtain electrical requirements. Besides, the line is terminated with 35 different artificial lines. After 35 measurements the mean sidetone value is determined (weighted).
In [ref.1] the FTZ requirements of 1988 are used to work out an application proposal. The acoustical measurement mentioned in that report shows a good sidetone suppression. Therefore the same balancing is used here also. However, due to the fact that several components in the sidetone balance bridge are changed, bridge balance has to be recalculated.
For optimum side tone suppression, see [ref.2], it must hold that:

$$R9*R2 = Ri(R3 + R8) \quad \text{with } Ri = \text{part of Zset,} \qquad (1)$$

$$Zbal = \frac{R3(R8 + R9)}{R9*R2} * ZLine = \frac{R3(R8 + R9)}{Ri(R3 + R8)} * ZLine = K * ZLine. \qquad (2)$$

Transformation from total set impedance to a simplified network is shown schematically in figure 3.1.

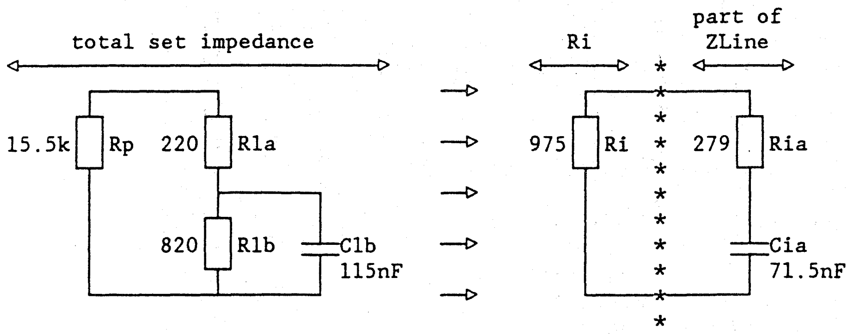


Figure 3.1: splitting set impedance into two parts
Rp = internal impedance of TEA1064A

As shown in figure 3.1 the set impedance is split up into a resistor Ri and a complex impedance which can be seen as a part of the line. With equation (2) the part of the line can be translated to Zbal if factor K is known.

The same K factor is used as in [ref.1]: K= 0.62.

Equation (2) rewritten gives: $R2 = \dfrac{R3(R8 + R9)}{R9*K}$. (3)

Combining equations (3) and (1) it follows that:

$$\left.\begin{array}{l} \dfrac{R3*R8}{K} + \dfrac{R9*R3}{K} = Ri*R3 + Ri*R8 \\ \\ \text{with:} \quad Ri = 975\Omega \\ \text{and taking:} \quad R3 = 3.92k\Omega \\ \qquad\qquad\quad R9 = 39\Omega \end{array}\right] R8 = 674\Omega.$$

A practical value is R8=681Ω. In [ref.1] R9 was chosen to be 27Ω, here R9 is chosen 39Ω to obtain a low attenuation of receiving signals. This will improve the noise performance in receiving condition.

With (3) and all known values it is obtained that:

$$R2 = \frac{3.92k\Omega(674 + 39)}{39*0.6242} = 114.8k\Omega.$$

A practical value is R2=115kΩ. Now only Zbal needs to be determined. Since the extra impedance introduced in parallel to the line and the scaling factor K are known this is a rather simple operation. In figure A.2 R11, C11 and C12 are based upon Zbal of [ref.1] again. R11a and C12a are the result of the parallel line termination caused by introducing a complex set impedance, see also figure 3.1.

DC setting:

In Germany the set resistance including bridge must be less than 470Ω in speech or DTMF-dial condition, or less than 325Ω in pulse dial condition. This resistance is measured with 'Speisebrücke A', a feeding bridge of 60V and 2530Ω series resistance (1kΩ internal, Rv=1530). This means that:

Vline < 9.4V at 20mA in speech or DTMF-dialling condition
Vline < 6.8V at 21mA in pulse dialling condition

In figure A.2 the line voltage is switched to a lower value during pulse dialling by means of T1 which is controlled via T2 by means of a microcontroller signal (Mute or NSA). T1 short circuits resistor R21 of 1kΩ and thus will decrease the voltage drop over it.
The current through R21 is determined by the TEA1085 which will divide the line current into a small constant current into the TEA1064A and a large current (remaining part of the line current) into the TEA1085 itself.
The current flowing through the TEA1064A (and thus through R21) is determined by resistor R20 and the DC reference voltage of 315mV typical over it (between SUP and SREF). Taking a value of 150Ω for R20 the current through the transmission circuit is 2.1mA and thus the voltage drop over R21, which is chosen 1kΩ will be 2.1V typical.
The reference voltage of the TEA1064A is set at the nominal value of 3.3V because then the maximum sending signal is limited to the required value of 6.3dBm even in worst case condition, see [ref.2].

During speech the line voltage can be given by:

Va/b = (Iline-1.5mA)*R9 + Vref + VR21 + VR20 + Vdiodes + Vinter,

in which:

Iline-1.5mA	= current through R9
Vref	= reference voltage TEA1064A of 3.3V ± 0.25V
VR21	= voltage drop over R21 with R21=1kΩ
VR20	= voltage drop over R20 which equals voltage drop SUP-SREF of TEA1085, typical 315mV. SUP-SREF = 275mV - 355mV
Vdiodes	= diode bridge voltage drop: 1.6V
Vinter	= voltage drop over interrupter: 0.3V
R9	= 39Ω
Iline	= 20mA.

As a result:

typical : Va/b = 8.3 V
worst case: Va/b = 8.9 V.

During pulse dialling the line voltage can be given by:

Va/b = (Iline-1.5mA)*R9 + Vref + VsatT1 + VR20 + Vdiodes + Vinter,

in which:

VsatT1 = the saturation voltage of T1: 0.1V
Iline = 21 mA.

As a result:

typical : Va/b = 6.4 V
worst case : Va/b = 6.7 V.

Frequency response and microphone gain setting:

Since the set impedance is complex (to obtain high BRL figures) a frequency dependent sending gain occurs. This is caused by the fact that the load at the line (pin LN) is not constant over frequency.
Frequency curves used to be measured at a 600Ω line termination in Germany. In that case the frequency curve was flattened (load at line more resistive) and a compromise could be made between BRL (Zset) and frequency response. The compromise towards Zset has become impossible due to more severe BRL requirements. Moreover, the line termination is changed to a complex impedance: Zref.

So, when the frequency response of the output stage is measured a load at the line of Zset//Zref is present. To compensate for the high frequency roll-off which will occur the network between GAS1 and GAS2 is used in figure A.2. The principle operation of this network is that the pole and zero occurring at the line due to the complex load impedance are cancelled by a zero and a pole respectively (pre-emphasis). The latter ones are created by R7a, R7b, R7c and C7c of figure A.2.

The gain of the TEA1064A used in the configuration of figure A.2 is set by means of R7a and R7b. For low distortion and accurate gain setting the internal impedance (3.48kΩ) of the TEA1064A between GAS2 and SLPE is short circuited for AC signals by means of C15=470nF. If 44.5dB of gain is required, see [ref.1], the value of the resistance between GAS1 and GAS2 can be calculated, see also [ref.2]:

$$R7a + R7b = 35.4k\Omega \qquad (4)$$

The GAS1-GAS2 network introduces:

$$\text{a pole: } Fp = 1/[2*\pi*C7c*R7c] \qquad (5)$$
$$\text{a zero: } Fz = 1/[2*\pi*C7c\{(R7a//R7b) + R7c\}]. \qquad (6)$$

The pole and zero occurring due to the complex line impedance can also be determined. For calculation of the actual load impedance at the line the TEA1064A internal impedance of 15.5kΩ has to be included. This gives the following result:

Fpole of lineload: 1740Hz (7)
Fzero of lineload: 7950Hz (8)

Now the values of the components between GAS1 and GAS2 can be calculated if one value is taken as a start.

Taking:

C7c = 10 nF

as a starting value and combining (5) and (8), because the pole generated by the GAS network should disable the zero caused by the line load, it follows that:

R7c = 2 kΩ.

Combining (6) and (7) and taking into account that R7a + R7b =35.4kΩ it follows that:

R7a = 25.5 kΩ
R7b = 10.0 kΩ.

High frequency roll-off and stability is obtained with C7a=470pF (f-3dB = 13.3kHz).
If for any reason extra components (e.g. EMC capacitor) are connected to the line they should be included in the load impedance. Most certainly they will shift Fpole (7) and Fzero (8) and thus must be included in the calculations for the GAS1-GAS2 components.
In chapter 7 the measurement of the microphone gain can be found.

Receiver gain setting:

In Germany receiver gain (Areq) is measured from source to telephone capsule outputs [ref.1]. The source is coupled to the line via Zref (was 600Ω). As an example the value for this gain is based upon [ref.1] and is taken -3dB. The gain from source to line (Aln) was measured to be -6.5 dB over the frequency range of 300-3400Hz.

The gain from the line to the input of the receiving amplifier (Ast) by the sidetone network is about -30.5dB. So the receiver amplification (ATS) must be set at:

ATS = Areq - Aln - Ast = 6.5dB + 30.5dB - 3dB = 34dB

The gain is set with resistor R4 of 143kΩ, see [ref.2]. High frequency roll-off is obtained with C4=150pF (f-3dB = 7.4kHz), see figure A.2. A low frequency roll-off can be obtained by the coupling capacitor C5 in combination with the receiving input impedance at IR of 20kΩ typical.
In chapter 7 the measurement of receiver gain can be found.

Start time:

For speech, thus not for ringing, the maximum allowed start-up time in Germany is 500ms. This means the telephone set must be fully operational within 500ms. The circuit of figure A.2 is completely operational after 150ms. This time was measured at a line current of 20mA with all capacitors discharged. An external start-up circuit as used with the TEA1064, see [ref.1], is not applied because the TEA1064A has an integrated start-up circuit.

Noise:

The noise output voltage at the line is required to be less than -75dBmp, with 0dBmp being 1mW in 600Ω psophometrically weighted (frequency mask). The circuit of figure A.2 generates only -79.5dBmp ± 0.5dBmp measured with a 200Ω load at the microphone inputs, so within the limits.

In receiving direction -78dBmp is required and -81.5dBmp ± 0.5dBmp was measured under the same conditions with an 68nF load at QR+. With respect to [ref.1] this noise figure is improved due to the fact that R9 is increased. Also R2 is decreased which results in a lower receive gain. Noise measurements were performed with line currents between 20mA and 60mA generated from a noise free voltage source with Zref in series.

4. The listening-in part

The TEA1085 is used for listening-in as well as for amplifying the ringing melody. The application depicted in figure A.3 is kept as close as possible to the standard application of [ref.3]. However, a few small changes had to be made.
The first change concerns the listening-in inputs LSI1 and LSI2. In the standard application LSI2 is connected via a capacitor to the ground reference Vee of the speech-transmission circuit TEA1064A. This because the outputs of the TEA1064A are referenced to Vee. In the presented application the listening-in input is referenced to Vslpe, this means LSI2 is connected via a capacitor to Vslpe. As a result the output of the DTMF generator and the listening-in input of the TEA1085 have the same ground reference.
The second change concerns the Larsen level limiter (LLL). The LLL prevents the set from howling during listening-in. However, during ringing no howling can occur. The ringing melody generated however, is of that loudness that the LLL detects howling via the microphone. As a result the gain of the TEA1085 will be reduced. This must be avoided because for ringing maximum volume is desired. To do such a pull-up resistor has to be connected to the capacitor C26 of the LLL. This pull-up may only be active during ringing and not during speech or dialling. To obtain this, the pull-up resistor can be (dis)connected under control of Vcc1 of the TEA1064A. This voltage is low during ringing and high during the other conditions. During pull-up a current of 30μA has to flow to C26, this leads to figure 4.1.

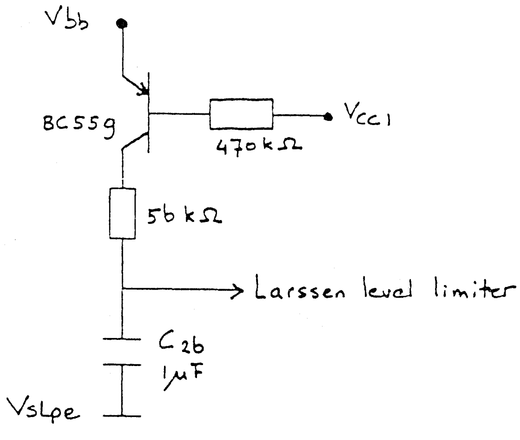


Figure 4.1: switched pull-up resistor

The third aspect that is different from the standard application is the charging of the capacitor of the dynamic limiter. When the TEA1085 is supplied via Vsup, the internal start circuit ensures that the dynamic limiter capacitor C28 is charged quickly. When the TEA1085 is supplied via Vbb the start circuit does not work and C28 is charged very slow. However C28 must be charged very quick. To obtain this a current must be applied to C28 during start-up. After pre-charging no external current may be applied because during ringing the dynamic limiter is essential to prevent

distortion. Either limitation in voltage or limitation in current can be expected.
When a ringing melody is generated the microcontroller has to switch the TEA1085 from standby into listening-in via the mute-input. Because this mute-input has a so called toggle function, the microcontroller has to generate a mute pulse with for instance a width of 5ms, see chapter 5. With this pulse the current source for the pre-charging of C28 can be controlled. The basic circuit that can be applied for pre-charging C28 is depicted in figure 4.2.

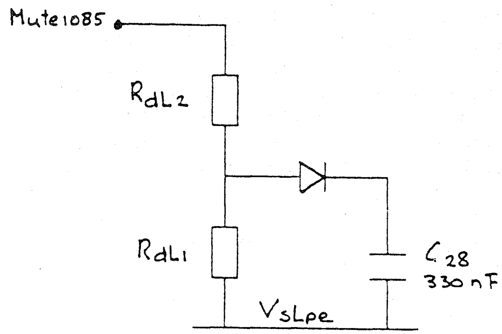


Figure 4.2: basic circuit for pre-charge of C28

The two resistors Rdl1 and Rdl2 divide the voltage of the mute pulse and the diode prevents discharging of C28 if the mute pulse is gone. The voltage over C28 reaches its maximum of 1.9 Volts when there is no limiting. When the voltage is lowered to 1.3 Volts (is 2 Vbe) the dynamic limiter becomes active. The mute pulse must be divided such that during charging the voltage over C28 doesn't exceed the 1.3 Volts. This must be done to prevent distortion in advance. Rdl1 is chosen 12kΩ to obtain an RC-time smaller then 5ms. Rdl2 must be smaller than Rdl1.
A fourth thing different from the standard application is the value of the capacitor C20. The standard application of the TEA1085 contains a C20 of 470μF. When this one is applied in the software controlled ringer, the start-up time of the set becomes too large. If C20 is too small the voltage over it is not smooth enough. A ripple will be on it due to the supply and demand on current. Or in other words, the capacitor will serve as a back-up for the power converter. A value of 220μF was found to be a good compromise between start-up time and residual supply ripple.
The ripple left on Vbb by this choice of C20 is hardly audible when the loudspeaker is used as a bridge tied load. This means, the loudspeaker is connected between the outputs QLS1 and QLS2 of the TEA1085.
The last thing different from the standard application is the added shunt resistor Rt. This resistor ensures that in the ringing condition the current sources at the input of the TEA1085 are not in the saturation region. As a result the total current consumption of the TEA1085 during ringing is less than without Rt. The resistor does not influence the characteristics of the speech-dialling mode.

5 The microcontroller and DTMF-generator

In the software controlled ringer concept the microcontroller and the DTMF-generator play a central role. The microcontroller represents the software control and the DTMF-generator the ringing. As mentioned in chapter 1, the PCD3346 and the PCD3312 are chosen. The reason for using the PCD3346 is the integrated EEPROM, see [ref.5]. Now first the PCD3312 is looked at and after that the PCD3346.

The PCD3312 is an I^2C controlled DTMF-generator, see [ref.4]. It is able to generate dual as well as single tones. A dual tone consists of a tone out of the so called low group and one out of the so called high group. Melody tones are generated at pin TONE at the same voltage as the high group, eg. 192mVrms typical.

The output TONE is biassed at half the supply voltage. Because of the ripple on the supply also the output signal will contain a ripple. To filter the ripple, the PCD3312 and the PCD3346 are supplied via a diode and an extra capacitor (Cbb'), see figure A.4.
The frequencies of the single tones equal musical notes. The notes used for generating the melody are chosen conform the requirements:

A: 782.1 Hz, B: 1044.5 Hz, C: 1318.9 Hz.

In the PCD3346 it is possible to program several melodies, for instance ABC-ABC-ABC-etc. The repetition rate at which a group of tones is generated, the so called warble frequency, is also programmable. The melody and its warble frequency are to be chosen by keyboard entries.
The PCD3346 has EEPROM on board, see [ref.5]. With this combination it is possible to program for instance the ringing volume and store it for a longer time, even when no constant supply is present (line or battery). An example of the pinning of the PCD3346 is:

- positive and negative supply (VDD, VSS)
- keyboard sensing (ROW1-6, COL1-6)
- crystal oscillator (XTAL1, XTAL2)
- chip enable (CE/T0)
- reset (RESET)
- I^2C-bus (SDA, SCL)
- power down/ pulse dialling (PD)
- mute for speech transmission circuits (Mute106x)
- mute for listening-in circuit (Mute1085)
- ringing or speech mode (T1)
- volume control (Vol1-4).

Because the PCD3346 is referenced to Vslpe the dial pulse output will not be low enough to keep the line interrupter in conduction and therefore the dial pulse cannot be taken directly from (PD). The power down signal therefore has to be shifted to Vee. The extra circuit at pin PD takes care of this, see figure 5.1. The output of this circuit is named DP.

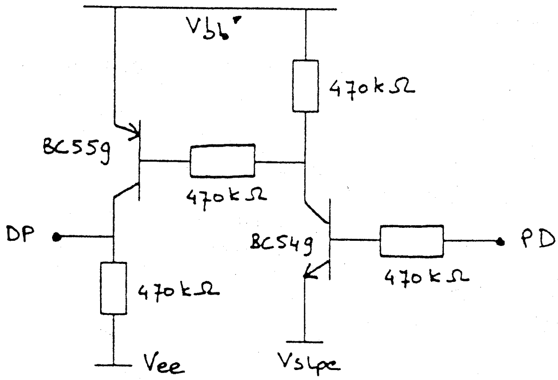


Figure 5.1: dial pulse shift circuit

For listening-in with the TEA1085, the microcontroller has to generate a pulse (Mute1085) because the mute function of the TEA1085 is a toggle. The pulse width is determined by software and 5ms is recommended. The pulse height equals the voltage difference between Vdd and Vss.
The PCD3346 has to distinguish the two conditions ringing and dialling/ speech. When during the ringing condition the set is supplied the PCD3346 has to start determining the ring frequency. After that, together with the DTMF-generator, it has to generate the melody. When the set is supplied during dialling-speech condition this must not take place. To inform the PCD3346 about the situation pin (T1) is created. Just like in chapter 4, the voltage Vcc1 is used as the detection signal. During ringing condition Vcc1 is low, during dialling-speech condition it is high.
One of the tasks of the microcontroller is detecting the ring frequency. For this purpose no extra pin is needed. With the circuit of figure 5.2 at pin (CE/T0), this pin gets a double function. During dialling-speech condition Vcc1 serves as the chip enable signal. During ringing the signal containing ring-frequency information, for instance a wave form shaped signal see chapter 2, can also be used as the chip enable.

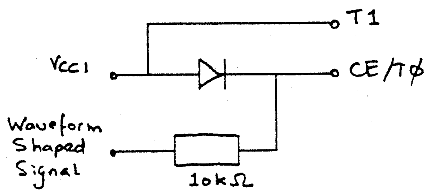


Figure 5.2: chip enable and frequency determination via pin CE/T0 and ring-speech detection via pin T1

The volume control is one of the essential parts of the software controlled ringer. As explained in chapter 1 four attenuators are used: two on the TEA1085 and two in the attenuation network, see chapter 6. The software of the PCD3346 must be suitable for programming the ringing volume during ringing via keyboard entries.

6. The adjustable attenuation network

The complete attenuator can be split up in four parts as depicted in figure 6.1.

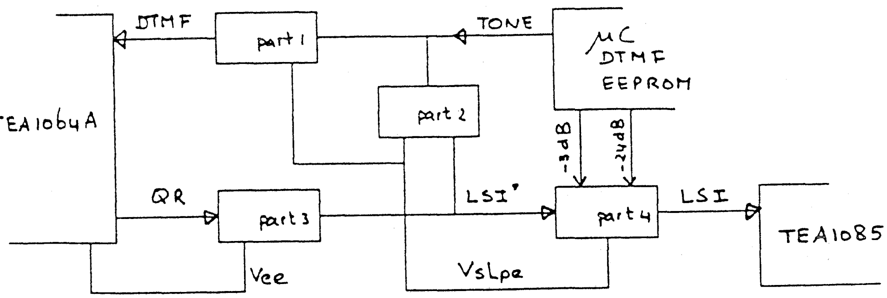
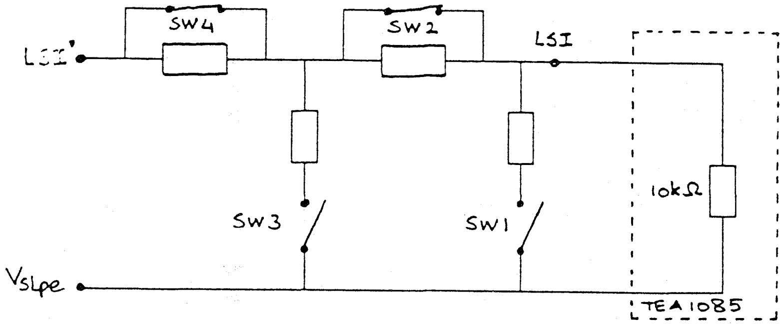


Figure 6.1: adjustable attenuation network

Part 1 adapts the output signal of the DTMF-generator (here named as TONE) to obtain the desired DTMF level on the line during DTMF dialling. Part 2 attenuates the signal from TONE to the maximum allowed level for port LSI1 of the TEA1085. Part 3 more or less has the same function for the signal coming from the earpiece. Part 4 consists of the attenuators of 3dB and 24dB. Both during ringing and during listening-in it is possible to adjust the attenuation of part 4 via the microcontroller.
The TONE output of the DTMF-generator, the DTMF input of the speech circuit TEA1064A and the LSI1 input of the TEA1085 are referenced to Vslpe. The earpiece output QR of the TEA1064A however is referenced to Vee. Therefore in part 3 a reference shift is needed.

The interaction of the various parts is kept small by designing with constant input and output impedances. In the solution of figure A.5 this is done.

First the design considerations of part 4 will be presented. After that the considerations of the other parts. The component names in this paragraph do not correlate with the component numbers of the figures in appendix A.
Part 4 consists of two attenuators of 3dB and 24dB. They consist of resistive voltage dividers which can be switched on and off. Figure 6.2 gives the general circuit of the two attenuators.



SW1	SW2	SW3	SW4	Att
0	1	0	1	0 dB
1	0	0	1	3 dB
0	1	1	0	24 dB
1	0	1	0	27 dB

0 = Open
1 = Closed

Figure 6.2: general circuit of the 3dB and 24dB attenuators

Before the values of the resistors are determined first the switching principle will be discussed.
When no attenuation is desired then switches SW1 and SW3 are open and switches SW2 and SW4 are closed. When an attenuation of 3dB or 24dB is desired then SW1 respectively SW3 are closed and SW2 respectively SW4 are opened. The switches used are the so called transmission gates or analog switches. Four of them are integrated in one circuit, the HCT4066 block. To drive this block two inverters are necessary. For the switches figure 6.3 applies.

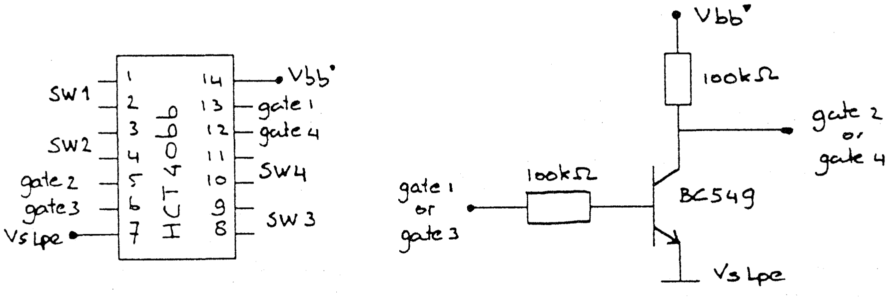


Figure 6.3: switches of the 3dB and 24dB attenuators
gate - low, then switch is open

In general for voltage dividers the diagram of figure 6.4 holds. Here R1 and R2 are the resistors of the voltage divider and Rp is the parallel impedance of the voltage divider for instance formed by the input impedance of an IC.

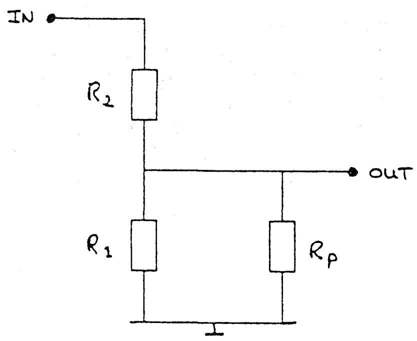


Figure 6.4: general voltage divider with load

For the attenuator of 3dB the parallel impedance is the input impedance of the TEA1085 at pin LSI1. For the 24dB attenuator the parallel impedance is either the input impedance at pin LSI1 or the input impedance of the 3dB attenuator. Because the attenuation has to remain the same with both loads it can be stated that the attenuators must be transparent, thus the input impedance at pin LSI' must be equal to the input impedance at pin LSI1 of the TEA1085 regardless the fact that attenuators are switched on or off. If A is the attenuation factor, it can be deduced that the following two relations must hold:

$$R1 = Rp * A/(1-A)$$
$$R2 = (1-A) * Rp$$

For the 3dB attenuator it follows that (E96-values between brackets):

A = 0.71, Rp = 10kΩ, R1 = 24.2 kΩ (24.3 kΩ), R2 = 2.92 kΩ (2.97 kΩ).

For the 24dB attenuator it follows that:

A = 0.063, Rp = 10kΩ, R1 = 673 Ω (681 Ω), R2 = 9.37 kΩ (9.43 kΩ).

These values can be found in figure A.5.

The attenuation of part 1 can be calculated when the desired DTMF level on the line and the microphone gain are known. The desired level on the line is -6dBm and the electrical microphone gain 44.5dB. The DTMF gain is 26dB smaller than the microphone gain and therefore 18.5dB. It follows that the attenuation of part 1 is around 0.25. With the same philosophy as with part 4 it can be deduced that:

A = 0.25, Rp = 20 kΩ, R1 = 6.67 kΩ (6.81 kΩ), R2 = 15.0 kΩ (15 kΩ).

These values can be found in figure A.5. The input impedance of part 1 has become 20kΩ. An other approach and tolerance considerations are given in [ref.6].

Part 2 and part 3 of the attenuation network are discussed together. Part 2, a simple voltage divider, has to attenuate the output signal of the DTMF-generator to approximately 22mVrms because for higher voltages the dynamic limiter of the TEA1085 will act upon the signal and this is not wanted. Part 3 has to shift the reference. The earpiece output QR+ of the TEA1064A is referenced to Vee while the input LSI1 of the TEA1085 is referenced to Vslpe. The circuit which does such is depicted together with part 2 in figure 6.5.

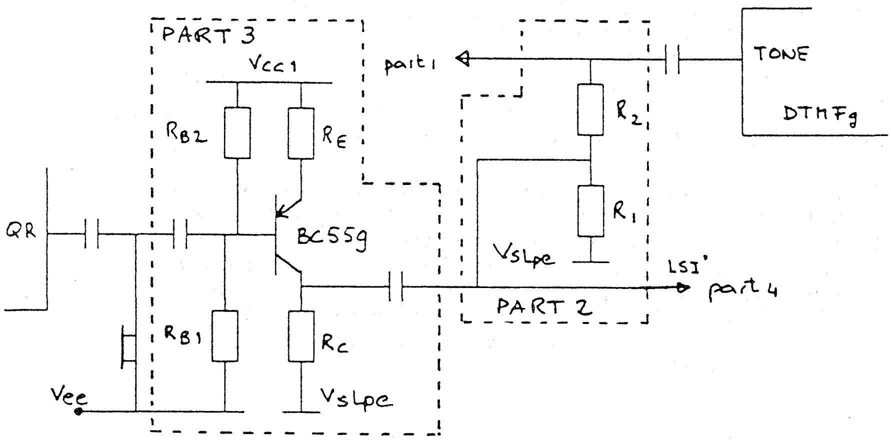


Figure 6.5: principle circuit of part 2 and part 3

For correct biassing of the transistor a capacitor is needed in front and at the end of the transistor stage. It also follows that:

RB2 ≃ RB1 but with RB2 ≤ RB1
RC ≃ RE but with RC ≤ RE.

This is needed to prevent the transistor from saturation. When for simplicity RC and the output impedance of part 2 is chosen to be 10kΩ, for part 2 it follows that:

R1 = 10kΩ / (1-3A)
R2 = 10kΩ / 3A.

With an attenuation of 22mVrms/192mVrms≃1/8 it follows that (chosen values between brackets):

A = 0.125, R1 = 14.8 kΩ (15.0 kΩ), R2 = 26.7 kΩ (27.4 kΩ)

These values can be found in figure A.5.

For part 3 the next values for resistors are calculated and chosen and can be found in figure A.5:

RB1 = 100 kΩ, RB2 = 47 kΩ,

RC = 10 kΩ, RE = 10 kΩ,

and thus: A = 0.33.

If more attenuation is desired a resistive voltage divider can be set in front of part 3. As an example the 1kΩ and 82.5kΩ are drawn in figure A.5.

Tolerances of the input and output impedances of the IC's of about 30% have only little influence on the attenuation network. It introduces a deviation in the attenuation of part 4 smaller than 0.5dB. If improvement is wanted, resistors have to be connected in parallel with the input and output impedances to become less dependent of their values.

7. Measurements

A set has been build according the diagrams of appendix A and several measurements have been done on the speech-transmission part and on the ringing function.

Speech-transmission:

Figures 7.1, 7.2 and 7.3 give the performances that are changed with respect to [ref.1]: balance return loss, microphone gain and receiver gain.

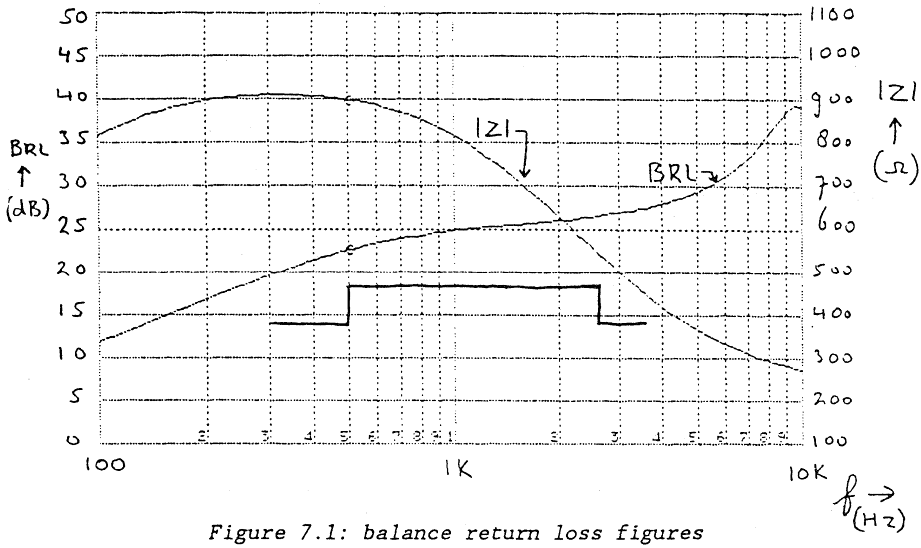


Figure 7.1: balance return loss figures

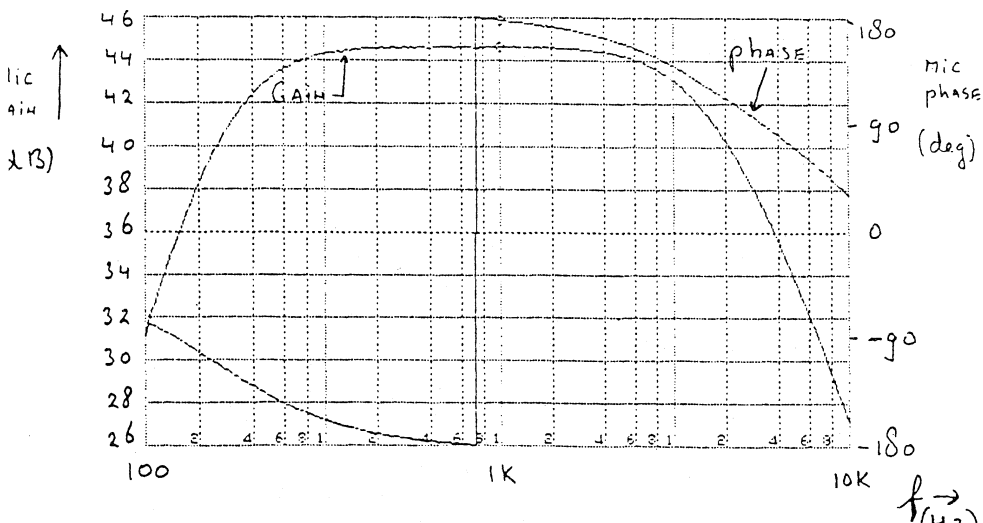


Figure 7.2: microphone gain versus frequency

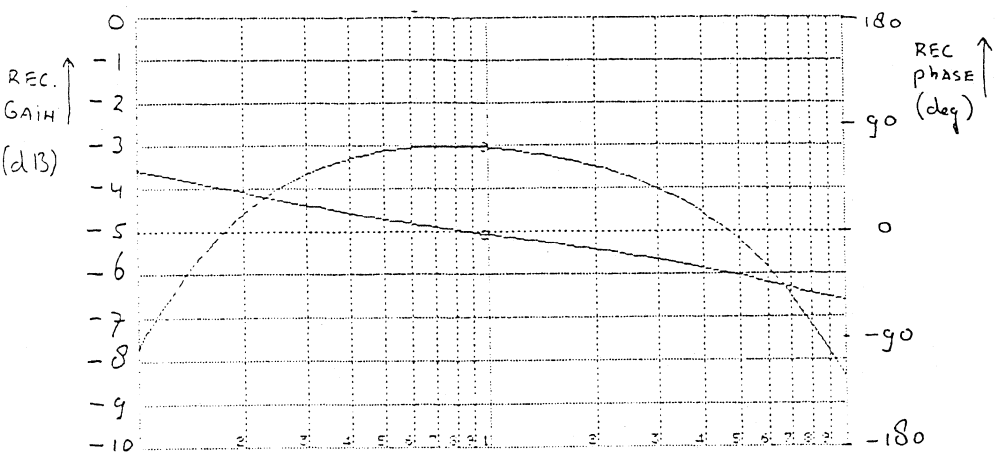


Figure 7.3: receiver gain versus frequency

Ringing:

The most important measurements for ringing are:

- input impedance
- output power
- start up-time
- volume control

All measurements for ringing were done with and without the voltage divider at the input, see appendix B. The ringing signal applied is 50Vrms at the frequencies 25Hz and 50Hz.

The input impedance during ringing contains a power converter, the bridge and the RC-combination, see figure A.1. The input impedance is measured by measuring the rms input current of the set at the ringing voltage of 50Vrms. It follows that

$|Zin| = 8.3\ k\Omega$ at $fr = 50$ Hz

$|Zin| = 10.6\ k\Omega$ at $fr = 25$ Hz.

The input impedance required is $|Zin| \geq 8\ k\Omega$ which is fulfilled.

For determining the sound pressure level, no acoustical measurements are done. Instead an electrical analogon is made. With a Philips loudspeaker AD2071-Z50 mounted on an IEC-baffle an electrical power of 550mW is needed to produce a sound pressure of 90dB(A) in open air at a distance of 0.5 meter. The values obtained in open air at this distance are 17dB(A) smaller than the values obtained in the special box, see appendix B.
The output signal is a warble of the three ground frequencies 782Hz, 1044Hz and 1318Hz. The sound pressure level therefore is an averaged value of the three frequencies. The sound pressure level can be calculated when the

output power is known. Then as a result table 7.1 follows.

	ring frequency	calculated SLoa(dB(A))		required SLoa(dB(A))
		SEL	BTL	
with voltage divider	25 Hz	69.9	64.9	60
	50 Hz	66.7	60.4	60
without voltage divider	25 Hz	75.6	76.0	71
	50 Hz	76.0	79.4	71

Table 7.1: sound pressure level
Sloa - sound pressure level in open air at a distance of 0.5 meter
SEL - loudspeaker connected as a single ended load
BTL - loudspeaker connected as a bridge tied load

As it can be seen the output power of the set is sufficient.

The measured start-up time (typical) is shown in table 7.2. The start-up time is mainly caused by charging of the capacitors of the set and determining the ring frequency. The conclusion is that the set starts up within the practical requirement of 200ms.

	fr(Hz)	time(ms)
with voltage divider	25	160
	50	150
without voltage divider	25	135
	50	90

Table 7.2: start-up time of the set during ringing

An essential part of the concept is the volume control via 3dB steps. The 16 available steps are tested for a set without the voltage divider at the input. Table 7.3 gives the results of the measurements for the loudspeaker connected as a bridge tied load. Although only the electrical power is measured the sound level can be calculated.

step (dB)	sound level (dBA)	step (dB)	sound level (dBA)	step (dB)	sound level (dBA)	step (dB)	sound level (dBA)
0	79.4	-12	71.8	-24	60.6	-36	47.9
-3	79.4	-15	68.8	-27	57.7	-39	45.0
-6	77.7	-18	65.6	-30	53.9	-42	41.8
-9	74.8	-21	62.6	-33	51.0	-45	38.9

Table 7.3: volume control

The first steps of the volume control are not totally effective because of two limiting effects. First, when the signal at the input LSI1 of the TEA1085 exceeds the 22mVrms, the output of the TEA1085 will not react upon variations at pin LSI. Second, the output signal is limited in voltage. At lower ring-voltages or with the voltage divider at the input of the set current limiting occurs. Measurements with the voltage divider confirm this. As a result the range of volume control varies. The minimum sound pressure level however is independent of the line input voltage. With this configuration the minimum value of the sound pressure level is about 39dB(A) for a bridge tied load and 36dB(A) for a single ended load. The TEL02 specifications require a minimum value between 33dB(A) and 48dB(A) at 0.5 meter distance.

8. Conclusions

The proposal for a software controlled ringer for German market described fulfills the essential requirements. This is obtained by using the Philips speech-transmission circuit TEA1064A, the listening-in circuit TEA1085, the microcontroller PCD3346 which has EEPROM on-board, the DTMF-generator PCD3312 and a power converter.
The speech-transmission requirements are fulfilled by a modified application of the speech-transmission circuit TEA1064A. The ringer requirements are fulfilled by adding some external circuitry to the listening-in circuit TEA1085 and by using dedicated software.
The number of IC's used can be reduced by the new series of microcontrollers that will become available, the PCD3351, which has a DTMF-generator on board.

References

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   Philips Components laboratory report ETT8802
   K. Wortel

2. Application of the versatile speech/transmission circuit TEA1064 in full electronic telephone sets
   Philips Components application report ETT89009
   F. v. Dongen, P.J.M. Sijbers

3. Preliminary data sheet TEA1085 listening-in circuit for line powered telephone sets
   Philips Components application note ETT/AN90008
   F. v.Dongen, P.J.M. Sijbers

4. Philips Components Data handbook IC03 1989
   IC's for telecom, bipolar, MOS, subscriber sets, cordless telephone

5. Philips Components Development data sheets PCD3346

6. The coupling network between the DTMF-generator PCD3311/12 and the transmission circuit TEA1060/61
   Philips Components laboratory report ETT8503
   J. Mulder

Appendix A. Diagrams of the software controlled ringer

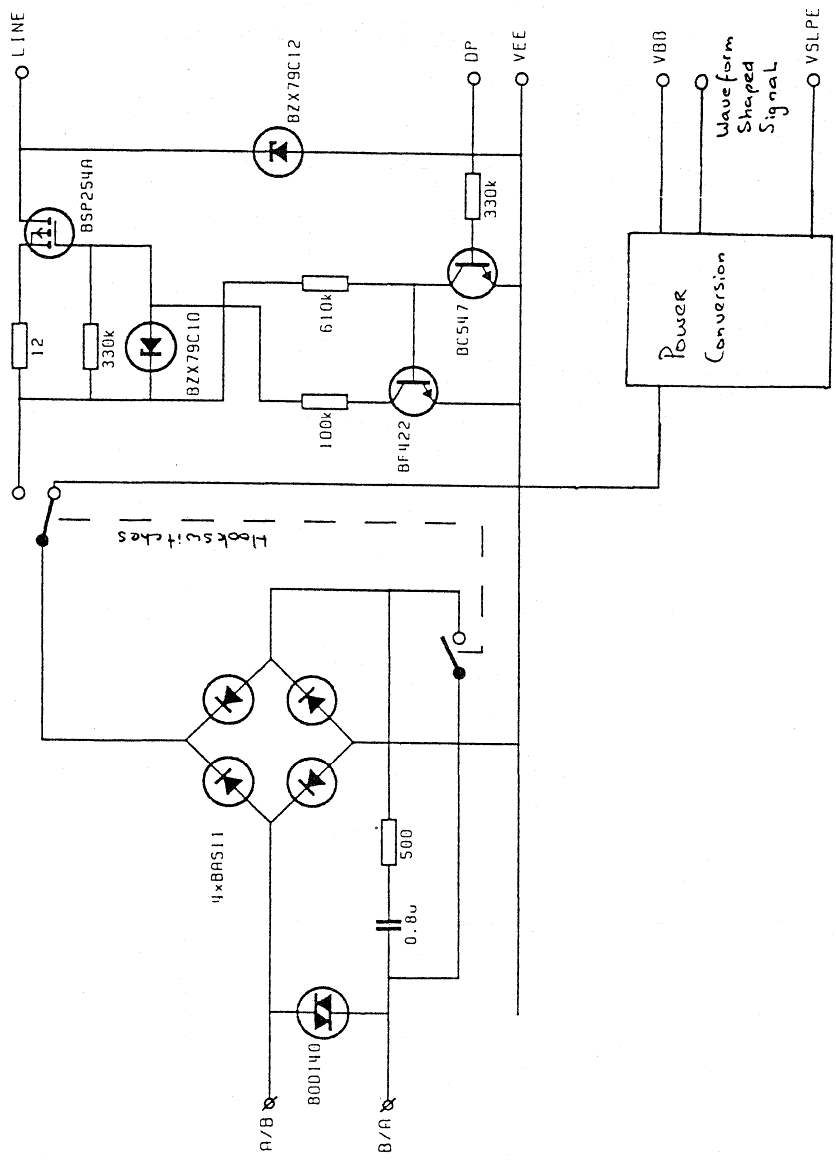

Figure A.1: the line interface

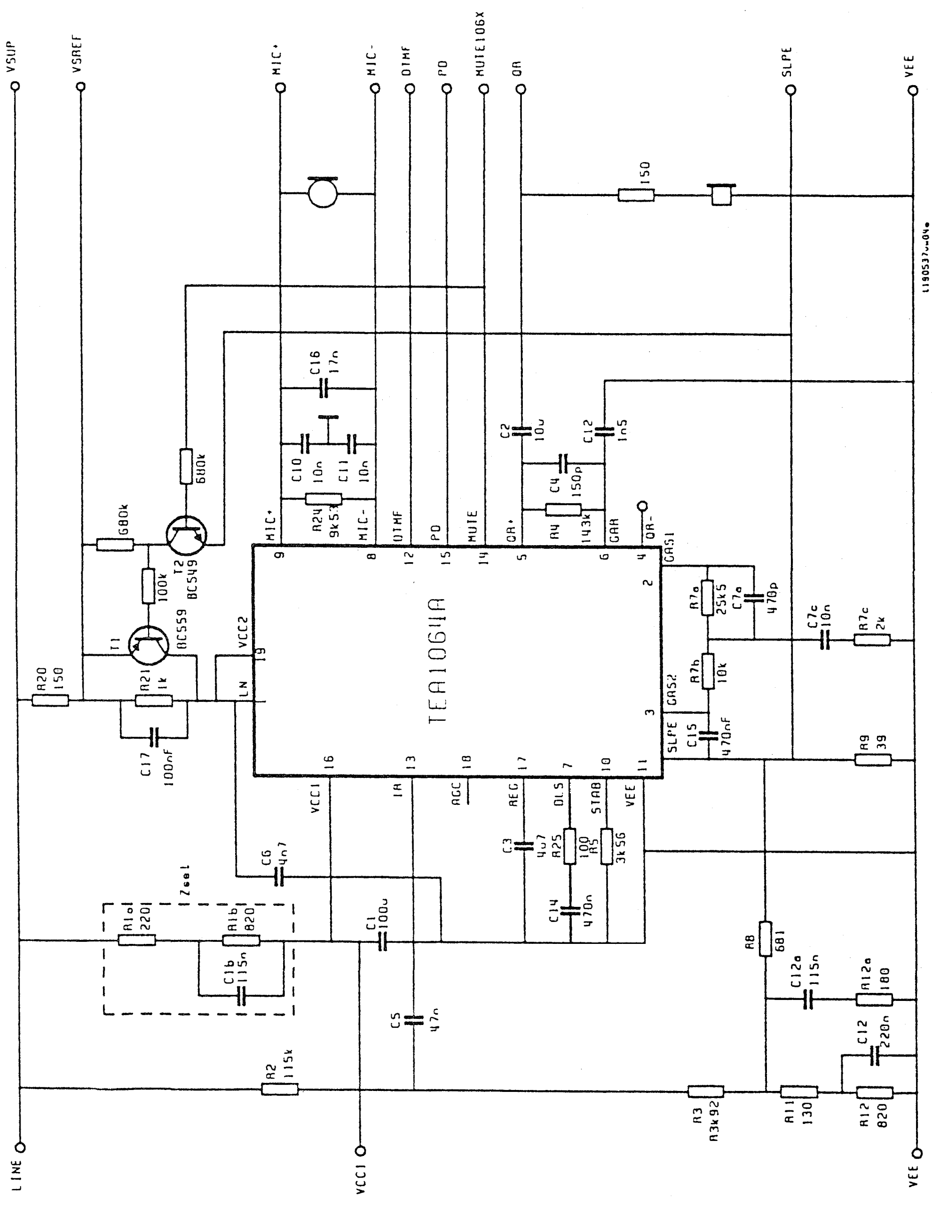

Figure A.2: the speech transmission circuit TEA1064A

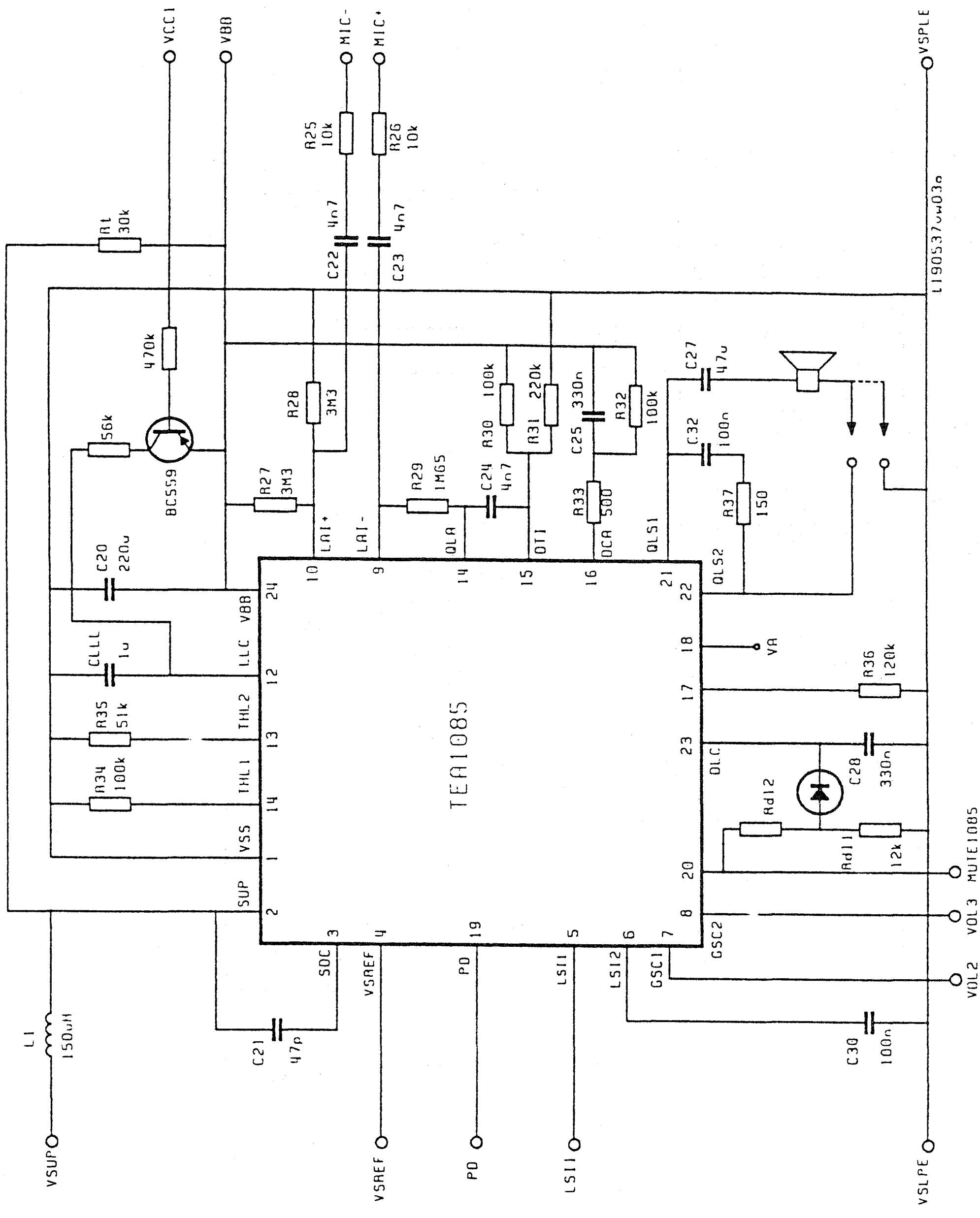

Figure A.3: the listening-in circuit

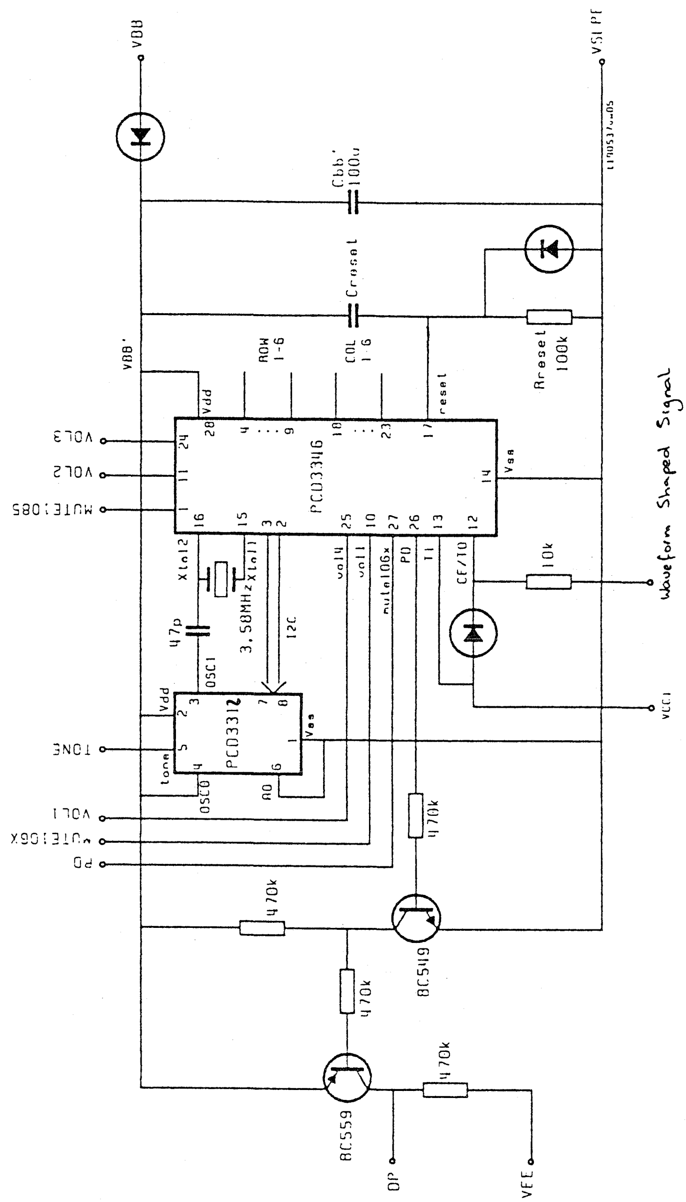

Figure A.4: the microcontroller PCD3346 and DTMF-generator PCD3312

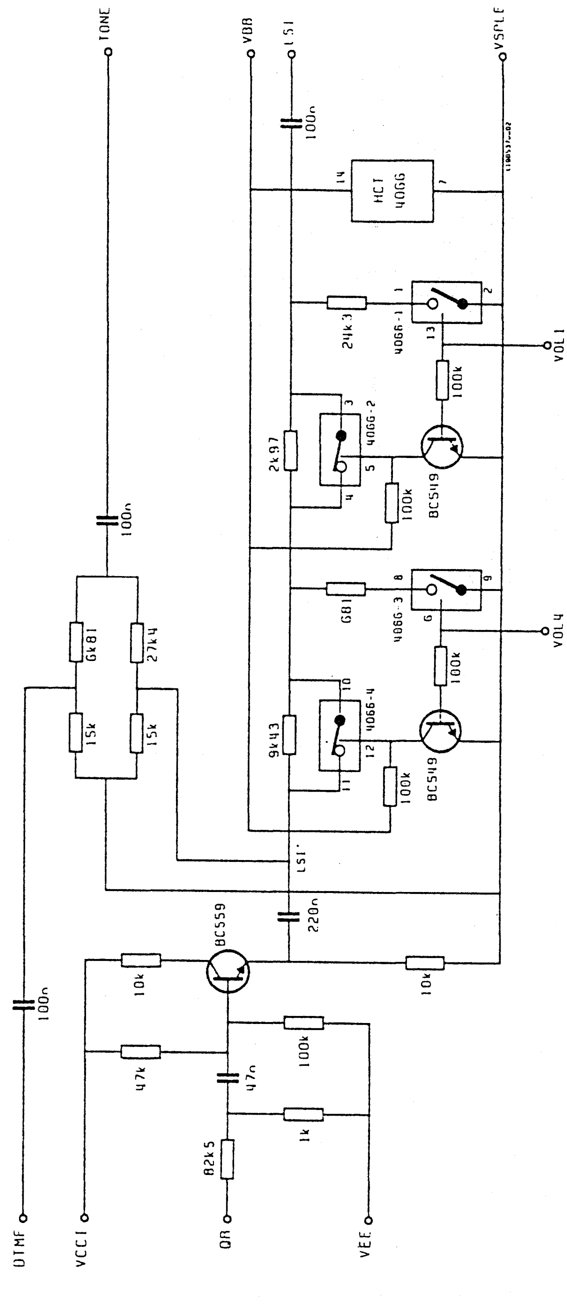

Figure A.5: the adjustable attenuation network

Appendix B. German TEL02 requirements

The specifications given in this appendix is a mixture of the specifications of the German 'Bundespost' given in TL5805-3112 (august 1987) and ETR2 (april 1989) and information from Valvo Hamburg. Only the essential ones are given here.

- No energy may be drawn from the line during inactive call.
- During ringing no external energy sources are allowed, all energy must be drawn from the line.
- In the ringing condition the set is connected to the line via an RC-combination as in figure B.1. The capacitor is for DC-blocking and the resistor is for current protection. Cs = 0.8 μF is required, Rs = 500 Ω is chosen.

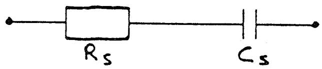

Figure B.1: RC-combination to the line

- The impedance of the ringer (Zin) including the RC-combination is according to

  $|Zin| \geq 8\ k\Omega$.

- The time-constant of the input must be smaller then 10ms.
- Some specifications have to be fulfilled with the voltage divider network of figure B.2 in front of the set.

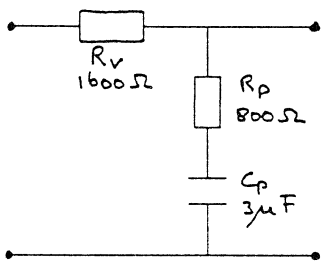

Figure B.2: voltage divider network

This network simulates parallel operation.
- At a ringing voltage (Vr) of

  Vr = 50 Vrms,

  with or without the network of figure B.2, the ringer must work for the ring-frequencies (fr) in the interval

$23\ Hz \le fr \le 54\ Hz$.

- The ringer may not work at the same Vr with the resistor Rv = 5 kΩ.
- The ringer may not work with

$Vr = 75\ Vrms$ and $fr \le 18\ Hz$, $fr \ge 60\ Hz$.

- The target specification for the maximum start-up and switch-off time is 100ms. For the moment as a practical requirement a maximum of 200ms is allowed.
- Ten ringing melodies must be programmable, composed out of the three base tones

800 Hz, 1067 Hz, 1333 Hz.

The tolerance of the generated tones is 5%.
- The follow-up of the tones is the so called warble frequency. It must be possible to vary the warble frequency between 2.5Hz and 25Hz.
- The acoustic sound level (SL) to be generated is

$SL \ge 88$ dB(A)
$SL \ge 77$ dB(A) with the network of figure A.2

This is measured for the highest warble frequency and for the complete ring-frequency range.
- The acoustic measurements take place in a special box. The sound level measured in it is 23dB(A) higher compared to the sound level measured in open air at a distance of 1 meter
- The volume of the ringing melody must be controllable in at least 10 steps to a minimum sound pressure between 50dB(A) and 65dB(A).
- Protection must be sufficient for a ring signal of 115Vrms, 50Hz during 15 seconds.

APPLICATION NOTE Nr	ETT/AN92010
TITLE	User's manual OM4723 DEMO board: PCD3330-1/ TEA1067/ TEA1083A featurephone application
AUTHOR	J. C. F. van Loon
DATE	July 1992

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Figures and Tables:

1. INTRODUCTION

Philips provides a wide range of IC's specially developed for telephony applications.
In order to demonstrate the application possibilities of the PCD3330-1 this featurephone DEMO board was developed, which realizes the following functions:

* Speech/transmission.
* Call progress monitoring.
* Ringer detection and generation
* Dialling features including:
    - Pulse, DTMF and Mixed mode dialling
    - Last number redial.
    - Repertory dialling.
    - On-hook dialling.

The following provides a brief description of the main components used in this featurephone:

*** PCD3330-1:**

The PCD3330-1 is a mixed mode multi-standard repertory dialler/ringer. The 13 repertory numbers and redial are stored in EEPROM so that memory retention is guaranteed for 10 years without using a battery backup. The various country specifications can be fulfilled by changing a few bytes in EEPROM which contain the different timing and dialling procedures. The incoming ringer frequency is measured and a ringer melody is generated via the special ringer melody output and supplied to the ringer output stage.

*** TEA1067:**

Low-voltage versatile speech/transmission IC performing all speech and line interface functions required in fully electronic telephone sets. The IC offers the possibility to connect a classical set in parallel. Other members of the TEA106X-family can be used too.

*** TEA1083A:**

Call progress monitor circuit.

*** BSP254A:**

High-voltage vertical DMOS FET for line-current interruption, flash and on-hook dialling.

*** PCF8581:**

With this external 128 bytes EEPROM IC, the internal EEPROM of the PCD3330-1 can be loaded with the correct country specifications. When done this PCF8581 can be removed.

This report gives an extensive description of the circuit, a guide line for setting parameters, and some brief application information as well.

2. HARDWARE DESCRIPTION

This printed circuit board contains all the electronic circuitry required for a featurephone.

It is provided with the PCD3330-1, a TEA1067 speech/transmission IC (other members of the TEA106X-family can be used too), a TEA1083A call progress monitor IC, and a PCF8581 EEPROM IC for storing new EEPROM parameters in the PCD3330-1.

Figure 1 shows the block-diagram of this circuit.

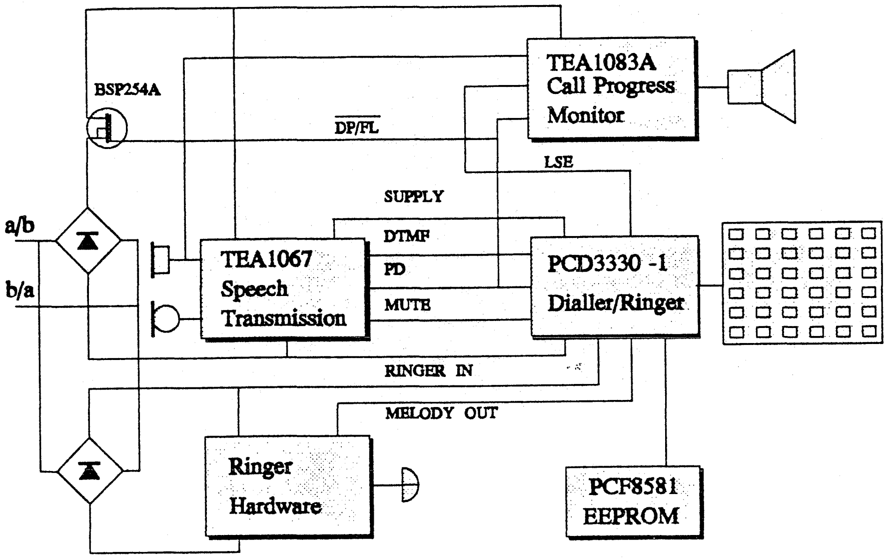


Figure 1. Block diagram of the printed circuit board

The printed circuit board contains the following parts:

* Power supply
* Dialling
* Transmission
* Call progress monitor
* PCD3330-1 interface
* Ringer interface
* PCD8581 EEPROM

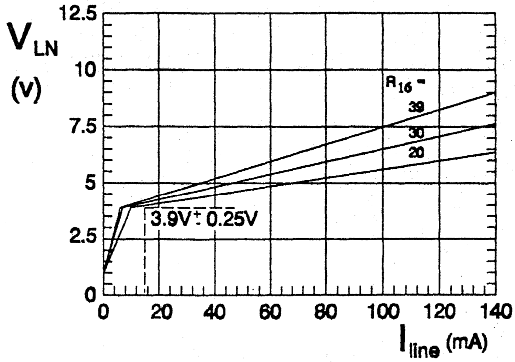
The in this printed circuit board mounted PCD3330-1 is loaded with the parameters which are underlined in Table 2 "Option bit status and location", for changing the internal parameters see chapter "PCF8581 EEPROM" or/and see chapter "EEPROM organisation and programming procedures".

2.1. POWER SUPPLY

The ICs are supplied with current from the telephone line; the PCD3330-1 is supplied via VCC of the TEA1067, and the TEA1083A via LN of the TEA1067.

For effective operation of the telephone set it must have a low resistance to d.c. and a high impedance to speech signals (300 to 3400 Hz). This is realized with the TEA1060-family by incorporating a voltage regulator in series with an artificial inductor in the IC. The internal voltage regulator generates a temperature compensated reference voltage. This internal voltage regulator must be decoupled by a capacitor C_{12} between REG and VEE. C_{12} has been chosen to give optimum start-up time of the circuit. It means that the voltage regulator starts-up after the smoothing capacitor C_{20} has been charged.
In Figure 2 the voltage V_{LN} at pin LN is given as a function of the line current I_{line} for three different values of R_{16}. $R_{16} = 20\Omega$ is mounted on the board.

(a) TEA1067



(b) Other TEA1060-family

* ± 0.2V for TEA1060/61/66

± 0.25V for TEA1068

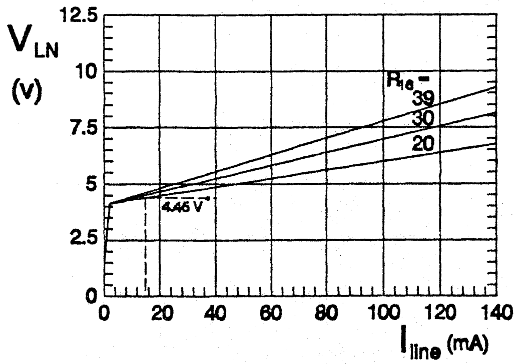


Figure 2. DC characteristics

With the connection R_{17} between IC_2-TEA1067 pins REG and SPLE the reference voltage V_{ref} ($V_{LN\text{-}SLPE}$) increases (see Figure 3). R_{17} = 20 kΩ is mounted on this board, i.e. V_{ref} = 4.7 V.

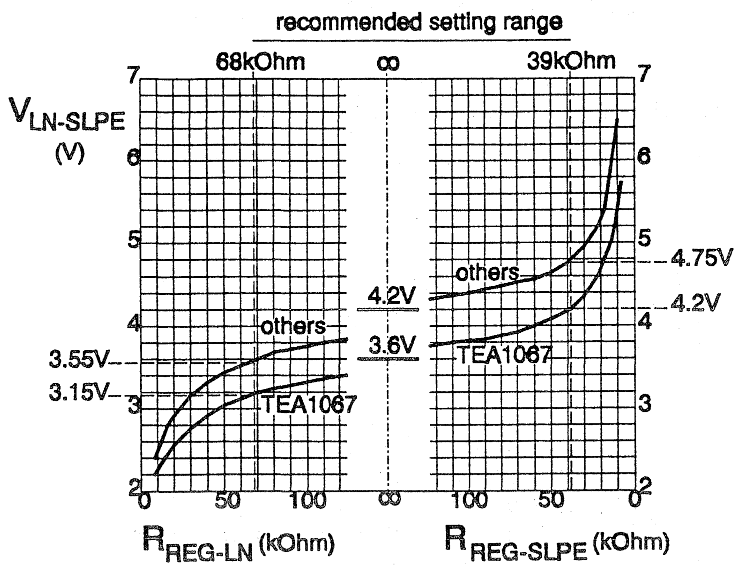


Figure 3. Internal reference voltage $V_{LN\text{-}SLPE}$ as a function of R_{17}

In the audio-frequency range the impedance of the set is determined by R_{20}, or more exactly by the value of $R_{20}//R_p$ (internal resistor, typically 16.2KΩ for the TEA1067 and 17.5 KΩ for the rest of the TEA106X-family with tolerance +/-20%).

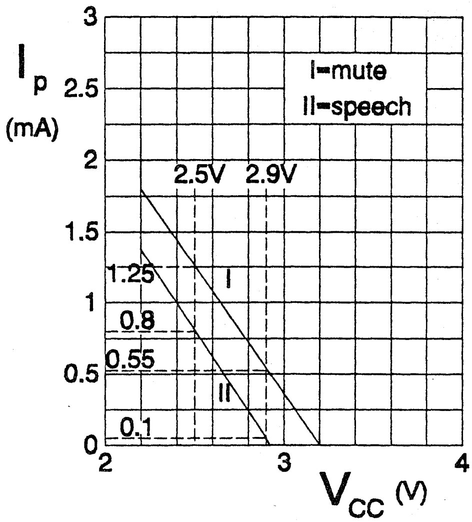
2.1.1. Regulated Line Voltage

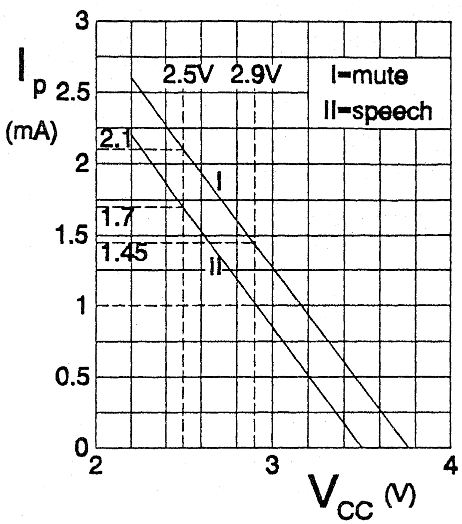
The supply voltage VCC is derived from the line via a dropping resistor R_{20} and regulated by the TEA1067. VCC may also be used to supply external circuits e.g. dialling and control circuits. Decoupling of the supply voltage is performed by a capacitor C_{20} between TEA1067 pins VCC and VEE.
The DC current drawn by the device will vary in accordance with varying values of the exchange voltage (V_{exch}), the feeding bridge resistance (R_{exch}), The DC resistance of the telephone line (R_{line}), and the DC voltage across the subscriber set including the polarity guard. Current (I_p) available from VCC for peripheral circuits depends on the external components used. If MUTE is LOW when the receiving amplifier is driven the available current is further reduced.
R_{16} (static resistance of the DC-characteristic) can change the slope of the DC characteristics V_{LN} v.s. I_{LINE}. It also shifts the low-voltage threshold current I_{TH}. Furthermore it determines microphone gain and DTMF gain, shifts the gain-control characteristic and in case its value exceeds 30Ω it decreases the maximum output swing on LN. Finally the side-tone will be affected. The preferred value 20Ω is used on this board. It is advised to follow the adjustment procedure given in the Reference [4] appendix II if R_{16} is to be changed.

2.1.2. Supply for Peripherals

The RC lowpass filter network R_{20}/C_{20} between the pins LN, VCC and VEE provides a smoothed voltage VCC for the Speech/Transmission IC itself and also a current Ip for the PCD3330-1. The ground reference for peripherals is pin VEE.

The supply voltage for peripherals VCC decreases with increasing I_p due to the series resistor R_{20}. This is shown in the following Figure 4. The supply possibilities can be increased by choosing a higher reference voltage V_{REF} (lower value for R_{17}).





(a) V_{LN} = 3.9V (typical for TEA1067, or other members with R_{REG-LN} = 75 KΩ)

(b) V_{LN} = 4.45V (TEA1067 with $R_{REG-SLPE}$ = 39KΩ and typ. for all other members)

Speech mode: V_{LN} = 1.4 V_{rms} (THD < 2%).
V_{QR+} = 150 mV_{rms} across 150Ω single ended load (THD<2%)

Mute mode: V_{LN} = 1 V_{rms}.

Figure 4. TEA1060-family typical I_p available from VCC at I_{LINE} = 15mA.

The voltage retention necessary for on-hook dialling, is guaranteed via resistor R_4, which is connected directly to the telephone line.

2.2. DIALLING

The dialling part mainly consists of:

1) PCD3330-1 mixed-mode multi-standard repertory dialler/ringer with on-chip DTMF generator, offering the user two dialling modes i.e. Pulse Dialling (PD) and Dual Tone Multi- Frequency (DTMF), and EEPROM for storing telephone numbers and options.

2) Keyboard.

2.2.1. Oscillator

The oscillator of the PCD3330-1 is controlled by a 3.58 MHz crystal connected to PCD3330-1 pins 9 and 10. Alternatively, a ceramic resonator may be used as a timing element.

2.2.2. Keyboard Connection

The 6 X 6 single-contact keypad matrix is directly connected to PCD3330-1 dialler sense column pins COL 1 to COL 6 and the scanning row output pins ROW 1 to ROW 6. Organization of the keyboard is shown in Figure 5.
During the STAND-BY mode of the dialler all column keyboard inputs and all row keyboard outputs are HIGH except ROW 5 which is LOW.

	COL1	COL2	COL3	COL4	COL5	COL6
ROW1	1	2	3	A/ TONE	M0	M1
ROW2	4	5	6	B/ DIS	M2	M3
ROW3	7	8	9	C/ PR	M4	M5
ROW4	*	0	#	D	M6	M7
ROW5	LNR	AP	FLA	HOOK	M8	M9
ROW6	STO	MEM	M	E-1	E-2	E-3

Figure 5. Keyboard organization

Function of the keys is:

0 to 9,* and #	Standard keyboard; in pulse dialling mode the valid keys are the 10 numeric keys (0 to 9), the 2 non-numeric keys (* and #) will result in switching to the DTMF mode (mixed mode dialling). In DTMF dialling mode the 10 numeric and the 2 non-numeric keys are valid.
A to D	These keys are only valid in DTMF dialling mode.
TONE	Pulse to DTMF switching key (mixed mode dialling). If the PCD3330-1 is set to the Pulse-Dial mode (PCD3330-1 pin 25 = LOW via jumper J_2) pressing pushbutton TONE changes the dialling mode to DTMF. After a switch-over from pulse to DTMF automatically an access pause is generated. A second touch of the button within one call is neglected.
DIS	The disconnect key will activate output notDP/FL for 800 ms. In this case the telephone set turns to the ON-HOOK state for this calibrated time. The disconnect function acts like CE with respect to redial.
PR	Program ringer key. With this key the ringer volume and repetition rate can be changed.
M0 to M9	One key abbreviated dialling: the 10 repertory numbers are directly accessible via push-buttons M0 to M9.
LNR	Last number redial: if the first key operated after going off-hook is the button LNR the dialler transmits the last number dialled before going on-hook (sliding-cursor and Atlanta access pause method are implemented).
AP	Access pause key: results in inserting an access pause in the telephone number during manual dialling and/or during programming. Length of the access pause can be selected via EEPROM bits.
FLAsh	FLASH/EARTH key: depending on the status programmed this key starts a FLASH or an EARTH procedure. At FLASH the output notDP/FL becomes active for a calibrated time. Dependent on EEPROM option, the time measures 100 ms, 115 ms, 270 ms or 600 ms. At EARTH the output EARTH becomes active for a calibrated time of 400 ms.
HOOK	Hook key (for on-hook dialling / loudspeaker on/off): as long as the handset stays on the cradle activation of this key activates outputs notDP/FL and LSE to operate the electronic hook contact and to switch on the loudspeaker so that dial tones can be heard via the TEA1083A. When the handset is not on the cradle (conversation mode) activation of this key switches the loudspeaker on/off (listening-in feature).
STOre	STORE key: this key provides the storage of the keyboard entries into one of the memory locations. Digits entered after pressing STORE are not transmitted and MUTE is not activated. When the STORE button is pressed the second time the number is actually stored into one of the memory locations. Then normal dialling can be started or the store procedure repeated.
MEMory	Two-key abbreviated dialling (MEM + digit): the repertory numbers M0 to M9 are also accessible via this two-key dialling procedure.

M	Mute key, each time this key is pressed and dialling is not active, the mute output goes to HIGH or LOW depending on the previous state.
E-1 to E-3	One key abbreviated dialling, three extra repertory numbers which are only directly accessible by push-buttons E-1 to E-3, these numbers can only be used when the repertory length is 16 digits.

2.3. SPEECH/TRANSMISSION

The speech/transmission part is built around the TEA1067 - speech/transmission IC. The TEA1067 is a member of the TEA1060-family of speech/transmission circuits for analog telephone sets. It has been developed to fulfil USA DC-requirements 6V at 20mA (RS470) with a normal diode bridge having 1.4V voltage drop. It enables parallel operation with classical telephone sets as at a line current below the low-voltage threshold current I_{TH} (typically 9 mA) the internal reference voltage is automatically adjusted to a lower value. At 1 mA a typical voltage drop of 1.6V is obtained. This means that the operation of the circuit with more telephone sets in parallel is possible with line voltages down to an absolute minimum of 3.0 V approximately (polarity guard included). Of course the sending and receiving amplifiers have reduced gain and output swing in the low voltage range. Furthermore the supply point for peripherals is degraded.

The internal voltage regulator of TEA1067 is decoupled by a capacitor C_{12} between pins REG and VEE. The TEA1067 has an internal current stabilizer working at a level determined by a resistor R_{28} connected between pins STAB and VEE. When the line current is more than 0.5 mA greater than the sum of the IC supply current (I_{CC}) and the current (I_P) drawn by the peripherals circuit connected to VCC, the excess current is shunted to VEE via V_{LN}.

In this paragraph, setting details for the TEA1067 are provided.
Please refer to the reference [4] in case another member of the TEA106X-family is used.

2.3.1. Microphone Amplifier

This section will provide the details for the microphone inputs and gain adjustment.

2.3.1.1. Microphone Inputs

The printed circuit board is ready for use with microphone capsules which require a symmetrical input (dynamic, magnetic and piezo electric microphones). The input impedance of the TEA1067 microphone amplifier is high (typ. 64 kΩ). If an asymmetrical input is required (electret) the MIC- input is used as a signal input via a capacitor. The MIC+ input should be connected to ground by a capacitor. The value of both capacitors is determined by the lower cut-off frequency required. For testing purposes only (low-ohmic drive) 2.2 µF is a suitable value.

In practice the so-called "motorboat" effect can occur (a low-frequency oscillation of the line voltage) in case the value of the MIC+ capacitor is too small compared with the MIC- capacitor (tolerance of capacitors). This can be prevented by choosing a lower value for the MIC- capacitor (one step down in the E12 series). Since the TEA1067 has a microphone gain that can be set between 44 and 52 dB (for R_{16} = 20 Ω) an external attenuation network is required if a sensitive type of microphone is used needing less than 44 dB of gain.

2.3.1.2. Gain Adjustment

The gain of the microphone amplifier can be calculated using the following formulas:

$$A_m = 1.356 * \left(\frac{R_{19} + 3.47K\Omega}{R_{16} * R_{28}}\right) * \frac{R_i * R_L}{R_i + R_L}$$

in which:

$R_i = R_{20}$ // [16.2 kΩ]
R_L = load resistance at LN during the measurement
(normally 600 Ω)

The following Figure 6 shows the gain of the microphone amplifier as a function of R_{19} for different values of R_{16} (static resistance of the DC-characteristic).

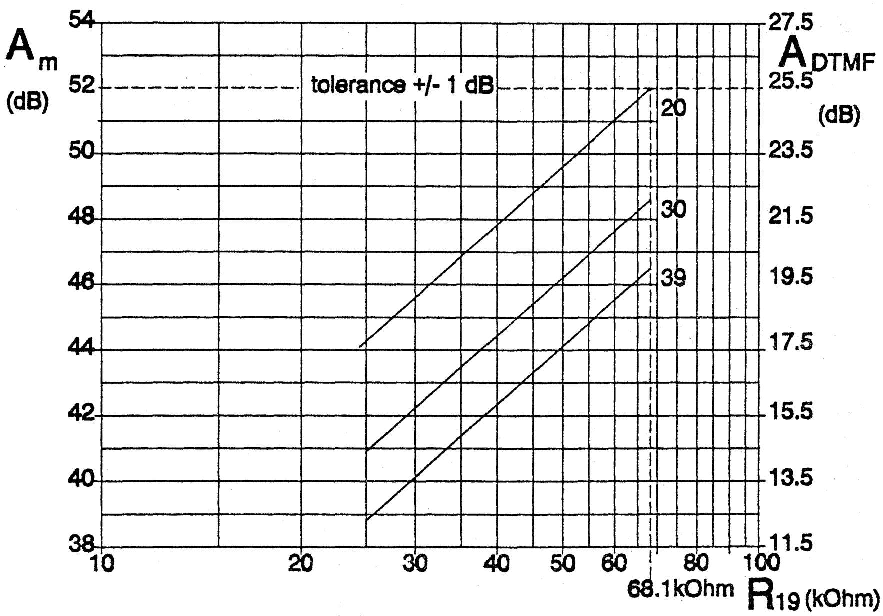


Figure 6. Microphone gain and DTMF gain as function of R_{19}

On the board the following resistors are mounted:

R_{16} = 20 Ω; R_{19} = 68.1 kΩ; R_{20} = 620 Ω; R_{28} = 3.65 kΩ.

This results in a gain of 52 dB (assuming a load resistance R_L of 600 Ω at the transmitter output LN).
In case of sufficient line current, symmetrical clipping of the output signal at IC_2 pin 1 (LN) can be obtained by using the zener diode D_{14} between IC_1 pin SUP and IC_2 pin SLPE.
To ensure the stability of the transmitting amplifier capacitor C_{11} is connected between GAS1 and SLPE. C_{11} can be adjusted to obtain a first-order low-pass filter with time constant R_{19}*C_{11} for the cut-off frequency. On the board C_{11} = 100 pF, R_{19} = 68.1KΩ, and $f_{(3dB)}$ = 23 kHz.

2.3.2. DTMF Amplifier

By applying a HIGH voltage to the MUTE input of the TEA1067, the device is switched over from the speech mode to the dialling mode. In this mode DTMF signals applied to the DTMF input are transmitted to the line. The microphone and receiver amplifiers are blocked now.
The Input impedance of the DTMF input is typically about 20KΩ. The voltage gain measured between the DTMF input and the transmitter output at LN is 26.5dB less than that of the microphone amplifier
The gain of the DTMF amplifier is:

$$20 \log A_{(DTMF)} = 20 \log A_m - 26.5 \text{ dB}$$

The DTMF gain depends on the values of R_{16}, R_{19}, R_{20}, R_{28} and R_L in the same way as the microphone gain. Thus once the microphone gain has been adjusted, the DTMF gain is predetermined. This means that the DTMF input level must be adjusted to the correct level by means of an external attenuator (R_{36}, R_{37}) at the DTMF input. The DTMF input accepts signals up to 170 mV (rms) for THD = 2% with internal soft limiting of the input stage.

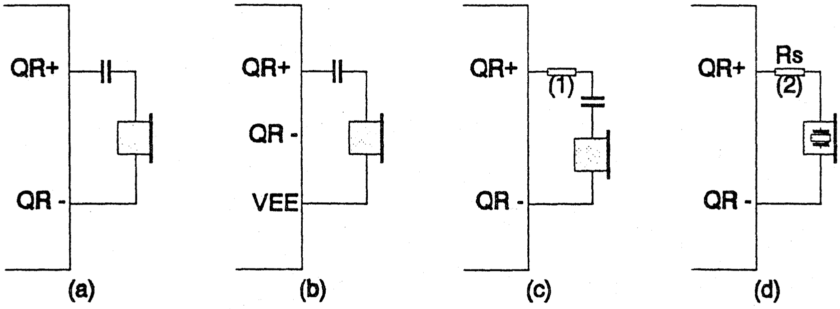
The DTMF amplifier is internally temperature compensated. However because of the asymmetrical input structure (single ended drive) some influence can be expected from the residual AC-line voltage being present on the supply pin VCC which depends on the performance of the low-pass filter components R_{20}/C_{20}, especially C_{20}. Therefore an electrolytic capacitor with low temperature coefficient should be used.

2.3.3. Receiver Amplifier

The input of the receiving amplifier is IC_2 pin 11-IR. The input signal IR is symmetrically soft-limited internally, to 17 mV for 2% THD and to 53 mV for 10% THD. Input impedance is typically 21 KΩ. The amplifier has two complementary class B outputs i.e. the non-inverting output QR+ at pin5, and the inverting output QR- at pin4. The output can be used either for single ended drive or for symmetrical drive, depending on the impedance, sensitivity and type of earpiece.

2.3.3.1 Earpiece Outputs

The TEA1067 can drive either dynamic, magnetic or piezoelectric earpieces. Earpieces with an impedance up to Z_T = 450 Ω (low-impedance dynamic or magnetic capsules) must be driven in single-ended mode. For impedances above 450 Ω differential drive is possible. A resistor in series with the earpiece must be mounted if required. (see the following Figure 7 for the details).



(a) Dynamic earpiece with $Z_T \geq 450\ \Omega$.
(b) Dynamic earpiece with $Z_T < 450\ \Omega$.
(c) Magnetic earpiece with $Z_T \geq 450\ \Omega$ resitor (1) may be connected to prefent distorsion.
(d) Piezo-electric earpiece, resitor (2) is requered to ensure sufficient phase margin.

Figure 7. Connection of the earpieces

As with the microphone amplifier, the performance of the receiving amplifier is dependent of the d.c. line voltage V_{LN}. The maximum output swing on the earpiece depends on the load and on the d.c. voltage drop across the circuit.

2.3.3.2. Gain Adjustment

The gain of the receiver amplifier between the input IR and the output QR+ can be calculated by means of:

single-ended drive: $$A_{TA}=1.314*\frac{R_{24}}{R_{28}}*\frac{Z_T}{Z_T+4\Omega}$$

differential drive: $$A_{TS}=2.628*\frac{R_{24}}{R_{28}}*\frac{Z_T}{Z_T+8\Omega}$$

in which: Z_T = earpiece impedance

The following Figure 8. shows the receiver gain as a function of R_{24}.

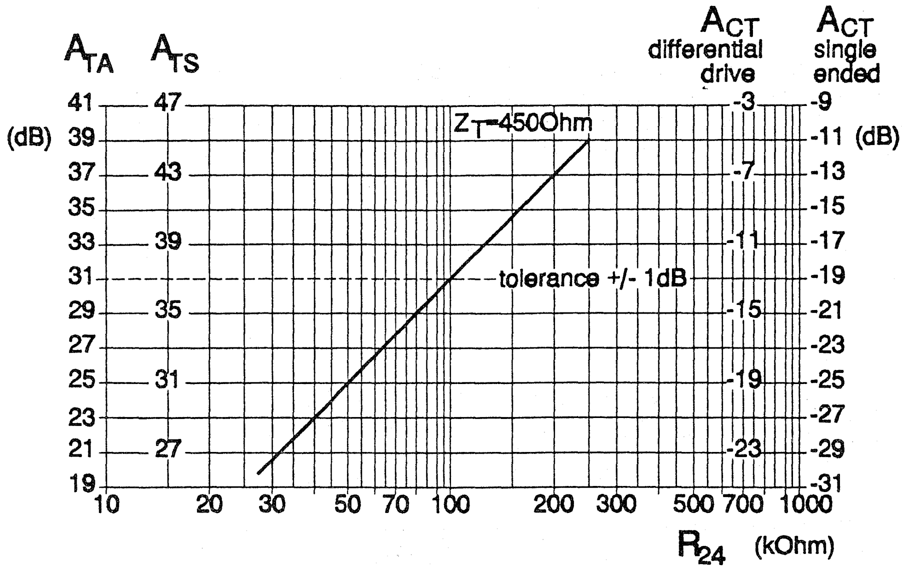


Figure 8. Gain of the receiving amplifier [A_{TA} and A_{TS}] and confidence tone as a function of R_{24}.

On the board : R_{24} = 100 kΩ, R_{28} = 3.65 KΩ resulting in a gain A_{TA} of 31 +/- 1dB. The gain of the receiving amplifier can be adjusted over a range of -11 and + 8 dB by means of R_{24}. The total receive gain from the line to the earpiece can be obtained by subtracting the attenuation of the anti-sidetone network (32 dB) from A_{TA} or A_{TS}, resulting in an overall gain of -1dB as the signal received on the line is attenuated by the anti-sidetone network before it enters the amplifier.

C_{14} and C_{15} prevent parasitic oscillations of the receiver amplifier while C_{14} can also be adjusted to obtain a first-order low-pass filter with time constant R_{24}*C_{14} for the cut-off frequency. The value of C_{15} should be 10 times C_{14}. A high-pass filter is formed by C_5 (together with the input impedance of pin IR) and by C_{13} (together with the earpiece impedance).

On the board the following components are mounted:

C_5 = 100 nF, C_{13} = 10 µF, C_{14} = 100 pF, C_{15} = 1 nF.

The cut-off frequency of the low-pass filter f_{3dB} = 16KHz.

2.3.3.3. Confidence Tone

During DTMF dialling, the dialling tones can be heard at a low level in the earpiece. The level of tones at the receiving output depends on the gain that has been set for the receiving amplifier and on the tone level applied to the DTMF input. For both single-ended and differential drive the gain A_{CT} from the DTMF input to the receiver output, is given by (see the aforementioned Figure 8):

$$20 * \log A_{CT} = 20 * \log A_T - 50 \text{ dB}$$

in which A_T is a general term for the receiver gain and this can be replaced by either A_{TA} or A_{TS} described in the previous section.

2.3.4. Automatic Gain Control

The gain Figures of the microphone amplifier and the receiving amplifier which were derived in the preceding section are applicable only when the AGC is inoperative; i.e. IC_2 pin17-AGC is open. When Jumper J_1 is closed, R_{18} is connected between pins AGC and VEE, the line current dependent gain control of both the microphone amplifier and the receiving amplifier becomes operative; the DTMF amplifier is not affected. Below a specific value of the line current $I_{line\text{-}start}$, the gain is equal to the values with the formula given before. If the current $I_{line\text{-}start}$ is exceeded, the gain of both of the controlled amplifiers decreases as a function of increasing DC line current. Gain control stops when another value of line current $I_{line\text{-}stop}$ is exceeded.

The gain control range of both amplifiers is typically 6 dB corresponding with a 5 km line length of 0.5 mm diameter copper twisted pair cable (DC resistance 176 Ω/km, average AC attenuation 1.2 dB/km).

If no automatic gain control is required jumper J_1 on the printed circuit board has to be removed.

The <u>optimum values of Resistor R_{18}</u> for the various values of the exchange supply voltage and its feeding bridge resistance, with a 1.4V diode bridge, R_{16} = 20 Ω and the increased line voltage V_{LN} = 4.45V at 15mA (R_{17} = 39KΩ) are as follows:

V_{EXCH}	R_{EXCH} 400Ω	R_{EXCH} 600Ω	R_{EXCH} 800Ω	R_{EXCH} 1000Ω	
36V	100	78.7	-	-	
48V	140	110	93.1	82	R_{18} (kΩ)
60V	-	-	120	102	

2.3.5 Anti-Sidetone Circuit

The anti-sidetone circuit takes care that the microphone signals available on the line output LN are suppressed sufficiently before they enter the receiving amplifier input IR. The anti-sidetone circuit takes the signal which is available at pin 18-SLPE and uses it to compensate the microphone signal at the input IR of the receiving amplifier. On the printed circuit board the TEA1060-family bridge is used to suppress the sidetone. For details of this bridge please refer to Reference [4].

Optimum suppression is obtained when the following two conditions are fulfilled:

a) $R_{16} * R_9 = R_{20} * (R_{10} + [R_{13} // Z_{bal}])$
b) $Z_{bal} = (R_{13}/R_{20}) * Z_{line} = k * Z_{line}$

Normally R_{20} and R_{16} are fixed and R_9, R_{10}, and R_{13} can be chosen by the designer. These resistors have to be chosen to meet the following criteria:

* compatibility with a standard capacitor from the E6 or E12 range for the capacitor used in Z_{bal}.

* $|Z_{bal} // R_{13}| << R_{10}$ to avoid influence of Z_{bal} on the receiver gain.

* $|Z_{bal} + R_{13}| >> R_{16}$ to avoid influence on the microphone gain.

* $(R_{10} + |R_{13} // Z_{bal}|) << 20$ kΩ to avoid the influence of the input impedance at IR (21 kΩ ± 4 kΩ) on the bridge attenuation.

In practice Z_{line} varies strongly with line length and type of cable. Therefore an average value has to be chosen for Z_{bal}.
On the printed circuit board Z_{bal} is optimised for a cable length of 5 km with a diameter of 0.5 mm copper with a DC resistance of 176 Ω/km and a capacitance of 38 nF/km resulting in an average AC attenuation of 1.2 dB/km.

For compatibility of the capacitor value in Z_{bal} with a standard capacitor from the E6 range (220 nF):

$$k = \frac{140\ nF}{220\ nF} = 0.64$$

For R_{10} a value of 3.92 kΩ has been chosen resulting in K = 0.64. So Z_{bal} and R_{13} and R_9 can be calculated resulting in the following practical values on the board e.g.

$R_{11} = 130\Omega$, $R_{12} = 820\Omega$, $C_6 = 220nF$, $R_{13} = 390\Omega$, and $R_9 = 130\ K\Omega$.

This results in a roughly equal sidetone level (acoustically measured) at zero Km line and with 10Km line with the line current dependent gain control activated. In case no AGC is used, the sidetone has to be optimized for a shorter line length in order to obtain equal (acoustical) side tone levels at zero Km and 10 Km line length. In practice normally a compromise is chosen between loudness of the set and sidetone level; It means that sending and receiving gain will be reduced somewhat.

The anti-sidetone network attenuates the signal from the line LN to the input IR of the receiver amplifier. The attenuation can be derived from:

$$\frac{V_{IR}}{V_{LN}} = \frac{R_t//R_{10}}{R_9 + (R_t//R_{10})}$$

where R_t is the input impedance of the receiving amplifier (typically 21KΩ). This attenuation is about 32 dB.

Detailed information about the application of the TEA1067 can be found in Reference [4].

2.4. CALL PROGRESS MONITOR.

The OM4723 contains a call progress monitoring application for a modern electronic telephone set with a common-line interface. The call progress monitoring function is realized using the TEA1083A call progress monitoring IC and the TEA1067 speech-transmission IC. Moreover, during the conversation mode, listening-in is possible with reduced performance (no howing limiting, no dynamic limiting of the loudspeaker signal). Detailed information about the TEA1083A can be found in the documentation as given in reference [5]

The TEA1083A is connected between the positive line terminal and pin SPLE of the TEA1067, so the transmission characteristics such as set impedance, gain settings, gain control etc. are not affected.

2.4.1. Supply and logic inputs

The TEA1083A is connected between the positive line voltage and pins LN and SLPE of the TEA1067. The line current split-up is controlled by resistor R_{15} which is in serie with pin LN of the TEA1067 and which is connected between pins SUP and SREF of the TEA1083A. Because in this way the current flow is controlled by the TEA1083A, always the maximum current will be available for the TEA1083A. Resistor R_{15} insures that the output stage of the TEA1067 get always enough bias current.

$I_{bias\ TEA1067} = 0.5 / R_{15} = 3.3$ mA.

This results in a supply current for the TEA1083A of:

$I_{SUP} = I_{LINE} - 3.3\ \text{mA} - I_{internal\ TEA1067} - I_{Peripherals}$

Because the TEA1083A is connected between the positive line voltage and pins LN and SLPE of the TEA1067, the logic inputs of the TEA1083A are all referred to SLPE. However the PCD3330-1 outputs are all referred to VEE. Therefore it is necessary to put diodes between the outputs of the PCD3330-1 and the inputs of the TEA1083A (D_{21} and D_{22}).

2.4.2. Power-Down/Pulse Dialling

During pulse dialling or register recall, the TEA1083A can be switched off by activating the Pin12-PD input via signal notDP/FL from the PCD3330-1. A 'HIGH' level applied at the PD terminal forces the TEA1083A into the power-down mode, thereby reducing the current consumption from the VCC and VBB supply points. A diode D_{21} is necessary because the negative supply reference of both IC's is different.

2.4.3. Loudspeaker enable LSE

The TEA1083A is provided with an enable input to enable or disable the loudspeaker amplifier. When the LSE terminal is 'HIGH' than the loudspeaker amplifier is enabled. The LSE input is controlled by the LSE output, pin 27 of the PCD3330-1. Diode D_{22} is necessary because the negative supply reference of both IC's is different.

2.4.4. Amplifier gain and volume control

The gain between inputs and output is fixed to 35dB. Volume control of the total monitoring device is obtained by attenuation of the input signal via R_{26} and potentiometer R_{25}.

The value of R_{26} and R_{25} on the PCB are both 2.2kΩ. R_{26} has to be chosen in accordance with the earpiece sensitivity and the required signal and maximum signal distortion.

2.5. PCD3330-1 INTERFACE

This section describes the interfacing from the PCD3330-1 to TEA1067, TEA1083A, and the line interrupter BSN254A. The most difficult part of the PCB is the on-hook dial facility and will be elucidated in paragraph 2.5.1.

2.5.1. On-hook Dialling

The status of the handset is monitored via the input HOOK; HOOK is LOW when the handset is on the cradle and HOOK is HIGH when the handset is lifted.

For on-hook dialling the line interrupter is also used as an electronic hook switch controlled by the repertory dialler IC_3-PCD3330-1. However, if the dialler is without supply (no back-up left), it must still be possible to use the telephone set. Therefore the PCD3330-1 dialler has a normal and an on-hook operation mode which can be selected by a EEPROM bit in the PCD3330-1 (see EEPROM program part).
The telephone set with on-hook dial function starts up as follows:

* Dialler is in STAND-BY mode (CE/RF = LOW), Voltage $V_{DD} > 2.0V$.
* Key HOOK is pressed interconnecting column 4 and row 5.
* T_{10}-BC558 is switched-on pulling CE/RF = HIGH.
* Dialler starts-up (HOOK = LOW).
* Dialler puts notDP/FL HIGH switching on T_2 and T_3 (notDP/FL = LOW in STAND-BY mode).
* IC_2-TEA1067 connected to the line will start-up.
* As long as the HOOK input stays LOW input CE will not be tested to detect on-hook.
* Telephone set is operational.

Switch-off via keyboard: Pressing key HOOK will result in dialler pin notDP/FL to go LOW, which the line disconnected. If HOOK is pressed with the handset off-hook (HOOK = HIGH) listening-in (with reduced performance) can be switched on and off (toggle). This is achieved via output LSE.

Switch-off via handset: First the handset has to be lifted from the cradle before it is possible to switch-off via handset. When lifted then putting down the hand set will result in HOOK to go LOW. As a result notDP/FL goes HIGH and CE goes LOW putting the PCD3330-1 into STAND-BY mode and switching off the set.

2.5.2. Pulse Dialling (selected with jumper J_2)

Signal MUTE mutes the speech and the call progress monitoring circuits.
Signal notDP/FL controls via T_3 the line interrupter T_2 so that pulse dialling is possible and also puts the TEA1067 and TEA1083A into power-down mode during these line breaks.

2.7. PCF8581 EEPROM

This PCF8581 is not necessary for normal operation of the board as a telephone set, but with this external EEPROM it is possible to load the internal EEPROM of the PCD3330-1 with all the telephone options necessary for the different countries and also to pre-program in a very fast way the repertory numbers for demonstration.

The contents of the PCF8581 EEPROM is loaded into the EEPROM of the PCD3330-1 when at off-hook jumper J_4 is closed. To exit this EEPROM loading mode jumper J_4 has to be removed again.

The contents of the internal EEPROM of the PCD3330-1 is sent via the SDA and SCL lines when at off-hook jumpers J_4 and J_3 are closed. This contents can be loaded into the external PCF8581 EEPROM, for evaluation.

For normal operation this PCF8581 can be removed.

2.8. PROTECTION

Unprotected speech/transmission ICs might be destroyed by the excessive current surges on the telephone lines if preventive measures are not taken. Protection on this printed circuit board is achieved by means of Break-Over-Diode (BOD) D_1 (BR211-240) between the a/b lines, a zener diode D_{14} between the positive line terminal and SLPE of 8.2 Volt, and current-limiting device (T_1, R_2). The interrupter T_2-BSP254A gate-source voltage is limited by zenerdiode D_7 of 6.8 V and the supply voltage of the PCD3330-1 is limited via D_{17}.

2.9. Electro Magnetic Compatibility (EMC)

In the presence of high-intensity electromagnetic field, it is possible for common-mode amplitude modulated R.F. signals to be introduced in the a/b lines. These common-mode signals can sometimes become differential-mode signals as a result of asymmetrical parasitic impedances to ground (for example through the hand of the subscriber holding the handset). Preventive measures have to be taken to avoid the possibility of these signals being detected and the low-frequency modulation appearing as unwanted signal at the earpiece or on the line.

The layout of this printed circuit board is designed with respect to electromagnetic compatibility. Basic protection is provided by means of small discrete capacitors to suppress the unwanted R.f. signals before they can enter the circuit. Capacitor types suitable for high frequencies must be used, for example ceramic types. Those are C_{18}, C_{21} at pin MIC- and MIC+ of the TEA1067, C_{31} and C_{32} at the microphone inputs points, C_{19} at the receiving input IR, C_9 at the transmitter output LN, C_{29} at pin SLPE, C_{33}, C_{34}, R_{21} and R_{22} at the telephone output points and C_1 and C_2 ant the A-B and B-A telephone set connection points. (see Appendix IV).

More detailed information in case a higher EMC (Electro Magnetic Compatibility) performance is required can be found in Reference [6].

2.5.3. DTMF Dialling (selected with jumper J_2)

The MUTE signal mutes the speech circuit and enables the DTMF amplifier of the TEA1067 and mutes the TEA1083A. DTMF signals from the PCD3330-1 can now be transmitted onto the line via the TEA1067.

2.5.4. Chip Enable Control

Pin CE is controlled by the following three cases:

1). When handset is lifted (HOOK also HIGH).

2). If HOOK key is pressed in STAND-BY mode.

3). If ringer signal is present (see ringer activation part below).

2.6. RINGER INTERFACE.

The ringer part can be divided into three parts:

Supply:
When a ringer signal is present on the a-b lines, this AC signal supplies the ringer hardware via capacitors C_4 and C_{28}, zenerdiodes D_{23} and D_{24}, resistor R_6 and diode bridge D_8, D_{11}, D_{12} and D_{25}. Capacitors C_4 and C_{28} block the dc-current flow via the ringer hardware, two capacitors are necessary because in this on-hook dialling application both connections can give dc-current flow. Zenerdiodes D_{23} and D_{24} are necessary because otherwise the ringer harware short-circuits the ac signals (DTMF-out) of the TEA1067. Resistor R_6 is presend to make the total impedance ok. Finaly there is a diode bridge D_8, D_{11}, D_{12} and D_{25} for rectifier the ac-ringer signal.
The PCD3330-1 is supplied via the voltage divider network D_{10} and D_{17}. Zenerdiode D_{10} together with zenerdiode D_{17} keeps the maximum voltage across the ringer output stage below 28 V and across the PCD3330-1 below 6 V.

Ringer frequency measurement:
To activate the ringer frequency measurement the ringer input frequency has to be supplied to the CE/FR input of the PCD3330-1. This is done by connecting the CE/RF input of the PCD3330-1 to one of the ac-connection points of the ringer rectifier bridge (here at diodes D_{11} and D_{12}) via resistor R_{14}. When the ringer input frequency is between the ringer detection LOW and HIGH the PCD3330-1 will go to the ringer melody generation mode.

Ringer output stage:
The ringer melody is amplified via T_5 and T_4 and supplied to the buzzer. The output volume control is achieved with T_6 and T_7 which are controlled from PCD3330-1 outputs RVOL1 and RVOL2.

3. FEATURES PCD3330-1

The PCD3330-1 is a mixed-mode multi-standard repertory dialler/ringer IC. The (maximum 13) repertory numbers, redial and various country specifications are stored in EEPROM. Therefore, different models can be created by changing the contents of some EEPROM bytes.
The two on-chip tone generators are used for DTMF dialling and for generating a melody during ringing.

The following features are implemented in the PCD3330-1:

- Pulse dialling
- DTMF dialling
- Mixed mode dialling
- Flash or register recall
- Connect a/b line to earth function
- Mute function
- Disconnect function
- Standard 4x4 keyboard for: 0 to 9 and *, #, A, B, C and D
- Function keys for: Flash, Hook, Mute, Tone, Disconnect, LNRedial, Memory recall, Store, Access Pause, 1 Key repertory dialling and Program Ringer
- On-hook dialling control
- Redial Cursor method (maximum 24 digits) stored in internal EEPROM
- Storage for 13 repertory dial numbers (16 digits each) or 10 repertory dial numbers (20 digits each) in internal EEPROM
- Access pause generation and termination: manually or by Atlanta procedure
- Ringer input frequency detection
- Three-tone ringer with 4 different ringer sequences
- Ringer melody generation with four signal speeds and four output volume steps, keypad controlled
- Country specifications which can be stored in EEPROM are:
    - Will * and/or # be transmitted when switching over to DTMF dial mode
    - Mark to space ratio (3:2 or 2:1)
    - 6 Tone time selections (60/90, 70/70, 80/80, 100/100, 100/140 or 140/140 ms)
    - 4 Flash time selections (100, 115, 270 or 600 ms)
    - Mute output type selection (M1, notM1, M2 or notM2)
    - Dial Pulse output selection (DP or notDP)
    - DTMF keys or Function keys selection
    - Access Pause time selection (1.5/1.0, 2.5/1.5, 3.0/3.5 or 6.0/6.0 s)
    - 10 Number repertory dialler selection (1 or 2 key)
    - Two repertory number programming procedures (General or Germany)
    - Repertory length 16 or 20 digits
    - Ringer input frequency detection selection
    - Ringer output selection (via DTMF or special -RT- output)
    - 4 Possible ringer melodies
    - 4 Possible ringer repetition rates
    - 4 Possible ringer volumes

4. APPLICATION EXAMPLES

This chapter is a tutorial, it describes the use of the OM4723 to a new user.
The full specification of the PCD3330-1 is given in the data sheet of the PCD3330-1 (see reference [1]).

4.1. MODES OF OPERATION

The PCD3330-1 has 4 operation modes:

* Standby
* Conversation
* Dialling
* Ringing

4.1.1. STAND-BY Mode

When both input **CE** and **HOOK** are LOW the PCD3330-1 is in the standby mode. The only current drawn is from a back-up supply (line powered via R_4) for on-hook dialling. During the standby mode all column keyboard inputs and all row keyboard outputs are HIGH except ROW 5 which is LOW.

4.1.2. CONVERSATION Mode

After the handset is lifted, or the HOOK button is depressed, **CE** is activated and V_{DD} rises to the working voltage. **MUTE** is inactive and speech or a dial tone can be heard. With the oscillator operating the chip is ready to accept keyboard entries.

* If the conversation mode is determined by the operation of the HOOK button then next entry of the HOOK button changes the PCD3330-1 to the standby mode.

* If the conversation mode is determined by the activation of the cradle contact (HOOK = HIGH) then going to the standby mode is only possible by de-activation of the cradle contact (HOOK and CE = LOW).

4.1.3. DIALLING Mode

The dialling mode starts after CE = HIGH (conversation mode) with the first valid keyboard entry when it initiates:

* a normal call of a newly dialled number
* a repertory or redialling cycle of previously entered and stored numbers

4.1.3.1. Pulse Dialling

The pulse-dial mode is selected with making input PD/DTMF = LOW via jumper J_2.

The keyboard entry initiates a recall of a previously stored number or is a simultaneous keying-in and pulsing-out activity, with storing for possible later recall.
If in the recalled number or at keying-in the keys A, B, C or D (option A to D keys selected) are used these digits are not transmitted.
If at keying-in the keys * or # are used this results in a switch over to DTMF dialling.
Normally, keying in is faster than pulsing-out (fed from the redial register).
Pulse sequences start with an inter-digit pause of 840 ms duration, followed by a sequence of pulses corresponding to the present digit in store. Each pulse starts with a mark (line break) followed by space (line make).
The pulse period is 100 ms with a mark-to-space ratio of 3:2 or 2:1 (mark to space ratio selection). After transmission of a digit, the next digit is processed, again starting with an inter-digit pause.
The pulses are available at the notDP/FL output and can be used to drive an external switching transistor in pulse dialling mode.
The transmission IC is put in the dialling mode by means of output MUTE.
The output MUTE has several programmable options, MUTE as M1, notM1, M2 and notM2. This hardware needs the notDP/FL and M1 programmable output configuration. Timing sequence for pulse dialling is shown in Figure 9.
After completion of the number string the circuit changes from dialling mode to conversation mode.

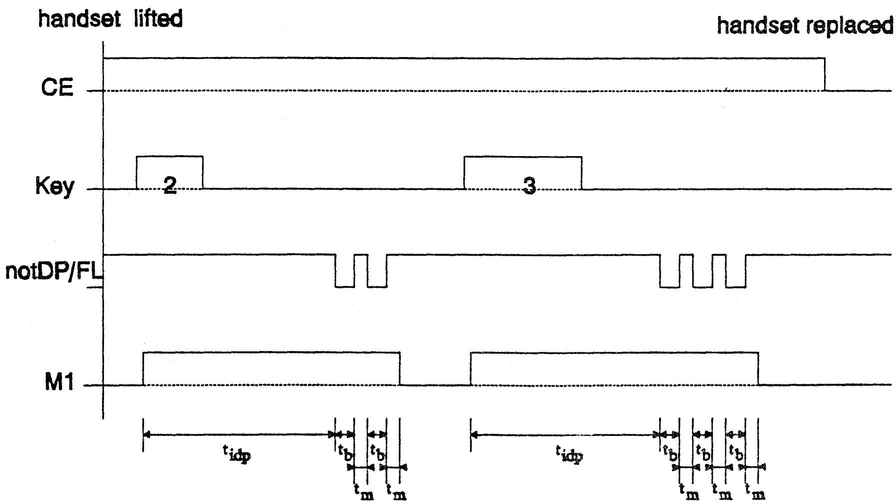


Figure 9. Timing diagram in pulse mode, showing notDP/FL and MUTE output.

4.1.3.2. DTMF Dialling

The Dual Tone Multi Frequency (DTMF) mode can be selected by making input PD/DTMF = HIGH via jumper J_2.

The PCD3330-1 converts keyboard inputs into data for the on-chip DTMF generator.
Tones are transmitted via output TONE with six programmable minimum tone burst/pause durations of 60/90, 70,70, 80/80, 100/100, 100/140 or 140/140 ms.
The maximum tone burst duration is equal to the key depression time.
With redial and repertory dialling tones are automatically fed at the programmed rate.
The mute is again depending on the selected MUTE-mode.
Again the MUTE output has several programmable options namely, M1, notM1, M2 and notM2.
This hardware needs the normal M1 programmable output configuration. Timing sequence for DTMF dialling is shown in Figure 10.

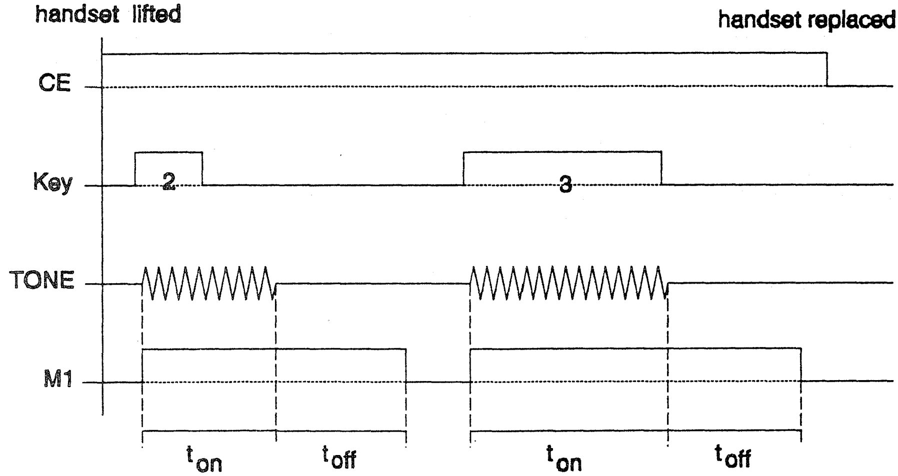


Figure 10. Timing diagram in DTMF mode, showing TONE and MUTE output.

4.1.3.3. DTMF dialling in pulse dialling mode (mixed mode dialling)

If the PCD3330-1 is set to the pulse dial mode (pin PD/DTMF is LOW), activation of push-buttons TONE, * or # changes the dialling mode to DTMF.
Its entry is stored in the working register and it generates automatically an access pause, after which the digits following are transmitted in the DTMF mode.
The digits entered after keys TONE, * or # are not transmitted in the redial mode.
The TONE key is never transmitted, whether * or # are transmitted depends on the selected option.
A second touch of the TONE key is ignored.
The * and # keys pressed after a switch over to DTMF dialling are all transmitted.

If the circuit is initially set to the DTMF mode (pin PD/DTMF is HIGH), activation of push-button TONE is ignored and the * or # are stored in the redial register and transmitted in DTMF mode.

4.1.3.4. Access pauses

If during entering a telephone number via keyboard for normal dialling or during repertory number programming the AP-key (access pause key) is pressed, then an access pause is stored in the redial or repertory dial register.

If during redial or repertory dial an access pause is detected a pause is inserted.
With this PCD3330-1 it is possible to select between four possible access pause times.

For pulse dialling	1.5 / 2.5 / 3.0 / 6.0 s
For DTMF dialling	1.0 / 1.5 / 3.5 / 6.0 s

4.1.4. Ringer function

The PCD3330-1 has a three-tone melody ringer with the following characteristics.

- Ringer output pin selection
- Ringer input frequency measurement
- Ringer melodies selection
- Ringer volume change during conversation and ringer mode
- Ringer repetition rate change during conversation and ringer mode

In figure 11 the timing diagram of the ringer function is given.

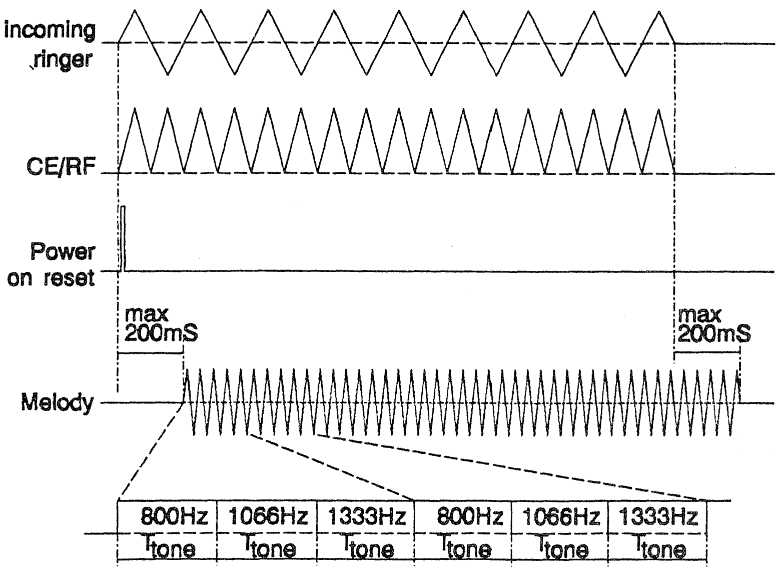


Figure 11. The timing diagram of the ringer function.

Ringer output pin selection
On this demo print the ringer is fed to the ringer output stage via the special ringer output pin RTO (via an EEPROM bit the TONE output pin can be selected).
One of the internal tone generators drives this output with a block with a peak-to-peak output voltage of V_{DD}-V_{SS} (because of the use of a PXE transducer).

Ringer input frequency measurement
The melody ringer becomes active for all incoming ringer frequencies higher then the ringer detection LOW frequency and lower then the ringer detection HIGH frequency supplied to the CE/RF input of the PCD3330-1. The ringer detection LOW and ringer detection HIGH frequencies are selected such that it is possible to use this PCD3330-1 for both single or double phase rectifier applications.

Ringer melodies selection
The ringer melody generator can select out of four different ringer melodies options (stored in EEPROM).

Ringer volume change during conversation and ringer mode
The ringer volume can be controlled by the port pins RVOL1 and RVOL2 and its value is stored in EEPROM.
During conversation mode the volume and repeat frequency can be changed by the following procedure, while during ringer mode for volume and repeat change only a simple key press is sufficient.

In conversation mode the procedure is as follows:

- put the set in conversation (supply necessary)
- depress PR (ringer program key)
- press one of the four acceptable volume keys (1 to 4) or one of the four acceptable repeat frequency keys (9, *, 0 or #)

The newly selected value for volume or repeat frequency is directly stored into EEPROM.

During ringer mode it is not necessary to press the PR-key first only a simple press at on of the four volume or one of the four repeat frequency keys is sufficient.

4.2. OPERATING PROCEDURE

This section provides the detailed operating procedure starting with the initialization.

4.2.1. Initialization

When the PCD3330-1 is started up the complete internal RAM is cleared to avoid an incorrect contents. By lifting the handset, operating the HOOK button, or detecting a ringer input signal the buffer capacitor at V_{DD} is charged to the operating voltage and CE is activated. Within the start-up time the oscillator runs and the program starts.

4.2.2. Dial Procedures

The dial procedures are categorized into:

* Manual dialing
* Last Number Redial (LNR)
* Repertory dialling

4.2.2.1. Manual Dialling

If the first push-button pressed is 0 to 9 in pulse dialling or 0 to 9, * and # in DTMF dialling mode, digits are entered into the work register and compared with the previous entries stored in the redial register. As long as the newly dialled digits are equal to those stored the contents of the redial register stays unaffected.

When the newly pressed digit is different from the one stored in the redial register the contents of the work register is copied to the redial register when going on-hook (or every other action equal to on-hook).

During the data entry the circuit starts immediately with transmission of the digit(s) and the minimum transmission time is unaffected by the speed of entry.
Transmission continues as long as further data input has to be processed.

Up to 24 digits can be stored in the work register ready for storing in the redial register.
After the work register overflows, a 10 digits First-In-First-Out register (FIFO) takes over as buffer and the contents of the work register is not copied to the redial register.
After transmitting the first digit of the FIFO register this position is automatically cleared to provide space for the storage of new data. In this way, the total number that can be transmitted is unlimited, provided the key-in rate is not excessive.
However, if the FIFO register overflows (more than 10 digits in store) further input is ignored.

4.2.2.2. Last number redial (1 to 24 digits)

Recalling the last number dialled stored in memory is done by pressing the LNR-key, and the procedures used are the "cursor" or the "Atlanta" procedure.

<u>Cursor method.</u>
If the first key pressed and released is the redial (LNR) button, the stored number in the redial register is recalled and transmitted.

If the first key entered is not the redial button, but numerical digits and these digits are equal to the digits held in store, the redial register is not cleared and dialling can be continued by pressing the redial button, but then the already dialled part is not redialled.
Redial is inhibited as soon an entry is unequal to the digit, at the same position, held in store.

<u>Atlanta procedure</u>
If the first key entered is the redial button, but this button is kept down, then only the first digit held in the redial register is transmitted. After releasing the redial button the remaining digits held in the redial register are dialled.
This is called the "Atlanta Access Pause" procedure.

4.2.2.3. Repertory Dialling

The PCD3330-1 includes a 13-number repertory dialler when a 16 digits length is chosen and a 10-number repertory dialler for 20 digits each (length is stored in EEPROM bit).
The 10 repertory numbers can be recalled with the M0 to M9 keys or by pressing the MEM key followed by a numeric digit from 0 to 9.
The extra 3 repertory numbers can only be recalled with the E-1 to E-3 keys.
The by M0 called number is identical to the MEM + 0 called number etc. till M9 which is equal to MEM + 9.
This maximum length of 16 or 20 digits includes the manually stored access pauses.

Chain dialling: Repertory numbers can be dialled-out after or before entering manual dialling, last number redial and by entering the memory locations in successive order.
During transmission of a number called from the memory location, the PCD3330-1 does not accept keyboard entries.
Dialling can be continued as soon as the number under transmission is completed.

Note that the last memory location which is transmitted is stored in the redial register.

4.2.2.4. Storing repertory numbers

The store mode starts after going off-hook and depressing the STO-key. With the PCD3330-1 a selection can be made between two store modes, the "General" and the "German".
Repertory numbers can be stored into EEPROM via the one key access or the two key access method and following the German or General storing procedures:

a) One-key access repertory number mode (M0 to M9 and E-1 to E-3).

General procedure	German procedure
- set in operation mode	set in operation mode
- depress STO (store key)	depress STO (store key)
- telephone number	location (M0 - M9/E-1 - E-3)
- depress STO (store key)	telephone number
- location (M0 - M9/E-1 - E-3)	STO (store key)

b) Two key access repertory number mode (MEM + 0 to 9).

General procedure

- set in operation mode
- depress STO (store key)
- telephone number
- depress STO (store key)
- depress MEM (location key)
- depress 0 to 9 (real location)

Memory clear: Cleaning of the memory location is possible via the same procedure as for storing a number, only no telephone number is entered.

4.2.3. Flash or Earth function

Whether the Flash or Earth function is activated by the FLA key is programmed in EEPROM.

If the FLASH function is selected a calibrated FLASH pulse (recall register) is generated on the notDP/FL output and the MUTE output is active.
The calibrated FLASH time is programmed for 100, 115, 270 or 600 ms in EEPROM.

If the EARTH ("Connect a/b to earth function") is selected, the EARTH output becomes high and the MUTE output is active.
The time of earth connection is 400 ms.

When the FLA key is pressed the telephone number entered before the FLA key is stored in the redial register (EEPROM).
- After dialling 1 - 2 - 3 - "FLA" - on-hook Redial is 1 - 2 - 3.
- After dialling 1 - 2 - 3 - "FLA" - 4 - 5 - 6 - on-hook Redial is 4 - 5 - 6

The FLA key pressing acts like CE with respect to redial.

4.2.4. Disconnect function

The DIS (disconnect) key is only available if the function key option is programmed.
Touching the DIS key activates output notDP/FL for 800 ms.
In this case the telephone set turns to the ON-HOOK state for this calibrated time, after which it comes back to the OFF-HOOK mode.

The disconnect function acts like CE with respect to redial.

4.2.5. Mute function (M-key)

When no dialling or programming is active, every time this M-key is pressed the MUTE output goes to the active or inactive state depending on its previous status.
When the MUTE output is in the active state and an other key is pressed then the MUTE output is switched back to the inactive state.

5. EEPROM organisation and programming procedures

5.1. EEPROM organisation

The dialling, memory, and ringer options and the telephone numbers are all stored in EEPROM.
By using EEPROM no special backup requirement are necessary such as battery, current from the line, or very big capacitors.
Table 1 describes the meaning of each EEPROM byte at a repertory length of 16 and 20 digits.
Table 2 describes the meaning of each bit of all the bytes that do not contain telephone numbers.

Table 1. EEPROM organisation table.

	Repertory length is 16 digits		Repertory length is 20 digits	
Function	**Length**	**byte places**	**Length**	**byte place**
Redial	13 bytes	0 till 12	13 bytes	0 till 12
M0 or MEM + 0	8 bytes	16 till 23	10 bytes	16 till 25
M1 or MEM + 1	8 bytes	24 till 31	10 bytes	26 till 35
M2 or MEM + 2	8 bytes	32 till 39	10 bytes	36 till 45
M3 or MEM + 3	8 bytes	40 till 47	10 bytes	46 till 55
M4 or MEM + 4	8 bytes	48 till 55	10 bytes	56 till 65
M5 or MEM + 5	8 bytes	56 till 63	10 bytes	66 till 75
M6 or MEM + 6	8 bytes	64 till 71	10 bytes	76 till 85
M7 or MEM + 7	8 bytes	72 till 79	10 bytes	86 till 95
M8 or MEM + 8	8 bytes	80 till 87	10 bytes	96 till 105
M9 or MEM + 9	8 bytes	88 till 95	10 bytes	106 till 115
E-1	8 bytes	96 till 103	not available	--
E-2	8 bytes	104 till 111	not available	--
E-3	8 bytes	112 till 119	not available	--
Options	4 bytes	120 till 123	4 bytes	120 till 123
Program Blocking	1 byte	127		

Table 2. Option bit status and location.

Function	EEPROM byte	bit7	bit6	bit5	bit4	bit3	bit2	bit1	bit0
Not sending *	120	X	X	X	X	X	X	X	0
Sending *	120	X	X	X	X	X	X	X	1
Not sending #	120	X	X	X	X	X	X	0	X
Sending #	120	X	X	X	X	X	X	1	X
Mark to space ratio 3:2	120	X	X	X	X	X	0	X	X
Mark to space ratio 2:1	120	X	X	X	X	X	1	X	X
Tone/pause 60/90 ms	120	X	X	0	0	0	X	X	X
Tone/pause 70/70 ms	120	X	X	0	0	1	X	X	X
Tone/pause 80/80 ms	120	X	X	0	1	0	X	X	X
Tone/pause 100/100 ms	120	X	X	0	1	1	X	X	X
Tone/pause 100/140 ms	120	X	X	1	0	0	X	X	X
Tone/pause 140/140 ms	120	X	X	1	0	1	X	X	X
Flash duration 100 ms	120	0	0	X	X	X	X	X	X
Flash duration 115 ms	120	0	1	X	X	X	X	X	X
Flash duration 270 ms	120	1	0	X	X	X	X	X	X
Flash duration 600 ms	120	1	1	X	X	X	X	X	X
Mute is M1	121	X	X	X	X	X	X	0	0
Mute is notM1	121	X	X	X	X	X	X	0	1
Mute is M2	121	X	X	X	X	X	X	1	0
Mute is notM2	121	X	X	X	X	X	X	1	1
Access Pause time for pulse dialling (Inter-digit pause not included).									
A.P. time 1.5 s	121	X	X	X	0	0	X	X	X
A.P. time 2.5 s	121	X	X	X	0	1	X	X	X
A.P. time 3.0 s	121	X	X	X	1	0	X	X	X
A.P. time 6.0 s	121	X	X	X	1	1	X	X	X
Access Pause time for DTMF dialling (Tone-off time not included).									
A.P. time 1.0 s	121	X	X	X	0	0	X	X	X
A.P. time 1.5 s	121	X	X	X	0	1	X	X	X
A.P. time 3.5 s	121	X	X	X	1	0	X	X	X
A.P. time 6.0 s	121	X	X	X	1	1	X	X	X
General program proc.	121	X	X	0	X	X	X	X	X
German program proc.	121	X	X	1	X	X	X	X	X
Repertory 16 digits	121	X	0	X	X	X	X	X	X
Repertory 20 digits	121	X	1	X	X	X	X	X	X
Ringer via RTO pin	122	X	X	X	X	X	X	X	0
Ringer via TONE pin	122	X	X	X	X	X	X	X	1

Table 2. Option bit status and location (con't).

Function	EEPROM byte	bit7	bit6	bit5	bit4	bit3	bit2	bit1	bit0
Ringer melody A	122	X	X	X	X	0	0	X	X
Ringer melody B	122	X	X	X	X	0	1	X	X
Ringer melody C	122	X	X	X	X	1	0	X	X
Ringer melody D	122	X	X	X	X	1	1	X	X
Ringer volume 1	122	X	X	0	0	X	X	X	X
Ringer volume 2	121	X	X	0	1	X	X	X	X
Ringer volume 3	122	X	X	1	0	X	X	X	X
Ringer volume 4	122	X	X	1	1	X	X	X	X
Ringer repetition 1	122	0	0	X	X	X	X	X	X
Ringer repetition 2	122	0	1	X	X	X	X	X	X
Ringer repetition 3	122	1	0	X	X	X	X	X	X
Ringer repetition 4	122	1	1	X	X	X	X	X	X
Ringer detection LOW 1	123	X	X	X	X	X	X	0	0
Ringer detection LOW 2	123	X	X	X	X	X	X	0	1
Ringer detection LOW 3	123	X	X	X	X	X	X	1	0
Ringer detection LOW 4	123	X	X	X	X	X	X	1	1
Ringer detection HIGH 1	123	X	X	X	X	0	0	X	X
Ringer detection HIGH 2	123	X	X	X	X	0	1	X	X
Ringer detection HIGH 3	123	X	X	X	X	1	0	X	X
Ringer detection HIGH 4	123	X	X	X	X	1	1	X	X
A to D keys	123	X	X	X	0	X	X	X	X
Function keys	123	X	X	X	1	X	X	X	X
Flash function	123	X	X	0	X	X	X	X	X
EARTH function	123	X	X	1	X	X	X	X	X
No on-hook dialling	123	0	X	X	X	X	X	X	X
on-hook dialling control	123	1	X	X	X	X	X	X	X

5.2. EEPROM programming procedures

The PCD3330-1 supports four EEPROM programming procedures:

1) LNR, as described for normal operation procedures.
2) Repertory numbers, as described for normal operation procedures.
3) Via pins 1 to 4, already described in the PCF8581 EEPROM part.
4) Via keyboard.

Method 3 is normally used by the setmaker before the set leaves his factory.
Method 4 is most suited for usage in the field (e.g. the shop where the set is purchased).

5.2.1. EEPROM programming procedures via keyboard

In the field all telephone options can be changed easily by a special program procedure:

- Depress the STO-key (this selects the program mode),
- depress the LNR-key (switches the program module to storing EEPROM options),
- depress the first key of a three digit access code (the 1),
- depress the second key of a three digit access code (the 6),
- depress the third key of a three digit access code (the 0),
- depress the LNR-key again (end the access code),

- press the byte number (last digits of the EEPROM byte number given in table 2),

- press the number of the bit to change (see table 2),

- press 0 or 1 (this changes the EEPROM bit contents),

- depress the LNR-key, which stores the correction into EEPROM, now select a new byte or go to end,

- end the routine by pressing the STO-key again.

If during this procedure a mistake is made correction is possible after proper access code by pressing the LNR-key and during access code only by STO-key.
In all cases the routine can be ended by pressing the STO-key.

Example:
Change the mark-to-space ratio from 3:2 to 2:1.
Then bit 2 of byte 120 has to be changed from 0 to 1.
The necessary action is as follows:

- Depress the STO-key,
- depress the LNR-key,
- depress the 1-key (first digit access code),
- depress the 6-key (second digit access code),
- depress the 0-key (third digit access code),
- depress the LNR-key again (end the access code),

- press the 0-key (last digits of EEPROM byte 120 is the 0),

- press the 2-key (bit 2 has to be changed),

- press the 1 (this changes the mark-to-space ratio to 2:1),

- press the LNR or STO-key.

STO-key will end the programming procedure, while after LNR-key a new byte which bit has to be changed can be selected.

6. LIST OF REFERENCES

1. Objective specification PCD3330-1.
   J.C.F. van Loon, May 1992.

2. ICs for Telecom Philips Data Handbook IC03a/b 1991.

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4. TEA1060 Family Versatile Speech/Transmission ICs for Electronic Telephone Sets Designers' guide.
   P.J.M. Sijbers.
   12 NC : 9398 341 10011, July 1987.

5. TEA1083A: Data sheet; TEA1083/TEA1083A, Deceember 1991.

   Application Note; User's manual PR4535X demo board with TEA1083A-TEA1064A call progress monitoring application.
   F.v.Dongen, June 1991.

6. Measures to meet EMC requirements for TEA 1060-family speech transmission circuits.
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   Philips PCALE Report nr: ETT89016 Oct.'89.
   12 NC : 9398 071 30011, Oct. 1989.

7. APPENDIX I : JUMPER SETTING

JUMPER J1 : closed for automatic gain control.

JUMPER J2 : selects Pulse or DTMF dialling mode.

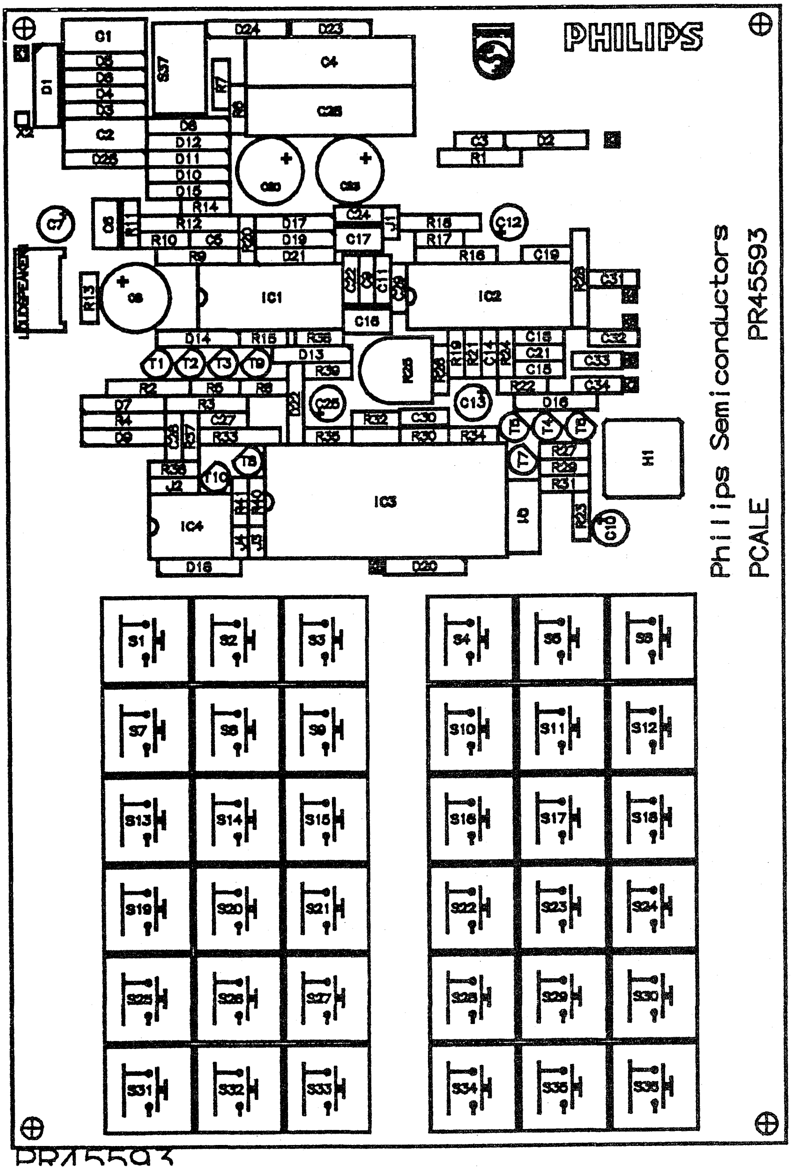
JUMPER J3 : closed for read/open for write internal EEPROM PCD3330-1.

JUMPER J4 : closed for selecting the test mode read/write EEPROM PCD3330-1.

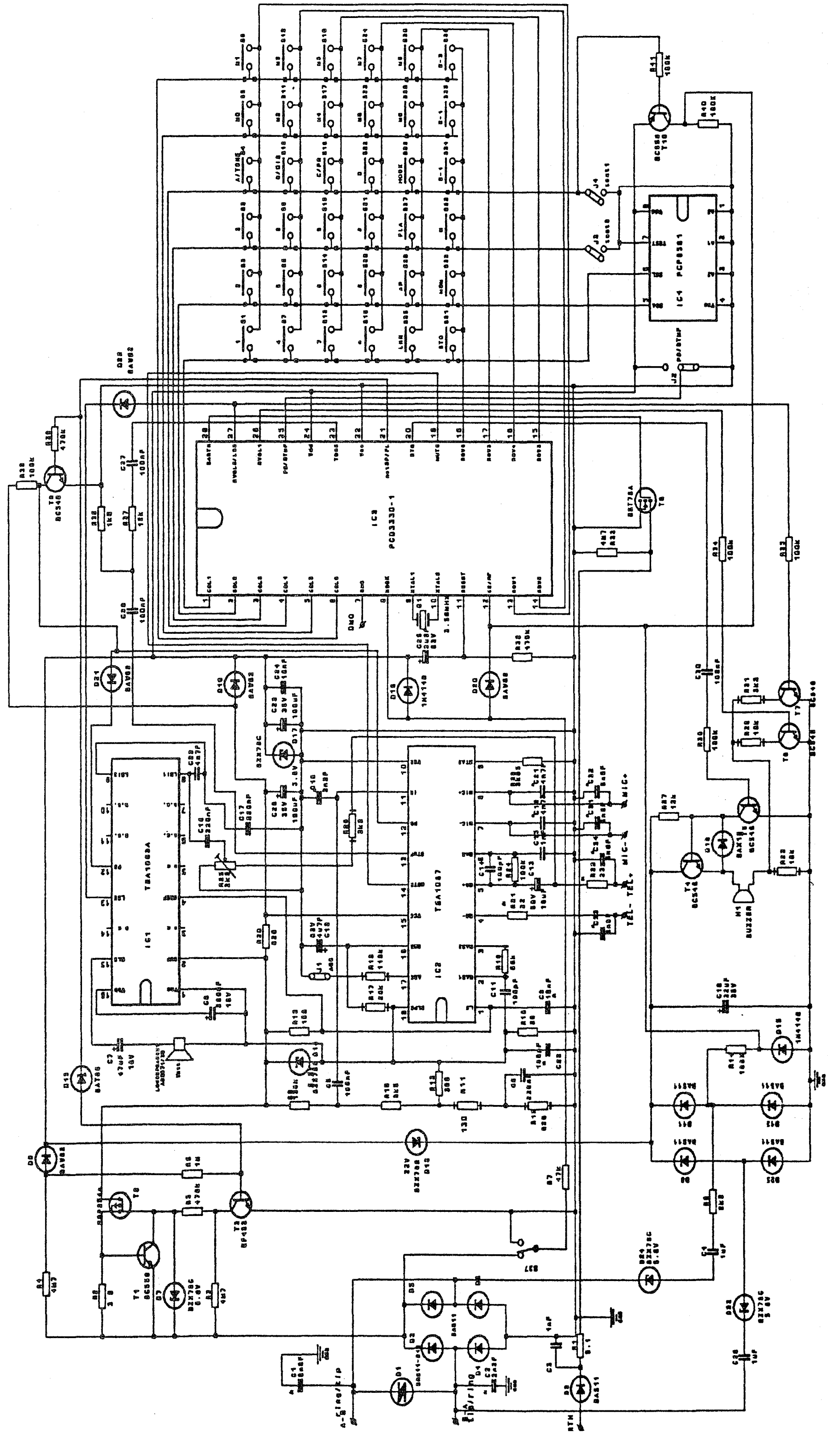
8. APPENDIX II : ADJUSTING PARAMETERS FOR TEA1067 and TEA1083A

Function	Components
Set impedance real or complex	R20
Re-balance anti side bridge (*R16=20 Ohm advised value)	R9/13/16, Zbal *Zbal: R11/12,C6
DC-characteristics * DC-slope * Max. line voltage * type of polarity guard	R17
Supply for periherals	R17, R20
Compromise BRL/Supply	
Gain control start/stop	R18
Microphone gain	R19, ext. comp.
Mic. ampl.: Upper cut-off freq.	C11
Mic. ampl.: Lower cut-off freq.	not mounted
DTMF input level adjustment	R37
Rec. gain: single ended or sym.	R24
Rec. ampl.: Upper cut-off freq.	C14, C15
Rec. ampl.: Lower cut-off freq.	C5, C13
TEA1083A output voltage adj.	R25, R26

9. APPENDIX III : DEMO-BOARD COMPONENTS LAYOUT



10. APPENDIX IV : CIRCUIT DIAGRAM OF THE DEMONSTRATION BOARD



APPLICATION NOTE Nr	ETT95007.0
TITLE	OM4757 Demonstration board PCD3332-3/TEA1064B-1062/TEA1093-1094
AUTHOR	Fred van Dongen
DATE	April 1995

CONTENTS

1. Introduction

The demonstration model OM4757 is a feature phone set based on applications of:

- An interface circuit with electronic hook switch between line and transmission IC.
- TEA1064B or TEA1062 transmission circuit. Both circuits contain speech and line interface functions for use in fully electronic sets. The TEA1062 is a basic transmission IC. The TEA1064B is more flexible and contains more functions like a dynamic limiter in sending direction, a Power Down input and an inverting receiver output stage.
- TEA1093 or TEA1094 handsfree circuit. Both IC's contain send and receive functions and a duplex controller with signal and noise monitors. The TEA1093 incorporates a supply and has to be applied in conjunction with a TEA106X in line powered applications. The TEA1094 can be supplied from an external source or from the line via a coil. See paragraph 3.3.2 concerning the consequences.
- PCD3332-3 multi-standard pulse/tone repertory dialler/ringer IC. The dial parameters can be set by diode options to specific country requirements. On-chip tone generators are used for dialling and ringer melody generation. A discriminator input enables the tone output at correct ringer frequencies. Repertory dialling is up to 13 numbers of 32 digits.
- A discrete ringer with PXE output stage

The OM4757 is provided with a microphone and a loudspeaker in the base for handsfree operations and a handset, with microphone and earpiece, connected with the base via the handset cord for handset operations. Provisions are made for use of an other handset or other capsules. The OM4757 includes a keypad.

Supply connections can be made via the line-jack and line-selection jumpers. The delivered handset is connected with the application via the handset-jack and selection jumpers.

The application of the OM4757 offers the following functions:

- Dialling
  - Pulse, DTMF, on-hook dialling
  - Repertory, recall, program functions
  - Flash or Earth register recall
  - Diode options for parameter selection
  - Confidence tone
- Transmission
  - Adjustable line voltage
  - Adjustable set impedance
  - Adjustable send and receive gains
  - Side tone reduction
  - Automatic gain control
  - Dynamic limiter of send signal (TEA1064B)
- Handset operation
- Handsfree operation
  - With TEA1093 powered from the line
  - With TEA1094 powered from an external supply or from the line

  - Adjustable send and receive gains
  - Loudspeaker amplifier with single (TEA1094) and double output (TEA1093)
  - Signal and noise envelope monitors with adjustable sensitivities and time constants.
  - Adjustable switching range and switching timing
- Listening-in operation
  - Voice switched
- Ringing
  - PXE buzzer
  - Volume control, tone selection
  - Diode options for parameter selection

This User's Manual is divided into three main parts: instructions for use, a brief description of the application and the settings of the application and how to change them.

'Application hints' and 'Bill of components' are added in an appendix.

For more details of the transmission IC's, handsfree IC's and controller is referred to the reports listed in chapter 7.

2. Use of the OM4757

2.1 External connections

Fig.7 illustrates the OM4757 with its components, modular jacks and test or connection terminals.

The application can be connected to the telephone line by a 6-pins LINE jack via a 2*6 jumper matrix or directly by the A_B-B_A terminals.

The capsules of the matched handset are connected with the application by a 4-pins HANDSET jack via the double 2*4 jumper matrix. The jumpers settings depend on the connections in the handset cord; they are in accordance with the delivered handset.

The purpose of the test or connection terminals is to measure the signals of the application. They are intended also to replace the mounted capsules by external ones, and to connect an external supply source for application of the TEA1094. The purpose of these terminals are summarised by:

Test / Connection terminal(s):	Purpose:	Remove:
A_B-B_A	Connection of telephone line	
EARTH	Terminal Earth function	
VEE / GND	System ground / TEA1093 ground reference	
P_LINE	Positive line level transmission circuitry	
VCC, VDD, VBB, VBBM	System supply levels	
VBB-VEE	External supply TEA1094	
LSP1-GND, LSP1-LSP2	External loudspeaker in SEL, BTL	Jumper J5
VBBM(+)-HFMIC(-)	External base microphone	Jumper J6
CE, CSI, MUTE, PD, DPN,		
LSP_ON, HANDS_N, DTMF	Test levels controller	
MIC_MUTE-GND	Test level of controller MUTE	
	Cannot be applied to MUTE the transmit channel of the TEA109X	
VR, BUZZ, VEE	Test levels ringer	
VR-BUZZ	External buzzer	Jumper J7

2.2 TEA1064B replaced by the TEA1062

The TEA1062 can be placed in the 16-pins shadow socket which is mounted next to the socket of the TEA1064B. The position of jumper J4 has to be as follows:

Transmission IC:	Position of J4:
TEA1064B	left
TEA1062	right

2.3 TEA1093 replaced by the TEA1094

For application of the TEA1093, jumper J1 is placed in the lower position and the jumpers J2 as well as J3 in the left-hand position. The TEA1093 can be replaced by the TEA1094 by using the same socket. The following possibilities are offered:

Handsfree IC:	Supply of handsfree IC:	Position of J1	J2	J3:
TEA1093	from line current	lower	left	left
TEA1094	from ext. source	upper	left	right
TEA1094	from line current	lower	right	right

The external voltage source, to supply the TEA1094, has to be connected between the terminals VBB (+) and VEE (-). Maximum voltage is 12 V.

2.4 Operation modes of the OM4757

The OM4757 offers 5 operation modes:

1. STANDBY mode
2. HANDSET mode
3. HANDSFREE mode
4. LISTENING-IN mode
5. RINGER mode

   A sixth operation mode within the handset, handsfree or listening-in mode is the

6. DIALLING mode

Actions to be taken to enter or to leave each mode, including the status (=) or changes (→) of the controller signals, are described in the following chapters. Extended dialling functions and programming operations are detailed described in [7].

2.4.1 STANDBY mode

Entering the STANDBY mode:

STANDBY mode is only entered if CE, CSI, LSP_ON and HANDS_N are low for a specific time.

Leaving the STANDBY mode:

STANDBY mode is cancelled if CE goes high. The following actions are to be taken by the µC:

- Start-up from µC
- Scan all I/O's for their status
- If handset is lifted: HANDSET mode. CSI → high.
- If key 'HOOK' is pressed and released: HANDSFREE mode. LSP_ON → high, HANDS_N → high.
- If non of the above applies: RINGER mode

2.4.2 HANDSET mode

Entering the HANDSET mode:

HANDSET mode can be entered from all modes by lifting the handset; CSI → high.

Leaving the HANDSET mode:

- Key 'HOOK' is pressed and released: LISTENING-IN mode. LSP_ON → high. Toggling between HANDSET mode and LISTENING-IN mode can be done by the HOOK-key.
- Key 'HOOK' is pressed while handset is put back on the cradle and key 'HOOK' is released: HANDSFREE mode. CSI → low, LSP_ON → high, HANDS_N → high.
- Handset is put back on the cradle: STANDBY mode. CSI → low.

2.4.3 HANDSFREE mode

Entering the HANDSFREE mode:

Handsfree mode can be entered from all other modes by pressing key 'HOOK'; CSI = low, LSP_ON = high, HANDS_N = high.

Leaving the HANDSFREE mode:

- Key 'HOOK' is pressed and released: STANDBY mode. LSP_ON → low, HANDS_N → low.
- Handset is lifted: HANDSET mode. CSI → high, LSP_ON → low, HANDS_N → low.

2.4.4 LISTENING-IN mode

Entering the LISTENING-IN mode:

Listening-in mode can be entered from the handset mode by pressing key 'HOOK': CSI = high, LSP_ON → high, HANDS_N = low. Hook-key serves as toggle switch.

Leaving the LISTENING-IN mode:

- Handset is put back on the cradle: STANDBY mode. CSI → low, LSP_ON → low.
- Key 'HOOK' is pressed and released: HANDSET mode. LSP_ON → low.
- Key 'HOOK' is pressed while handset is put back on the cradle and key 'HOOK' is released: HANDSFREE mode. CSI → low, LSP_ON → high, HANDS_N → high.

2.4.5 RINGER mode

Entering the RINGER mode:

Ringer mode can only be entered from the STANDBY mode after detection of an incoming ringer signal; CE → high, CSI = low, LSP_ON = low, HANDS_N = low.

- The keys 'VOL+/VOL–' can be used to control the ringer sound level.
- The keys '1, 2, 3' are available for ringer tone selection.

Leaving the RINGER mode:

RINGER mode is left:

- If CE remains low for time out: STANDBY mode
- If handset is lifted: HANDSET mode. CSI → high.
- If key 'HOOK' is pressed and released: HANDSFREE mode. LSP_ON → high, HANDS_N → high.

2.4.6 DIALLING mode

Dialling operations are possible in handset, handsfree and listening-in mode. To select the dialling modes (refer also to TABLE 2 and Fig.7:

- PULSE dialling : diode switch PTS = ON
- DTMF dialling : diode switch PTS = OFF
- EARTH recall : diode switch F_E = ON

  the EARTH pin will be short circuited to the A_B or B_A line inputs during 400ms.
- FLASH operation : diode switch F_E = OFF

  The line current will be interrupted during a time selected by the FTSA and FTSB. diode switches.

  The FLASH function or EARTH recall can be activated by pressing the key 'FLASH'.
- Keys M1 up to M10 can be used to store 10 numbers and to access one of the stored numbers. Diode switch MLA has to be in the ON-position.

3. Description of the application

A blockdiagram of the application of the OM4757 is shown in Fig.1. The description is referred to Fig.8 till Fig.11 showing the circuit diagrams of the ringer, electronic hookswitch/interrupter and controller PCD3332-3 respectively the transmission part of the TEA1064B or TEA1062 and TEA1093 or TEA1094. The settings are summarized in chapter 4 and the characteristics of the application in chapter 5.

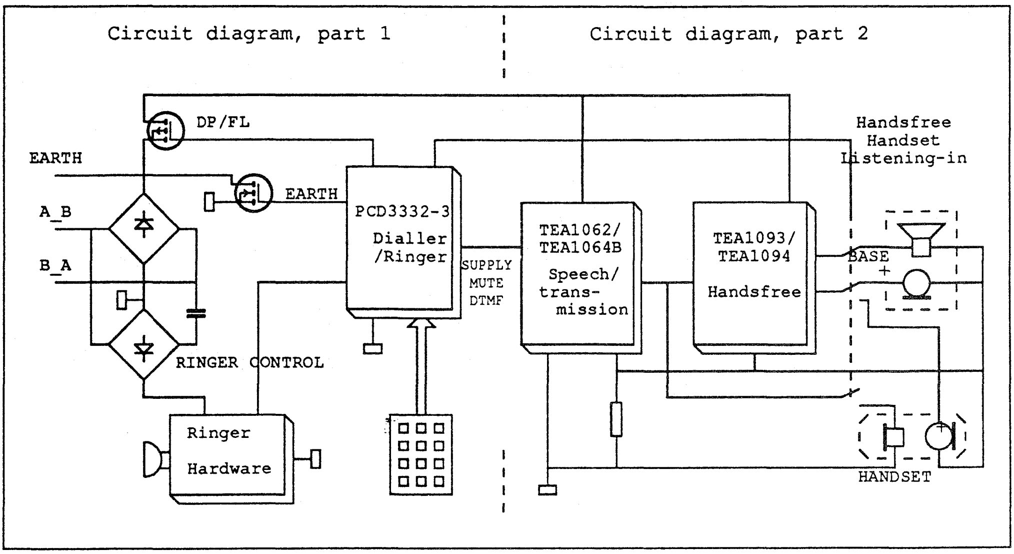


Fig.1 Block diagram of the OM4757 application

3.1 Polarity guard and protection

Two diode bridges are applied, consisting of D104 up to D108 for the transmission circuitry and D109 up to D112 for the ringer stage, to ensure proper functioning independent of the polarity of the line voltage and ringer signal.

Protection is achieved by break-over diode D100, type BR211-220, between the A_B-B_A terminals, by the current limiting components R100 and T100, by a 12 V zener diode between LN and VEE of the transmission IC and by the zener diode D120 between VDD and VSS of the controller PCD3332-3.

The voltage across the ringer output stage is limited by the zener diodes D117 and D120.

3.2 Electronic hookswitch / interrupter

The electronic hookswitch and line current interrupter consist of T102, P-channel enhancement D-MOS BSP304A, and inverter T103. The inverter is controlled by the DPN_FLN open drain output of the processor. During hook-off, when the handset is lifted, or when the 'HOOK-key' is activated, DPN_FLN is high resulting in a conducting T102.

Conducting of T103 is initiated by the high ohmic resistor R102 and is taken over by R110 connected to the positive line wire P_LINE. The PD inputs of the transmission and handsfree IC's are kept passive via D111 and R108 and the separation diodes D4 and D7, Fig.8.

Interruption of the line current is achieved by a low DPN_FLN level; the PD's are than active.

Components R100 and T100 limit the current through T102 when the current exceeds about 150mA. This current limiter only provides protection against current surges. It is not designed for continuous limitation of the line current.

Series diode D114, between T102 and P_LINE, is applied to get a fast trailing edge of the CE pulse after hook-on or line breaks.

3.3 Handset / Handsfree application

3.3.1 Transmission IC's TEA1064B / TEA1062

The application can be used with the TEA1064B or TEA1062 transmission circuit. Because of the different pinning of the IC's two IC sockets are mounted on the board; refer to chapter 2.2. Jumper J4 has to be set in accordance with the applied transmission IC, TEA1064B or TEA1062, to get the proper DC setting.

Jumpers J1, J2 and J3 are intended to select the different supply configurations of the handsfree IC; refer to chapter 2.3 and see next paragraph.

Both IC's are from the TEA106X-family. The difference of the TEA1064B and TEA1062 concerns the PD input, dynamic limiter DLS input and the second receiver output QR- which are present by the TEA1064B and not by the TEA1062.

Detailed information of the transmission IC's can be found in the data sheets, [1], and in the reports [2], [3] and [8].

3.3.2 Handsfree IC's TEA1093 / TEA1094

The TEA1093 as well as the TEA1094 can be applied. The difference between TEA1093 and TEA1094 concerns the supply part, resulting in a different supply configuration, and the loudspeaker amplifier outputs. SREF, SUP, PD, VA and LSP2 are available in the TEA1093, not in the TEA1094. Moreover, the TEA1094 has no VBB voltage stabilizer; the VBB-pin is in this case the supply input. The 28-pins IC socket on the board is suitable for both types. Jumpers J1, J2 and J3 have to be placed in accordance with the applied HF circuit.

TEA1093

The TEA1093 application is connected between LN and SLPE of the transmission IC to have most of the line current available for the loudspeaker amplifier. The transmission IC consumes 3 mA bias current.

The ground reference (GND) of the TEA1093 application has to be connected to TEA106X-SLPE by means of jumper J1 (lower position on the board). Jumper J2 has to be open while jumper J3 has to connect R_{sref2} with TEA106X-LN (both J2 and J3 in left-hand position).

TEA1094

The TEA1094 has to be supplied via the VBB pin. Two applications possibilities are offered: supply of the TEA1094 from an external source or from the line via the mounted coil L1.

External supply: The ground reference of the TEA1094 has to be connected to VEE while an external voltage source has to be connected between VBB(+) and VEE(-). The maximum voltage is 12 V.

Because of the external VBB supply and no supply connections exists between TEA1094 and transmission IC, the OM4757 operates from low line currents of ≥8 mA.

J1 has to be placed in the upper position, J2 has to be open (left-hand position) and J3 has to connect the TEA106X-LN with P_LINE (J3 in right-hand position).

Line supply: The TEA1094 is connected between LN (=P_LINE) and SLPE via an LC filter. The supply principle is different compared with a TEA106X / TEA1093 combination because of the separated supply of the TEA106X and TEA1094.

This application operates without loss of performance when the minimum line current, I_{line}-min, is sufficient to supply the:

- TEA106X via VCC
- Output stage of the TEA106X to stabilize the line voltage <u>and</u> to generate the required send level on the line across set and line impedance
- TEA1094: stand-by current and supply current loudspeaker amplifier:

  IBB ≥ IBB0-max + Vslp1-peak-max / (π. R_{loudsp}) (A)
- Controller.

The current in the output stage of the TEA106X is reduced by the supply current of the TEA1094 taken from LN. More over, the transmit line level versus line current is shifted by more than IBB0 because of IBB0 and the modulation current into the supply components of the TEA1094.

Coil L1 reduces the modulation current; it is an iron-core type designed on 500 mH at 50 Hz. The L-value decreases as function of frequency and DC current. The data of L1 is given in chapter 5.

To give an idea of I_{line}-min in transmit mode: to generate 6 dBm, across 300Ω in total (Z_{set} // Z_{line}), requires a line current of about 16 mA.

In only receive mode ≥15 mA is required to generate 20 mW output power across 50 Ω.

Coil L1 has to be connected to P_LINE by J2 (right-hand position), J3 has to connect TEA106X-LN with P_LINE (J3 in right-hand position) while J1 has to connect TEA1094-GND with TEA106X-SLPE (J1 in lower position)

Remarks / Warnings concerning TEA1094 connected between LN - SLPE and supplied from the line:

- **Line currents of <16 mA reduce the performances.**
- **Do not connect TEA1094-GND with VEE. Coil L1 is not suitable for this application.**
- **The LC supply network can causes LF instabilities or oscillations. A damping resistor, consisting of the DC resistance of the coil or by R_{vvb} (short circuited on the PCB) seems to be required.**

Detailed information of the handsfree IC's can be found in [1], [4], [5] and [6].

3.3.3 DC settings

Transmission IC's:

The stabilized voltage between LN and VEE is set to ≈4.5 V at 15 mA for the TEA1062 as well as the TEA1064B (selected by J4) to get the proper DC levels of the TEA1093. The voltage between SUP and VBB has to be ≥700 mV.

The voltage at P_LINE measures ≈5.1 V at 15 mA in case the TEA1093 is applied and 4.5 V with the TEA1094. R_{sref2} is applied to prevent disturbances of the TEA1093-supply at clipping levels of the transmission signal at LN.

The line voltage across the A_B-B_A terminals is a result of the DC level at P_LINE and the voltage drops across D114, interrupter T102, R100 and the diode bridge. It measures 7.2 V with the TEA1093 and 6.6 V with the TEA1094 at 15 mA.

Handsfree IC's:

TEA1093: The stabilized VBB level is 3.6V-typ. VBB can be increased by Rva2 and decreased by Rva1. Take into account that an increase of the VBB level requires a modification of the level at SUP, by means of the DC setting of the transmission IC (R17 or R18), to keep the voltage difference between SUP and VBB at ≥700 mV.

TEA1094 external supply: VBB - VEE is the supply input. The set operates from 8 mA line current.

TEA1094 line supply: VBB is not stabilized. It is determined by the voltage difference between LN - SLPE, set by the transmission IC, and the voltage drop across R_{vvb} (short circuited on the board) and coil L1. VBB measures 4.1V at stand-by and 3.8 V at 20 mW output power across 50 Ω at 20 mA line current and receive mode.

3.3.4 Set impedance / side tone

The set impedance, between 300 Hz and 3400 Hz, is mainly determined by the network between P_LINE and VCC. On the PCB is mounted R1a = 619 Ω, R1b = 0 and C1b is open delivering a set impedance of about 600 Ω. To achieve complex set impedance the network has to be modified.

The side tone bridge, R11 R12 and C12, is optimized for 600 Ω set impedance in combination with a 5 km cable of 0.5mm diameter copper twisted pair (176 Ω, 38 nF per km). AGC is set with R6 = 110 kΩ. This gives optimum results for an exchange of 48 V and 600 Ω in combination with the 0.5 mm diameter cable (1.2 dB attenuation per km).

3.3.5 Transmission handsfree / handset / listening-in

The OM4757 includes a handset with electret microphone and dynamic earpiece and a base microphone and loudspeaker. Selection between handsfree, handset and listening-in capsules, is performed by the analogue multiplexer/demultiplexer IC5, type 74HC4053P which is supplied by VCC of the transmission IC. The multiplexer /demultiplexer is controlled by the signals HANDS_N and LSP_ON from the PCD3332-3. See chapter 2.4: Operation modes of the OM4757.

Handset / listening-in mode

The electret microphone of the handset is supplied from VBB via a smoothing filter. The microphone signal is attenuated by R_{mic2} with respect to R25 and R26 and offered to the mic input of the TEA109X via the selection switch IC5. The signal from Rmic2 is amplified by the TEA109X with 15 dB (R_{gat} = 95.3 kΩ), 10 dB attenuated by the couple network R20, R21 and R22 between TEA109X and TEA106X and 44dB amplified by the transmission circuit (R7 = 27.4 kΩ). This results in a overall gain of 49 dB from microphone (R_{mic2}) to the line terminals at 600 Ω line load and 600 Ω set impedance.

The earpiece is selected by T1 which is controlled by IC5. The receive gain of the TEA106X is set to -4.5 dB by means of R4 (68k1). Due to series resistors R27 and R28 and the on-resistance of T1 the overall gain from line to earpiece results in -6dB.

The loudspeaker amplifier is disabled in this mode by a low level at pin DLC/MUTER of the TEA109X, unless LISTENING-IN is selected.

During listening-in the loudspeaker amplifier is operational. The overall gain from line to loudspeaker is 24.5 dB, at R_{vol} at maximum position, as a result of the -4.5 dB from line to QR and 29 dB gain from RIN to LSP1. The receive gain of the TEA109X is set by R_{gar} = 221 kΩ.

Handsfree mode

The gain from selected base microphone (Rmic1) to the line, at 600 Ω load, is the same as in the handset mode; 49 dB.

The base microphone can be muted by applying a high level between MIC_MUTE and GND.

The overall gain from line to loudspeaker is 24.5 dB, at R_{vol} at maximum volume (R_{vol} = 0), as a result of the -4.5 dB from line to QR and 29 dB gain from RIN to LSP1. The receive gain of the TEA109X is set by R_{gar} = 221 kΩ.

Every increase of R_{vol} with 950 Ω results in 3 dB lower loudspeaker gain.

The MIC_MUTE of the transmit channel of the TEA109X is wired to the logic MUTE- output of the controller. Confidence tones are audible, via the loudspeaker, in HF/LI mode during DTMF dialling.

Duplex controller

Switching range: The switching range is proportional to the ratio between the resistors R_{swr} and R_{stab}. R_{stab} is fixed to 3.65 kΩ. R_{swr} can be varied between 3.65 kΩ and 1.45 MΩ resulting in a switching range of 0 dB up to 52 dB.

The volume setting affects the switching range such that the sum of the microphone and loudspeaker amplifier gain is kept constant. Therefore, volume control has a maximum range equal to the switching range.

In order to prevent howling, the loop-gain has to be below 0 dB with a margin of 10 dB up to 20 dB. This guarantees good balance return loss figures and stable operation.

On the board the switching range is set to 40 dB by Rswr = 365 kΩ, providing a howling margin of 10 dB. If AGC is used, the gain from MIC+/- to QR of the TEA106x is lowered with approximately 12 dB for line lengths below 5 km. In that case the howling margin is in the order of 22 dB.

Envelope detectors: The sensitivity of the detectors can be adjusted with R_{tsen} and R_{rsen}. They can best be adjusted such that currents in the order of 10 μA_{rms} are flowing through them at nominal signals. The DC blocking capacitors C_{tsen} and C_{rsen} form a high pass filter together with R_{tsen} and R_{rsen}. On the board R_{rsen} is set to 4.75 kΩ giving a dial tone detector threshold of about 60mV between RIN1 and RIN2. At the telephone line this means the dial tone detector level is set to 100mV (receive gain TEA106x is -4.5 dB). Capacitor C_{tsen} in series with Rtsen blocks DC and forms a high pass filter (225 Hz). R_{tsen} = 3.92 kΩ on the board and C_{tsen} to 180 nF.

With AGC active R_{tsen} has to be increased to 6.7 kΩ

The charge and discharge times of both signal envelope detectors are proportional to the value of the capacitors C_{tenv} and C_{renv} giving a maximum rise slope in the order of 85 dB/ms maximum, and a maximum fall slope of 0.7 dB/ms.

The noise envelope timing is proportional to the value of the capacitors C_{tnoi} and C_{rnoi}. On the board 4.7 µF is mounted at TNOI providing a maximum rise slope of 0.07 dB/ms maximum, and a maximum fall slope in the order of 0.7 dB/ms.

Switch over timing: The switch over time from Tx to Rx or vice versa is proportional to the value of capacitor C_{swt} connected to pin SWT. On the board 220 nF is mounted for C_{swt}, resulting in switch over times of around 13 ms.

The switch over time from Idle mode to Tx or Rx is also proportional to the value of C_{swt}. With the value chosen, switch over will take 4 ms.

The timing from Tx or Rx to idle mode is determined by resistor R_{idt} in conjunction with C_{swt}. On the board 2.2 MΩ is mounted for R_{idt} resulting in an idle mode timing of about 2 s.

3.4 Controller PCD3332-3

A single contact 6 x 5 matrix keypad is connected with the corresponding COL and ROW I/O's. The keypad includes 10 memory keys, M1 to M10, for direct access of the stored numbers (MLA = ON). The diode functions, see chapter 4.1, can be selected by means of the diode switches S127 and S128.

Output EARTH drives T101 to perform the Earth Recall function (F_E = ON). Open drain output DPN_FLN drives the electronic hook switch to perform pulse dialling and flash function (F_E = OFF). The position of the cradle switch determines the CSI level during stand-by (CSI = low) and handset mode (CSI = high).

CE_FDI is connected to the ringer bridge and the positive line wire to detect the operation mode of the controller. Diode D114 is connected between interrupter T102 and LN of the transmission IC to prevents too slow decay of the CE pulse edge after hook-on or line breaks.

A pull-up resistor R119 is connected to ROW5. Pressing the 'HOOK-key' connects ROW5 with COL4. CE will be activated via COL4 and T107 to wake-up the controller.

Supply of the PCD3332-3 is obtained from VCC via a back-up diode during transmission. In ringer mode, the supply is obtained from the ringer signal.

The supply during stand-by mode is described in next paragraph.

3.5 Ringer circuit

During stand-by the VDD capacitor is kept charged by the relative high exchange voltage via the high ohmic resistor R104 to speed up the initialization of the controller.

The controller enters the ringer mode when CE becomes high and CSI is kept low via the cradle switch, for frequencies between 20 Hz and 57 Hz or between 14 Hz and 75 Hz depending on the position of the RFS diode switch (see chapter 4.1).

The ringer circuit is build up by discrete components. The PXE capsule is from Murata, type PKM34EW-1224. Supply is delivered by the ringer signal from the exchange via a diode bridge, series resistor and capacitors and two low voltage zeners D102 and D103 to reduce the load from the ringer circuit on the telephone line during transmission.

The ringer is enabled at HF_RTE = high while volume control is performed by the VOL1 and VOL2 levels. Maximum sound pressure from the ringer capsule is obtained when both VOL1 and VOL2 are low.

The blocking diode D136 is mounted to prevent ringing actions during DTMF dialling in the HF mode. See Appendix A.

4. Settings of the application

The functioning of the application of the OM4757 depends on the positions of the jumpers, diode switches and the DC and transmission settings. Fig.7 shows the jumper and diode switch positions. They are listed below as a starting position of your experiences of the applied TEA1064B / TEA1093 combination; see paragraphs 2.2, 2.3 and 3.3 for replacement of the TEA1064B and the TEA1093.

4.1 Jumpers / Diode switches

Line jumpers: Make selections according interconnections of line cord.

Handset jumpers: Positions according to Fig.2.

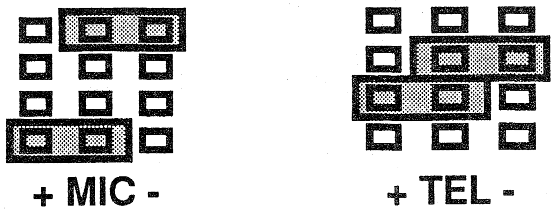


Fig.2 Handset jumpers

TABLE 1 Jumper settings

Jumper	Position	Selected / Connected
J1	Lower	TEA1093
J2	Left-hand	TEA1093
J3	Left-hand	TEA1093
J4	Left-hand	TEA1064B
J5	Present	Base loudspeaker
J6	Present	Base HF-microphone
J7	Present	Base buzzer

TABLE 2 Diode switch setting

Diode switch		Position	Selected	Note
S127-1	APT	OFF	Access pause 2s	ON : 4 s
S127-2	PTS	OFF	DTMF dialling	ON : Pulse dialling
S127-3	MLA	ON	Memory access M1-M10	OFF : M4 = MRC
S127-4	M/S	OFF	Mark/Space 2:1	ON : 3:2
S127-5	F_E	OFF	Flash	ON : Earth recall
S127-6	RDS	OFF	No ringer delay	ON : Delay 100 ms/25 Hz
S128-1	n.c.			Not applied
S128-2	FTSA	OFF	Flash time 95 ms	ON : 115 ms

TABLE 2 Diode switch setting

Diode switch		Position	Selected	Note
S128-3	FTSB	OFF	Flash time 95 ms	ON : 270 ms
S128-4	TBTA	OFF	Tone burst 70 ms/70 ms	ON : 100 ms/100 ms
S128-5	TBTB	OFF	Tone burst 70 ms/70 ms	ON : 85 ms/85 ms
S128-6	RFS	OFF	Ringer freq. range: 20 Hz-57 Hz	ON : 14 Hz-75 Hz

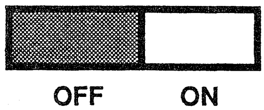


Fig.3 Switch positions on the PCB

4.2 DC, transmission, ringer settings

TABLE 3 DC, transmission and ringer settings

Setting		Adjustment	Remarks
	Basic settings	R5 {3.65 kΩ} Rstab {3.65 kΩ}	Do not modify
DC			
	Standby current	≈R102 // R104	Hook-on
	Voltage LN-SLPE	R17 / R18	630 mV is in series with LN
	DC-slope	R9 {20 Ω}	Include R100, R-on of T102
	Supply point VCC	C1 {100 µF}	Voltage drop: R1a, R1b
	Supply point VDD	C112 {100 µF}	From VCC via D123
	Artificial inductor	Creg {6.8 µF}	
	Bias current TEA106X	Rsref1 {100 Ω}	I = 315 mV/Rsref1 (A)
	Voltage VBB	Rva1 or Rva2	V(SUP-VBB) ≥ 700 mV
	HS microphone supply	R25, R26, Rmic2	From VBB via Rvbbm
	HF microphone supply	Rmic1	From VBB via Rvbbm
	Stability	L2 {150 µH}, Csref {4.7 nF}, Cvvb {470 µF}	I-L2 up to >100 mA
Transmission			
	Set impedance	R1a {619 Ω}, R1b {0}	Z-complex: R1a, R1b, C1b
	Side tone	R2, R3, R8, R11, R12, C12	
	AGC	R6 {110 kΩ}	
	Microphone gain(s)	R7 {27.4 kΩ-TEA106X}, Rgat {95.3 kΩ-TEA109X}	Attenuation: R20, R21 and R22
	Sensitivity HS mic	Rmic2 wrt R25, R26	
	Sensitivity HF mic	Rmic1	

TABLE 3 DC, transmission and ringer settings

Setting	Adjustment	Remarks
Freq. curve / stability	High pass: Cmic - Zin-MIC Low pass: Cmicx - Rmicx Low pass: Cgat - Rgat	Zin-MIC = 20 kΩ Also stability
Dynamic limiter Tx	C9 {470 nF}	Stability R10
Earpiece gain	R4 {68.1 kΩ}	Attenuation: R27, R28, R-on T1
Freq. curve / stability	Low pass: C4 - R4 Stability: C7 = 10 * C4 High pass: C5 - Zin-IR, C2 - Z-earpiece	 Zin-IR = 10 kΩ
Loudspeaker gain	R4, Rgar {221 kΩ}	A-BTL = A-SEL + 6 dB
Freq. curve / stability	Low pass: Cgar - Rgar High pass: Crin - Zin-RIN	Include C4 - R4 Zin-RIN = 20 kΩ Include Clsp1 - Z-loudspeaker
Volume control	Rvol {10 kΩ}	3dB reduction for each 950 Ω
Dynamic limiter Rx	Cdlc {470 nF}	
DTMF gain	R7	Attenuation: R122, R123
Duplex controller		
Switching range	Rswr {365 kΩ}	
Dial tone	Rsen {4.75 kΩ}	
Sensitivity	Rtsen {3.92 kΩ} Ctsen {180 nF} Crsen {150 nF}	 High pass with Rtsen High pass with Rrsen
Signal envelope	Ctenv {470 nF}, Crenv {470 nF}	
Noise envelope	Ctnoi {4.7 µF}, Crnoi {4.7 µF}	
Switch over timing	Cswt {220 nF}	
Idle mode timing	Ridt {2.3 MΩ}	
Ringer		
Ringer voltage VR	D117 {18 V}+D128 {5.6 V}	Include D136
Minimum sound level	R106 wrt R111	
Maximum sound level	R114 // R115 // R106 wrt R111	

5. Characteristics of the application

Line supply

VA_B (V)	Iine (mA)	Application
7.2	15	TEA1064B or TEA1062 / TEA1093 combination
11.5	100	TEA1064B or TEA1062 / TEA1093 combination
6.6	15	TEA1064B or TEA1062 / TEA1094 combination
10.9	100	TEA1064B or TEA1062 / TEA1094 combination

Standby current: <20 µA Vexchange = 48V

Transmission

BRL:	≥22 dB	300 Hz - 3400 Hz, Zref: 600 Ω R1a =619 Ω, R1b = 0 Ω
BRL	>18 dB	300 Hz - 3400 Hz, Zref: 220 Ω + 825 // 115 nF R1a = 220 Ω, R1b = 825 Ω, C1b = 115 nF

DTMF level: -6 dBm//600 Ω

Gain (dB)	Mode	Conditions (1 kHz)
49	HS-send	vline: 0 dBm, Zset: 600 Ω
-4.5	HS-receive	vline: 0 dBm // 600 Ω
48	HF/LI-send	vline: 0 dBm, Zset: 600 Ω
24.5	HF/LI-receive	vline: -23 dBm//600 Ω; max. volume control

Output power (P) of loudspeaker amplifier at 50 Ω loudspeaker, receive conditions, VBB = 3.6 V-typ, 1 kHz and maximum volume control:

HF-IC:	P > 0 @ Iline(mA):	P = 20 mW @ Iline (mA):	Pmax (mW)	@ Iline (mA)
TEA1093	>12	19	≈30	22
TEA1094	>9	15	≈33	19
External VBB = 5 V:				
TEA1094			≈50	8

Duplex controller TEA109X

Switching range	40 dB	
Dial tone detector threshold:	60 mV	(line treshold: 100 mV)
Rise slope signal envelope detector:	85 dB/ms	
Fall slope signal envelope detector:	0.7 dB/ms	

Rise slope noise envelope detector:	0.07 dB/ms
Fall slope noise envelope detector:	0.7 dB/ms
Switch over time Tx <--> Rx:	≈ 13 ms
Switch over time to Idle mode:	≈ 2 s

Ringer (RFS = OFF)

Freq.range:	20 Hz - 57 Hz
Start level:	20 Vrms at 22 Hz, 55 Hz
Start time:	120 ms at 45 Vrms / 22 Hz, 55 Hz (VDD level present)

Applied capsules

Handset:	Ericsson type RLGN40201/8B6
Electret microphone:	sensitivity -44.5 dBV/Pa at 1 kHz, 2 kΩ load
Dynamic earpiece:	sensitivity 49dBPa/V, 150W
Electret HF microphone:	sensitivity -44 dBV/Pa at 1 kHz, 2 kΩ load
Loudspeaker:	Philips type AD2071/Z50, 50 Ω
PXE ringer capsule:	Murata PKM34EW-1224

Supply coil L1

Coil with iron core: L = 500 mH 30 % at 50 Hz and 500 mVrms, R ≈ 32 Ω
Design from WWC - Philips Tilburg; code nr.: 8222 289 43362

6. Electromagnetic Compatibility

Measures are taken to reduce the sensitivity of the application for common mode amplitude modulated RF signals generated in the A_B lines.

The board is provided with a ground reference layer connected to VEE of the transmission IC. Basic protection of the TEA1064B / TEA1093 / PCD3332-3 combination is provided by:

- The line connection by means of the capacitors C100 {2.2 nF} and C101 {2.2 nF}
- Across the EARTH-switch T01 by means of C103 {2.2 nF}
- Across the VDD supply of the PCD3332-3 by means of C110 {10 nF}
- The TEA106X / TEA1093 supply with L2 {150 µH} and Rsref {4.7 nF}
- At the TEA106X receiver input with C28 {2.2 nF} and receiver output by means of the filter elements R27, C33 and R28, C32 {10 Ω - 10 nF combinations}
- At the handset microphone input by means of R25, C34 and R26, C35 (1 kΩ - 10 nF combinations} and C37 {2.2 nF}
- The microphone inputs of the TEA106X with C29 and C30 {2.2 nF both} and gain adjustments pins by means of C8 {100 pF} and C6 {100 pF}

7. References

[1] DATA HANDBOOK IC03 'Semiconductors for Telecom Systems'

[2] Application of the speech-transmission circuit TEA1062 P.T.J.Biermans, Report nr.: ETT89008

[3] Application of the speech-transmission circuit TEA1064 F.van Dongen / P.J.M.Sijbers, Report nr.: ETT89009

[4] Application of the TEA1093 handsfree circuit C.H.Voorwinden / K.Wortel, Report nr.: ETT/AN93015

[5] Tentative device specification -TEA1094 handsfree circuit- March 1994, version 2.1

[6] Application of the TEA1094 handsfree circuit C.H.Voorwinden, Report nr.:ETT/AN94004

[7] Objective specification PCD3333-3: In preparation

[8] TEA1060 family, versatile speech/transmission IC's for Electronic telephone sets Designers Guide. P.J.M Sijbers / Release 7-87, 12nc 9398 341 10011

APPENDIX 1 Application hints

This appendix describes some circuit proposals to improve or to adapt the offered applications of the OM4757.

1.1 Preventing ringing actions during Handsfree

The ringer circuit is also selected during HANDSFREE by the combined controller output HF_RTE. A ringer signal can be heard when a DTMF-key is pressed due to the generated DTMF tone and the DC coupling between VDD and VR. D136 in Fig.8 is mounted on the OM4757 to prevent this coupling, but because of the present charge of C105 (22 µF) it takes some seconds before the ringing sounds stops.

Disabling T106 by an inverted level from LSP_ON (= LFE), at HF_RTE active, gives the DTMF signal no change to be heard by the buzzer. The principle is shown in Fig.4. Two resistors and a PNP transistor have to be added. Diode D136 can be removed.

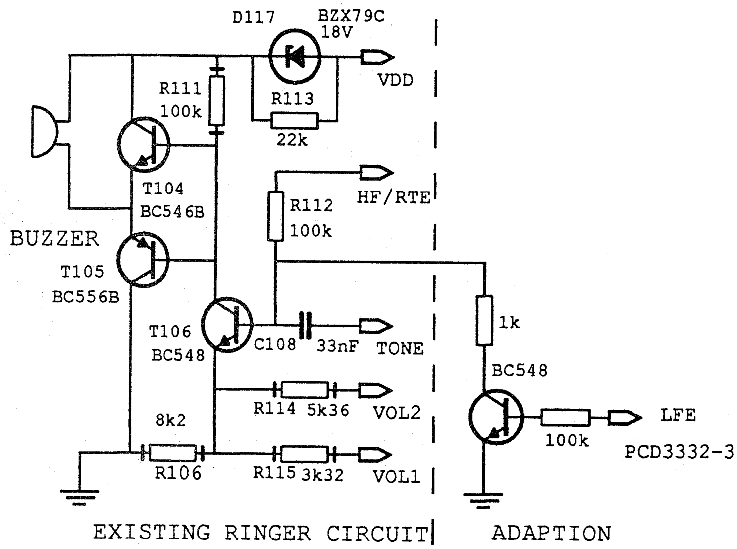


Fig.4 Improvement of ringer circuit

1.2 Volume control loudspeaker signal

Volume control of the loudspeaker signal is performed by potentiometer R_{vol}. When digital volume control is desired a solution is offered in Fig.5. Refer also to [4].

With a 4 bit digital volume control 16 volume levels can be set by 3 dB steps, from 0 dB to 45 dB attenuation. The switches can be either MOSFETs or analogue switches, for instance Philips HCT4066 type, controlled by the VOL+ /VOL- keys of the keyboard.

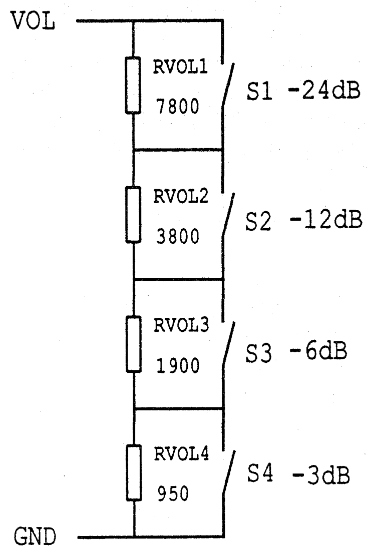


Fig.5 Digital volume control

As shown in Fig.8 and Fig.9 are the controller outputs VOL+/VOL- applied to control the ringer volume via pull-up resistors. A modification of the applied interface between controller and ringer is required if the ringer signal as well as the loudspeaker signal has to be controlled by the same VOL+/VOL- outputs and keys.

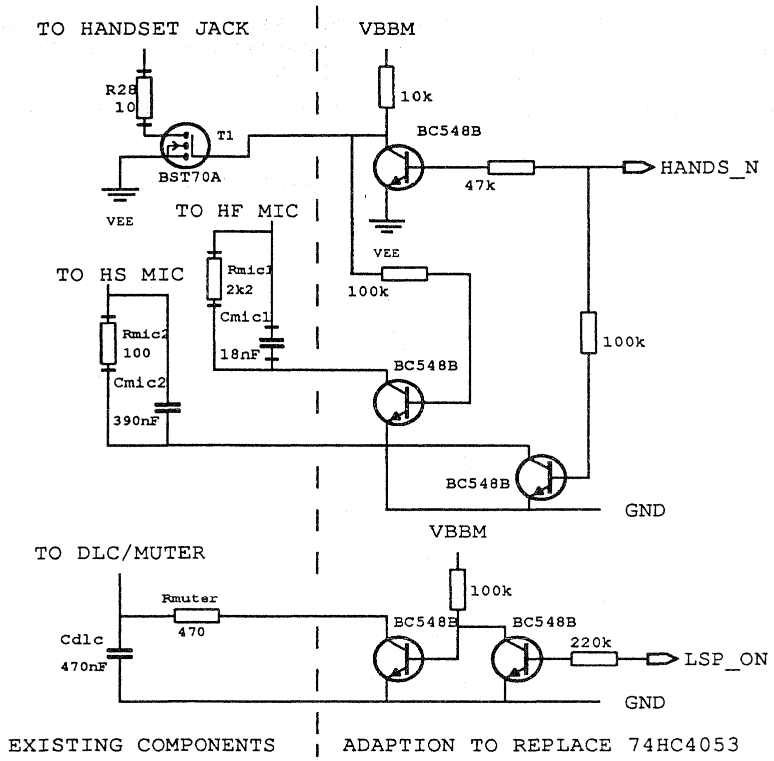


Fig.6 Replacement of the multiplexer/demultiplexer 74HC4053 by discrete components

1.3 Replacement of the mux/demux 74HC4053 by discrete components

The 74HC4053 (IC5) selects the HS or HF capsules depending on the required mode under control of the PCD3332-3. The following actions are performed:

TABLE 4 Handset and Handsfree actions

Mode	Control signal	Action
Handset	HANDSET_N = low	Handset mic connected to TEA109X-MIC input Earpiece connected to VEE by T1
	LSP_ON = low	Loudspeaker amplifier TEA109X disabled: DLC/MUTER input = low
Handsfree	HANDSET_N = high	Handsfree mic connected to TEA109X-MIC input Earpiece disconnected from VEE
	LSP_ON = high	Loudspeaker amplifier TEA109X enabled

Proposal Fig.6 shows a 'discrete' solution to perform the same actions by the control signals.

The 'inverted' HANDS_N controls T1, to select the earpiece. The networks Rmic1-Cmic1 and Rmic2-Cmic2 (Fig.10) are disconnected from GND. In this proposal they will be connected to GND by means of the transistor switches which are controlled by HANDS_N and the 'inverted' HANDS_N.

The mute function of the DLC/MUTER input is activated by LSP_ON via the 'VEE-GND' level shifter.

Note: Avoid DC paths between SLPE (TEA109X-GND) and VEE, in parallel with R9 (20 Ω), which could disturb the working of the TEA106X.

APPENDIX 2 Bill of materials

REF	PART NO.	VALUE/TYPE	SERIES	RATING	TOL
C1	2222-037-51101	100uF	C037	50V	20%
C2	2222-037-51109	10uF	C037	50V	20%
C4	2222-630-03561	560pF	C630	100V	10%
C5	2222-370-21333	33nF	C370	100V	10%
C6	2222-638-10101	100pF	C638-NP0	100V	
C7	2222-370-21562	5n6F	C370	100V	10%
C8	2222-638-10101	100pF	C638-NP0	100V	
C9	2222-370-11474	470nF	C370	63V	10%
C12	2222-370-11224	220nF	C370	63V	10%
C100	2222-368-55222	2n2F	C368	400V	10%
C101	2222-368-55222	2n2F	C368	400V	10%
C102	2222-368-45155	1u5F	C368	250V	10%
C103	2222-630-03222	2n2F	C630	100V	10%
C104	2222-368-45155	1u5F	C368	250V	10%
C105	2222-037-50229	22uF	C037	35V	20%
C106	2222-370-21473	47nF	C370	100V	10%
C107	2222-368-55472	4n7F	C368	400V	10%
C108	2222-370-21333	33nF	C370	100V	10%
C109	2222-370-11104	100nF	C370	63V	10%
C110	2222-370-21103	10nF	C370	100V	10%
C111	2222-370-11104	100nF	C370	63V	10%
C112	2222-037-51101	100uF	C037	50V	20%
C113	2222-035-69228	2u2F	C035	100V	20%
C114	2222-368-55222	2n2F	C368	400V	10%
C1B	2222-370-XXXXX	XnF	C370	63V	10%
C20	2222-370-XXXXX	XnF	C370	63V	10%
C21	2222-370-11124	120nF	C370	63V	10%
C22	2222-370-11124	120nF	C370	63V	10%
C25	2222-370-11683	68nF	C370	63V	10%
C28	2222-630-03222	2n2F	C630	100V	10%
C29	2222-630-03222	2n2F	C630	100V	10%
C30	2222-630-03222	2n2F	C630	100V	10%
C32	2222-370-21103	10nF	C370	100V	10%
C33	2222-370-21103	10nF	C370	100V	10%
C34	2222-370-21103	10nF	C370	100V	10%
C35	2222-370-21103	10nF	C370	100V	10%
C36	2222-370-21472	4n7F	C370	100V	10%
C37	2222-630-03222	2n2F	C630	100V	10%
CDLC	2222-370-11474	470nF	C370	63V	10%
CGAR	2222-638-58181	180pF	C638-N750	100V	
CGAT	2222-655-09391	390pF	C655	500V	10%
CLSP1	2222-134-55479	47uF	C134	16V	20%
CMIC	2222-370-21333	33nF	C370	100V	10%
CMIC1	2222-370-21183	18nF	C370	100V	10%
CMIC2	2222-370-11394	390nF	C370	63V	10%
CREG	2222-134-66688	6u8F	C134	25V	20%
CRENV	2222-370-11474	470nF	C370	63V	10%
CRIN1	2222-370-21333	33nF	C370	100V	10%
CRIN2	2222-370-21333	33nF	C370	100V	10%
CRNOI	2222-134-66478	4u7F	C134	25V	20%

CRSEN	2222-370-11154	150nF	C370	63V	10%
CSREF	2222-370-21472	4n7F	C370	100V	10%
CSWT	2222-370-11224	220nF	C370	63V	10%
CTENV	2222-370-11474	470nF	C370	63V	10%
CTNOI	2222-134-66478	4u7F	C134	25V	20%
CTSEN	2222-370-11184	180nF	C370	63V	10%
CVBB	2222-037-65471	470uF	C037	16V	20%
CVBBM	2222-037-51101	100uF	C037	50V	20%
D5	9331-178-10153	12V	BZX79C		
D7	PN-BAW62	BAW62			
D100	PN-BR211-220	BR211-220			
D101	9332-979-90153	BAS11			
D102	9331-442-40153	2.7V	BZX79B		
D103	9331-442-40153	2.7V	BZX79B		
D104	9332-979-90153	BAS11			
D105	9332-979-90153	BAS11			
D106	9331-177-70153	8.2V	BZX79C		
D107	9332-979-90153	BAS11			
D108	9332-979-90153	BAS11			
D109	9332-979-90153	BAS11			
D110	9332-979-90153	BAS11			
D111	9332-979-90153	BAS11			
D112	9332-979-90153	BAS11			
D113	9332-979-90153	BAS11			
D114	PN-BAT85	BAT85			
D115	9330-839-90153	1N4148			
D116	9330-839-90153	1N4148			
D117	9331-178-50153	18V	BZX79C		
D118	9330-839-90153	1N4148			
D120	9331-177-30153	5.6V	BZX79C	0.5W	5%
D123	PN-BAT85	BAT85			
D124	9330-839-90153	1N4148			
D125	9330-839-90153	1N4148			
D126	9330-839-90153	1N4148			
D127	9330-839-90153	1N4148			
D128	9330-839-90153	1N4148			
D129	9330-839-90153	1N4148			
D130	9330-839-90153	1N4148			
D131	9330-839-90153	1N4148			
D132	9330-839-90153	1N4148			
D133	9330-839-90153	1N4148			
D134	9330-839-90153	1N4148			
D4	PN-BAW62	BAW62			
D136	9332-979-90153	BAS11			
H1	KBS-20DB-2P-0	BUZZER	Murata PKM34EW-1224E		
IC1	PN-TEA1064B	TEA1064B			
IC2	PN-TEA1062	TEA1062			
IC5	PN-74HC4053P	74HC4053P			
IC100	PN-PCD3332-2	PCD3332-2			
IC3/IC4	PN-TEA1093	TEA1093/1094			
L1	8222-289-43362	400mH	WWC - Philips Tilburg		
L2	LHL06-151K	150uH	TAIYO		
LOUDSPE	2403-257-238xx	50Ω	AD2071/Z50		

MIC1	PN-MCE100	MCE100			
Q1	PN-XTAL-3.58MHz	3.58MHz			
R2	2322-151-71304	130K	MR25	0.4W	0.5%
R3	2322-151-73922	3K92	MR25	0.4W	0.5%
R4	2322-151-76813	68K1	MR25	0.4W	0.5%
R5	2322-151-73652	3K65	MR25	0.4W	0.5%
R6	2322-151-71104	110K	MR25	0.4W	0.5%
R7	2322-151-72743	27K4	MR25	0.4W	0.5%
R8	2322-151-73921	392	MR25	0.4W	0.5%
R9	2322-151-72009	20	MR25	0.4W	0.5%
R10	2322-151-71001	100	MR25	0.4W	0.5%
R11	2322-151-71301	130	MR25	0.4W	0.5%
R12	2322-151-78251	825	MR25	0.4W	0.5%
R17	2322-151-72003	20K	MR25	0.4W	0.5%
R18	2322-151-73923	39K2	MR25	0.4W	0.5%
R100	2322-151-73928	3.92	MR25	0.4W	0.5%
R101	2322-151-74704	470K	MR25	0.4W	0.5%
R102	2322-156-11006	10M	MRS25	0.6W	1%
R103	2322-194-13222	2K2	PR02	2W	5%
R104	2322-156-15625	5M62	MRS25	0.6W	1%
R105	2322-151-74704	470K	MR25	0.4W	0.5%
R106	2322-151-78252	8K25	MR25	0.4W	0.5%
R107	2322-151-74704	470K	MR25	0.4W	0.5%
R108	2322-151-71004	100K	MR25	0.4W	0.5%
R109	2322-151-71005	1M	MR25	0.4W	0.5%
R110	2322-151-74704	470K	MR25	0.4W	0.5%
R111	2322-151-71004	100K	MR25	0.4W	0.5%
R112	2322-151-71004	100K	MR25	0.4W	0.5%
R113	2322-151-72213	22K1	MR25	0.4W	0.5%
R114	2322-151-75622	5k62	MR25	0.4W	0.5%
R115	2322-151-73322	3K32	MR25	0.4W	0.5%
R116	2322-156-12215	2M21	MRS25	0.6W	1%
R117	2322-151-74704	470K	MR25	0.4W	0.5%
R118	2322-151-74704	470K	MR25	0.4W	0.5%
R119	2322-151-76814	681K	MR25	0.4W	0.5%
R120	2322-151-71005	1M	MR25	0.4W	0.5%
R121	2322-151-71005	1M	MR25	0.4W	0.5%
R122	2322-151-71213	12K1	MR25	0.4W	0.5%
R123	2322-151-71822	1K82	MR25	0.4W	0.5%
R1A	2322-151-76191	619	MR25	0.4W	0.5%
R1B	Rxxx-xxx-xxxxx	0	MR25	0.4W	0.5%
R20	2322-151-71002	1K	MR25	0.4W	0.5%
R21	2322-151-74752	4K75	MR25	0.4W	0.5%
R22	2322-151-74752	4K75	MR25	0.4W	0.5%
R25	2322-151-71002	1K	MR25	0.4W	0.5%
R26	2322-151-71002	1K	MR25	0.4W	0.5%
R27	2322-151-71009	10	MR25	0.4W	0.5%
R28	2322-151-71009	10	MR25	0.4W	0.5%
R29	2322-151-71002	1K	MR25	0.4W	0.5%
RGAR	2322-151-72214	221K	MR25	0.4W	0.5%
RGAT	2322-151-79533	95K3	MR25	0.4W	0.5%
RIDT	2322-156-12215	2M21	MRS25	0.6W	1%
RMIC1	2322-151-72212	2K21	MR25	0.4W	0.5%

RMIC2	2322-151-71001	100	MR25	0.4W	0.5%
RMUTER	2322-151-74701	470	MR25	0.4W	0.5%
RMUTET	2322-151-74753	47k5	MR25	0.4W	0.5%
RPD	2322-151-71003	10K	MR25	0.4W	0.5%
RRSEN	2322-151-74752	4K75	MR25	0.4W	0.5%
RSREF1	2322-151-71001	100	MR25	0.4W	0.5%
RSREF2	2322-151-71001	100	MR25	0.4W	0.5%
RSTAB	2322-151-73652	3K65	MR25	0.4W	0.5%
RSWR	2322-151-73654	365K	MR25	0.4W	0.5%
RTNOI	2322-151-71003	10K	MR25	0.4W	0.5%
RTSEN	2322-151-73922	3K92	MR25	0.4W	0.5%
RVA1	2322-151-XXXXX	X	MR25	0.4W	0.5%
RVA2	2322-151-XXXXX	X	MR25	0.4W	0.5%
RVBB	Rxxx-xxx-xxxxx	0	MR25	0.4W	0.5%
RVBBM	2322-151-71002	1K	MR25	0.4W	0.5%
RVOL	PN-10K-LOG	10k	MRC12	0.2W	20%
S127	PN-Dip-switch_6				
S128	PN-Dip-switch_6				
T1	9337-105-20112	BST70A			
T100	9331-977-30116	BC558			
T101	PN-BSN304A	BSN304A			
T102	PN-BSP304A	BSP304A			
T103	9334-196-50112	BF420	HV		
T104	9332-055-20112	BC546			
T105	9332-055-40112	BC556			
T106	9331-976-40112	BC548			
T107	9331-977-30116	BC558			
D6	9332-979-90153	BAS11			
D137	9332-979-90153	BAS11			
D138	9332-979-90153	BAS11			
D139	9332-979-90153	BAS11			
R30	2322-151-74753	47k5	MR25	0.4W	0.5%
R124	2322-151-71004	100K	MR25	0.4W	0.5%
R125	2322-151-71004	100K	MR25	0.4W	0.5%
R126	2322-151-71004	100K	MR25	0.4W	0.5%
C115	2322-366-75154	150nF	C366	63V	10%

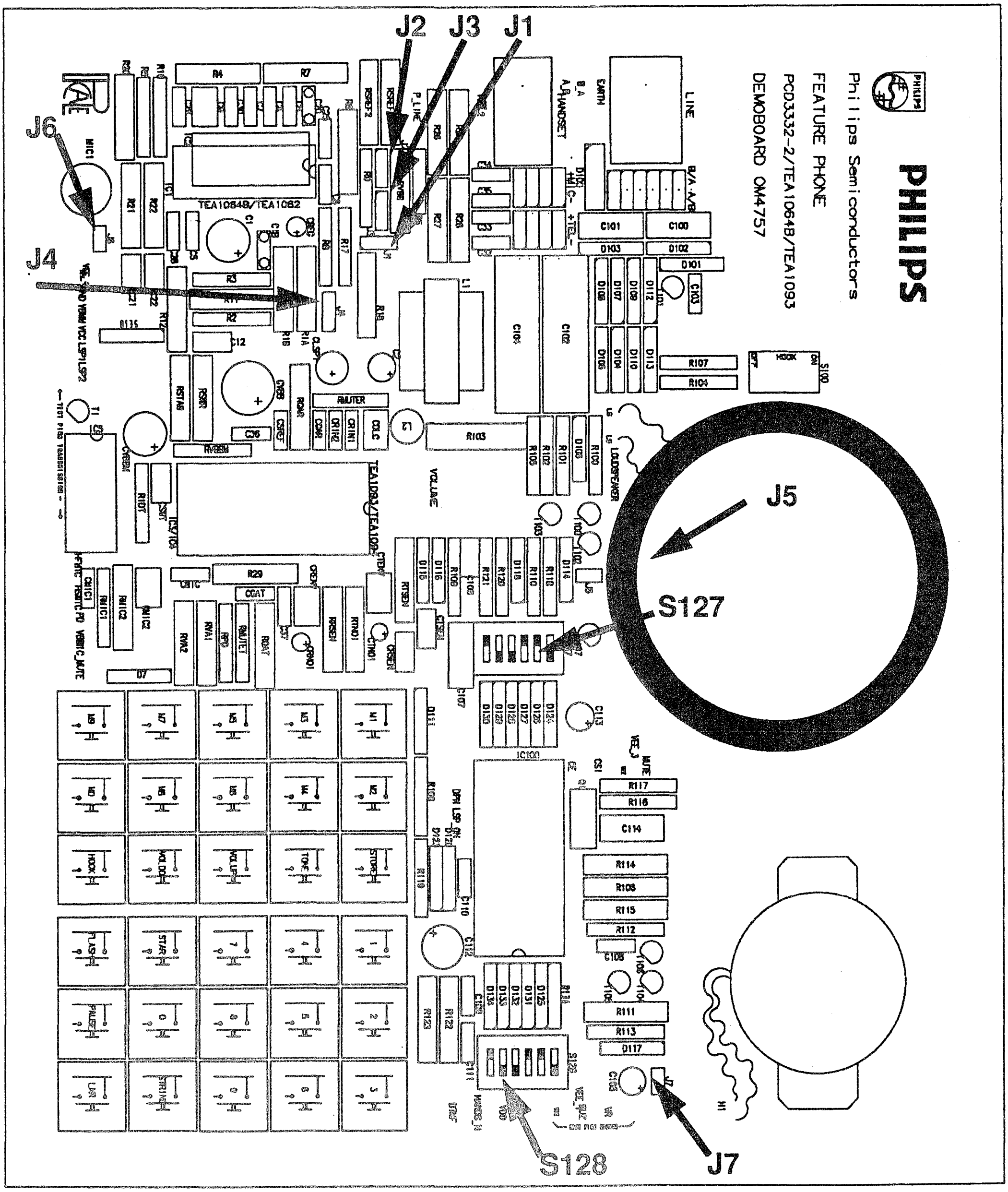


Fig.7 Components view of the OM4757

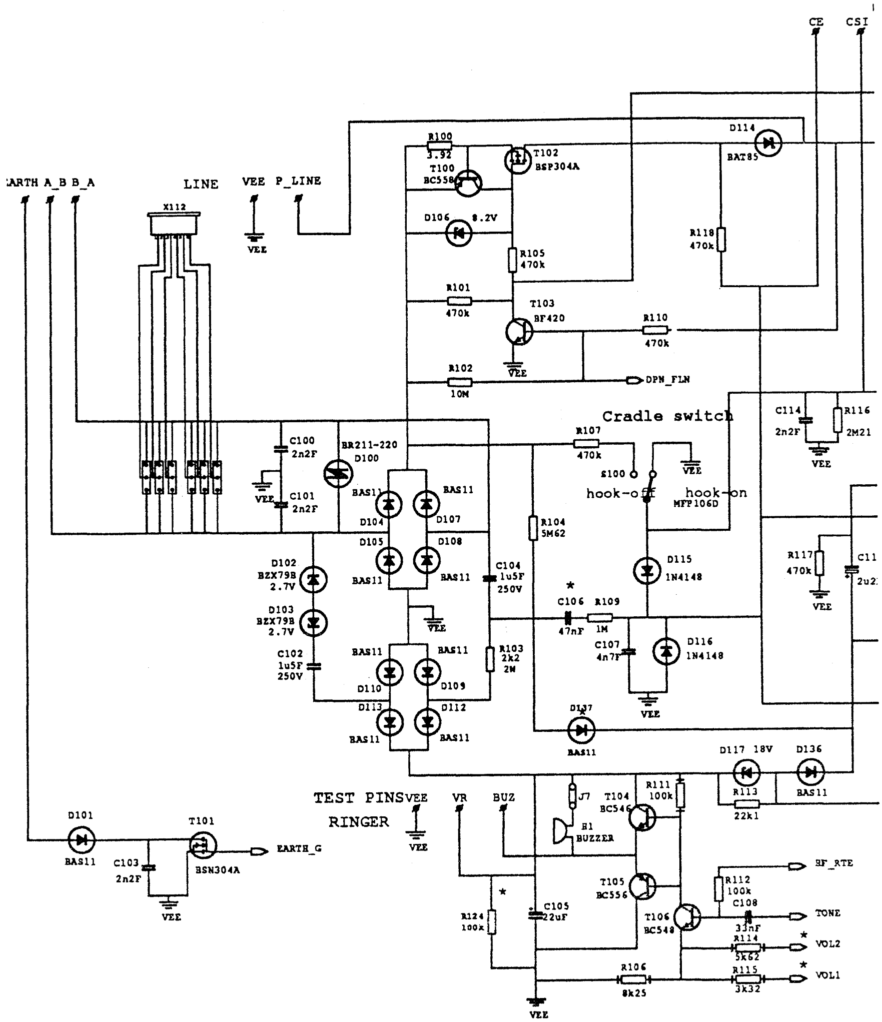


Fig.8 Circuit diagram, part 1: Electronic hook-switch / Controller / Ringer (left)

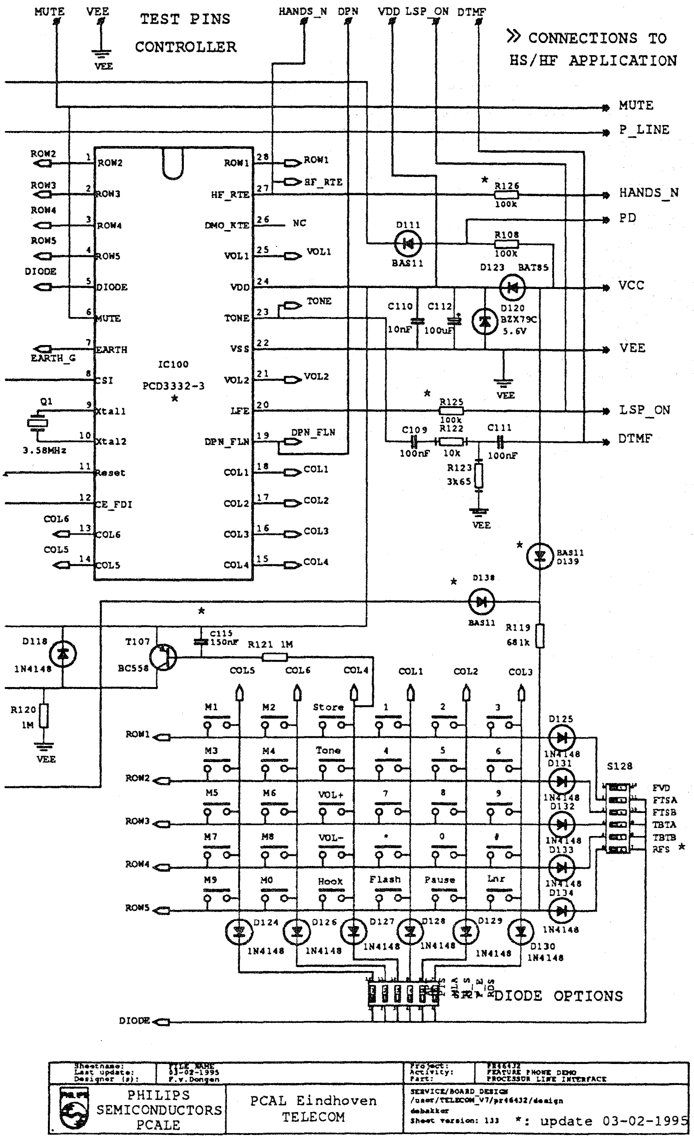
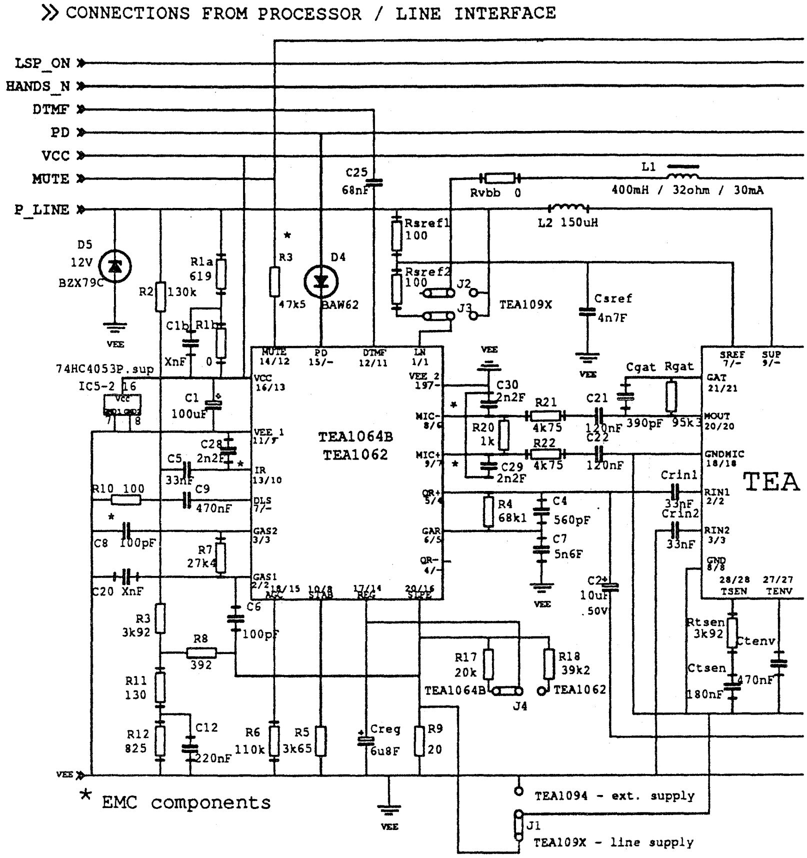


Fig.9 Circuit diagram, part 1: Electronic hook-switch / Controller / Ringer (right)



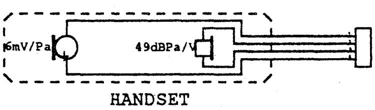
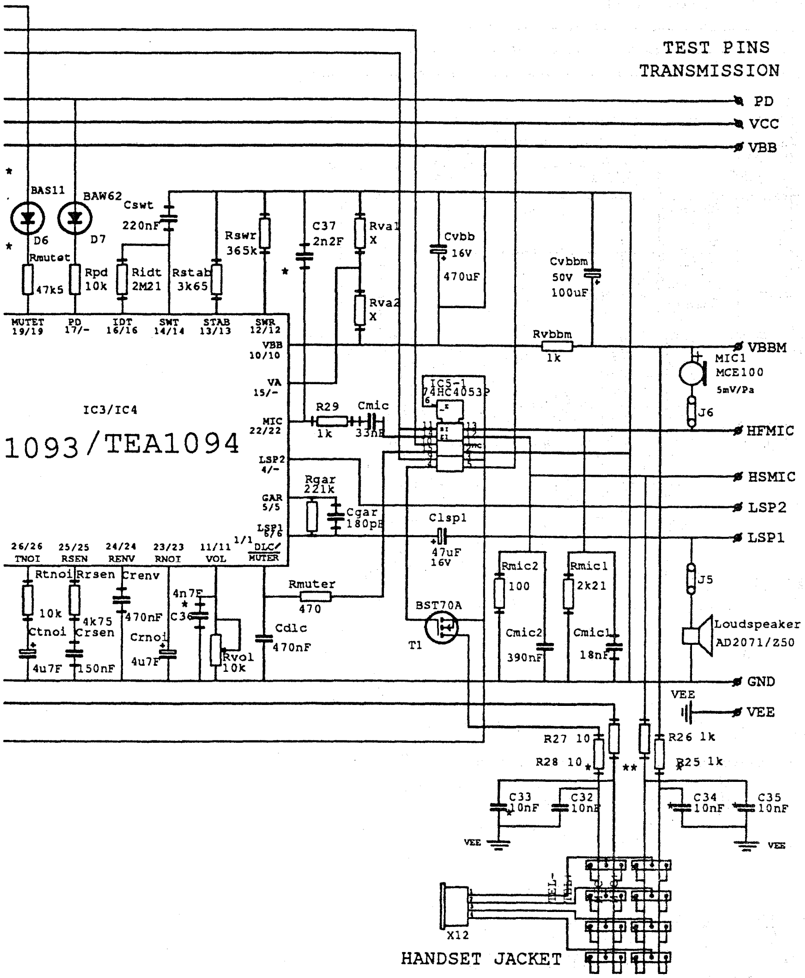


Fig.10 Circuit diagram, part 2: Transmission / TEA106X / TEA109X (left)



Sheetname: Last update: Designer (s):	FILE NAME 03-02-1995 F.v.DONGEN	Project: Activity: Part:	PR46432 FEATURE PHONE DEMO HF application
PHILIPS SEMICONDUCTORS PCALE	PCAL Eindhoven TELECOM	SERVICE/BOARD_DESIGN /user/TELECOM_V7/pr46432/design debakker Sheet version: 37	*: update 03-02-1995

Fig.11 Circuit diagram, part 2: Transmission / TEA106X / TEA109X (right)

APPLICATION NOTE Nr	ETT/UM95004.0
TITLE	OM4775 User Manual Basic Phone Demonstration Board OM4775: UBA1702/A, TEA1062/1064B and PCD3755A
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DATE	November 1995

CONTENTS

1. Introduction.

This user manual goes with the demonstration board OM4775. In order to demonstrate the application possibilities of the UBA1702/A this demoboard has been developed. It features:

- speech/transmission functions
- various dialling modes (pulse, DTMF and mixed-mode)
- ringer functions.

In this manual only the most important parameter settings of both the UBA1702/A and TEA1062/1064B are explained (chapter 2.1 and 2.2). For more details the respective designers guide / application report (see references) can be consulted. For the PCD3755A an objective specification (see also references) is available, the specific EEPROM organization and programming procedure are described in chapter 2.3. The software for this microcontroller is derived from the PCD3330-1 with a few modifications. For details like number storage please refer to the objective specification of the PCD3330-1.

Specific operating details including jumper settings are given in chapter 3, some notes on EMC in chapter 4 and the references can be found in chapter 5.

Throughout this manual a reference is made to the circuit diagram of the board (see appendix). A component lay-out of the board is also included in the appendix.

2. Function Partitioning and Application Description.

This chapter describes which IC is dedicated to a certain function and in which way set parameters can be adjusted to specific (country) requirements.

2.1 UBA1702/A: line interrupter driver and ringer circuit.

This IC contains the interrupter switch driver -which is required for pulsedialling and flash operation-, line current detector and limiter, pin controlled line voltage limiter, ringer interface / outputstage and stabilized microcontroller supply.

Two versions are available:

- UBA1702, intended to drive a PMOST interrupter (e.g. BSP254A) and
- UBA1702A, for a bipolar pnp interrupter (e.g. MPSA92).

Both versions and the belonging interrupter device can be used in the demoboard (board locations IC2 and T1), however care must be taken for the correct connections of the interrupter device.

2.1.1 DC settings

The **voltage limit** between SPO and VEE to protect the transmissioncircuit can be set by R29 (= R_{ZPA1}) and R30 (= R_{ZPA2}). When both are omitted (default setting) the voltage is limited to about 12 V. This voltage setting is recommended for the TEA106X family. When R_{ZPA1} is replaced by a short (R_{ZPA2} still open) the voltage is decreased to 9 V and in case R_{ZPA2} is shorted (R_{ZPA1} open) the voltage is increased to 18 V. Intermediate values can be achieved by applying appropriate resistors. The way these intermediate values can be achieved (by applying appropriate resistors) is the same for the various other adjustments of the UBA1702/A.

The **current limit** can be influenced by R7 (= R_{CLA}). An open results in the lowest limit (about 45 mA), the highest limit can be achieved by a short (default setting), which results in a limit of approx. 120 mA.

Default setting for **current detection** is 3 mA. This threshold can be increased by connecting a resistor on position R9 (= R_{CDA}). A linecurrent greater than the threshold results in a logic 'high' on pin CDO.

2.1.2 Dial mode operation

The **set resistance** during the make period of pulse dialling can be decreased by making input pin MSI 'high' (DMO-mode), in this way the voltage between SPO and VEE is limited to a very low value. This value is default about 2.7 V (R27 = R_{MSA} open) and can be decreased by replacing R_{MSA} by a short.

2.1.3 Ringer mode operation

The **ringer input impedance** for large ringer signals is mainly determined by R13 and C13. Almost at the same time an AC ringersignal is applied a square wave signal with twice the ringer frequency is available on pin RFO.

The ringer melody generated by the microcontroller is made audible by the piezo if the voltage on pin VRR crosses a certain upper **ringer threshold** of about 11 V (default setting). The ringer output stage is switched off again if a lower threshold has been reached (hysteresis). These thresholds can be influenced by R_{RTA}: decreasing the resistance of R19 (= R_{RTA1}; leaving R_{RTA2} open), results in decreasing the ringer threshold level, decreasing the resistance of R20 (= R_{RTA2}; leaving R_{RTA1} open) results in an increased ringer threshold level.

The ringermelody generated by the microcontroller (available on TONE output) is AC coupled by C7 to the RMI pin of the UBA1702/A.

2.2 TEA1062/1064B Transmission Circuit.

This IC regulates the DC line voltage, takes care for proper line termination, contains the amplifiers for the transmit and receive path and the hybrid for two to four wire conversion.

Two sockets are mounted on the PCB: a sixteen pins socket for the TEA1062 and a twenty pins socket for the TEA1064B. Major differences between these two transmission circuits are the additional power down function, inverting receiver output stage (pin QR-) and the dynamic limiter combined with a microphone mute (pin DLS/MMUTE) incorporated in the TEA1064B.

2.2.1 DC settings

At a line current of 15 mA, the **DC-voltage** between pins LN and VEE is 4 V for the TEA1062 and 3.5 V for the TEA1064B. This voltage can be increased by applying a resistor R3. This results in a voltage of 4.5 V for the TEA 1062 with R3 = 39 kΩ and in 4.4 V for the TEA1064B with R3 = 20 kΩ.

In case of the TEA1064B two different **supply structures** are possible:

- Reference is VEE. VBB supply for the UBA1702/A (and consequently the microcontroller) is derived from VCC. This structure is also applicable for the TEA1062.
- Reference is SLPE (TEA1064B only). VBB supply is derived directly from the line by means of R1A (392 Ω).

Jumper settings (JVCC1 and Jref) determine the supply structure.

In case a TEA1064B is used the possibility exists to connect an additional back-up capacitor C1 by means of jumper J1. This capacitor is required to 'survive' long line interrupts: e.g. flash durations of 600 ms. In case of TEA1062 this capacitor has little use because of the lacking power down function: the capacitor is discharged by the transmission circuit instead of being for the benefit of the microcontroller supply.

2.2.2 Set impedance.

The **set impedance** is mainly determined by the resistor R between the pins LN and VEE in combination with the internal resistance of the TEA106X:

$$Z_{set} = \frac{R * R_i}{R + R_i} \tag{1}$$

in which R_i = 16.2 kΩ.

On the PCB resistor R is split into R1 and R2//C2 to have the possibility to create a complex impedance. R2 is chosen 0 Ω and R1 is 620 Ω, providing a Zset of approximately 600 Ω in the frequency range of 300 Hz up to 3400 Hz.

2.2.3 Side tone

In order to obtain optimum **side tone suppression**, the following conditions should be met:

$$R_{12} * R_{14} = R * (R_{10} + R_{11} // Z_{bal}) \tag{2}$$

$$\frac{Z_{bal}}{Z_{bal} + R_{11}} = \frac{Z_{line}}{Z_{line} + R} \tag{3}$$

in which: Z_{bal} = the network consisting of R15, R21 and C17

Z_{line} = the load at the line

R = R1 + R2//C2

Equation (2) can be met if R11//Z_{bal} << R10 and if R12, R and R10 are real (not complex).

Equation (3) can be met if:

$$Z_{bal} = \frac{R_{11}}{R} * Z_{line} = k * Z_{line} \tag{4}$$

In practice, Z_{line} will vary strongly with line length and cable type. Consequently, an average value for Z_{bal} has to be chosen. On the board Z_{bal} is optimised for a 0.5 mm copper cable of 5 km with an attenuation of 1.2 dB/km, a DC resistance of 176Ω/km and a capacitance of 38 nF/km. The anti side tone circuit attenuates the received signal from the line with approximately 32 dB.

2.2.4 Microphone amplifier

Resistor R26 (8.2 kΩ) between MIC- and MIC+ can be used to lower the 64 kΩ input impedance of the TEA1062/1064B. By means of C18, C24 and R26 a high pass filter is created.

The **microphone gain** from the MIC+/- inputs to the line depends on the value of resistor R22 between pins GAS1 and GAS2 in the following way:

$$A_{MIC+,MIC-\ to\ LN} = 1.356 * \frac{R_{22} + 3470\Omega}{R_{12} * R_{16}} * \frac{Z_{set} * Z_{line}}{Z_{set} + Z_{line}} \tag{5}$$

On the board, resistor R22 is set to 27.4 kΩ, giving a microphone gain of approximately 44.5 dB (Z_{line} = 600 Ω) from MIC+, MIC- to the line. If another gain setting is required, the value of R22 may be varied between 27.4 kΩ and 68 kΩ (gains setting 44 dB up to 52 dB).

Capacitor C14 between GAS1 and SLPE should be at least 100 pF for stability reasons.

2.2.5 DTMF amplifier

The DTMF amplifier of the TEA106x is enabled if a high level (> 1.5 V) at microcontroller interface pin MUTE is applied. In mute condition, the microphone amplifier and the earpiece amplifier are muted and only a confidence tone can be noticed at the earpiece output.

The **DTMF gain** (gain from the DTMF input to the line) is 26.5 dB lower than the microphone gain. Therefore, with R22 = 27.4 kΩ the DTMF gain equals: 44.5 dB - 26.5 dB = 18 dB.

The gain from the DTMF input to QR+ (confidence tone) is 50 dB lower than the earpiece gain from IR to QR+ (see paragraph 2.2.6). With R25 = 82.5 kΩ the gain from DTMF input to QR+ output equals: 29.5 dB - 50 dB = -20.5 dB.

2.2.6 Earpiece amplifier

The receiving signal coming from the line is attenuated approximately 32 dB in the sidetone bridge before it enters pin IR. The **earpiece gain** (gain from IR to QR+ (QR+ = QR for the TEA1062)) is dependent on resistor R25 between QR+ and GAR in the following way:

$$A_{IR\ to\ QR} = 1.314 * \frac{R_{25}}{R_{16}} * \frac{Z_{ear}}{Z_{ear} + 4\Omega} \qquad (6)$$

in which: Z_{ear} = earpiece impedance

R16 = resistor at pin STAB which is 3.6 kΩ

On the PCB, resistor R25 = 82.5 kΩ is mounted, giving a receiver gain of 29.5 dB from IR to QR+. The overall gain from the line to QR+ will therefore equal -2.5 dB. If another gain setting is required, the value of R25 may be varied between 28 kΩ and 100 kΩ corresponding to a gain setting from line to QR+ from -12 dB up to -1 dB. Additionally, the TEA1064B has an inverting output (QR-) for BTL drive (gain increases 6 dB).

Capacitor C21 should be at least 100pF for stability and is also used to form a low pass filter. The mounted value is 560pF providing a cut-off frequency in conjunction with R25 of about 3.4 kHz. Capacitor C27 should be 10 times larger than C21 and therefore 5.6 nF is mounted.

2.2.7 Automatic gain control

The built-in **automatic gain control** (AGC) circuit of the TEA106X attenuates the gain of the microphone and earpiece amplifiers at high line currents. With AGC active the loop gain from MIC+, MIC- to QR+ is kept more or less constant over the range from 0 up to 5km. For longer lines, this loop gain will increase.

With R6 = 120 kΩ the gain of the microphone amplifier and the earpiece amplifier is reduced at line currents larger than 25 mA. At 63 mA, the maximum attenuation occurs (6dB) and above this value the gains are kept constant.

2.3 PCD3755A microcontroller.

The microcontroller has several functions: it generates the timing and control signals for pulse dialling and flash operation (available on output pins DP/FL. MUTE and DMO), the DTMF signals for DTMF dialling (output pin TONE), ringer frequency check (input pin RFO) and the ringer melody combined with volume setting (output pins TONE and RVOL1-3). The PCD3755A features 128 bytes of EEPROM. By using EEPROM no special backup measures are necessary such as battery, current from the line (trickle current) or very big capacitors.

2.3.1 Reset

To ensure proper startup, the microcontroller must be initialized to a defined starting condition by a reset. Default the reset is generated internally, but by applying R35 and C31 an external (passive) reset can be generated.
To improve the reset performance in case a TEA1064B is applied it is recommended to decrease the value of C5 (C_{VDD}). This also results in a faster start up.

2.3.2 EEPROM organization

The dialling, memory, and ringer options and the telephone numbers are all stored in EEPROM.
Table 1 describes the meaning of each EEPROM byte at a repertory length of 16 and 20 digits.
Table 2 describes the meaning of each bit of all the bytes that do not contain telephone numbers.

Table 1. EEPROM organisation table.

	Repertory length is 16 digits		**Repertory length is 20 digits**	
Function	**Length**	**byte places**	**Length**	**byte place**
Redial	13 bytes	0 till 12	13 bytes	0 till 12
M0 or MEM + 0	8 bytes	16 till 23	10 bytes	16 till 25
M1 or MEM + 1	8 bytes	24 till 31	10 bytes	26 till 35
M2 or MEM + 2	8 bytes	32 till 39	10 bytes	36 till 45
M3 or MEM + 3	8 bytes	40 till 47	10 bytes	46 till 55
M4 or MEM + 4	8 bytes	48 till 55	10 bytes	56 till 65
M5 or MEM + 5	8 bytes	56 till 63	10 bytes	66 till 75
M6 or MEM + 6	8 bytes	64 till 71	10 bytes	76 till 85
M7 or MEM + 7	8 bytes	72 till 79	10 bytes	86 till 95
M8 or MEM + 8	8 bytes	80 till 87	10 bytes	96 till 105
M9 or MEM + 9	8 bytes	88 till 95	10 bytes	106 till 115
E-1	8 bytes	96 till 103	not available	--
E-2	8 bytes	104 till 111	not available	--
E-3	8 bytes	112 till 119	not available	--
Options	4 bytes	120 till 123	4 bytes	120 till 123
Program Blocking	1 byte	127	1 byte	127

Table 2. Option bit status and location.

Function	EEPROM byte	bit7	bit6	bit5	bit4	bit3	bit2	bit1	bit0
Not sending *	120	X	X	X	X	X	X	X	0
Sending *	120	X	X	X	X	X	X	X	1 *)
Not sending #	120	X	X	X	X	X	X	0	X
Sending #	120	X	X	X	X	X	X	1	X *)
Mark/space ratio 3:2	120	X	X	X	X	X	0	X	X *)
Mark/space ratio 2:1	120	X	X	X	X	X	1	X	X
Tone/pause 60/90 ms	120	X	X	0	0	0	X	X	X
T/p 70/70 ms	120	X	X	0	0	1	X	X	X
T/p 80/80 ms	120	X	X	0	1	0	X	X	X
T/p 100/100 ms	120	X	X	0	1	1	X	X	X
T/p 100/140 ms	120	X	X	1	0	0	X	X	X
T/p 140/140 ms	120	X	X	1	0	1	X	X	X
Flash time 100 ms	120	0	0	X	X	X	X	X	X *)
Flash time 115 ms	120	0	1	X	X	X	X	X	X
Flash time 270 ms	120	1	0	X	X	X	X	X	X
Flash time 600 ms	120	1	1	X	X	X	X	X	X
Mute is M1	121	X	X	X	X	X	X	0	0 **)
Mute is notM1	121	X	X	X	X	X	X	0	1
Mute is M2	121	X	X	X	X	X	X	1	0 *)
Mute is notM2	121	X	X	X	X	X	X	1	1
<u>Access Pause time for pulse dialling (Inter-digit pause not included).</u>									
A.P. time 1.5 s	121	X	X	X	0	0	X	X	X
A.P. time 2.5 s	121	X	X	X	0	1	X	X	X
A.P. time 3.0 s	121	X	X	X	1	0	X	X	X
A.P. time 6.0 s	121	X	X	X	1	1	X	X	X
<u>Access Pause time for DTMF dialling (Tone-off time not included).</u>									
A.P. time 1.0 s	121	X	X	X	0	0	X	X	X
A.P. time 1.5 s	121	X	X	X	0	1	X	X	X

Table 2. Option bit status and location (con'd).

Function	EEPROM byte	bit7	bit6	bit5	bit4	bit3	bit2	bit1	bit0
A.P. time 3.5 s	121	X	X	X	1	0	X	X	X
A.P. time 6.0 s	121	X	X	X	1	1	X	X	X
Gen. program proc.	121	X	X	0	X	X	X	X	X
Germ. program proc.	121	X	X	1	X	X	X	X	X
Repertory 16 digits	121	X	0	X	X	X	X	X	X
Repertory 20 digits	121	X	1	X	X	X	X	X	X
M1/M2 mute	121	0	X	X	X	X	X	X	X
Microphone mute	121	1	X	X	X	X	X	X	X
Ringer melody A	122	X	X	X	X	0	0	X	X
Ringer melody B	122	X	X	X	X	0	1	X	X
Ringer melody C	122	X	X	X	X	1	0	X	X
Ringer melody D	122	X	X	X	X	1	1	X	X *)
Ringer volume 1	122	X	X	0	0	X	X	0	X
Ringer volume 2	122	X	X	0	1	X	X	0	X
Ringer volume 3	122	X	X	1	0	X	X	0	X
Ringer volume 4	122	X	X	1	1	X	X	0	X
Ringer volume 5	122	X	X	0	0	X	X	1	X
Ringer volume 6	122	X	X	0	1	X	X	1	X
Ringer volume 7	122	X	X	1	0	X	X	1	X
Ringer volume 8	122	X	X	1	1	X	X	1	X
Ringer repetition 1	122	0	0	X	X	X	X	X	X
Ringer repetition 2	122	0	1	X	X	X	X	X	X
Ringer repetition 3	122	1	0	X	X	X	X	X	X
Ringer repetition 4	122	1	1	X	X	X	X	X	X
Lo ringfreq det 16Hz	123	X	X	X	X	X	X	0	0 *)
Lo ringfreq det 20Hz	123	X	X	X	X	X	X	0	1
Lo ringfreq det 32Hz	123	X	X	X	X	X	X	1	0
Lo ringfreq det 40Hz	123	X	X	X	X	X	X	1	1

Table 2. Option bit status and location (con'd).

Function	EEPROM byte	bit7	bit6	bit5	bit4	bit3	bit2	bit1	bit0
Hi ringfreq det 35Hz	123	X	X	X	X	0	0	X	X
Hi ringfreq det 60Hz	123	X	X	X	X	0	1	X	X
Hi ringfreq det 70Hz	123	X	X	X	X	1	0	X	X
Hi ringfreq det 120Hz	123	X	X	X	X	1	1	X	X *)
A to D keys	123	X	X	X	0	X	X	X	X
Function keys	123	X	X	X	1	X	X	X	X
Flash function	123	X	X	0	X	X	X	X	X *)
Earth function	123	X	X	1	X	X	X	X	X
No keytone	123	X	0	X	X	X	X	X	X
Keytone active	123	X	1	X	X	X	X	X	X *)
No on-hook dialling	123	0	X	X	X	X	X	X	X *)
on-hook dial control	123	1	X	X	X	X	X	X	X

*) recommended / default setting.
**) due to a software bug this setting in combination with DMO operation is not recommended.

2.3.3 EEPROM programming procedure

This procedure is only active if EEPROM Program Blocking byte (number 127 of the internal EEPROM) is set to "FF" hex. If this byte is "00" hex it is not possible to do the programming procedures described in this chapter. Byte 127 of the EEPROM can only be set by a factory EEPROM programming procedure.

In the field all telephone options can be changed easily by a special program procedure:

- depress the STO-key (this selects the program mode),
- depress the LNR-key (switches the program module to storing EEPROM options),
- depress the first key of a three digit access code (the 1),
- depress the second key of a three digit access code (the 6),
- depress the third key of a three digit access code (the 0),
- depress the LNR-key again (end the access code),
- press the byte number (last digit of the EEPROM byte number given in table 2),
- press the number of the bit to change (see table 2),
- press 0 or 1 (this changes the EEPROM bit contents),
- depress the LNR-key, which stores the correction into EEPROM, select a new byte or go to end,
- end the routine by pressing the STO-key again.

If during this procedure a mistake is made correction is possible after proper access code by pressing the LNR-key and during access code only by STO-key.

In all cases the routine can be ended by pressing the STO-key.

Example:

Change the mark-to-space ratio from 3:2 to 2:1.

Then bit 2 of byte 120 has to be changed from 0 to 1.

The necessary action is as follows:

- depress the STO-key,
- depress the LNR-key,
- depress the 1-key (first digit access code),
- depress the 6-key (second digit access code),
- depress the 0-key (third digit access code),
- depress the LNR-key again (end the access code),
- press the 0-key (last digit of EEPROM byte 120 is the 0),
- press the 2-key (bit 2 has to be changed),
- press the 1 (this changes the mark-to-space ratio to 2:1),
- press the LNR or STO-key.

STO-key will end the programming procedure, while after LNR-key a new byte which bit has to be changed can be selected.

2.3.4 Dialling modes

Three different dialling modes can be distinguished:

I.) Pulse dialling
II.) DTMF dialling
III.) Mixed mode dialling

I.) Pulse dialling (input pin PD/DTMF = LOW)

The keyboard entry initiates a recall of a previously stored number or is a simultaneous keying-in and pulsing-out activity, with storing for possible later recall. If in the recalled number or at keying-in the keys A, B, C or D (option A to D keys selected) are used these digits are not transmitted. If at keying-in the keys * or # are used this results in a switch over to DTMF dialling. Normally, keying in is faster than pulsing-out (fed from the redial register).

Pulse sequences start with an inter-digit pause of 840 ms duration, followed by a sequence of pulses corresponding to the present digit in store. Each pulse starts with a mark (line break) followed by space (line make). The pulse period is 100 ms with a mark-to-space ratio of 3:2 or 2:1 (mark to space ratio selection). After transmission of a digit, the next digit is processed, again starting with an inter-digit pause. The pulses are available at the DP/FL output and this output is used to drive the UBA1702/A. The transmission IC is put in the dialling mode by means of output MUTE. Output MUTE has several programmable options, MUTE can be configured as M1, notM1, M2 and notM2. M2 has to be used for the demo board. After completion of the number string the circuit changes from dialling mode to conversation mode.

II.) Dual tone multi frequency (DTMF) dialling (input PD/DTMF = HIGH)

The PCD3755A converts keyboard inputs into data for the on-chip DTMF generator. Tones are transmitted via output TONE with six programmable minimum tone burst/pause durations of 60/90, 70/70, 80/80, 100/100, 100/140 or 140/140 ms. The maximum tone burst duration is equal to the key depression time. With redial and repertory dialling tones are automatically fed at the programmed rate.

The mute is again depending on the selected MUTE-mode. Again the MUTE output has several programmable options: M1, notM1, M2 and notM2. M2 has to be used for the demo board.

III.) DTMF dialling in pulse dialling mode (mixed mode dialling)

If the controller is set to the pulse dial mode (input PD/DTMF is LOW), activation of push-buttons TONE, * or # changes the dialling mode to DTMF. This entry is stored in the redial register and it generates automatically an access pause, after which the digits following are transmitted in the DTMF mode. The digits entered after keys TONE, * or # are not transmitted in the redial mode. The TONE key is never transmitted, whether * or # are transmitted depends on the selected option. A second touch of the TONE key is ignored. The * and # keys pressed after a switch over to DTMF dialling are all transmitted.

If the controller is initially set to the DTMF mode (pin PD/DTMF is HIGH), activation of push-button TONE is ignored and the * or # are stored in the redial register and transmitted in DTMF mode.

2.3.5 Ringer input frequency measurement

The ringermelody generator becomes active for all incoming ringer frequencies higher then the ringer detection LO frequency and lower then the ringer detection HI frequency supplied to the CE/RF input of the PCD3755A. The ringer detection LO and ringer detection HI frequencies can be selected in such way that it is possible to use the PCD3755A for both single or double phase rectifier applications. The UBA1702/A generates an RFO logic signal with twice the ringer signal frequency.It is possible to select one out of the four ringer detection LO and one out of the four ringer detection HI frequencies options which are given below:

- Ringer detection LO: 16 Hz
- Ringer detection LO: 20 Hz
- Ringer detection LO: 32 Hz
- Ringer detection LO: 40 Hz

- Ringer detection HI: 35 Hz
- Ringer detection HI: 60 Hz
- Ringer detection HI: 70 Hz
- Ringer detection HI: 120 Hz

The lowest ringer signal detection frequency of this board is therefore 16/2 = 8 Hz, the highest 120/2 = 60 Hz.

2.3.6 Ringer melody selection

The ringer melody generator can select one out of four different ringer melody options (stored in EEPROM) which are given below:

- Ringer melody A	738 Hz	826 Hz	925 Hz
- Ringer melody B	800 Hz	1067 Hz	1333 Hz
- Ringer melody C	1455 Hz	1621 Hz	1810 Hz
- Ringer melody D	1995 Hz	2223 Hz	2510 Hz

3. Operating details.

In this chapter an overview of the function of the keys on the keyboard is given, as well as the procedure for number storage, ringer volume and ringer melody repetition rate control and the various jumpersettings are described.

3.1 Key function

The function of the keys is:

0 to 9, * and #	Standard keyboard; in pulse dialling mode the valid keys are the 10 numeric keys (0 to 9), the 2 non-numeric keys (* and #) have no effect on the dialling. In DTMF dialling mode the 10 numeric keys and the 2 non-numeric keys are valid.
A to D	If selected (EEPROM bit), these keys are only valid in DTMF dialling mode.
TONE	If selected, pulse to DTMF switching key (mixed mode dialling).
DIS	If selected (EEPROM bit), disconnect key will activate output DP/FL for 800 ms. In this case the telephone set turns to the ON-HOOK state for this calibrated time.
PR	If selected (EEPROM bit), program ringer key. With this key the ringer output volume and ringer repetition rate can be changed.
M0 to M9	One key abbreviated dialling, the 10 repertory numbers are directly accessible via push-buttons M0 to M9.
LNR	Last number redial.
AP	Access pause key, results in inserting an access pause in the telephone number.
FLASH	FLASH key.
ST	STORE key.
MEM	Two-key abbreviated dialling (MEM + digit), the repertory numbers M0 to M9 are also accessible via this two-key dialling procedure.
M	Mute key, each time this key is pressed and dialling is not active, the mute output goes to HIGH or LOW depending on the previous state.
E-1 to E-3	One key abbreviated dialling, three extra repertory numbers which are only directly accessible by push-buttons E-1 to E-3; these numbers can only be used when the repertory length is 16 digits (programmable in EEPROM).

3.2 Number storage

See chapter 7.2 of [Ref. 8].

3.3 Ringer volume change during conversation and ringer mode.

The ringer volume is controlled by the pins RVOL1, RVOL2 and RVOL3 and its value is stored in EEPROM. The output volume can be changed:

- via the EEPROM programming procedure (see chapter 2.3.3)
- during conversation mode, when the function keys option is chosen, with a special key sequence
- during active ringer by a simple key press.

In conversation mode the procedure is as follows:

- put the set in conversation (supply necessary)
- depress PR (ringer program key)
- press one of the eight numeric keys (1 to 8); key 1 corresponds with minimum volume and 8 with maximum volume.

The last selected value is directly stored into EEPROM

During active ringing the PR key has not to be pressed, the procedure is as follows:

- activate the ringer (only then this volume correction is possible)
- press one of the eight numeric keys (1 to 8).

This value is directly stored into EEPROM.

3.4 Ringer repetition rate change during conversation and ringer mode

The generated melody consists of three frequencies. These frequencies are generated successively in a selected repeat frequency (warble).
There are four repeat frequencies, this repetition rate can be changed by:

- the EEPROM programming procedure (see chapter 2.3.3)
- during conversation mode, when the function keys option is chosen with a special key sequence
- in ringer mode by a simple key press.

In conversation mode the procedure is as follows:

- put the set in conversation (supply necessary)
- depress PR (ringer program key)
- press one of the four acceptable repeat frequency keys (9, *, 0 or #)

in which:

Key	Frequency	Tone time
9	7 Hz	47.6 ms
*	11 Hz	30.3 ms
0	15 Hz	22.2 ms
#	20 Hz	16.6 ms

The last selected value is directly stored into EEPROM.

In ringer mode the PR key has not to be pressed and the procedure is as follows:

- activate the ringer (only then this repetition rate correction is possible)
- press one of the four acceptable repeat frequency keys (9, *, 0 or #), the repetition rates are equal to the previous case.

The last selected value is directly stored into EEPROM.

3.5 Jumper settings

The various jumpers on the demoboard have the following function:

J2, J4-J8:	select the proper connection for the lineconnector.
J9-J24:	select the proper connection for the handset.
J3 (Jref):	selects the groundreference of VDD. Left hand position: SLPE (TEA1064B only), right hand position: VEE.
JVCC1:	selects supply point VBB. Left hand position: LN (TEA1064B only), right hand position: VCC.
J1 (Jvbb):	connects additional supply capacitor for long line interrupts (TEA1064B only).
JMSI1:	selects DMO-mode.
J25 (Jdls):	selects microphone mute (disables microphone amplifier, TEA1064B only).
J26:	selects between DTMF dialling mode (left hand position) and pulse dialling mode (right hand position).

Note: it is recommended to use the DMO mode (JMSI1 jumperconnection applied) only with additional supply capacitor (C1), therefore jumper J3 has to be in left hand position and jumper J1 connected (TEA1064B only).

4. Electromagnetic Compatibility.

Measures have been taken to reduce the sensitivity of the telephone set for common mode amplitude modulated RF signals generated in the A/B lines and the handset connections. The board has been provided with a ground reference layer connected to VEE of the UBA1702/A and TEA1062/1064B. This reference layer has maximum effect if the logic reference is VEE as well (jumper settings J3: VEE and JVCC1: VCC). If SLPE is chosen as logic reference the EMC performance of the board is not optimal (TEA1064B only).

Basic measures consisting of additional components are:

- C9 (2.2 nF) across the VCC supply of the transmission circuit
- C11 (1 nF) across the receive input
- C22 and C23 (both 2.2 nF) across the microphone input
- C25 and C28 (both 2.2 nF) across the microphone handset connections
- C29 (2.2 nF) across the earpiece connections
- R28 (22 Ω) in series with the earpiece

and additionally:

- C15 and C16 (both 2.2 nF) across the lineconnection. However to ensure proper operation of the ring-frequency detection at high ringer voltages a resistor R4 (= R_{RPI}; about 100 kΩ) has to be added.

These EMC components are in the circuit diagram indicated with a '⋆'.

More detailed information in case a further improved EMC performance is required can be found in [Ref. 7].

5. References.

[Ref. 1] Philips Semiconductors Data Handbook
Semiconductors for telecom systems - IC03
Philips Semiconductors, 1995

[Ref. 2] UBA1702/A Line Interrupter Driver and Ringer IC
Preliminary specification
October 1995

[Ref. 3] Application of the UBA1702/A Line Interrupter Driver and Ringer Circuit
Application Note by H. Derks
Report number: AN95023

[Ref. 4] TEA1060 family versatile speech/transmission ICs for electronic telephone sets
Designers' guide by P.J.M. Sijbers
July 1987, 12NC 939834110011

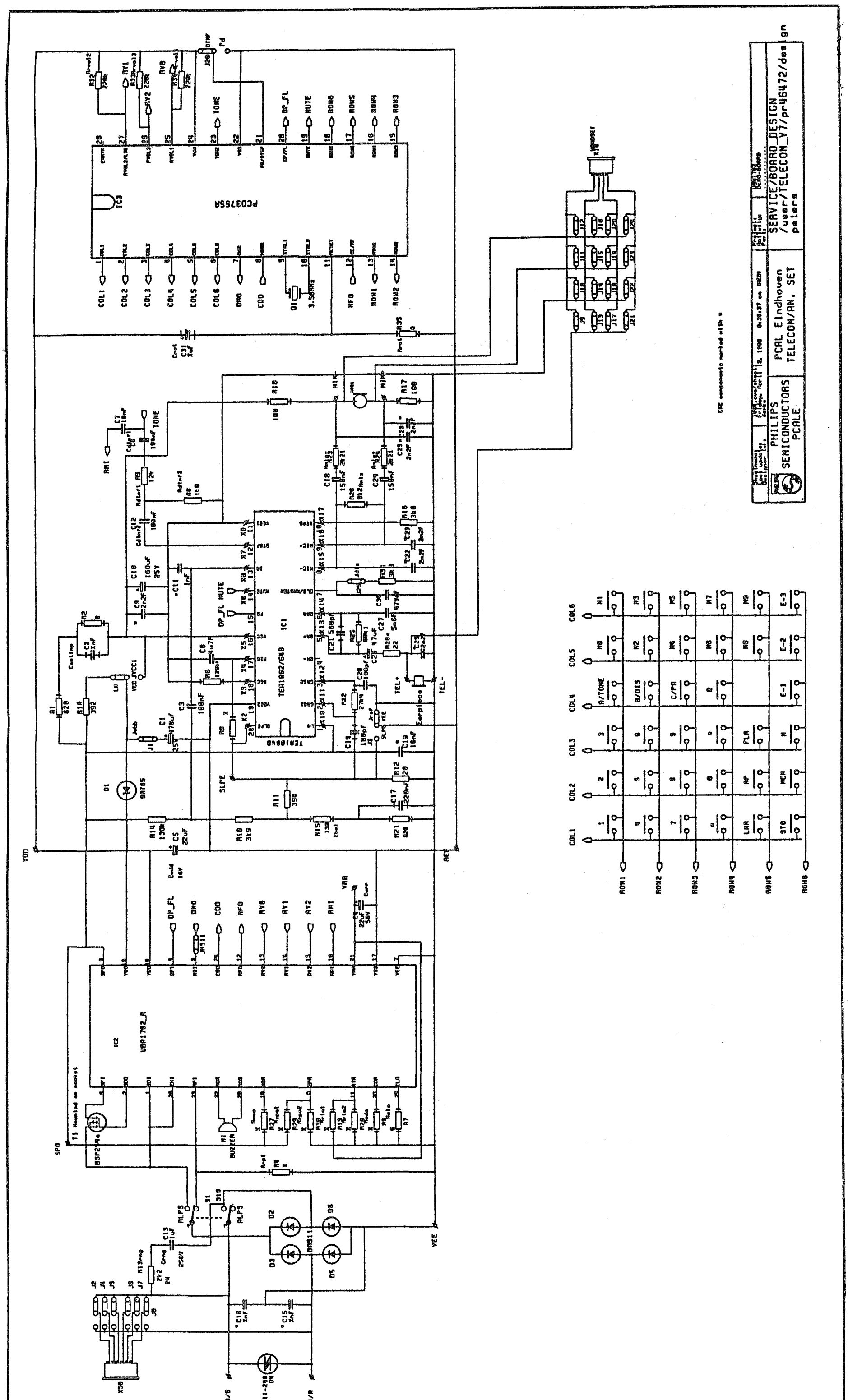
[Ref. 5] Application of the speech-transmission circuit TEA1062
Application Note by P.T.J. Biermans
October 1989, PCALE reportnumber: ETT89008

[Ref. 6] Application of the versatile speech/transmission circuit TEA1064 in full electronic telephone sets
by F. van Dongen and P.J.M. Sijbers
August 1989, PCALE reportnumber: ETT89009

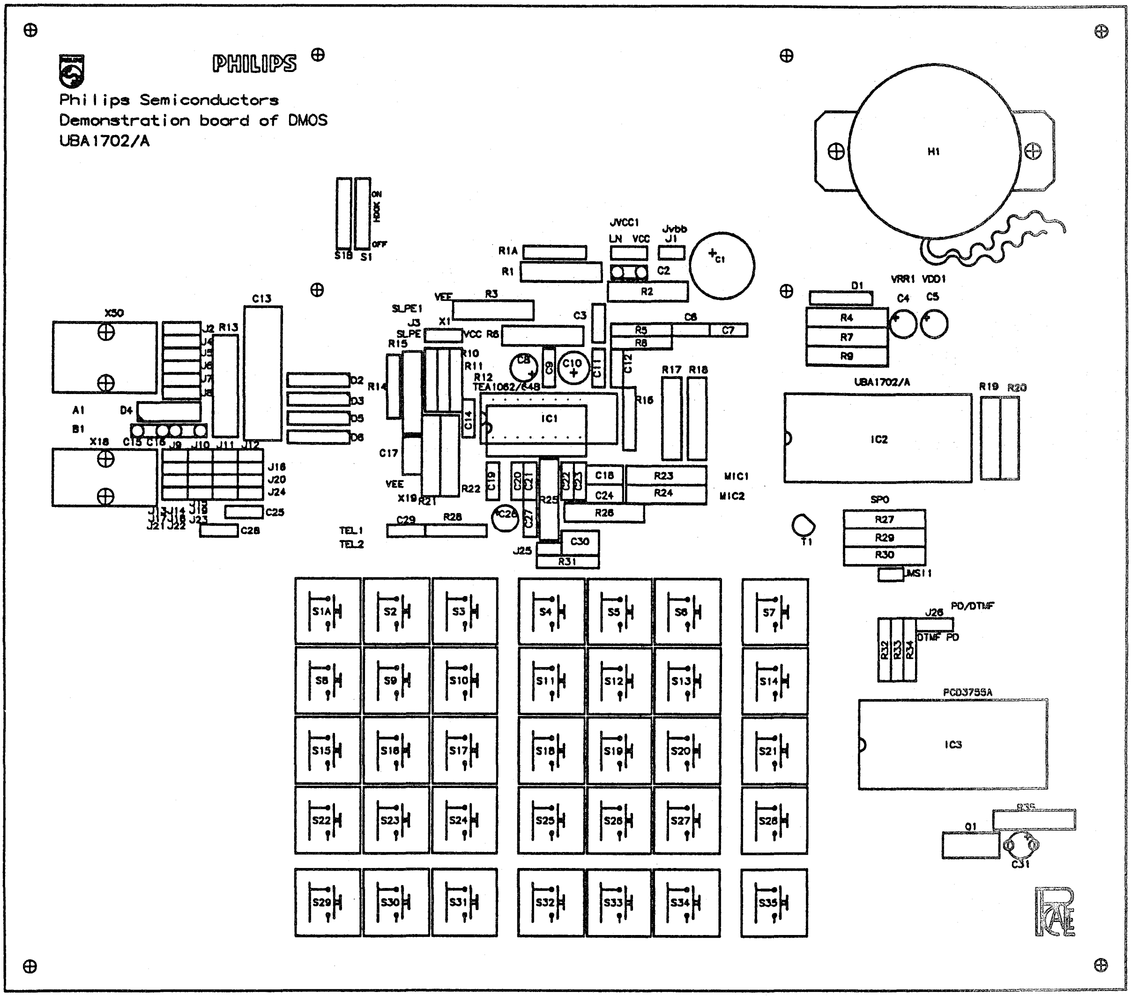
[Ref. 7] Measures to meet EMC requirements for TEA1060-family speech transmission circuits
Laboratory Report by M. Coenen and K. Wortel
October 1989, PCALE reportnumber: ETT89016

[Ref. 8] PCD3330-1: A multi-standard repertory dialler/ringer with EEPROM (programmed PCD3353A)
Objective specification; July 1993

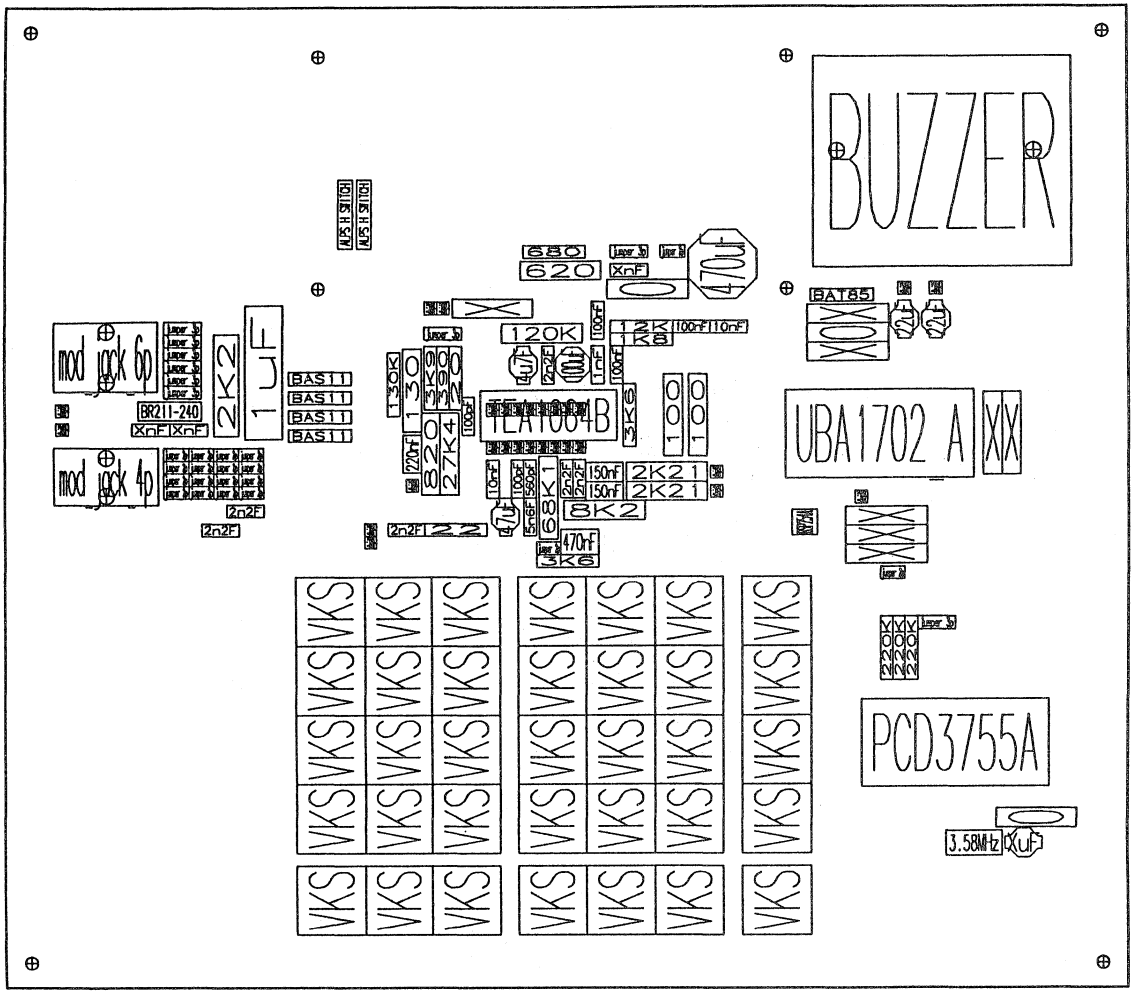
[Ref. 9] PCD3755A: Single-chip 8-bit microcontroller with on-chip DTMF generator, 8k OTP and 128-bytes
EEPROM; Objective specification; December 1993


PCD3755A
TEA1062/68
UBA1702_A
PHILIPS SEMICONDUCTORS PCALE
PCAL Eindhoven TELECOM/AN. SET
SERVICE/BOARD_DESIGN /user/TELECON_V7/pr46472/design peters

APPENDIX B. COMPONENT LAY-OUT WITH POSITION NUMBER.



8222_411_46472_2A1C


BUZZER
ALPS H SWITCH
ALPS H SWITCH
680
620
470uF
BAT85
mod jack 6p
2K2
1 uF
BAS11
BAS11
BAS11
BAS11
BR211-240
XnF
XnF
120K
130K
130
3K9
3K9
820
27K4
220nF
100pF
TEA1064B
12K
1K8
100nF
10nF
1nF
2n2F
3K6
100
100
UBA1702 A
mod jack 4p
2n2F
2n2F
2n2F
22
10nF
100pF
560pF
5n6F
68K1
68
150nF
150nF
2K21
2K21
8K2
470nF
3K6
VKS
220K
220K
220K
PCD3755A
3.58MHz

APPLICATION NOTE Nr	ETT/AN93017
TITLE	User's Manual of the OM4736 Demo kit: A MULTI-STANDARD TELEPHONE SET with PCA1070 and PCD3353A/008
AUTHOR	J. C. F. van Loon, P. J. M. Sijbers
DATE	December 1993

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TABLES:

FIGURES:

1 INTRODUCTION

Philips provides a wide range of dedicated IC's for telephony applications.
A new member is the PCA1070 which is a Multi-standard, Programmable Analog CMOS Transmission IC (abbreviated: **PACT**).
To demonstrate the flexibility of PCA1070 together with a µC the feature phone DEMO kit OM4736 has been designed.

The OM4736 demo kit consists out of:
- Printed circuit board PR45642 with the electronics of a feature phone .
- A handset with dynamic microphone and earphone + handset cord.
- I^2C-bus interface board for connection to the Centronics port (parallel port) of an IBM compatible PC.
- Evaluation software with an I^2C-bus control program allowing programming of the transmission parameters via the Centronics output (parallel port) of an IBM compatible PC (on a 3.5 inch diskette).

Functionality of the feature phone:
- Speech/transmission with privacy switch.
- Ringer detection and generation.
- Dialling features including:
  # Pulse, DTMF and Mixed mode dialling.
  # Last number redial.
  # Repertory dialling.
- Settings of programmable parameters are stored in an EEPROM.
- Changing of programmable parameters via keyboard.

Main components used are:

*** PCA1070:**
The PCA1070 is a CMOS integrated circuit performing all speech and line interface functions required in fully electronic telephone sets. It needs a minimum number of external components. The transmission parameters are programmable via I^2C-bus. This makes the IC adaptable to nearly all country requirements in the world, and to a various range of speech transducers, without changing the (few) external components.
These parameters are stored in the EEPROM of a µC and are loaded into the PCA1070 during the start-up phase of the transmission IC (hook-off).
PCA1070 also allows adaptation to the connected telephone line, by reading the line current via I^2C-bus and processing it in a µC.

*** PCD3353A/008:**
The PCD3353A/008 is a mixed mode multi-standard repertory dialler/ringer. The 10 repertory and redial numbers are stored in the (on-chip) EEPROM so that memory retention is guaranteed for 10 years without using a backup battery. Some dialling and ringer specifications can be fulfilled by changing a few bytes in EEPROM containing the different ringer and dialling procedures. The incoming ringer frequency is measured and a ringer melody is generated by the ringer melody generator and supplied to the ringer output stage. Also the transmission parameters of the PCA1070 are stored in the EEPROM. They will be send to the PCA1070 when necessary. All the parameters stored in the EEPROM can be changed via simple keyboard programming routines.

This report gives a brief description of the application and serves as a user-guide for operation of the DEMO kit.

For details about PCA1070 see Ref.[1]. The PCD3353A/008 is described in Ref.[2].
For specification of the I²C-bus see Ref.[3].
The circuit diagrams of the feature phone board and the I²C-interface board are given in Fig. 7. The components view of the feature phone board is given in Fig. 8 and the top/bottom copper views are given in Figs. 9, 10.

2 GENERAL INFORMATION OF THE DEMO KIT.

The feature phone demo board is marked with PR45642. In this report the name "board" will be used. When the provided handset is connected to the board, a complete telephone set is at your service. The set is ready to be connected to a telephone line. The transmission, dialling and ringer parameters have been set to default values (see Table 5) and must be adapted to the local requirements (keyboard programming routines are given in chapter 5). It is recommended to read the following instructions before connecting the set to the telephone line.

2.1 Telephone line connection/hookswitch.

The telephone line (or telephone line simulator) has to be connected via the modular-jack X32. To make the board universally applicable every pin of the modular-jack can be connected to the A/B terminals on the board by means of jumper J1 to J6. Default jumpers J3 and J4 have been mounted.

Before connecting a base cord, the connections should be adapted to the local standard.

The selection between hook-on/hook-off can be done by the double pole switch S1 (indication: "RING" or "SPEECH").

The terminals LINE+ and LINE- (after the polarity guard) and SPEECH and RING (after the hookswitch) are available for measurement purposes.

2.2 Handset connection.

The handset (or measuring equipment) can be connected via the modular-jack X33. To make the board universally applicable every pin of the modular-jack can be connected to the mic/tel terminals on the board by means of jumpers J7 to J22. Default jumpers J15 and J22 (mic) and J8 and J13 (tel) have been mounted for use with the provided handset.

In case a handset with an electret microphone is used, the circuitry on the printed circuit board has to be adapted slightly (see 3.3.5.1 microphone inputs).

For testpurposes the MIC+/MIC- terminals and the TEL+/TEL- terminals are available as well as terminal VP (supply for electret microphones).

2.3 Handset.

The provided handset is equipped with dynamic speech transducers. The DC resistance of both the microphone and the earpiece is 250 Ω.
The handset is specified according to the German requirements (FTZ 121 TR 8, July 1989 and FTZ 12 R 21, January 1989).
More details about the handset can be found in Ref.[4].

2.4 I²C-interface connector.

An I²C-interface connector X11 is available. It is intended for connection of external I²C devices. For example an external EEPROM (like PCF8582A) in which the transmission, dialling and ringer parameters are stored, can be used to load quickly the correct bitcode into the EEPROM of PCD3353A/008. Another possibility is to connect the I²C-bus interface board that allows direct programming of the PCA1070 by means of a personal computer via its centronics output (see Chapter 6 for more details).

2.5 Special jumper options

Three jumpers on the board PR45642 allow activation of the following special functions:

- AGC When closed the AGC (Automatic Gain Control) is dynamic, i.e. in conversation mode every 200 ms the line-current is measured and when necessary the gains of the microphone and the earpiece amplifiers are corrected (see paragraph 3.3.11 Line current control).
  When open the AGC is static, which means that this procedure only takes place during the start-up phase of the set.
  Default this jumper is open.

- EE-WRITE If an external EEPROM with hex-address code A0 (1010 0000) is connected to the I²C-bus, and the jumper EE-WRITE is closed, the PCD3353A/008 will go in read mode and will copy the data stored in the external device into its internal EEPROM after a hook-on/hook-off action, or a reset action. Then when the jumper is opened, the µC will load the data into the PCA1070.
  Default this jumper is open.

- PACT TEST When closed, the transmission parameters that are changed in keyboard programming mode and are displayed on the LCD, are directly loaded into the PCA1070 so that the effect can be observed immediately. The contents of the EEPROM of the µC are not changed unless the <PROG> key is pressed. This feature makes evaluation rather easy.
  When open, the transmission parameters that are stored in the EEPROM of the µC are loaded into PCA1070 only after hook-on/hook-off.
  Default this jumper is closed.

3 HARDWARE DESCRIPTION OF THE FEATURE PHONE BOARD.

A blockdiagram of the electronic feature phone on the PR45642 printed circuit board is shown in Fig. 1. In this chapter a brief description and/or explanation of some parts will be given, with reference to the circuit diagram in Fig.7.

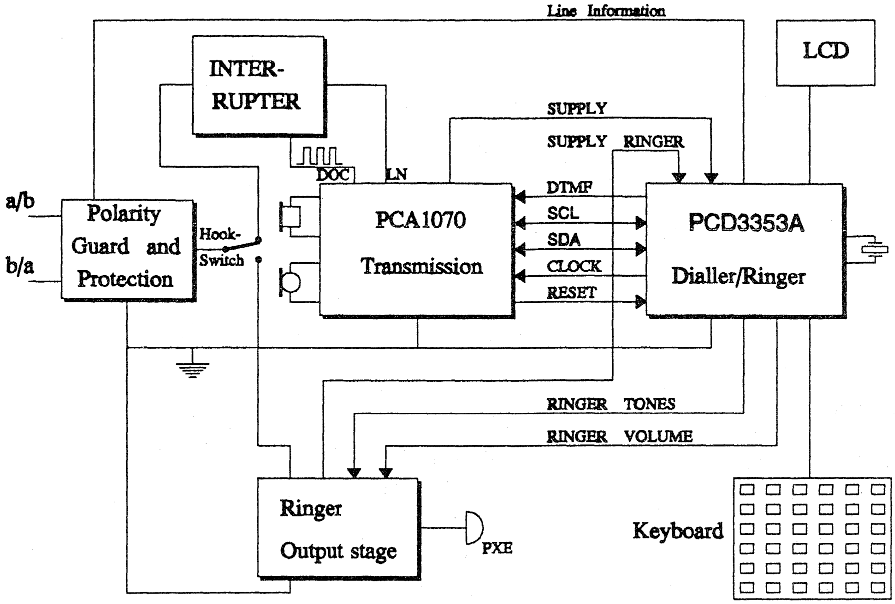

Figure 1 Block diagram of the printed circuit board

3.1 Polarity guard and protection

The diode bridge D2, D3, D4, D5 ensures that the set functions properly independent of the polarity of the line voltage applied to it.

Unprotected speech/transmission ICs might be destroyed by excessive current surges on the telephone lines if preventive measures are not taken. Protection is achieved by means of Break-Over-Diode (BOD) D_1 (BR211-240) between the a/b lines, an 11V zener diode D_9 between LN and VSS of the PCA1070, and current-limiting circuit (T_1, R_2).

The supply voltages of PCD3353A/008 and PCA1070 are limited by zener D8 and diode D11.

During ringing the supply voltages of the ringer output stage and the µC are limited by zenerdiodes D_7 and D_8.

Line information
By means of 2 diodes D16, D17 the incoming line signal (ac ringer signal or dc) is rectified to provide information about the ringer frequency and the line status (hook-on/hook-off) to the chip enable input (CE/T0) of the µC.

3.2 Interrupter

The interrupter consists of the BSD254A n-channel depletion MOSFET (T2) with a pull-up resistor R5. The interrupter transistor is directly controlled by the DOC output of the PCA1070. When the set goes off-hook a voltage is applied between the SPEECH and ground terminals. As pin DOC is default open the drain-gate voltage of T2 is zero and the transistor will conduct.

Interruption of the line current is done by making pin DOC "LOW"by setting the PCA1070 control bit DPI to "1" via I^2C-bus; then the gate voltage becomes negative with respect to the source and the transistor will stop conducting.

Transistor T1 and resistor R2 limit the current through T2, as T1 will switch on when the current through R2 exceeds about 150 mA (Vbe/R2). This current limiter is only intended to provide protection against current surges and is not to be used (for reasons of power dissipation) for continuous limitation of line current.

The board is prepared for an alternative interrupter circuit, using the BSP254A p-channel enhancement MOSFET. The circuit diagram is shown in Fig. 2.
If this circuit is mounted, the following components must be removed: T2, R5, R2, T1.

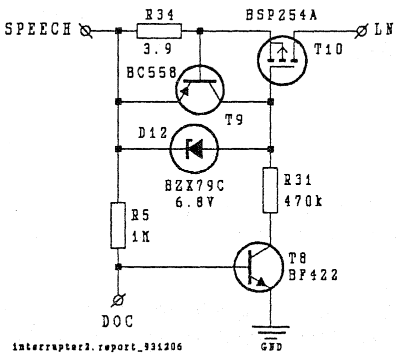

Figure 2 Interruptor with p-channel MOSFET and current limiter

3.3 PCA1070 APPLICATION

Only the main features of the PCA1070 are explained here. Complete information can be found in the data sheet (see Ref.[1]).
How to program (under keyboard control) the transmission parameters into PCA1070 is described in Chapter 5.

3.3.1 DC-setting

For DC the PCA1070 (between pins LN and VSS) is a voltage limiter with a slope of 20 V/A. The DC voltage V_{SLPE} can be programmed via the I^2C bus between 3.1 and 5.9 V.

The voltage drop between the a/b terminals of the set is determined by the DC voltage V_{SLPE}, the voltage drop across R7, T2, R2 and the polarity guard.

3.3.2 Set impedance

For AC (300-3400 Hz) the line termination is mainly determined by the programmable set impedance Zs consisting of two programmable resistors in series, Ra and Rb, with Rb shunted by a capacitor C. The value of C is not programmed directly but via the pole frequency: $C = 1/(2*\pi*fpole)$.

Tax pulse filter
In case a tax pulse filter is connected in series with a/b connections, the total set impedance is affected. This can be corrected by reprogramming the impedance Zs of PCA1070. PCA1070 allows correction of the overall set impedance for capacitive loads between pins LN and VSS of up to about 40nF. It is recommended to use a capacitor value of 33nF in the series LC filter which is normally connected between the a/b lines.

3.3.3 Supply for peripheral circuits

In conversation mode the PCD3353A/008 and the liquid crystal display get their supply voltage via diode (D11) from the VDD pin of the PCA1070. This voltage is derived from the line voltage via a first order external low-pass filter (R24, C17) in such a way that the line termination impedance is not deteriorated.

On the board the 2 supply capacitor system is used. Capacitor C17 (470uF) is used to decouple the supply voltage at pin VDD of the PCA1070 during conversation mode. It has a relatively large value to ensure that sufficient power is available to supply the PCA1070 and (via diode D11) the µC during line breaks.
Capacitor C18 (47uF) is used to decouple the supply pin VDD of the µC. It has a relatively low value to ensure fast start-up during ringing condition. In ringing condition diode D11 prevents that C17 is charged.

With jumper J23 diode D11 can be shorted. In this way the 1 supply capacitor system can be evaluated. If this is done, the values of the supply capacitors have to be adapted to ensure correct and fast start-up during ringing condition. Also the speech/ring detection input (T1) of the µC must be connected to another point (e.g. to VP of PCA1070 via a levelshifter) to ensure correct operation.

3.3.4 Power Control

3.3.4.1 Reset conditions

The PCA1070 has an internal reset circuit that monitors the supply voltage VDD. If VDD is below the threshold level (1.2V +/- 0.2V) the circuit is in reset-mode. In this mode the current consumption is low, the internal reset is active and writes the default values into all registers (See Ref.[1]).

The reset for the PCD3353A/008 is provided by the PCA1070. The sense input VMC of PCA1070 is connected to VDD of the µC. If VMC is below the threshold level, 2V±0.2V in conversation mode and 2.15V±0.25V in ringer mode, the output RMC is "1".

3.3.4.2 Start-up and switch-off behaviour

The behaviour of PCA1070 during hook-off, hook-on (switch-off and ringing) is described for this application with 2 supply capacitors.

HOOK-OFF PROCEDURE:
After switching the hookswitch from "RING" into "SPEECH" position, line current will be applied to the line input LN. The supply capacitors C17 and C18 will be charged.
The internal reset signal will change from "1" to "0" when VDD passes the threshold (1.2V+/-0.2V) and the circuit becomes partly active (the line interface part is kept in power down mode, so all of the line current is available to charge the supply capacitors).
The supply voltage of the µC is monitored by PCA1070 via sense input VMC. When VMC passes the reset level for µC (in start-up mode 2V+/-0.2V), the reset output for the µC (pin RMC) changes from "1" to "0" and the PCA1070 is switched into normal operating (speech) mode. As both T0 and T1 of the µC are high, the µC will select the speech mode and will load the transmission parameters from its EEPROM into PCA1070. This is done with bitcode DST (Dc Start Time) is "high" to activate the DC start circuitry in PCA1070. This ensures short DC settling time. After 100ms the transmission parameters are loaded once more into PCA1070 with DST bit is "low" to switch off the DC start circuit. PCA1070 is then in normal operating mode.

In case of an incoming call, the µC supply capacitor C18 may be charged already by the ringer signal at the moment when the handset is lifted. In this case the reset of the µC (pin RMC) is already low when the line current is applied to the line input LN. PCA1070 becomes active now as soon as the internal reset (VDD=1.2V±0.2) changes from "1" to "0". The µC loads the transmission parameters with DST="1" into PCA1070 as soon as both T0 and T1 become high (directly after hook-off). After 100ms the transmission parameters are loaded once more with DST="0".

HOOK-ON PROCEDURE:
After switching the hookswitch from "SPEECH" into "RING" position, the voltage at pin T0 will change from "1" to "0". The µC will recognize a line-break and will switch PCA1070 into power down mode via I^2C-bus (PD=01). If T0 stays "0" longer than the reset-delay time (typ. 160 ms), the µC will switch PCA1070 in normal operating mode via I^2C-bus (PD=00) and the µC will go into standby mode. So the supply decoupling capacitor at VDD of PCA1070 will be discharged fast until the internal reset of PCA1070 takes place. The internal

reset signal will change from "0" to "1" when VDD passes the threshold (1.2V+/-0.2V) and the PCA1070 will go into reset mode (line interface part in power down and all programmable parameters are reset to default values).

The supply capacitor at VDD of the µC (connected to VMC of PCA1070) will also discharge (but slowly). The reset output for the µC (pin RMC) will change from "0" to "1" when VMC passes the threshold level (2.15V+/-0.25V in this case, because PCA1070 is already in reset mode).

RINGER CONDITION:
When a ringer signal is applied to the a/b terminals (hookswitch in "RING" position), the supply capacitor at VDD of the µC (also connected to VMC of PCA1070) is charged. No line current is applied to pin LN, so SLPE is low and VDD of PCA1070 is low.
The voltage on VMC will increase and when VMC passes the level for µC reset (2.15V+/-0.25V) output RMC changes from "1" to "0".
As T0 is "1" and T1 is "0" the µC will select the ringer mode and will generate ringer tones if the frequency of the ringer signal is within the pre-programmed frequency range (see paragraph 3.4.4).

3.3.4.3 Dial pulse input/pulse dialling

The interruptor is directly controlled by pin DOC (open drain output). Pin DOC is controlled by the µC via I^2C-bus by setting bitcode DPI. DPI="1" results in pin DOC="0"; DPI=0 (default) results in DOC=high-ohmic.
During the break periods, the PCA1070 is switched to power down by setting the control bits PDx="01" via I^2C-bus. Also the PCA1070 control bit DST is set to DST="1" during pulse dialling for fast switching behaviour (see Ref.[2] for a detailed description of the dialling procedure).

3.3.5 Microphone channel

3.3.5.1 Microphone inputs/supply for electret microphone

The PCA1070 has symmetrical microphone inputs and accepts input signals up to 70 mVpeak for THD=2% (with VDD>=2.5V). Its input impedance is typically 100 kΩ. Dynamic, magnetic, piezoelectric and electret (with built-in FET source follower) microphones can be used. Some possible microphone arrangements are shown in Fig. 3.

The feature phone board is ready for use with microphone capsules which require a symmetrical input. If necessary R14 can be used to lower the terminating impedance.

Since PCA1070 has a microphone gain that can be programmed via I^2C-bus between 30 and 51 dB, an external attenuation network (R12, R13, R14) is required if a very sensitive type of microphone is used needing less than 30 dB of gain.

For use with electret microphones components R8 and/or R9 can be mounted for the supply, and DC-blocking capacitors C7, C8 must be mounted.

In case asymmetrical drive of the microphone inputs is used, the MIC+ input should be used as a signal input. Care should be taken that both inputs MIC+ and MIC- see equal impedances

to the common, otherwise residual line signals being present on the supply point VDD will cause inaccuracy in gain and sometimes (with a large DC-blocking capacitor connected to MIC+) even low-frequency hicking may occur.
In practice a low-ohmic (e.g. 50 Ohm) sinewave generator can be connected via a DC blocking capacitor to pin MIC+; a capacitor with the same value has to be connected between pin MIC- and VSS.

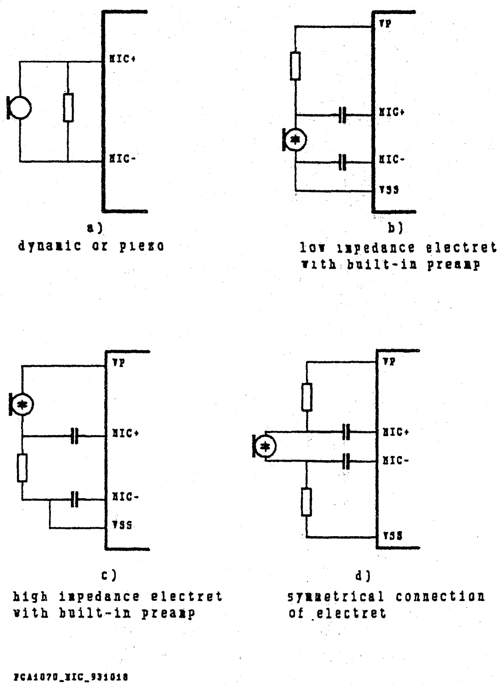

Figure 3 Microphone arrangements.

3.3.5.2 Gain of microphone channel

The total gain can be programmed between 30 dB and 51 dB in 1 dB steps by means of a programmable amplifier ("**send prog-amp**") which can be controlled via the I^2C-bus. The "send prog-amp" has an allowed gain range between 4 and 25 dB.

Total gain of the microphone channel is (with the assumption that the load presented to the a/b terminals of the set is equal to the programmed set impedance):

$G_M = 26 + Gma$ [dB]

Default $G_M = 26 + 15 = 41$ [dB]

where Gma = Gain "send prog-amp"

3.3.6 Dynamic limiter

To prevent distortion of the transmitted speech signal, the gain of the microphone pre-amplifier is reduced when peaks of the signal on the line exceed an internally determined threshold. The threshold levels can be selected via I²C-bus by bitcode **DLT;** if DLT = "0" the AC peak-to-peak line voltage on pin LN is default typ. 3.5V(p-p) and if DLT = "1" the level is typ. 2.6V(p-p).
The gain control is done in 1dB steps. Therefore the mean peak to peak output level on the line is slightly below the specified threshold value. The gain control range is typically 12 dB.

3.3.7 DTMF amplifier

3.3.7.1 DTMF input

The PCA1070 has an asymmetrical DTMF input. Its input impedance is typ. 200 kΩ//45pF. A DC blocking capacitor is integrated onto PCA1070. For EMC reasons, the DTMF input is connected internally via a highohmic resistor (typ. 200 kOhm) to the bias voltage VDD/2. Pin DTMF input can sink/source currents up to typ. 10 µA.
In this application the DTMF input is coupled directly (without) DC blocking capacitor to the tone output of the PCD3353A/008.

The DTMF input is enabled by setting the Send Mute control bit (SM) to "1" via I²C-bus. In this application this is done automatically by the PCD3353A/008 when a key is pressed in DTMF dialling mode.

3.3.7.2 DTMF gain

The gain between DTMF input and the line LN can be programmed between 1 dB and 21 dB in 1 dB steps by setting the gain (Gma) of the "send prog-amp" via I²C-bus. Recommended range in DTMF mode for Gma:-5 to 15 dB.

Total gain of the DTMF channel between DTMF input and the line LN is:

$G_{DTMF} = Gma + 6$ [dB]

Default $G_{DTMF} = 15 + 6 = 21$ [dB]

3.3.8 Receiving amplifier

3.3.8.1 Earpiece outputs/hearing protection.

The outputs may be used to connect dynamic, magnetic or piezoelectric earpieces with single ended or differential drive. Earpiece arrangements are shown in Fig. 4.
The load select bit RFC is default 1 to guarantee stable operation in case of a capacitive load (piezoelectric earpiece). With a resistive load (dynamic capsule) RFC should be set to "0" via I²C-interface to obtain optimum performance w.r.t. distortion and bandwidth.

Two levels for hearing protection can be selected via I²C-interface with control bit HPL.
If HPL="0" the output level is limited to typ. 2.3 Vp-p. When HPL="1" the output level is limited to typ. 5.9 Vp-p.

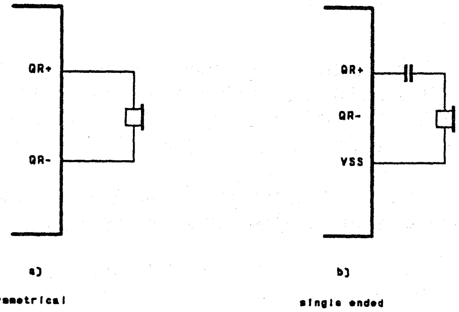

Figure 4 Earpiece arrangements.

The electrical hearing protection levels can be adapted to the electrical/acoustical characteristics of the particular earpiece by using an attenuation (resistor) between the earpiece outputs and the earpiece.

3.3.8.2 Gain setting receive channel.

The gain of the receive channel is defined between the line connection LN and the earpiece outputs QR+, QR-.
The LN terminal accepts receive signals up to typically 1V(rms) for THD=2%.

The gain of the receive channel can be programmed between -19dB and +11dB (symmetrical drive) in 1 dB steps by means of a programmable amplifier "**rec prog-amp**" which can be controlled via the I^2C-bus (allowed range Gra:-19dB to +11 dB).

Total gain of the receiving channel is:
Symmetrical drive

G_{RS} = Gra [dB]
Default G_{RS} = -6 [dB]

Asymmetrical or single ended drive

G_{RA} = Gra - 6 [dB]
Default G_{RA} = -6 - 6 [dB] = -12 [dB]

Where Gra = Gain of the "rec prog-amp"

3.3.9 Confidence tone

The confidence tone gain (between DTMF input and earpiece outputs QR) can be programmed between -40dB and -19dB (symmetrical drive of earpiece) by setting the gain (Gra) of the "rec-prog-amp" (allowed range Gra:-25 to 0 dB).
The confidence tone gain (DTMF to QR outputs) is:
With symmetrical drive of earpiece

$$G_{CTs} = Gra - 19 \quad dB$$
$$\text{Default } G_{CTs} = -6 - 19 = -25 \quad dB$$

At low gain settings (Gra<-10dB), the confidence tone gain will be slightly higher than the calculated value. This is caused by a residual signal.

3.3.10 Sidetone balance

The PCA1070 has an on-chip anti sidetone circuit. An internal balance impedance Zoss can be programmed via the I^2C bus to match the external line impedance Zline for optimal side tone suppression.
Zoss = Rsa+[Rsb//Cs]; (fp=1/(2π.Rsb.Cs))

<u>Tax pulse filter</u>
In case a tax pulse filter is connected in series with a/b connections, the total impedance as seen from the telephone set is affected. Therefore the internal balance impedance Zoss has to be reprogrammed. PCA1070 allows correction of the Zoss for capacitive loads between pins LN and VSS of up to about 40nF. It is recommended to use a capacitor value of 33nF in the series LC filter which is normally connected between the a/b lines.

3.3.11 Line current control

The DC line current can be read via I²C interface. This gives information can be used to change several parameters with line current (for example line loss compensation, sidetone balance and DC characteristic).
In this application it is used for automatic line loss compensation (also knows as Automatic Gain Control or AGC). The gain control characteristic of both the microphone and earpiece amplifiers can be programmed under keyboard control. As shown in Fig. 5 a start current (dc_i_0), 5 current values (dc_i_1 to dc_i_5), a stop current (dc_i_6) and the desired step resolution of the gain control (dc_i_d_G) can be programmed. The values of dc_i_0 to dc_i_6 that are depicted in Fig. 5 are those as shown on the LCD. They do not match exactly the real line current because the ROM code of the PCD3353A/008 has been defined according to the first objective specification of PCA1070. The specification has been modified slightly thereafter (Ref.[1]). A correction table is given in Table 2.
The automatic line los compensation can be disabled by setting the step resolution dc_i_d_G to 0dB.

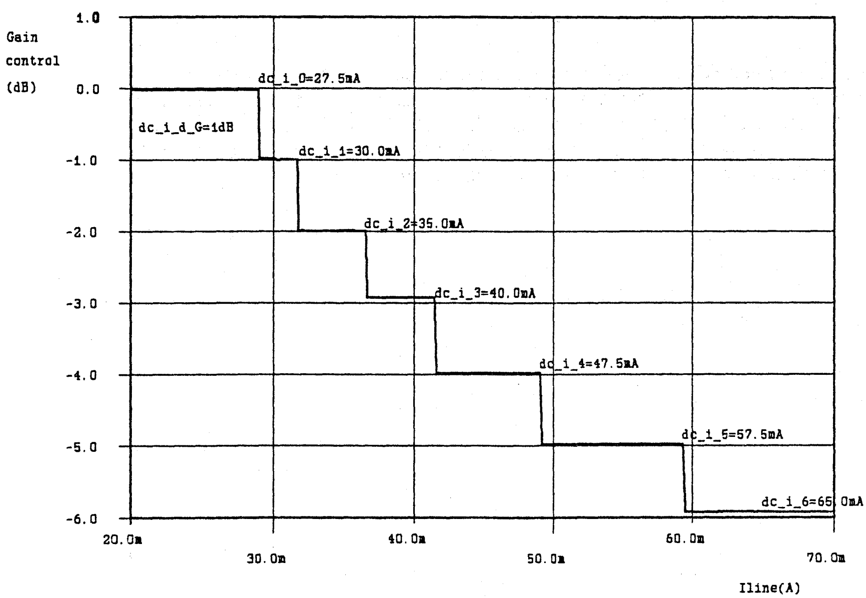

Figure 5 AGC characteristic.

3.4 PCD3353A/OO8 CONTROLLER

In this chapter a brief description of the functions is given, full specification of the PCD3353A/008 is given in (Ref [2]).

The six column inputs and six scanning row outputs are connected to a 6 x 6 matrix keyboard. Except for the normal dialling, redialling and repertory dialling keys this matrix contains also five special PCA1070 programming keys and a ringer programming key.

All keyboard actions are shown on the LCDisplay, for easy dialling, programming and evaluation.

3.4.1 Features

- Pulse, DTMF and Mixed mode dialling.
- Flash or register recall.
- Mute function (privacy switch).
- Access pause generation and termination.
- Redial (maximum 24 digits) stored in internal EEPROM.
- Storage for 10 repertory dial numbers (18 digits each) in internal EEPROM.
- Three-tone ringer with 4 volume steps, 4 speeds, 4 different sequences, and 1f/2f input frequency selection.
- Selectable mark to space ratio (3:2 or 2:1) stored in EEPROM.
- PCA1070 control via the I^2C-bus.
- PCA1070 transmission parameters can be programmed and stored via keyboard.
- Special PCA1070 evaluation mode is available that allows direct programming of the PCA1070 transmission parameters via keyboard (jumper option PACT_TEST).
- Possibility to program all PCA1070 parameters from a PC via the I^2C-bus.
- Programmable AGC characteristic of microphone and receiving amplifiers. (Dynamic or static with jumper option AGC).
- Display control to simplify the dialling, programming and evaluation.
- On-chip oscillator uses low-cost 3.58 MHz crystal or PXE resonator.
- On-chip voltage reference for stabilized supply and temperature independent DTMF output.
- On-chip filtering for low DTMF output distortion (CEPT CS 203 compatible).
- Reset is generated via PCA1070 (On-chip power-on reset of PCD3353A/008 is preset to low value (typ. 1.2V)).
- Supply voltage range 1.8 to 6.0 V (2.5 to 6.0 V in EEPROM erase/write and DTMF and ringer mode).
- Possibility to copy data from an external memory into the internal EEPROM (jumper option EE-WRITE).

3.4.2 Operating modes

The PCD3353A/008 has six modes of operation:

- Standby: both T0 and T1 are LOW and the PCD3353A/008 is in the standby mode (very low current).

- Conversation: both T0 and T1 are HIGH and the PCD3353A/008 is in operating mode and the AGC function can be set to static or dynamic by jumper option AGC.
  From this state the micro-computer can go to the dialling, keyboard programming mode or PACT evaluation mode.

- Dialling: T0 and T1 are HIGH and dialling is started either by direct key press or by recalling an autodial number from redial or repertory.

- Keyboard programming: T0 and T1 are HIGH and by pressing the PROG-key the keyboard programming mode is activated and one of the programming modes can be selected by pressing the appropriate key.

- PACT evaluation: this is a special option of the keyboard programming mode.
  It can be activated with jumper PACT_TEST (see paragraph 2.5).

- Ringer: T1 is LOW and at T0 the AC ringer voltage is supplied. If the AC frequency at T0 is within the programmed frequency range the three tone melody generator is activated.

A special programming mode is available:
- EEPROM programming: by means of jumper EE-WRITE, the PCD3353A/008 can be put in read mode and it will to copy data from an external memory with hex-address code A0 (1010 0000) into its EEPROM (see paragraph 2.5).

3.4.3 Dialling functions

Pulse dialling. The dialling frequency is 10 Hz (100 ms) with selectable break and make times of 60/40 (3:2) or 66/33 (2:1) ms.
Actual dialling, muting and setting of the power down bits and the DST bit is done in the PCA1070 by sending the appropriate control bits via I^2C-bus to the PCA1070 under control of the PCD3353A/008 (see also paragraph 3.3.4.3).

DTMF dialling. The minimum tone burst/pause duration is 100/100 ms.
Muting and gain corrections are send from the PCD3353A/008 via the I^2C-bus to the PCA1070, The DTMF signal is send directly to the DTMF input of PCA1070.

Mixed mode dialling. This is only possible when the controller is set to the pulse dial mode, activation of push-button TONE, * and # changes the dialling mode to DTMF.

FLASH results in a timed line current break of 100 ms.

The MUTE can be used as a privacy switch. It can be used when no dialling is active. It is a toggle function. Every time when the Mute key is pressed the send and receive mute of the PACT are switched on or off, depending on the previous state.

Last number redial is activated by pressing the LNR-key as the first key after hook-off.
Up to 24 digits can be stored in the redial register.
After the work register overflows, a 10 digits First-In-First-Out register (FIFO) takes over as buffer and the contents of the work register are not copied to the redial register.
However, if the FIFO register overflows (more than 10 digits) further input is ignored.

10-number repertory dialling, 18 digits each, which is accessible by a one-key procedure (M0 to M9). Repertory numbers can be dialled-out after or before entering manual dialling, last number redial and by entering the memory locations in successive order.
During transmission of a number called from the memory location, the controller does not accept keyboard entries.

Access pause. If during entering a telephone number via keyboard for normal dialling or during repertory number programming the AP-key (access pause key) is pressed, then an access pause (of 3 seconds) is stored in the redial or repertory dial register.

3.4.4 Ringer function.

The melody ringer of the PCD3353A/008 µC can be programmed to become active for incoming signals at the T0 input with frequencies between 40 and 120Hz (2f method) or 20 to 60Hz (1f method).

Since in this application a double phase rectifier is used (2f), the frequency range 40 to 120 Hz has been preset so that the ringer becomes active for all incoming ringer frequencies at the a/b lines between 20 and 60 Hz.
In an application where a single phase rectifier is used (1f), the frequency range (at T0) must be programmed to 20 to 60 Hz.
The ringer melody generator can select one of four melodies.
The generated melody is built up of three frequencies. These frequencies are generated successively with a repetition rate (speed) which can be adapted too.

Volume change in ringer mode can be done by pressing keys 1 (low) to 4 (high) at the time when a ringer burst is present.

All ringer options such as (1) volume, (2) melody, (3) speed and (4) input frequency can be changed during conversation mode. See Table 4 and the detailed specification of PCD3353A/008 (Ref.[2]).

3.5 Ringer output stage/volume control

A push-pull output stage (T4, T5, D10, R15, R16) drives a PXE buzzer.
Volume control is done by means of switching a resistor network in series with the buzzer. The resistors are switched with transistors T6, T7 and are driven from port pins RVOL1 and RVOL2 from the µC.

4 ELECTROMAGNETIC COMPATIBILITY

In the presence of high-intensity electromagnetic field, common-mode amplitude modulated R.F. signals can be introduced in the a/b lines. These common-mode signals can become differential-mode signals as a result of asymmetrical parasitic impedances to ground (for example through the hand of the subscriber holding the handset). Preventive measures have to be taken to avoid these signals being detected and the low-frequency modulation appearing as unwanted signals at the earpiece or on the line.

The layout of this printed circuit board is designed with respect to electromagnetic compatibility (EMC). Basic protection is provided by means of discrete capacitors to suppress the unwanted r.f. signals before they can enter the circuit. Capacitor types suitable for high frequencies must be used, for example ceramic types. Those are C1, C2 at the a/b line connections, C5, C6, C15, C16 and the handset in/outs and C14 at the LN pin of the PCA1070.

The EMC capacitors at the a/b terminals (C1, C2) and the LN pin (C14) slightly influence the balance return loss and the sidetone characteristics at higher frequencies.

5 PROGRAMMING PROCEDURES VIA KEYBOARD

The PCA1070 variables, dialling options and ringer parameters which are stored in the EEPROM of the µC can be changed via the keyboard. For programming of the PCA1070, five special programming keys (<DCVOLT>, <GAIN MIC>, <GAIN REC>, <SET IMP>, <SIDE TONE> are available. For programming of the dialling options the <LNR> key is used and for the ringer parameters the special <RING> key is available.
The keyboard programming modes can be activated by pressing key <PROG> and this is indicated at the LCD by the label <STORE>. Subsequently one the programming modes can be chosen.

In the PACT evaluation mode (selectable by mounting jumper J25: PACT_TEST) the PCA1070 values that are changed by keyboard are send directly to the PCA1070. The EEPROM contents, however, will only be changed after ending the program procedure by pressing the program key <PROG>.
The programming routines are described in shortform in the next paragraphs and are described in depth in Ref.[2]. Default settings and setting ranges are given in Table 5.

5.1 PCA1070 transmission parameters

5.1.1 DC setting/Automatic Gain Control

The DC-voltage drop ($V_{SLPE\text{-}VSS}$) and the gain control characteristic (see Fig. 5). A start current (dc_i_0), 5 current values (dc_i_1 to dc_i_5), a stop current (dc_i_6) and the desired step resolution of the gain control (dc_i_d_G) can be changed and stored as follows:

- Press the PROG key,
- Press the DC-VOLT key,
- Program the correct DC voltage value with the + or - keys,
- Press the DC-VOLT key again,
- Program the start current dc_i_0 (+/- keys),

- Press the DC-VOLT key again,
- Program the 1st current dc_i_1 (+/- keys),
- Press the DC-VOLT key again,
- Program the 2nd current dc_i_2 (+/- keys),
- Press the DC-VOLT key again,
- Program the 3rd current dc_i_3 (+/- keys),
- Press the DC-VOLT key again,
- Program the 4rd current dc_i_4 (+/- keys),
- Press the DC-VOLT key again,
- Program the 5th current dc_i_5 (+/- keys),
- Press the DC-VOLT key again,
- Program the stop current dc_i_6 (+/- keys),
- Press the DC-VOLT key again,
- Program the step resolution of the gain control dc_i_d_G (+/- keys),
- Press the PROG key; all the programmed values are stored into EEPROM.

5.1.2 Sending channel

The microphone gain, the threshold of the dynamic limiter and the DTMF gain can be changed and stored as follows:
- Press the PROG key,
- Press the GAIN-MIC key,
- Program the send prog-amp to set the microphone gain (+/- keys),
- Press the GAIN-MIC key again,
- Program the DLT bit (Dynamic Limiter Threshold) to on or off (+/- keys),
- Press the GAIN-MIC key again,
- Program the send prog-amp to set the DTMF gain (+/- keys),
- Press the PROG key; all the programmed values are stored into EEPROM.

5.1.3 Receiving channel

The receive gain, the hearing protection level, the load select bit and the confidence tone gain can be changed and stored as follows:
- Press the PROG key,
- Press the GAIN-REC key,
- Program the rec prog-amp to set the receive gain (+/- keys),
- Press the GAIN-REC key again,
- Program the Hearing Protection Level (HPL) to on or off (+/- keys),
- Press the GAIN-REC key again,
- Program the load select bit (RFC) to on or off (+/- keys),
- Press the GAIN-REC key again,
- Program the rec prog-amp to set the confidence gain (+/- keys),
- Press the PROG key; all the programmed values are stored into EEPROM.

5.1.4 Set impedance

Set-impedance values Ra, Rb and fpole values can be changed and stored as follows:
- Press the PROG key,
- Press the SET-IMP key,
- Program the set impedance value Ra (+/- keys),
- Press the SET-IMP key again,
- Program the set impedance value Rb (+/- keys),
- Press the SET-IMP key again,
- Program the set impedance value Fpole (+/- keys),
- Press the PROG key; all the programmed values are stored into EEPROM.

5.1.5 Sidetone

The sidetone balance impedance values Rsa, Rsb and Cs values can be changed and stored as follows:
- Press the PROG key,
- Press the SIDE-TONE key,
- Program the sidetone value Rsa (+/- keys),
- Press the SIDE-TONE key again,
- Program the sidetone value Rsb (+/- keys),
- Press the SIDE-TONE key again,
- Program the sidetone value Cs (+/- keys),
- Press the PROG key; all the programmed values are stored into EEPROM.

5.2 Dialling (options) features

Dialling option programming part for changing from pulse to tone dialling, select another mark-to-space ratio for pulse dialling and selecting the correct DST action during pulse dialling for old or final PCA1070 samples.
This is done as follows:
- Press the PROG key,
- Press the LNR-key (selects the dialling option programming mode),
- Press key 1 for pulse dialling,
- 2 for DTMF dialling,
- 3 for mark-to-space of 3:2 (60/40 ms),
- 4 for mark-to-space of 2:1 (66/33 ms),
- 5 DST action for old PCA1070 samples (obsolete),
- 6 DST action for final PCA1070 (always to be used).
- Press the PROG key, all programmed values are stored into EEPROM.

Storing repertory numbers can be done via the following one key access method:
- Press PROG key,
- Press one of the ten location keys M1 to M10,
- Enter the telephone number (maximum 18 digits inclusive AP),
- Press the PROG key, the new number will be stored into EEPROM.

If no telephone number is entered then the old one is cleared.

5.3 Ringer parameters

The ringer output volume, melody, speed and also input frequency detection range can be changed by the following program procedure:

- Press PROG-key.
- Press Ring-key (ringer program key). The LCD shows the ringer status (see Fig. 6).
- Select parameter to be changed (key 1 to 4)

  Where: Key-1 Volume (u=..)
  Key-2 Melody (m=..)
  Key-3 Speed (S=..)
  Key-4 Input frequency (F=..)
- Select value (See Table 4)

  This will automatically change the value for the previously selected parameter (volume, melody, speed or input frequency). The LCD gives direct the new values of the ringer parameters (See Fig. 6).
- Now the routine can be repeated by selecting again a(nother) parameter and a value.
- Press the PROG-key; the program cycle is now ended and the new values will be stored in EEPROM and the LCD is cleared.

Figure 6 Display status in the ringer programming mode.

If the above routine is interrupted by an on-hook/off-hook action the procedure is ended and the old programmed values stay in the EEPROM.

6 PROGRAMMING VIA PC

6.1 The I²C-bus interface board

The I²C-interface board can be connected to the centronics output (parallel port) of most IBM compatible PCs. It can be plugged directly onto the centronics connector or connected via a 25 core flat cable.
Note that a galvanic separation between the mains supply and the telephone line is needed. Therefore it is recommended to use a laptop or palmtop PC with battery supply or to use a mains isolating transformer.

The I²C-bus jacks of both the I²C-interface board and the PCA1070 evaluation board have to be connected via a 4-core flat cable.
Note that the maximum allowed length of this cable is 3 meter.

The centronics output is driven by the I²C-bus control program on the 3.5 inch diskette. For explanation of the use of this program see paragraph 6.2.

Note that programming via keyboard is not possible any more when the I²C-interface board is connected to the I²C connector. Also the display does not operate correct any more. It is recommended to remove the PCD3353A/008 from its socket during programming via PC.

6.2 Software description

The OM4736 demo kit contains also a 3.5 inch diskette with the control program which drives the centronics output (parallel port) of an IBM compatible PC and via the I²C-bus interface board, controls the PCA1070 on the feature phone board.
The use of this diskette and its routines is very easy.

Put the diskette in the IBM compatible PC or copy this program to a sub-directory on the harddisk. This makes loading of the separate programs much faster.

Now type the word 'PACT' followed by a <return>. A batch program will be started and a menu will pop-up on the monitor with six selectable programs and a special command for leaving this batch program.
These programs consist of 3 programs in 2 versions (for use with a monochrome or a color monitor) and a stop program.

Programs 1 and 2: Here a table is given with the default values of all programmable parameters from PCA1070. Instead of the bitcode, the numerical (default) values are given. For example the DC-voltage is given as DC voltage = 4.7 V. It can be programmed between 3.1V and 5.9V in steps of 0.4 volt. When you want to change a value, simply go to the value you want to change, by means of the cursor keys. This parameter will be high-lighted now. Change the values with the plus and/or minus key and finally press S or s to send the corresponding bitcode to the PCA1070.

Programs 3 and 4: Here the same table as above is given but now the bit codes are shown. Like programs 1 and 2 each bit code can be changed by moving the cursor to the bitcode you want to change. Change the value by pressing the space bar and send the new values to the PCA1070 by pressing S or s. The difference with the previous program is that here all the bits (also those which are not in use) can be changed. E.g. for the send prog-amp and the receive prog-amp, the maximum possible setting range is between +25 and -25 dB while the maximum bitcode can be between +31 and -31dB.

Programs 5 and 6: This programs reads the bit code of the line current.

All the above programs can be left by pressing the <esc> key.

S or s: They stop the batch program and will go back to DOS.

7 USING THE OTP µC PCD3755A

The PCD3755A is an 8-bit microcontroller with on-chip DTMF generator, 8k OTP and 128-bytes EEPROM. The instruction set is based on that of the MAB8048 and is software compatible with the PCD33xxA family. For details please see Ref.[6].

The PCD3353A/008 may be replaced by the PCD3755A. However for correct operation of the ringer function pull-up resistors (100k) must be connected at pins RVOL1 (pin 21) and RVOL2 (pin 25).

8 REFERENCES

[1] Philips Semiconductors data sheet,
'PCA1070, Multi-standard, Programmable Analog CMOS Transmission IC',
Preliminary specification PCA1070,
Dated: 25 August 1993.

[2] Philips Semiconductors Report nr. ETT/AN92011.
'Software specification to control the PCA1070 PACT_IC (PCD3353A/008)',
by J.C.F. van Loon.

[3] Philips Semiconductors publication, January 1992,
'The I^{2}C-bus and how to use it (including specifications)',
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[4] AKG WIEN-TELECOM
'Technische Lieferbedingungen zu IHA 90 D',
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[5] Philips Semiconductors Application Note ETT/AN93010,
'User's manual of the OM4737 evaluation kit for PCA1070' by J.C.F. van Loon and P.J.M. Sijbers.

[6] Philips Semiconductors data sheet,
'Single chip 8-bit microcontroller with on-chip DTMF generator 8k OTP and 128-bytes EEPROM',
Objective Specification PCD3755A,
Dated: August 1993.

TABLE 1: JUMPER FUNCTION AND DEFAULT SETTINGS.

Jumper	Function	Default setting
J1	Line -a/b connection	open
J2	Line -a/b connection	open
J3	Line -a/b connection	closed
J4	Line -a/b connection	closed
J5	Line -a/b connection	open
J6	Line -a/b connection	open
J7	Handset: Tel- connection	open
J8	Handset: Tel- connection	closed
J9	Handset: Tel- connection	open
J10	Handset: Tel- connection	open
J11	Handset: Tel+ connection	open
J12	Handset: Tel+ connection	open
J13	Handset: Tel+ connection	closed
J14	Handset: Tel+ connection	open
J15	Handset: Mic+ connection	closed
J16	Handset: Mic+ connection	open
J17	Handset: Mic+ connection	open
J18	Handset: Mic+ connection	open
J19	Handset: Mic- connection	open
J20	Handset: Mic- connection	open
J21	Handset: Mic- connection	open
J22	Handset: Mic- connection	closed
J23	One-two C supply selection	open
J24	EE-write activation	open
J25	PACT-test activation	closed
J26	AGC activation	open

TABLE 2: CORRECTION TABLE LINE CURRENT CONTROL.

Iline (mA)	
Value on LCD	Switching current
10.0	11.25
12.5	13.75
15.0	16.25
17.5	18.75
20.0	21.25
22.5	23.75
25.0	26.25
27.5	28.75
30.0	31.25
32.5	33.75
35.0	36.25
37.5	38.75
40.0	41.25
42.5	43.75
45.0	46.25
47.5	48.75

Iline (mA)	
Value on LCD	Switching current
50.0	51.25
52.5	53.75
55.0	56.5
57.5	59.5
60.0	62.5
62.5	65.25
65.0	68.75
67.5	70.25
70.0	72.75
72.5	76.25
75.0	78.75
77.5	81.25
80.0	83.75
82.5	86.5
85.0	89.5
87.5	92.5

TABLE 3: CORRECTION TABLE SIDETONE IMPEDANCE

Rsa (Ohm)		Rsb (Ohm)		Cs (nF)	
Value on LCD	Real value	Value on LCD	Real value	Value on LCD	Real value
136	134	465	465	4	5
155	153	636	637	53	55
192	193	709	710	60	58
216	221	802	803	67	69
239	246	892	893	74	76
268	277	1001	1003	83	85
305	295	1257	1259	94	96
340	341	1409	1410	103	105
385	369	1569	1572	115	121
421	443	1770	1773	131	134
491	492	1975	1978	148	145
		2212	2216	164	166
				184	186
				205	207
				230	232
				258	259

TABLE 4: PROGRAMMING TABLES RINGER PARAMETERS

Value	μC port pins		Volume
	RVOL1	RVOL2	
1	0	0	minimum
2	1	0	
3	0	1	
4	1	1	maximum

Value	Melody		
	f1 (Hz)	f2 (Hz)	f3 (Hz)
1	738	826	925
2	800	1067	1333
3	1455	1621	1810
4	1955	2223	2510

Value	Speed	
	Frequency (Hz)	Tone time (ms)
1	7	47.6
2	11	30.3
3	15	22.2
4	20	16.6

Value	Input frequency (at pin T0)		
	Lower limit (Hz)	Higher limit (Hz)	Method
1	20	60	1f
2	40	120	2f

TABLE 5: SETTINGS OF PROGRAMMABLE PARAMETERS.

PRESS KEYS	DISPLAY DEFAULT	SETTING RANGE (with +/- keys)	STEP RESOLUTION
PROG, DC VOLT	dc_u=4.7	3.1 to 5.9 V	0.4 V
DC VOLT	dc_i_0=27.5	See Table 2	
DC VOLT	dc_i_1=30.0	See Table 2	
DC VOLT	dc_i_2=35.0	See Table 2	
DC VOLT	dc_i_3=40.0	See Table 2	
DC VOLT	dc_i_4=47.5	See Table 2	
DC VOLT	dc_i_5=57.5	See Table 2	
DC VOLT	dc_i_6=65.0	See Table 2	
DC VOLT	dc_i_d_G=0dB	1 to 3 dB	1 dB
PROG			
PROG, GAIN MIC	G_mic=15dB **1)**	4 to 25 dB	1 dB
GAIN MIC	dlt=off	off=3.5Vp-p on=2.6Vp-p	
GAIN MIC	G_dtmf=-5dB **2)**	-5 to +15 dB	1 dB
PROG			
PROG, GAIN REC	G_rEc= -6dB **3)**	-19 to +11 dB	1 dB
GAIN REC	hPl=on	on=5.9Vp-p off=2.3Vp-p	
GAIN REC	rFc=off	on=C-load off=R-load	
GAIN REC	G_conF=-25dB **4)**	-25 to 0 dB	1 dB
PROG			
PROG, SET IMP	rA=200	0 to 600 Ohm	100 Ohm
SET IMP	rb=800	0, 600 to 1000 Ohm	100 Ohm
SET IMP	Fp=1915	828 to 5859 Hz, 12kHz	See Ref. 1
PROG			

PRESS KEYS	DISPLAY DEFAULT	SETTING RANGE (with +/- keys)	STEP RESOLUTION
PROG, SIDE TONE	rSA=491	See Table 3	
SIDE TONE	rSB=1257	See Table 3	
SIDE TONE	CS=131	See Table 3	
PROG			
PROG, RING	F=2 S=4 m=4 u=4	1: Volume: u=1 to 4 2: Melody: m=1 to 4 3: Speed: S=1 to 4 4: Frequency: F=1,2	4=max, 1=min See TABLE 4 See TABLE 4 1=20 to 60 Hz (1f) 2=40 to 120 Hz (2f)
PROG			
PROG, LNR	pulse 3-2 dSt=nEu	1: pulse 2: DTMF 3: mark/space=3/2 4: mark/space=2/1 5: dSt=0Ld 5) 6: dSt=nEu	
PROG			

Notes: 1) Gain of the send prog-amp is indicated on LCD. Total gain is 26 dB higher.
2) Gain of the send prog-amp is indicated on LCD. Total gain is 6 dB higher.
3) Gain of the rec prog-amp is indicated on LCD. This is the overall gain with symmetrical drive of the earpiece. In case of single ended drive, the overall gain is 6 dB lower.
4) Gain of the rec prog-amp is indicated on LCD. Total gain is about 19 dB less (see also paragraph 3.3.9.)
5) The setting "dSt=0Ld" is obsolete.

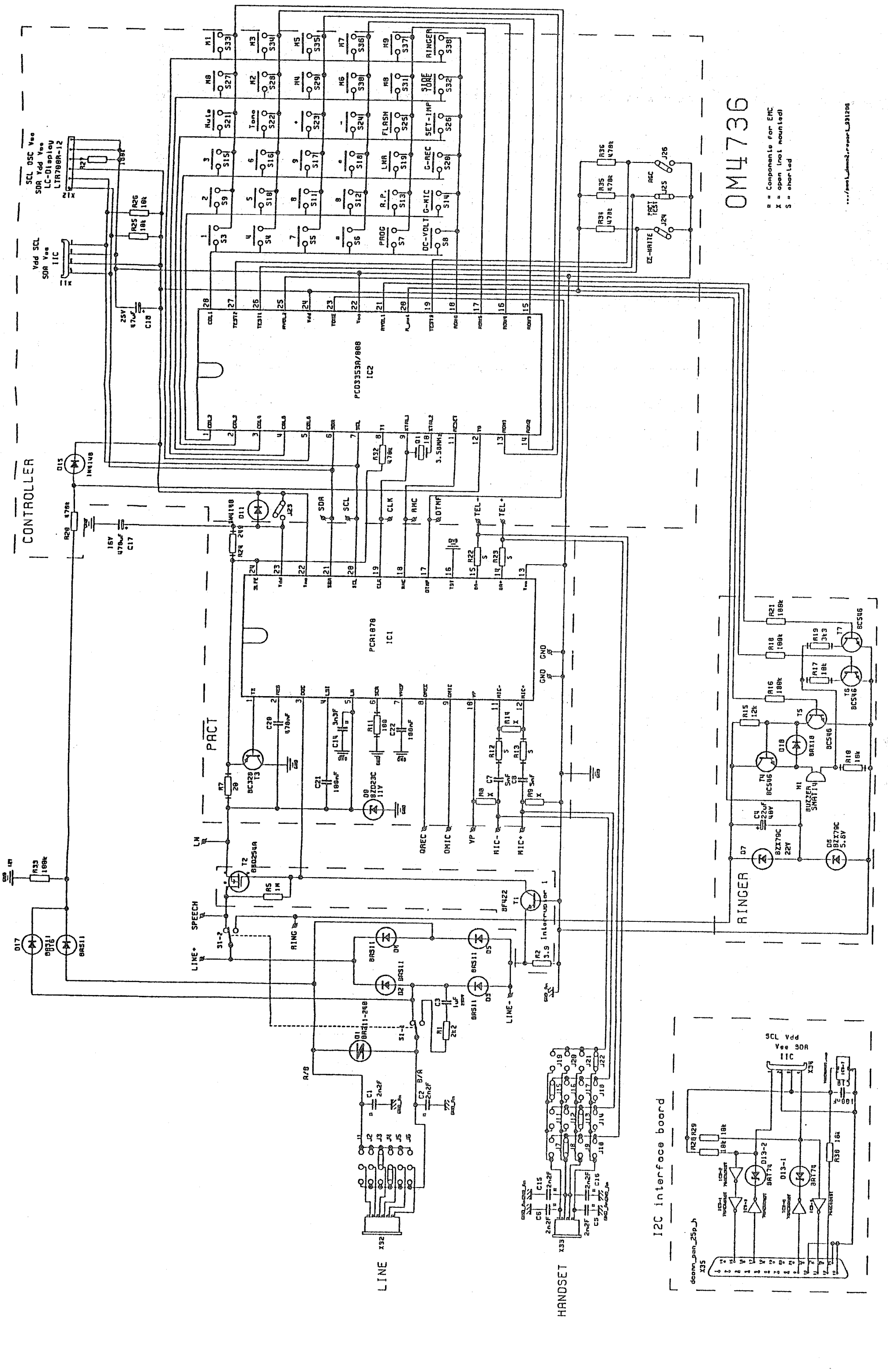

Figure 7 Circuit diagram of feature phone board PR45642 with PCA1070 + PCD3353A/008 and the I^{2}C-interface board

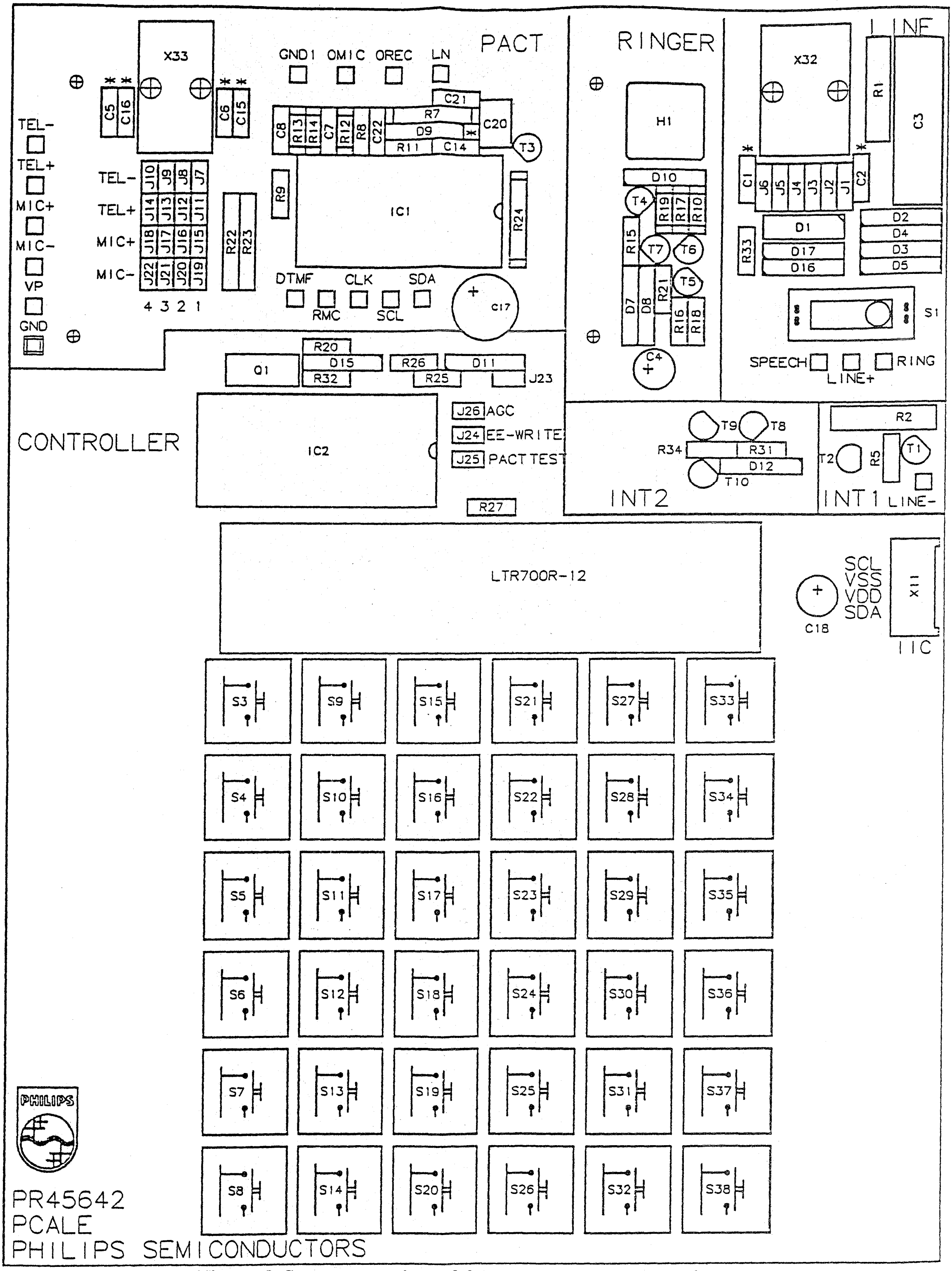

Figure 8 Components view of feature phone board PR45642.

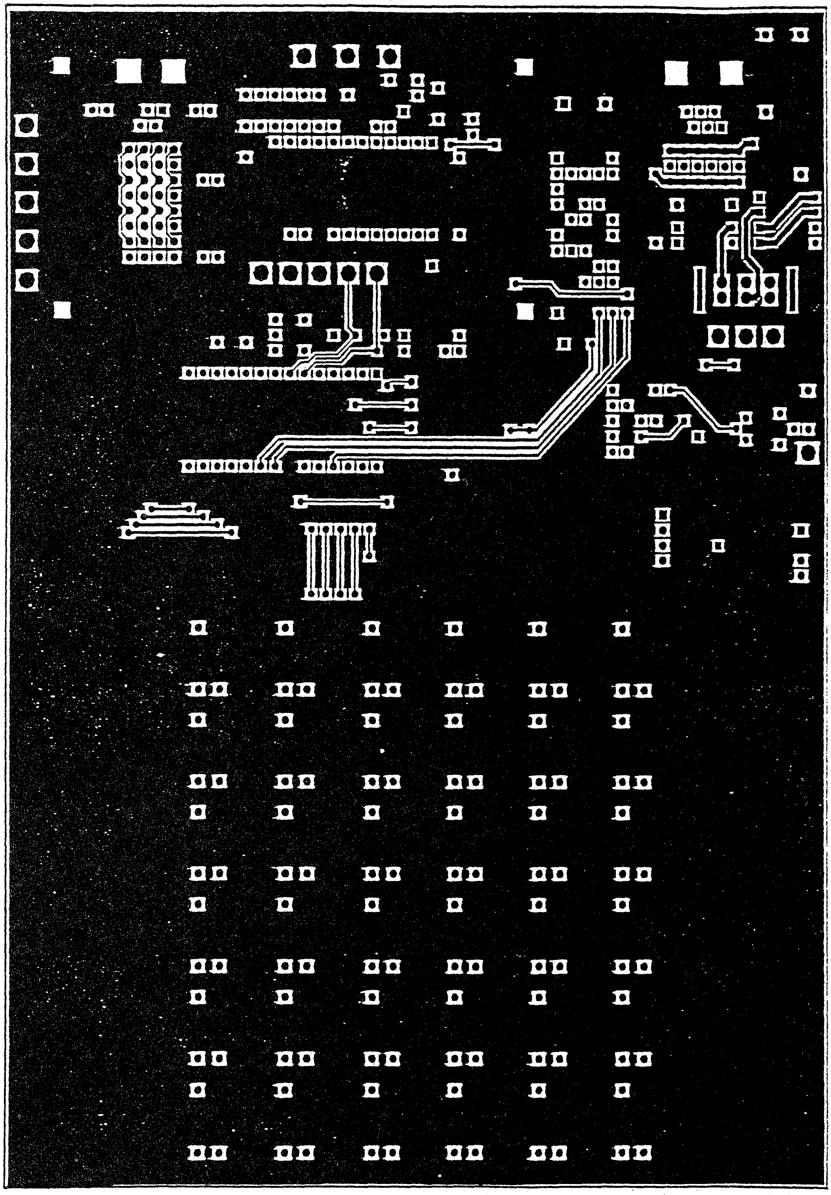

Figure 9 Top copper view of feature phone board PR45642

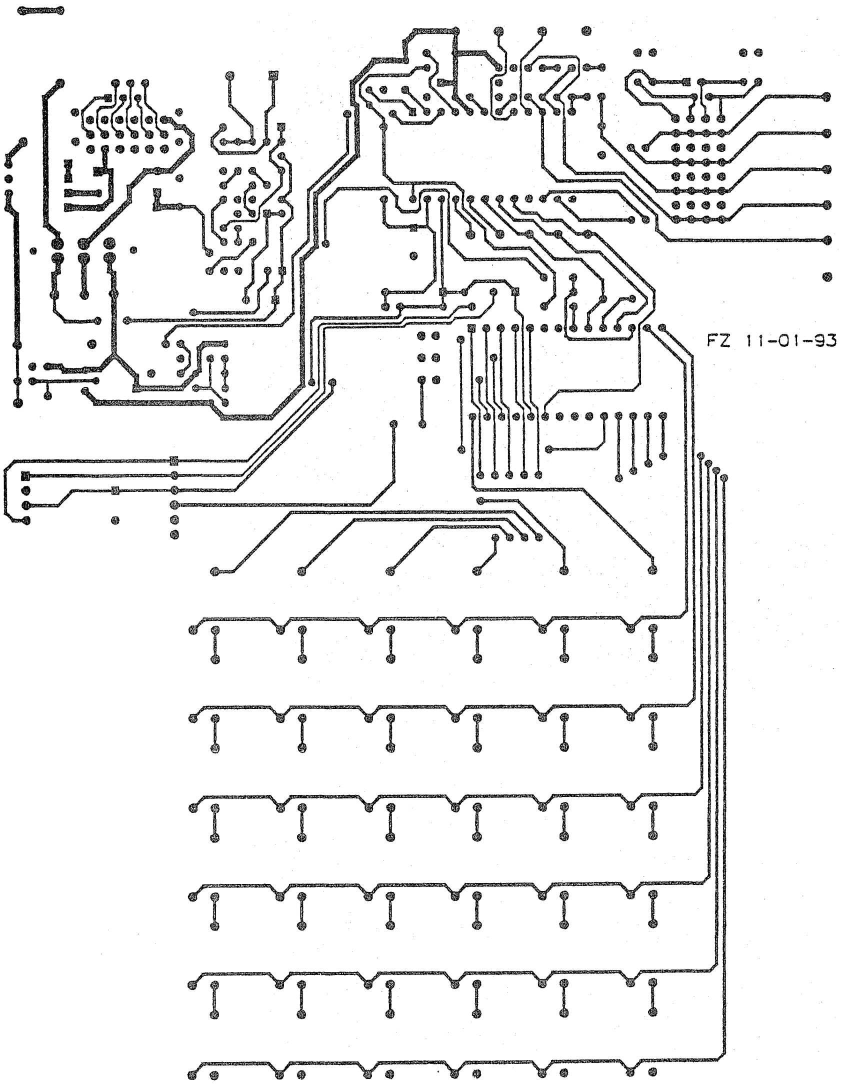

Figure 10 Bottom copper view of feature phone board PR45642.

APPLICATION NOTE Nr	AN94069 (Supersedes ETT/AN94002)
TITLE	High-end telephones with PCA1070, TEA1093 and PCD3755A
AUTHOR	J. C. F. van Loon, P. A. M. v. d. Sande, P. J. M. Sijbers
DATE	September 1994

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APPENDICES

Note: Table of contents of Appendix D can be found on pp: D1, D2.
List of Figures and Tables of Appendix D can be found on pp: D3.

List of figures

1 Introduction

This report describes hardware and software design considerations for a handsfree telephone set with the PCA1070 speech/transmission circuit (programmable via I^2C-bus), the TEA1093 handsfree add-on circuit and the PCD3755A one time programmable (OTP) microcontroller. The Philips PCD3755A OTP supports the PCD335x family of µC. This µC family is recommended for use with PCA1070. The application described in this report has been evaluated on PCALE printed circuit board PR46311 using a preprogrammed PCD3755A. Measurements performed on this board are included in this report.

Appendix A gives the setting range and the default settings of programmable parameters. In Appendix B an overview of the features of the PCD335x µC family is given.
Appendix C gives a list of abbreviations used in this report.
In Appendix D a complete functional description of the SW for µC PCD3755A to control PCA1070 + TEA1093 is given.

Details about PCA1070 can be found in Ref. [1] and [2]. Details about TEA1093 can be found in Ref. [3] and [4]. For details about PCD3755A and PCD335x see Ref. [5] and Ref. [6]. For specification of the I^2C-bus see Ref. [7].

2 Block diagram and application diagram

Fig. 1 shows the block diagram of a high end telephone set with PCA1070, TEA1093 and PCD3755A. Figs. 2 and 3 show the circuit diagram of this application with a preprogrammed μC (the PCD3755A). For a functional description of the software used in the PCD3755A is referred to Appendix D.

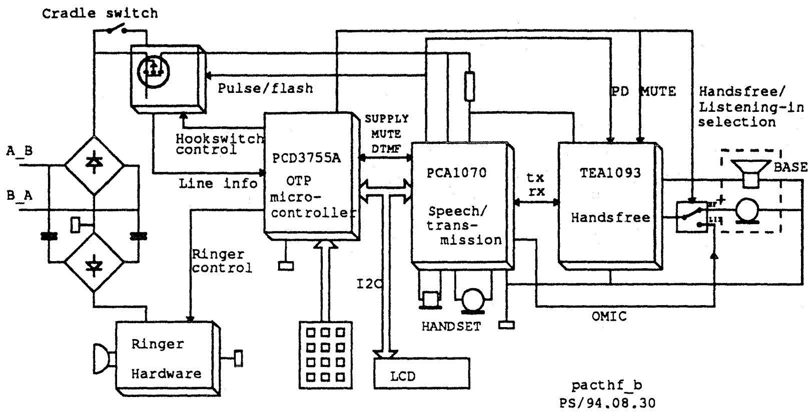


Figure 1 Block diagram of PCA1070 + TEA1093 + PCD3755A application

Functionality of the feature phone:

- Speech/transmission operating modes with privacy switch.
  - Handset (= HS)
  - Handsfree (= HF)
  - Voice switched listening in (= LI1) (anti-howling by means of voice switching)
  - Basic listening in (= LI2) (no anti-howling)
- Ringer detection and generation.
- Dialling features including:
  - Pulse (with programmable "make"-resistance or NSA function), DTMF and Mixed mode dialling.
  - Last number redial.
  - Repertory dialling.
- Settings of programmable parameters are stored in an EEPROM.
- Changing of programmable parameters via keyboard.
- Hold function for switching a call to a parallel phone.

The PCD3755A preprogrammed OTP µC has four functions:

- Control of the normal feature phone functions such as: pulse/tone dialling, redial/repertory dialling, software controlled ringer function, display and operating modes (handset/handsfree/listening-in).
- Setting of the transmission parameters of the PCA1070 in handset, handsfree, both listening-in modes. These parameters are stored in the on-chip EEPROM.
- Setting of the dialling and ringer parameters which are also stored in the EEPROM.
- Changing of all the programmable parameters via keyboard to show the flexibility of the total application.

The hardware aspects of the application are described extensively in Chapters 3, 4, 5 and 6. In Chapter 7 measurement results of this application are given.

For some countries a 12kHz or 16kHz tax pulse counting system is used. For these countries an external tax pulse filter has to complete the design of Figs 2 and 3. Details can be found in Ref. [8]. The application is intended for countries which require voltage regulation for DC termination of the telephone line. For countries (e.g. France) where a DC current regulation must be used in stead of a DC voltage stabilizer, a possible realization is described in Ref. [9].

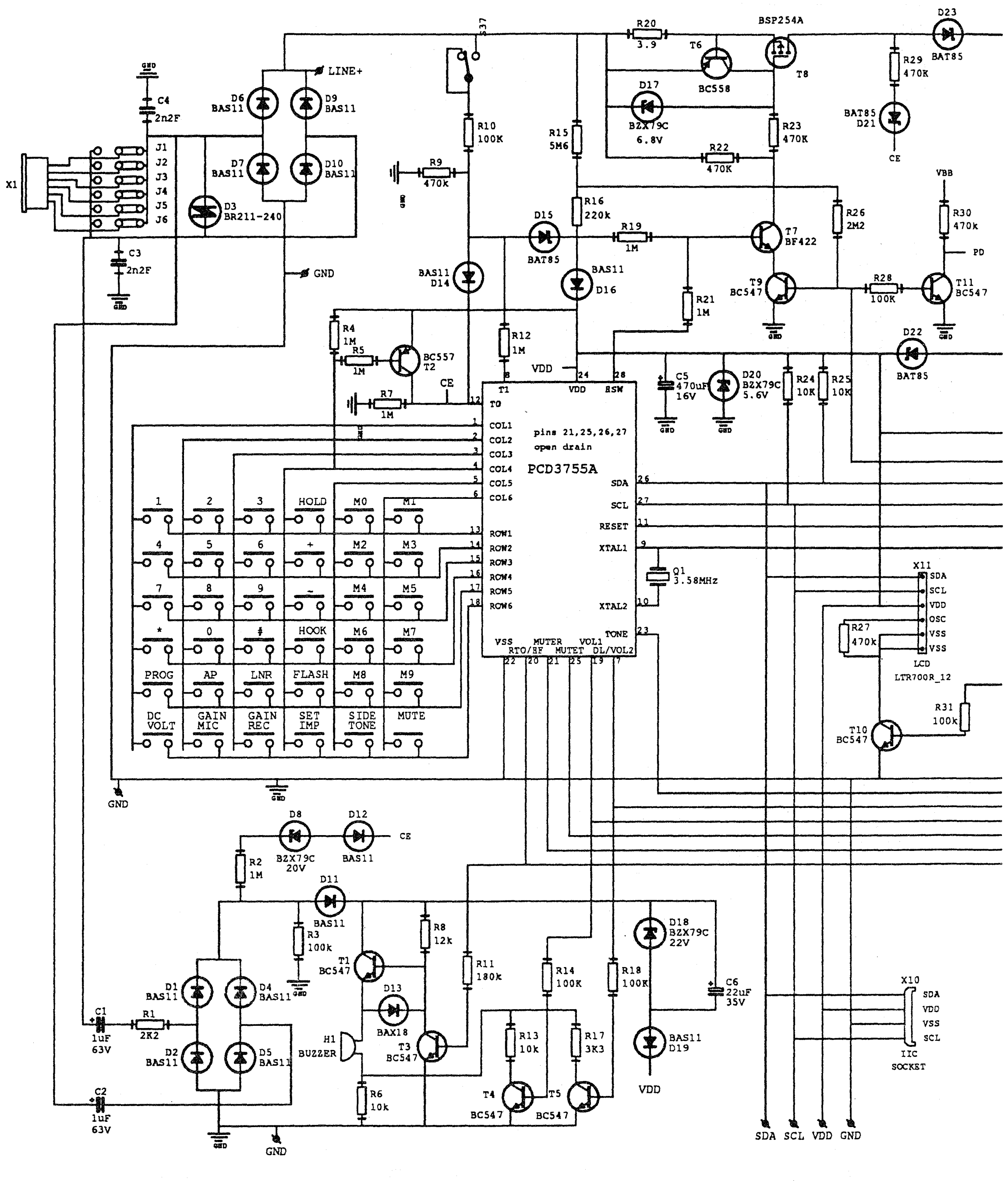
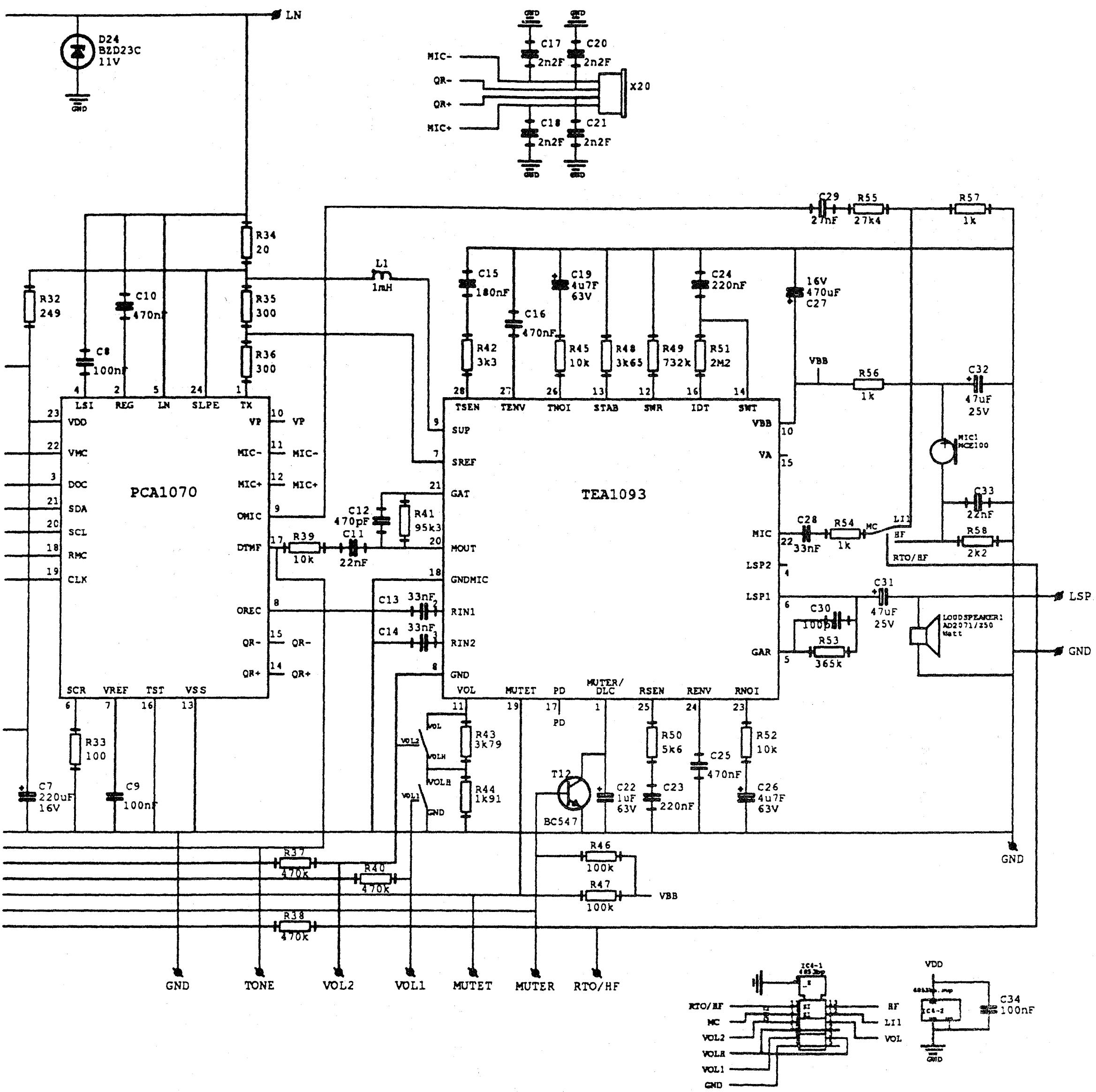


Figure 2 Application diagram PCA1070+TEA1093+PCD3755A (left hand part)



PCD3755A + PCA1070 + TEA1093 APPLICATION
CIRCUIT DIAGRAM
Piet van de Sande
06-09-94

Figure 3 Application diagram PCA1070+TEA1093+PCD3755A (right hand part)

3 Interface between PCA1070 and TEA1093

Details and design considerations of the interfacing of PCA1070 and TEA1093 are described. Throughout this chapter the application diagram of Figs. 2 and 3 is used as a reference.

3.1 Supply management

The principle arrangement of current and supply management of the PCA1070 TEA1093 combination is shown in Fig. 4.

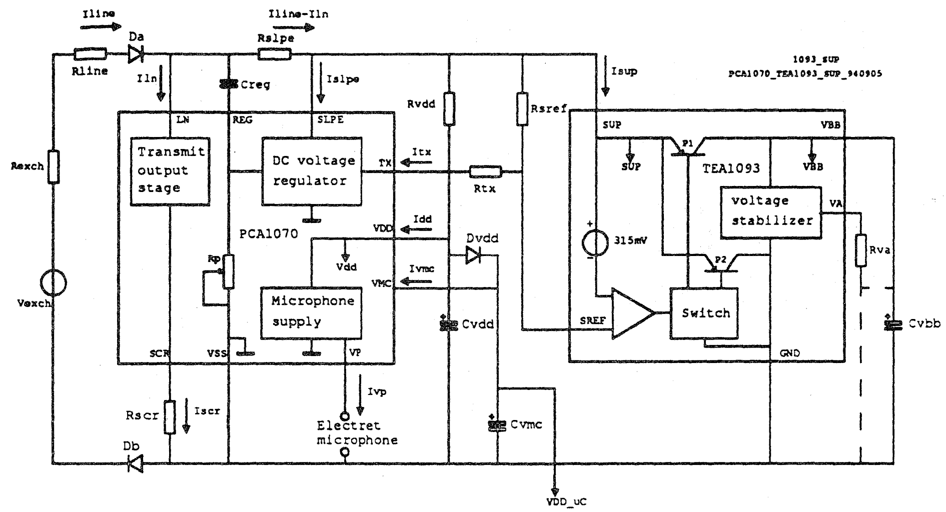


Figure 4 Supply arrangement of PCA1070 and TEA1093

In comparison with the double reference structure of a TEA106x/TEA1093 combination, the PCA1070/TEA1093 combination has only one common ground. This is considered to be a major advantage with respect to interfacing to a micro-controller.
The PCA1070 incorporates a line interface block consisting of a DC and an AC stage (Ref.[2]). The AC stage creates the set impedance and is the output of the transmit channel. It consumes about 3mA of DC current via pin LN if the line current is larger than 16mA typically. Below 16mA, the bias current in the AC stage is automatically reduced to ensure maximum possible transmit output swing under all line current conditions. The AC stage can work with such a low current since it acts as a voltage source with the set impedance in series. In the basic application of the PCA1070, the residual line current flows into the DC stage (via an external pnp which is driven from pin TX) which stabilizes the DC line voltage and powers the IC itself and its peripherals via an RC smoothing filter (Rvdd, Cvdd) (Ref. [2]).

As shown, the TEA1093 is connected between SLPE and TX of the PCA1070. Via resistor Rsref=R35=300Ω, the DC stage of the PCA1070 is biassed at only 315mV/300Ω ≈ 1mA. This current is sufficient for correct operation. Rtx=R36 is added in order to decrease the DC voltage at pin TX. This ensures a correct clipping behaviour of the PCA1070 DC stage at high speech levels. No external pnp transistor is needed at pin TX of PCA1070. The interface structure has virtually no effect on the basic transmission parameters of the PCA1070 in the normal operating range.

The TEA1093 needs a bias current I_{SUP0}=5.5mA typical from SUP and VBB. Excess line current will be shunt to ground via the voltage stabilizer at VBB, or can be used to power the loudspeaker amplifier stage.

The current available to power the loudspeaker amplifier of TEA1093 is:

$$Ilsp = Iline - I_{LN} - I_{SLPE} - I_{TX} - I_{DD} - I_{VMC} - I_{SUP0} - I_P$$

Where:

I_{LN}	=	Bias current of the AC sending output stage PCA1070 (≈3mA if I_{LINE}>16mA typically)
I_{SLPE}	=	Bias current of the DC line interface PCA1070 (I_{SLPE}≈0.35mA)
I_{TX}	=	Bias current in output transistor of DC line interface PCA1070 (315mV/Rsref≈1mA).
I_{DD}	=	Internal supply current PCA1070 (typ. 2.3mA at VDD=2.5V)
I_{VMC}	=	Internal supply current of pin VMC (I_{VMC}≈4μA)
I_{SUP0}	=	Total current consumption of TEA1093 from SUP and VBB (I_{SUP0}=5.5mA typical).
I_P	=	Current to peripheral circuits supplied from VDD, VBB or VP (in this case μC, LCD and electret base microphone)

Taking into account an extra current consumption of I_P = 0.7mA for peripherals like the micro-controller (0.35mA in operating condition (no DTMF tones)), the LCD (assume 50μA) and the base microphone (electret type MCE100; assume 0.3mA), the current Ilsp equals:

$$Ilsp = Iline - 12.85mA \text{ (with } Iline>16mA)$$

Although the PCA1070 reduces its bias current for the AC stage below typically 16mA of line current, no acoustic output signal from the loudspeaker can be expected below line currents of about 12mA. However, handset operation (with relaxed performance) is possible down to about 6mA (see also Chapter 7 for more details).

To guarantee stability of the supply management circuitry under all possible line conditions, an inductor (1mH) in series with pin SUP of TEA1093 is used (see Fig. 3).

3.1.1 DC characteristics

The DC voltage drop over the set is adjustable by programming the voltage at pin SLPE of the PCA1070.

Vab = Vbridge + Vinter + Vschottky + Vdcslope + Vslpe

Vbridge	=	voltage drop over the diode bridge, 1.5V at 20mA
Vinter	=	voltage drop over the interruptor, 0.28V at 20mA (14Ω)
Vschottky	=	voltage drop over schottky diode D23, 0.35V at 20mA
Vdcslope	=	voltage drop over slope resistor R34, 0.34V at 20mA (20Ω)
Vslpe	=	programmable voltage drop of PCA1070; 4.7V nominal (3.1V - 5.9V setting range)

At Iline = 20mA this results in:

Vab = 1.5 + 0.28 + 0.35 + 0.34 + 4.7 = 7.2V

Via the I^2C bus of the PCA1070 this voltage drop can be programmed between 6V up to 8.4V to suit country specific requirements. However the VBB of TEA1093 cannot be programmed. In the application of Figs. 2 and 3 the supply voltage of TEA1093 is fixed at its default value of VBB = 3.6V. For correct operation of the TEA1093 supply structure the DC voltage drop between SUP (=SLPE in this case) and VBB needs to be at least 0.4V. This means that the programming range of V_{SLPE} of PCA1070 is limited to a minimum value of V_{SLPE}=4.3V. Therefore the minimum tip-ring voltage at Iline = 20mA is Vab = 6.8V typically.
In case a lower line voltage is required, then a resistor between pin VA and VBB must be connected to lower VBB of TEA1093 (see paragraph 3.1.2).

The fixed supply voltage of TEA1093 at VBB=3.6V allows an output power into a 50Ω loudspeaker of 20mW at I_{SUP}=16.5mA (Ilsp=I_{SUP}-I_{SUP0}=11mA typically). In this application this means that a line current of Iline = Ilsp + 12.85mA = 23.85mA is needed.
In case more output power of TEA1093 is needed, the supply voltage of TEA1093 must be increased. This can only be done by means of a resistor between pin VA and GND. The maximum permitted value for VBB depends on the programmed voltage setting of PCA1070.

<u>NSA function</u>
In some countries the DC requirements are different for loop (= normal speech) condition and dialling condition. In practice this can mean that during pulse dialling the DC voltage drop of the set must be switched to a lower voltage to suit the PTT requirement for the "make" resistance of the set. This function is known as NSA in Germany or MUTE2 or DMO in other countries. With the PCD3755A SW described in Appendix D this is realized under software control by reprogramming the DC voltage V_{SLPE} during the dialling period. The DST control bit is set to "1" during dialling to ensure short DC settling time.

3.1.2 Optimum DC settings TEA1093

The optimum DC setting of TEA1093 for correct operation of the TEA1093 and allowing maximum power of the loudspeaker amplifier is shown in Table 1.

Programmed DC voltage at SLPE (V)	Optimum DC voltage at VBB (V)	Corresponding value of Rva_vbb, (Ω)	Corresponding value of Rva_vss, (Ω)	Calculated line voltage Vab (V) at 20mA
3.1 (1)	--	--		5.6
3.5	3.0 (2)	110k		6.0
3.9	3.3	220k		6.4
4.3	3.6			6.8
4.7	4.1		100k	7.2
5.1	4.5		56k	7.6
5.5	4.9		39k	8.0
5.9	5.3		27k	8.4

(1) Only to be used to lower the line voltage during pulse dialling (NSA function)
(2) Relaxed performance of TEA1093; lowest advised VBB setting is 3.2V !

TABLE 1 DC settings for maximum power in loudspeaker

In this table a voltage drop of 0.6V between SUP (=SLPE) and VBB of TEA1093 is chosen for Vslpe >= 3.9V. This allows about 200mVpeak of AC signal on SUP for maximum efficiency. In practice this value will hardly ever be reached as an average value. At larger AC swing, the efficiency of the TEA1093 power supply starts decreasing (down to 50% for extremely large signals).

3.1.3 Considerations about automatic adjustment of DC settings

Where the PCA1070 can easily be adapted to country requirements (e.g. EEPROM code µC), the VBB voltage of the TEA1093 needs to be adjusted by hand by soldering a resistor R_{VA-VBB} or R_{VA-VSS} on the board. Therefore the programmability of the PCA1070 cannot be fully exploited (programmable country parameters).

There are two possibilities to realize programmable VBB setting of the TEA1093 (application hints):
- Use simple HEF switches to switch between different Rva_vbb and Rva_vss and control these switches via the micro controller.
- Use an analog control circuit which compares VBB and SUP and controls at VA such that the DC difference between SUP and VBB is in the order of 0.6V.

The slope of the line voltage is in the order of 35Ω (20Ω PCA1070, 14Ω interruptor). Some countries however allow much larger slopes. With the micro controller, such a larger slope can be easily realised by monitoring the line current register in the PCA1070 via I^2C-bus and reprogramming the SLPE voltage depending on line current. The previously mentioned solutions may be used to adapt the supply voltage of the TEA1093 automatically. Another possibility is (application hint):

- Use a resistor between pin VA of TEA1093 and a slightly negative voltage depending on line current (sense resistor//capacitor in negative rail). With increasing line current, more current is drawn from VA resulting in a higher VBB.

3.2 Transmit channel

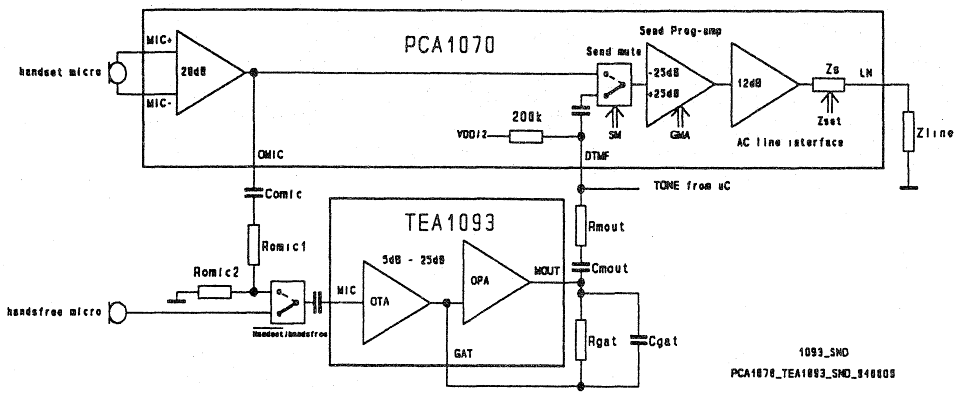


Figure 5 Transmit channel TEA1093 and PCA1070

Handset operation is performed via the amplifiers of the PCA1070 only. For details see Ref. [2]

Handsfree operation is performed via the additional amplifiers in the TEA1093 which interface to the PCA1070. The microphone amplifier output of the TEA1093 (MOUT) interfaces to the DTMF input of the PCA1070. The DTMF input is selected for simplicity since interfacing to MIC+ and MIC- of PCA1070 would require extra switches.

DTMF signals from the micro-controller can still enter the PCA1070 DTMF input directly (DC blocking capacitor inside PCA1070). Rmout=R39 is applied in series with MOUT since this output of the TEA1093 is low ohmic. Signals coming from MOUT are only slightly attenuated by Rmout=R39 and the DTMF input impedance (200kΩ resulting in -0.4dB). The TONE output of the micro-controller switches to a high ohmic state when not active. Capacitor Cmout=C11 blocks DC between the TEA1093 and the micro-controller.

Listening-in operation can be performed in two ways:

- With the handset microphone signal passing the TEA1093, so called voice switched listening-in (LI1 mode). In this case the voice switches in TEA1093 are used to prevent howling (half-duplex).
- With the handset microphone signal directly interfacing to the line via PCA1070, so called basic listening-in (LI2 mode). No anti howling precautions are used here.

The switch in front of the input of the TEA1093 microphone amplifier (MIC) selects either the signal from the handsfree microphone or the signal from the PCA1070 microphone pre-amplifier (OMIC). This switch is controlled by the micro-controller and is combined with the ringer tone output (pin RTO/HF). This combination is possible since both functions are never needed simultaneously. With micro-controller output RTO/HF being high, the handsfree microphone is selected. For voice switched listening-in the signal from OMIC is selected.

The signal from OMIC of the PCA1070 is transferred to the MIC input of the TEA1093 in case of voice switched listening-in operation. The signal from OMIC is first attenuated by Romic1=R55 and Romic2=R57 in order to obtain comparable input levels at MIC of the TEA1093 in both handsfree and voice switched listening-in mode. The amount of attenuation needed is dependent of the microphone types (sensitivity) in the handset and the base, the gain setting of the TEA1093 and the gain settings of the send prog-amp in HS and HF conditions. Comic=C29 in series with Romic1=R55 blocks DC and can be used to form a high pass filter.

3.2.1 Handset and base microphones

For this application example a handset has been used which is specified according to the German requirements (FTZ 121 TR 8, July 1989 and FTZ 12 R 21, January 1989). It has a dynamic microphone with a DC resistance of 250Ω. More details about the handset can be found in Ref.[10]. An electrical gain of ≈48dB is needed to obtain a SLR ≈+3dB.

The base (electret) microphone (e.g. type MCE100 is used here) is supplied from VBB of the TEA1093 via an RC smoothing filter (R56 and C32) and loaded with R58 of 2k2. C33 in parallel with R58 filters out speech signals above 4kHz. C28 forms a first order high pass with the TEA1093 MIC input with a 300Hz cut-off frequency.
The MCE100 microphone has a sensitivity of 0.5mV/μbar with a 2k2 load, which equals 5mV/Pa. Reference vocal level at 25mm (lips) equals 89.3dBspl with a corresponding sound pressure level of 65dBspl at 0.5m of distance (see Ref. [11]). However, while using handsfree, people raise their voices with about 5dB so the sound pressure at the MCE100 microphone will be 70dBspl which equals 63mPa. Thus the microphone produces a voltage of 5mV•63mPa = 315μV. To obtain a line signal of 100mV (rule of thumb value) a microphone gain is needed of 50dB. The TEA1093 gain is therefore set to its maximum value of 25dB. The gain of the PCA1070 from DTMF input to the telephone line is programmable up to 31dB so a maximum gain setting for handsfree operation can be obtained of 56dB.

For voice switched listening-in an attenuation of 29dB between OMIC of PCA1070 and the base microphone input MIC of TEA1093 has been set.

3.2.2 Gain setting

The gain in handset (HS) mode, handsfree (HF), voice switched listening-in (LI1) and basic listening-in (LI2) can be programmed in 1dB steps by means of a programmable amplifier ("send prog-amp") which can be controlled via the I²C-bus.
The "send prog-amp" has a gain setting range between -25 and +25dB. However to prevent overdrive in the transmit channel the practical gain setting range depends on the operating mode (See Ref.[2] for details about maximum signal levels).

Total gain of the handset microphone channel between MIC+, MIC- and LN of PCA1070 is

In HS and LI2 mode:

$$G_{M\_HS} = 32 + Gma + 20\log(Z_{LINE}/(Z_{LINE}+Zs)) \quad [dB]$$

where Gma = Gain "send prog-amp"
Z_{LINE} = load impedance presented to LN
Zs = programmed set impedance

With Zline=Zs this results in:

$$G_{M\_HS} = 26 + Gma \quad [dB]$$

Recommended gain setting range GM_HS is between 30 and 51dB.

In LI1 mode:

$$G_{M\_LI1} = 32 + Gatt1 + Gswitch + Gtx\_1093 + Gatt2 + Gma + 20\log(Z_{LINE}/(Z_{LINE}+Zs)) \quad [dB]$$

where Gatt1 = 20•log(Romic2 / (400 + Romic1 + Romic2)
with Romic2=1kΩ, Romic1=27.4kΩ --> Gatt = -29.2dB
Gswitch = gain of the switch in front of TEA1093 (≈-0.3dB)
Gatt2 = gain between MOUT and DTMF
= 20•log(200k/(Rmout+200k))
with Rmout=10k --> Gatt2 ≈ -0.4dB
Gtx_1093 = 20•log(0.674•Rgat/Rstab)
with Rgat=95.3kΩ, Rstab=3.65kΩ--> Gtx_1093 = 24.9dB

With Zline=Zs this results in:

$$G_{M\_LI1} = 21 + Gma \quad [dB]$$

Recommended gain setting range G_{M_LI1} is between 23 and 46dB.

Total gain between of the base microphone channel between pin MIC of TEA1093 and LN of PCA1070 is

In HF mode:

$$G_{M\_HF} = Gswitch + Gtx\_1093 + Gatt2 + Gma + 12 + 20\log(Z_{LINE}/(Z_{LINE}+Zs)) \quad [dB]$$

With Zline=Zs, Gswitch=-0.3dB, Gtx_1093=24.9dB and Gatt2=-0.4dB this results in:

$$G_{M\_HF} \approx 30 + Gma \quad [dB]$$

Recommended gain setting range GM_HF is between 32 and 55dB.

Total gain of the DTMF channel between pins DTMF and LN

G_{DTMF} = Gma + 12 + 20•log(Z_{LINE}/(Z_{LINE}+Zs))

With Zline=Zs this results in:

G_{DTMF} = 6 + Gma

Recommended gain setting range G_{DTMF} is between +1 and 21dB (for VLN <= 1.6Vp-p).

3.2.3 Muting

In HS and LI2 mode the handset microphone can be muted by setting PCA1070 control bit SM to "1" via I²C-bus. In HF and LI1 mode the transmit channel of the TEA1093 can be muted by applying a logic high at pin MUTET. If the transmit channel of the TEA1093 is muted, automatically the TEA1093 is forced into receive mode. The loudspeaker amplifier of the TEA1093 can be muted by pulling pin DLC/MUTER lower than 200mV.
The correct muting sequences in all operating conditions are performed by the µC (see Appendix D).

3.2.4 Noise in HF and LI1 modes

The noise generated at the output MOUT of the TEA1093 at 25dB of gain equals -90dBmp in transmit mode. With the gain of the PCA1070 send prog-amp gain also set to 20dB (G_{M_HF}=50dB) this will result in a noise level at the line of ≈-64.4dBmp (PCA1070 noise neglected, since much lower). In idle mode the transmit gain of the TEA1093 will drop 23dB (with 46dB switching range) so -113dBmp at MOUT could be expected. However, the TEA1093 has a noise floor at MOUT of about -107dBmp. Thus in idle mode the noise contribution of TEA1093 to the line will be -107dBmp + 25dB = -82dBmp. Therefore at G_{M_HF}=50dB the total noise at the line will be determined by noise of PCA1070 mainly.
In voice switched listening-in mode (LI1) the same noise figures can be expected since the microphone signals from the handset are passed through the TEA1093. Measurement figures for noise at the line can be found in Chapter 7.

3.2.5 Software controlled dynamic limiter

During handset (HS) mode, the dynamic limiter of PCA1070 prevents overdrive of the microphone channel. In LI1 mode and HF mode this dynamic limiter is not active because the DTMF input of PCA1070 is used to transmit the speech signals to the line. If necessary a dynamic limiter which operates under SW control may be used. An external detector is needed then. This is described in Ref. [2]. The SW of the PCD3755A which is described in this report (Appendix D) is prepared for use with SW controlled dynamic limiter. For this purpose pin DL/VOL2 has a double function: sense input for the dynamic limiter or volume control. In case the dynamic limiter function is used, only 2 volume steps can be programmed (pin VOL1).

3.3 Receive channel

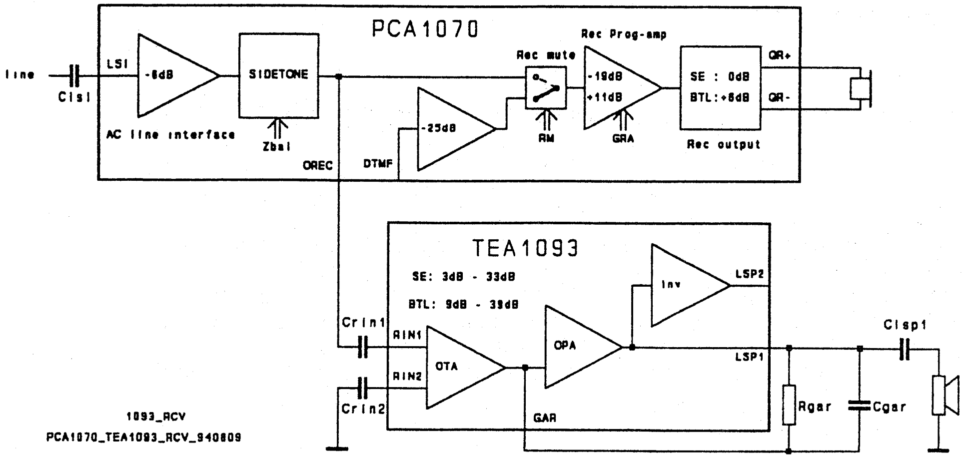


Figure 6 Receive Channel TEA1093 and PCA1070

Handset operation is done via the amplifiers of the PCA1070 only. For details is referred to Ref. [2].

The receive signal for the TEA1093 loudspeaker amplifier is derived from pin OREC of the PCA1070. At this pin receiving signals from the line are present, attenuated with 6dB. The signal at pin OREC is superimposed with residual signals of the internal clock (224kHz) of the PCA1070. These are filtered sufficiently in the loudspeaker amplifier of TEA1093. Crin1=C13 is implemented at the RIN1 input of the TEA1093 to block DC and to form a high pass for frequencies above 300Hz. Crin2=C14 at pin RIN2 is used for reasons of symmetry. Crin2 as well as the pin RIN2 itself of the TEA1093 are not essential any more since there is no shifted reference like in the application with TEA106x.

At the output of the receiver amplifier of the TEA1093 the loudspeaker is connected via a DC blocking capacitor Clsp1=C31. The speaker can be connected as a single ended load (SEL) or bridge tied load (BTL). Due to offset between LSP1 and LSP2, the DC blocking capacitor Clsp1=C31 is essential also in BTL drive. In BTL configuration 6dB more gain and signal swing is available. However, about twice the DC current from VBB of the TEA1093 is needed for the same output signal. In practice, a rule of thumb is that in case of a 50Ω loudspeaker BTL drive becomes interesting at high line currents above 35mA. The application described in this report is intended to realize HF with sufficient performance also at lower line currents. Therefore SEL is used.

3.3.1 Handset earpiece and loudspeaker

A handset has been used which is specified according to the German requirements (FTZ 121 TR 8, July 1989 and FTZ 12 R 21, January 1989). It has a dynamic earpiece with a DC resistance of 250Ω. More details about the handset can be found in Ref.[10].
An electrical gain of -3dB is needed to obtain a RLR ≈ -9dB.

At 0.5m distance of the loudspeaker again the reference vocal level of 65dBspl is wanted at 45dBA of room noise and 70dBspl is wanted at 55dBA of room noise (CCITT recommendation). Adding a 5dB margin to this value for safety and flexibility, 75dBspl at 0.5m is required. The speaker AD2071/Z50 produces 90dBspl at 0.5m at 1kHz while consuming 0.55W. If this speaker is to produce 75dBspl at 0.5m at 1kHz, it will consume 15dB less power meaning 17.4mW. Since the AD2071/Z50 is about 50Ω at 1kHz this represents a voltage swing of 933mV. Assuming a line signal of -20dBm (Ref. [12]): 70dBspl at 0.5m at 440Hz at -20dBm), an overall receiver gain of 21.5dB is needed.

3.3.2 Gain setting

The gain of the handset receive channel can be programmed in 1dB steps by means of a programmable amplifier "rec prog-amp" which can be controlled via the I^2C-bus. The gain of the loudspeaker amplifier cannot be programmed and must be set to a fixed value with Rgar=R53 at TEA1093.

Total gain of the receive channel is in HS, LI1 and LI2 modes between line and earpiece (QR+, QR-) with symmetrical drive (= BTL)

G_{RS_ear} = Gra [dB]

where Gra = Gain of the "rec prog-amp"

Recommended gain setting range G_{RS_ear} is -19dB to +11dB.

Total gain of the receive channel in HF mode between line and loudspeaker (LSP1) with asymmetrical or SEL drive:

G_{RA_lsp} = Grx_1070 + Grx_1093

where Grx_1070 = gain between LN and OREC; -6+20•log(Rrin1/(Rrin1+1000) [dB]
with Rin1=20kΩ (input impedance pin RIN1) --> Grx_1070=-7dB
Grx_1093 = 20•log(0.435 • Rgar / Rstab) [dB]

G_{RA_lsp} = -7 + Grx_1093

With Rgar = 365kΩ, Rstab = 3.65kΩ ---> G_{RA_lsp} = 25.8 ≈ 26 [dB]

This means that even an extra safety margin (4.5dB) with respect to the required value of 21.5dB is obtained.

3.3.3 Noise

The noise in receiving direction of PCA1070 may be represented by an equivalent noise voltage source of ≈-76dBmp[600Ω] at the line. This results in ≈-83dBmp at pin OREC (with 20kΩ load at pin OREC). The equivalent input noise level of the loudspeaker amplifier of TEA1093 is about -98dBmp which is negligible w.r.t. the noise coming from the line. In handset mode the noise level at the earpiece will be -79dBmp (with GRS_ear = -3dB). In handsfree receive mode the total noise level at the loudspeaker will be ≈-50dBmp (with Grx_1093=+33dB, G_{RA_lsp} ≈ 26dB. In handsfree idle mode the noise at the loudspeaker will be reduced because the voice switch reduces the gain of the receiving channel with the half the switching range (Gswr = 46dB in this case). A noise level of ≈-73dBmp can be expected at the loudspeaker.

3.3.4 Digital volume control

Digital volume control (better keyboard controlled volume setting) is performed at pin VOL of the TEA1093 via resistors R43 and R44 and the switches connected in parallel with the resistors. Maximum volume occurs if both switches are conducting connecting pin VOL to GND of the TEA1093. Every 950Ω increase of series resistance connected to pin VOL corresponds to 3dB volume reduction. With R44 of 1k91 a step of 6dB can be made, with R43=3.79kΩ 12dB. In total four levels of volume are adjustable: 0dB, -6dB, -12dB, -18dB. By changing R43 and R44 other step sizes can easily be realised. Under all volume control conditions, TEA1093 keeps the sum of gain for the transmit and receive channel constant. This means that the switching range is automatically adapted to the right level.

3.4 Handsfree settings

Switching range

This paragraph describes the basic handsfree adjustment mechanism for the TEA1093 for good switching behaviour. Not all details are mentioned since they are already described in the TEA1093 application report (Ref. [4]).

The total loop gain in handsfree mode can be expressed as follows:

$$Gloop = G_{M\_HF} + Gstb + G_{RA\_lsp} + Gac - Gswr$$

where $G_{M\text{-}HF}$ = gain of base microphone channel (set to 50dB)
Gstb = gain of sidetone bridge (= minus sidetone suppression)
G_{RA_lsp} = receive gain between line and the loudspeaker (26dB)
Gac = electro acoustic coupling loudspeaker and microphone (-40dB)
Gswr = switching range of the TEA1093

The sidetone suppression is strongly dependent on the line length and the programmed sidetone parameters within PCA1070. For this paragraph a minimum sidetone suppression is assumed of 10dB (Gstb = -10dB) (see also Chapter 7). If automatic gain control is used or multiple anti sidetone by line current control (see Ref. [4] and [2], this figure should be adapted.

The electro acoustic coupling is strongly dependent on the distance between microphone and loudspeaker and the cabinet used. For the PR46311 evaluation printed circuit board a maximum coupling of about -40dB has been measured (see Chapter 7).

So, now the loop gain can be calculated:

Gloop = 50 - 10 + 26 -40 - Gswr = 26dB - Gswr.

In practice, the loop gain should be far below 0dB in order to have sufficient howling margin. On the printed circuit board PR46311 a switching range of 46dB is set which provides a minimum howling margin of 20dB. The switching range is set by the ratio of R49 and R48 at pin SWR and STAB respectively (R48 has a fixed value of 3k65).

Dial tone detector level

Now that the loop gain is known, sensitivities can be set. First step is to select a value for Rrsen=R50 at RSEN of the TEA1093. This resistor sets the receive sensitivity and the dial tone detector threshold level:

Vdialtone = 12.7μA * Rrsen

With Rrsen set to 5k6, a dial tone detector level at RIN1 of 70mVrms applies and this corresponds to 140mVrms at the line.

Transmit sensitivity

For the transmit sensitivity (Rtsen=R42) the following applies (see Ref. [4] for more details).

$$20\log(Rtsen) = 20\log(Rrsen) - G_{M\_HF} - Gstb - Grx\_1070 + Gtsen + 1/2\ Gloop$$

where Gtsen = internal gain from MIC to TSEN of TEA1093 (fixed to 40dB)

With G_{M_HF} = 50dB; Gstb = -10dB; Grx_1070 = -7dB; Gloop = -20dB

$$20\log(Rtsen) = 20\log(Rrsen) - 50 + 10 + 7 + 40 - 10 = 20\log(Rrsen) - 3dB$$

With Rrsen 5.6kΩ, Rtsen = 3.96kΩ. A practical value for Rtsen = 3.3kΩ is selected. The capacitors in series with Rtsen and Rrsen block DC and form a high pass filter. Crsen=C23 is set to 220nF resulting in a cut off frequency of 130Hz. Ctsen=C15 is set to 180nF resulting in a cut off frequency of 268Hz. The transmit channel sensitivity has a higher cut off frequency since low frequency room noises (tapping fingers on the table etc.) should not influence the switching behaviour.

Timings

The signal envelope timing, noise envelope timing, switchover timing and idling timing have been set according to recommended values as given in Ref. [4].

Idle mode

With Rtnoi=R45=10kΩ and Rrnoi=R52=10kΩ the speech/noise threshold of both channels is raised from 4.3dB up to 7.6dB. This improves steady idle mode selection (see Ref. [4]).

4 Interface for the PCD3755A micro controller

This paragraph describes the interface for the PCD3755A OTP micro controller. The interfacing can also be used for the PCD335x µC (PCD3351A/52A/53A) family when the I/O options are defined in the same way as PCD3755A. By defining some I/O pins differently some pull-up resistors may be saved.

4.1 PCA1070 and TEA1093 control

The PCD3755A is supplied either from VDD of the PCA1070 via a schottky diode D22 in case of off-hook or from the telephone line via a high ohmic resistor (R15) in case of on-line. The PCA1070 monitors the VDD voltage of the PCD3755A and provides a RESET signal in case VDD has become too low (e.g. set unplugged).

The PCD3755A controls the PCA1070 via the I²C bus. Parameters like DC setting, gain setting, impedance setting etc. are transferred from the EEPROM of the µC immediately after hook-off.

The extra socket X10 can be used to program the PCA1070 with a personal computer (see Ref. [13] and [14].

The PCD3755A controls the TEA1093 via normal I/O wires (parallel). For transmit and receive mute an open drain structure at the PCD3755A is chosen (4 open drains are available at pins 21, 25, 26, 27). For transmit mute a simple pull-up (R47) from MUTET of the µC to VBB of the TEA1093 is used. For receive mute an extra transistor T12 is used to force pin DLC/MUTER sufficiently low in mute condition. Also a pull up is applied (R46) from MUTER of the µC to VBB of the TEA1093.

Volume control is achieved via the VOL1 and DL/VOL2 pins of the PCD3755A (if pin DL/VOL2 is configured as VOL2 output). Via an analog switch of the HC4053 type R43 and/or/nor R44 can be connected to pin VOL of the TEA1093. The resistors in series with the VOL1 and DL/VOL2 pins of 470k are essential since these pins are combined with the volume setting of the ringer. Without these resistors, protection diodes inside the HC4053 short circuit the volume control wires to VDD of the PCA1070 during ringing. Same holds for the RTO/HF pin of the PCD3755A.

The diode D23, which is in series with the telephone line, is meant to assure a proper operation of CE/TO whenever a line interruption from the central office occurs. Of course this diode, although from the Schottky type, adds a voltage drop to the DC setting of the total application.

4.2 LCD display

The socket X11 connected to the I²C bus wires is meant for an LCD display. A chip-on-glass Liquid Crystal Display module Type LTR700R-12 (see Ref. [15]) has been chosen for this application.
The negative supply line is switched off by transistor BC547 in case the supply voltage of the PCA1070 is low. This prevents that current is lost in on-hook condition.

4.3 Electronic hookswitch and on-hook dialling

HOOK-OFF PROCEDURE:
There are two methods to start-up the application on PR46311:
- by lifting the handset (cradle contact S37)
- by pressing the key 'HOOK'

- If the handset is lifted switch S37 will close and pin T0 (=CE/T0) and T1 of the PCD3755A will go high. At the same time, T7 is switched on which activates the hookswitch. The transistor in series with T7, T9, is controlled by the PCA1070 pin DOC and is normally on (open drain). With T0 being high the PCD3755A will start-up, scan its I/O's and notice that pin T1 is high meaning that the handset was lifted. Pin HSW (28) of the PCD3755A is pulled high as soon as the PCD3755A is fully operational. This way of starting up the set is guaranteed even when the supply voltage of the PCD3755A is low (e.g. just plugged in the set). Of course the start-up time will be longer in this case.

- If key 'HOOK' is pressed transistor T2 starts conducting and will pull T0 (CE) high. The PCD3755A will start-up, scan its I/O's and notice that the key was pressed. Immediately after start-up the µC will pull pin HSW high which turns on the interruptor switch. T0 will be kept high via R29 and D21 after releasing the 'HOOK'-key. Of course, when VDD of the PCD3755A is low, this feature will not work.

ON-HOOK RESISTANCE:
Via resistor R35, which is connected directly to the positive line terminal, a small current will flow keeping the supply voltage VDD of the PCD3755A at about 4V in on-hook condition. All components connected to VDD should be in power down to save current consumption. All I/O pins with port option "standard I/O" to which an external load is connected must be set low (by the SW) to prevent that current is lost. In this case this applies to pins VOL1,DL/VOL2 and RTO/HF.
The current which is available from the telephone line in on-hook condition is dependent on country requirements. In Germany less than 12µA (5MΩ at 60V) may be drawn from the telephone line. With R35 set to 5M6 this requirement is easily fulfilled. As soon as the set is in normal speech mode, the PCD3755A is powered from VDD of the PCA1070 via diode D22.

PULSE DIALLING:
The structure of T7 and T9 forms an AND function. Both T7 and T9 must be on to switch on the interruptor. The reason for choosing this structure is that pulse dialling can still be performed via the PCA1070 (I^2C-bus). This enables a correct timing of power down of PCA1070 and TEA1093 and line interrupt (must be synchronous). In order to power down the TEA1093 an extra transistor is connected to the DOC pin of the PCA1070. This transistor directly interfaces to the TEA1093 PD pin.

HOOK-ON PROCEDURE IN HS MODE:
After switching the cradle switch (S37) into on-hook position (or unplugging the set), transistor T7 will switch off and the electronic hookswitch T8 will be disabled. Then the voltage at pin T0 will change from "1" to "0". The µC will recognize a line-break and will switch PCA1070 into power down mode via I^2C-bus (PD=01). If T0 stays "0" longer than the

reset-delay time (typ. 160 ms), the µC will switch PCA1070 in normal operating mode via I^2C-bus (PD=00) and the µC will go into standby mode. So the supply decoupling capacitor at VDD of PCA1070 will be discharged fast until the internal reset of PCA1070 takes place. The internal reset signal will change from "0" to "1" when VDD passes the threshold (1.2V+/-0.2V) and the PCA1070 will go into reset mode (line interface part in power down and all programmable parameters are reset to default values).

HOOK-ON PROCEDURE IN HF MODE:
In HF mode the cradle switch (S37) is in on-hook position. When key "HOOK" is pressed, the µC will switch pin HSW to "0" and the electronic hookswitch will be disabled.
The PCA1070 will stay in normal operating mode and the µC will go into standby mode. So the supply decoupling capacitor at VDD of PCA1070 will be discharged fast until the internal reset of PCA1070 takes place. The internal reset signal will change from "0" to "1" when VDD passes the threshold (1.2V+/-0.2V) and the PCA1070 will go into reset mode (line interface part in power down and all programmable parameters are reset to default values).

4.4 Ringer control

A software controlled ringer concept is used in this application. If a ring voltage occurs at the telephone line, capacitors C1 and C2 block DC and the diode bridge consisting of D1,2 and D4,5 rectifies (double) the AC ring signal. The ringer part is high ohmic for speech signals with amplitudes in the normal range. The double rectified sine wave after the diode bridge is used to create a chip enable signal for the PCD3755A. Diode D11 and the resistor of R3=100kΩ to GND ensure that this chip enable signal can be used by the PCD3755A for frequency determination (anti tinkle during pulse dialling). After diode D11 zener diode D18 and the zener diode at VDD of the PCD3755A D20 clip the double rectified signal to 27V peak. The capacitors across these diodes smooth the waveform such that a suitable DC voltage of 27V occurs. The diode D18 (BAS11) in series with the zener diode D17 ensures that there is no feedback from VDD of the PCD3755A to the ringer supply during on-hook (otherwise current might be lost).

During ringing, T0 goes high and diode D14 (between cradle and T0) prevents that the hookswitch is turned on. Since transistor T7 needs only a little current and voltage at its base to turn on, diode D14 must be a low leakage type (not Schottky).

The PCD3755A generates the ringer tones at pin RTO/HF if the frequency of the ringer signal is within the pre-programmed frequency range. Via the RTO/HF output of the PCD3755A the output stage for the PXE buzzer, consisting of T1, D12, T3 and R8, is driven. Via VOL1 and DL/VOL2 pins of the PCD3755A transistors T4 and T5 can be used for volume control. Of course pin 7 must be configured as VOL2 output (see Appendix A).

5 E.M.C

In comparison with the double reference structure of a TEA106x/TEA1093 combination, the PCA1070/TEA1093 combination has only one common ground. This is considered to be a major advantage with respect to EMC performance and makes p.c.b. easier.

Basically the measures described in Ref. [2] for "basic protection" are used in this application. The layout of the printed circuit board PR46311 has been designed with respect to electromagnetic compatibility (EMC). Basic protection by means of an "RF-guard" as described in Ref. [2] is provided and it consists of discrete (ceramic) capacitors to suppress the unwanted r.f. signals before they can enter the circuit. Capacitors used are C3, C4 at the a/b line connections, C17, C18, C20, C21 at the handset in/outs.

The EMC capacitors at the a/b terminals (C3, C4) slightly influence the balance return loss and the sidetone characteristics at higher frequencies.

The LINE connector X1 and the Handset connector X20 are placed close together and the EMC-capacitors are mounted close to these connectors.

6 Protection

The transmission IC can be destroyed by excessive current surges on the telephone lines if no proper measures are taken.

A break over diode is needed to protect the circuit against spikes. This break over diode (BR211-240) is connected between the a,b lines in front of the polarity guard.

According to the PCA1070 specification Ref. [1] the voltage on most of the pins may not exceed the 7V. Some pins are allowed to have 12V.

The voltage on LN (also LSI, REG, SLPE, TX and DOC) may not exceed 12V. To prevent this a voltage regulator diode (BZD23C,11V) is connected between LN and VSS. This zener has also a function during start-up at high line currents (>40mA). Pin DOC will not exceed 0.7V as transistor T9 will limit this voltage at its base.

In this application with a P-most (BSP254A) used for pulse-dialling and line-interrupt, resistor R20 in the LINE+ line measures the current. If this current exceeds a limit determined by 0.7V/R20, the extra transistor T6 pulls the voltage on the gate of the P-most up. This current limiter is only intended to provide protection against current surges and is not to be used (for reasons of power dissipation) for continuous limitation of line current.

In ringer condition the zener D20 of 5V6 is applied to the VDD of the µC. This zener in combination with the Schottky diode D21 also limits the supply voltage of the PCA1070.

7 Measurement results

In this Chapter measurement results of the application of Figs. 2 and 3 are given. The evaluation board PR46311 was used for this purpose. All measurements were performed with a preprogrammed PCD3755A OTP μC (SW as described in Appendix D).

The settings of the programmable parameters are according to Appendix A unless otherwise noted. Test conditions and component values are as follows (unless otherwise noted): I_{LINE}=20mA, V_{SS}=0V, f=1000Hz, I_{VP}=0mA, Tamb=25°C, Z_{LINE}=220Ω+(820Ω//115nF). The handset Ref. [10] and base microphone MCE100 are connected.

7.1 DC characteristics

Fig. 7 shows measurements of the DC voltage at the telephone line versus line current with the programmed SLPE voltage as a parameter. To fulfil country specific DC mask requirements, this plot can advise to select a specific SLPE voltage.
Below 6mA of line current the low voltage region is entered. In this region performance of the total set is relaxed to allow parallel operation. Of course, handsfree operation is not possible in this region. As described in paragraph 3.1, handsfree operation is possible down to about 12mA.
Lower SLPE settings are easily programmed (Vslpe=3.9V, 3.5V or 3.1V) but the fixed VBB setting of the TEA1093 (at 3.6V) limits Vslpe to 4.3V minimum (see paragraph 3.1.1).

Fig. 8 shows different DC voltages occurring in the set for a programmed SLPE voltage of 4.7V and a VBB of 3.6V. In this plot it is clear that the voltage drop between Vln and Vab is about 2V at 20mA. If it is necessary to decrease this voltage drop, diode D23 (BAT85) can be skipped (this however makes a different design of the CE sense circuitry necessary), a different interruptor can be used or the diode bridge may be replaced by an active polarity guard.

Figs. 9 and 10 show the current management of the application. Iscr represents the bias current in the AC stage (≈I_{LN}). This current is measured by measuring the voltage drop across the 100 ohm resistor (R33) connected to SCR. The actual current flowing into pin LN (DC bias of AC stage) however is a little higher (few %) but the curve gives a good impression of the behaviour below 16mA line current.
Current Irvdd represents the current through resistor Rvdd=R32 which is connected between LN and VDD of the PCA1070. This current represents the current consumption of the PCA1070 + PCD3755A + LCD. In Fig. 10 it can be seen that at higher Vslpe settings, meaning a higher VDD, the current Irvdd is higher.
Irslpe represents the current flowing through the slope resistor Rslpe=R34 of 20 ohm. It can be noticed that the slope of this current changes at Iline=16mA. This of course is due to the reduction of current consumption of the AC stage below 16mA.
Isup represents the current flowing into pin SUP of the TEA1093. Also this curve changes its slope around 16mA for the same reason Irslpe does. If Isup exceeds the current consumption of the TEA1093 (5.5mA typ.) the handsfree part is fully functional. In the plots this happens around 12mA.

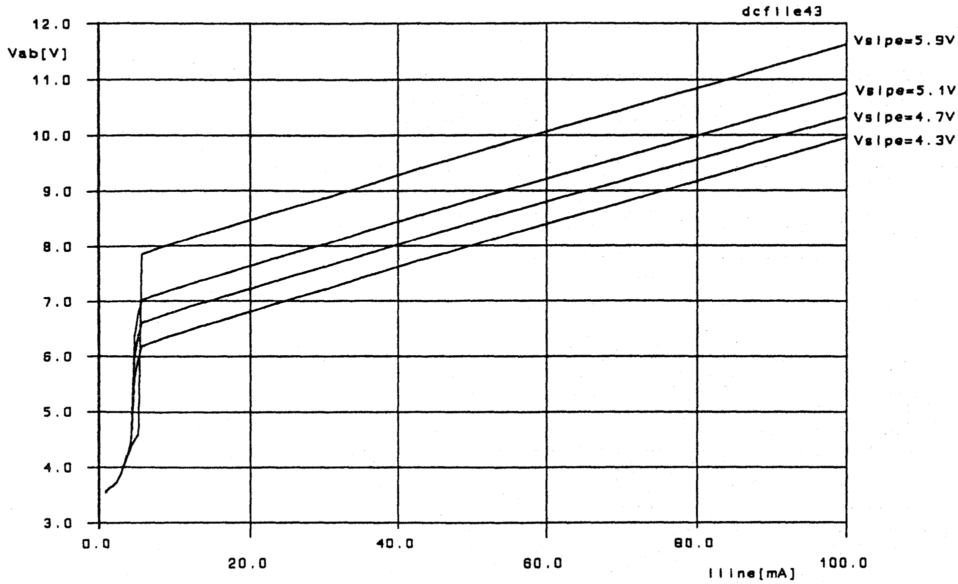


Figure 7 Line voltage versus line current, parameter programmed Vslpe

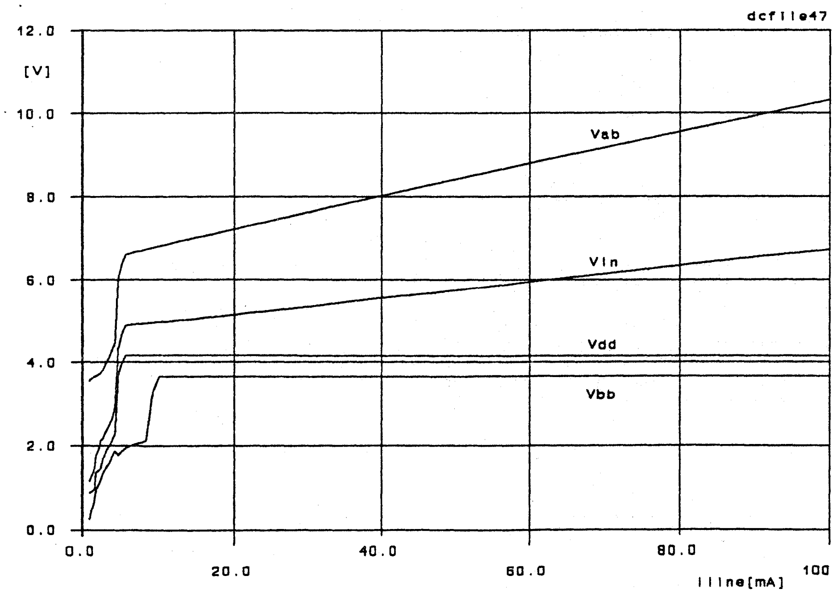


Figure 8 DC voltages versus line current with Vslpe=4.7V; VBB=3.6V

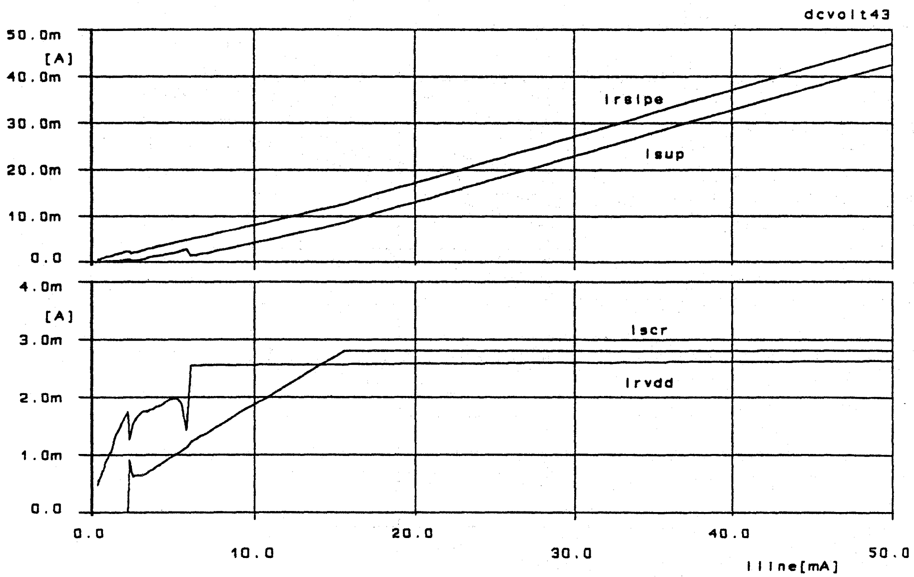


Figure 9 DC current management versus line current with Vslpe=4.3V;VBB=3.6V

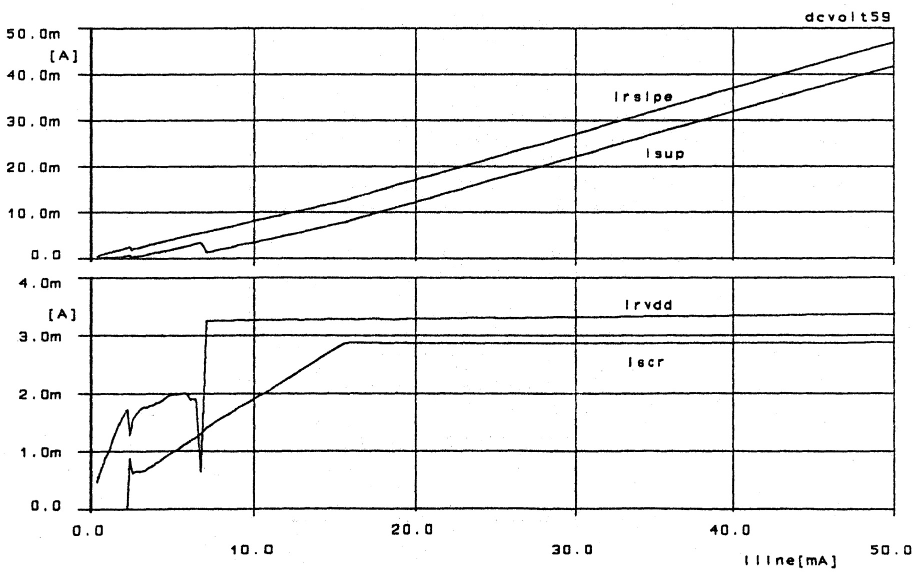


Figure 10 DC current management versus line current with Vslpe= 5.9V; VBB=3.6V

7.2 Gain setting and frequency shaping

Fig. 11 shows the gain of the transmit channel for HF mode as a function of frequency. The input signal has been applied across R58. Output voltages have been measured at pin MOUT of the TEA1093, pin DTMF of PCA1070 and the a/b line. The TEA1093 has been forced into TX mode by muting the RX channel (MUTER=high). The cut-off frequencies (-3dB) are at about 215Hz and 3.9kHz which covers the telephony frequency band. Note that the PCA1070 has no high frequency roll-off with the complex line load (ZD in this case). This is due to the voltage source type output stage.

Fig. 12 shows the overall HF transmit gain. By programming the send prog-amp gain of the PCA1070 in steps of 5dB the different curves were obtained.

Fig. 13 shows the gain of the handset transmit channel (at various points) in voice switched listening-in mode (LI1) between MIC+, MIC- of PCA1070 to the a/b line. Cut-off frequencies are about the same as in handsfree mode.

Fig. 14 shows the gain of the handset microphone channel between MIC+,MIC- of PCA1070 and the line versus gain setting of the send prog-amp.

Fig. 15 shows the receive path from the telephone line towards pin OREC of the PCA1070. Also the overall receive gain from the telephone line to the loudspeaker is shown with different volume settings (6dB steps). The curves has been measured with the TEA1093 forced into receive mode (MUTET=high). The upper curve has been measured at maximum volume setting (VOL1=1, VOL2=1). Due to the resistance of the 4053 switches the maximum gain is some 2dB reduced with respect to the maximum possible value when pin VOL of TEA1093 would be shorted to GND. The frequency characteristics fully covers the telephony frequency band. Roll-off frequencies are 218Hz and 4.6kHz.

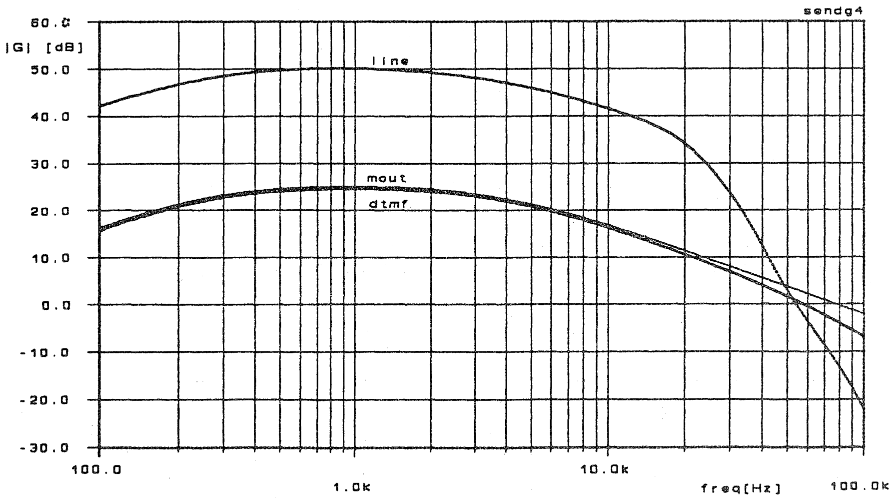


Figure 11 Gain of the HF transmit channel versus frequency (MIC of TEA1093 to MOUT, DTMF and line)

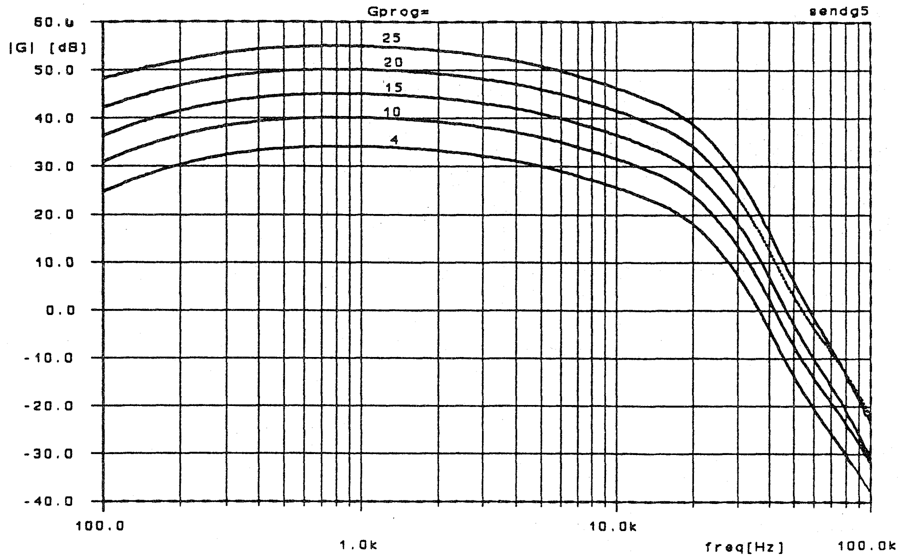


Figure 12 HF transmit gain from MIC of TEA1093 to telephone line (parameter is gain of send progamp)

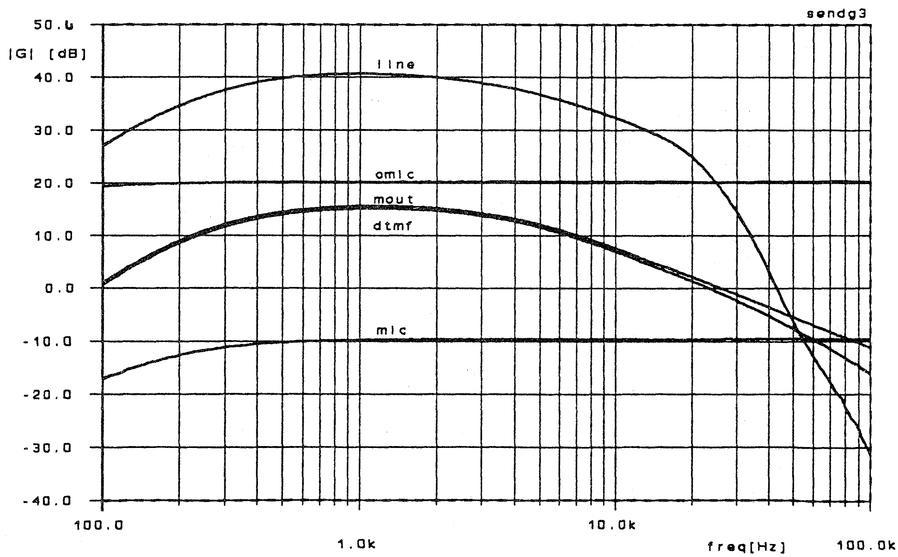


Figure 13 Gain transmit channel from MIC 1070 to line in LI1 mode (set to 41dB with G_{MA}=20dB)

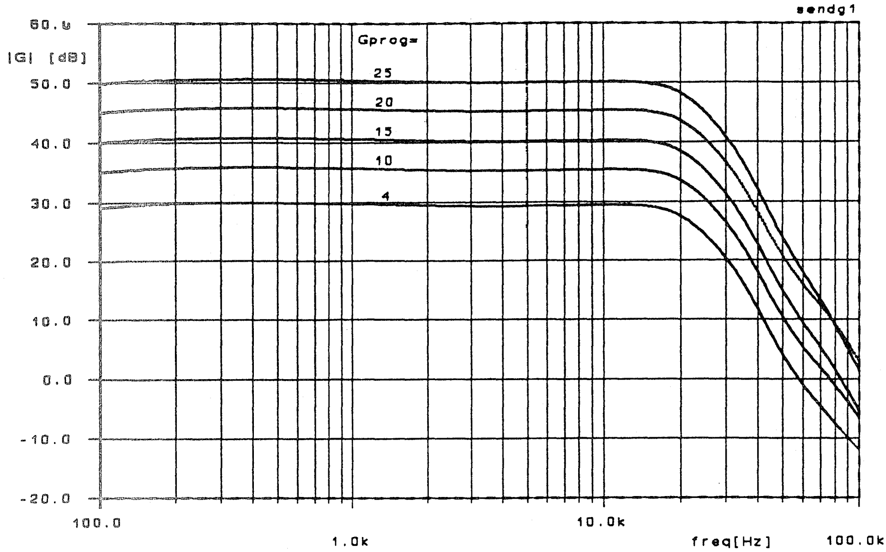


Figure 14 Gain of the HS transmit channel (MIC+, MIC- from PCA1070 to line; parameter is gain of send prog-amp)

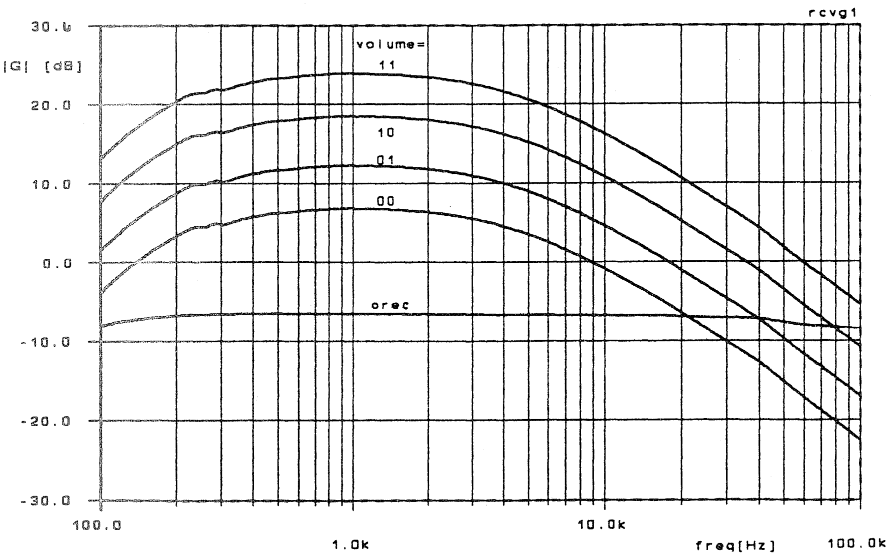


Figure 15 Gain of the receive channel gain from line to loudspeaker (with different volume settings)

7.3 Balance return loss

To eliminate the influence of the transducers in the handset on BRL, they have been replaced by 150Ω resistors during the measurement.

In Fig. 16 the BRL is shown with a programmed set impedance Zs=200Ω+(800Ω//104nF) according to the German requirements. The reference impedance is Zref=220Ω+(820Ω//115nF). Because EMC capacitors are used in this application at the a and b lines, the BRL at high frequencies is somewhat lower than for the PCA1070 alone. However BRL requirements are met with quite a large margin. The influence of TEA1093 on overall BRL is negligible.

In Figs. 17 and 18 the BRL is shown with a programmed set impedance of 600Ω. The reference impedance is Zref=600Ω. Fig. 18 shows a reduced BRL at higher frequencies, but BRL requirements of up to 14dB are met. Although Ra=600Ω and Rb=0 yields a higher BRL, it is recommended to use the alternative setting with Ra=0 and Rb=600Ω for reasons of stability when the sidetone balance impedance has values for Rsa<=300Ω (See Ref.[1] and Ref.[2]). Setting Ra=600Ω, Rb=0Ω should only be used when Rsa>(Ra/2)=300Ω

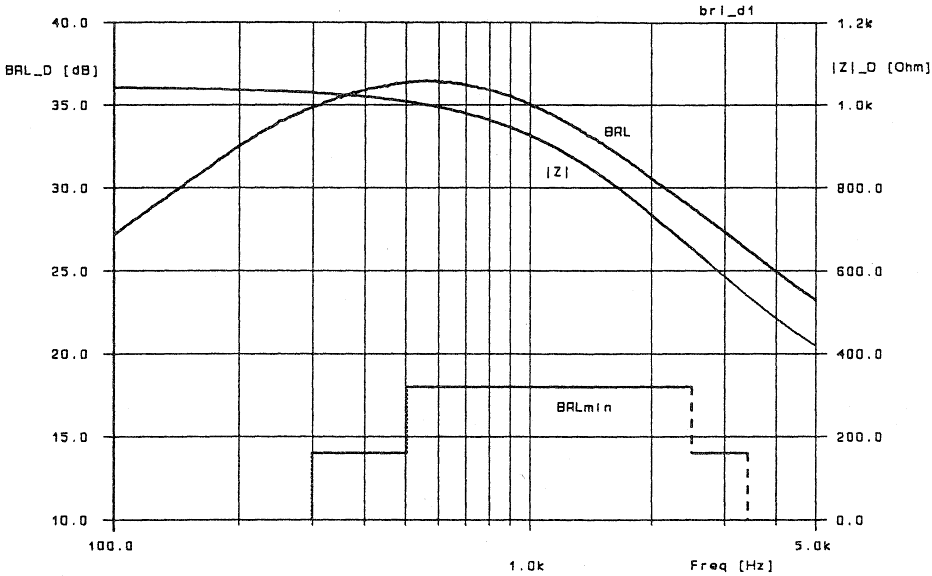


Figure 16 Balance return loss against Zref for Germany (Settings Ra=200Ω, Rb=800Ω, fp=1915Hz)

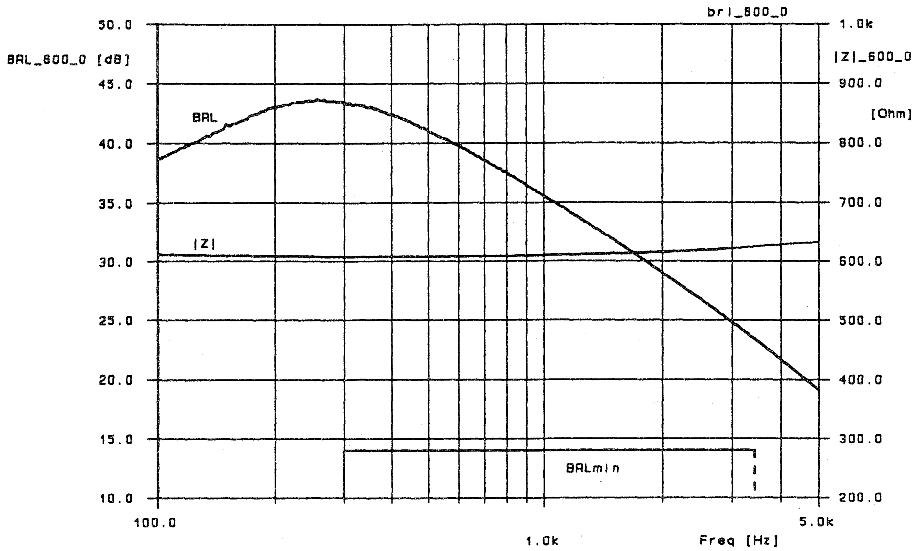


Figure 17 Balance return loss against Zref=600Ω (Settings Ra=600Ω, Rb=0, fp=12kHz)

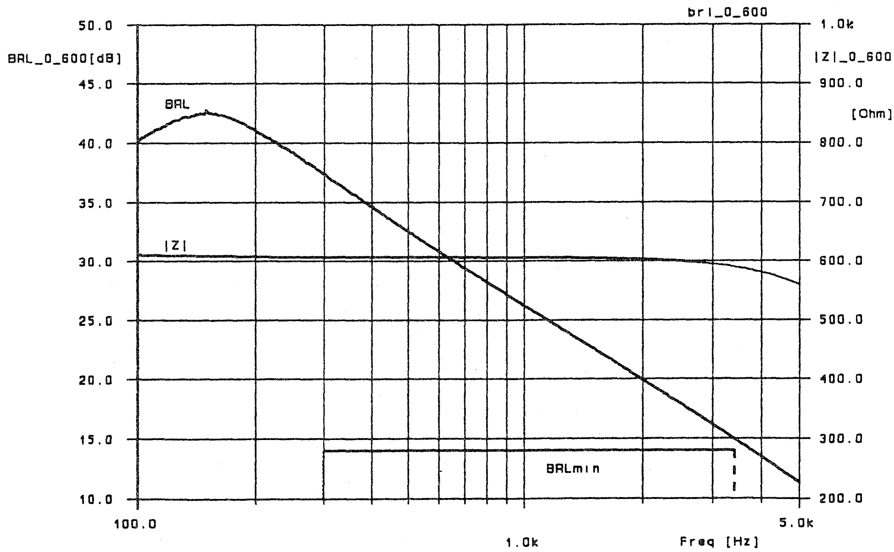


Figure 18 BRL against Zref=600Ω (Settings Ra=0, Rb=600Ω, fp=12kHz)

7.4 Sidetone characteristics

The frequency characteristic between the HS microphone inputs and output OREC of PCA1070 is shown in Fig. 19. This represents the electrical sidetone gain. Because the handset microphone gain has been set to G_{M_HS}=41dB and the receive gain between line and OREC is fixed to -6dB, the sidetone gain would be +35dB in case the sidetone suppression would be zero. During the measurement the line has been loaded with an artificial line (0km and 2km 0.4mm diameter) of 260Ω/km and 44nF/km terminated with 220Ω+(820Ω//115nF). The sidetone suppression has been optimized for 2km line length by programming the sidetone balance impedance of the PCA1070 to Zoss= 221Ω + (1410Ω//121nF). This yields a sidetone suppression in the telephony band of values better than 25dB (peaks of more than 40dB suppression can be observed). However at 0km line a suppression of some 13dB is obtained.
For safe calculation of the loop gain in handsfree a value of 25dB of gain from MIC+/MIC- to OREC has to be taken into account. This equals 10dB electrical sidetone suppression (Gstb=-10dB).

With the above given setting of sidetone balance impedance in PCA1070, rather large differences in sidetone suppression for different line load conditions occur. In handset mode this is a normal behaviour. However in HF mode it means that the loop gain will vary strongly and therefore the switching behaviour of the voice switches may be affected. PCA1070 allows reprogramming of the sidetone balance in HF mode in such a way that less variation of sidetone suppression for different line loads occurs (avoid optimum balancing for one specific line length). In this way less variation of loop gain is obtained. Of course the average sidetone suppression is worse when compared to HS mode and this will result in less gain margin (see paragraph 3.4). Fig. 20 shows the sidetone gain between line LN and OREC (= Gstb-6dB) for 0 to 10km line (176Ω/km, 38nF/km) terminated with 600Ω. The sidetone balance impedance of PCA1070 is programmed to Zoss= 155Ω + (465Ω//103nF). The maximum sidetone suppression is now 22dB and the minimum is 7dB. With the switching range of 46 dB the gain margin of the total loop gain is still better than 16dB.

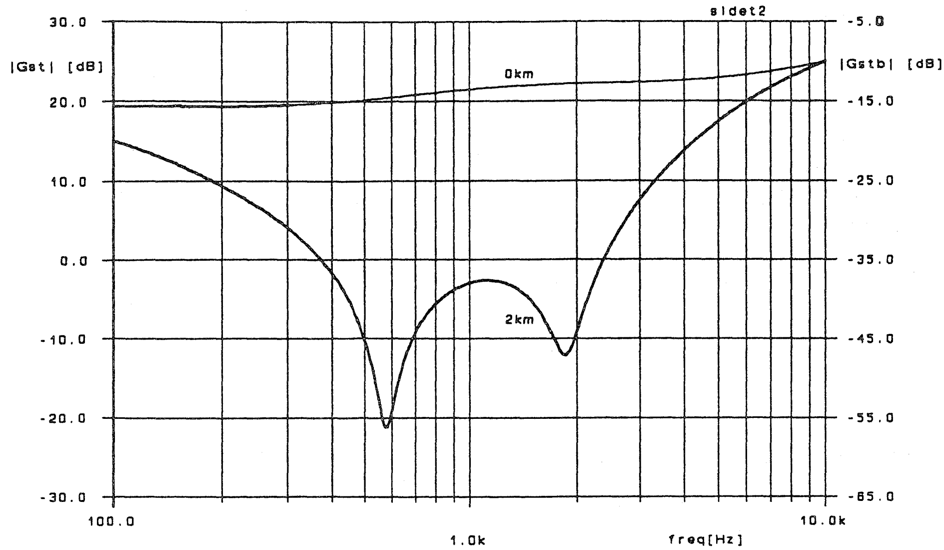


Figure 19 Electrical sidetone (HS mic inputs to OREC) for 0km and 2km 0.4mm terminated with ZD=220Ω+[820Ω//115nF] (Settings Zbal=Zoss=221Ω+[1410Ω//115nF])

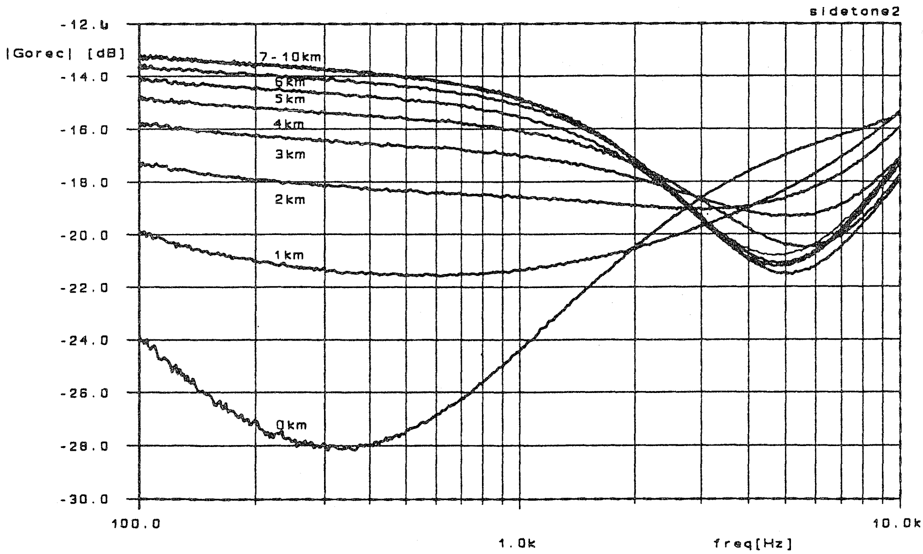


Figure 20 Electrical sidetone (HS mic inputs to OREC) for 0km to 10km 0.5mm cable (176Ω/km, 38nF/km) terminated with 600Ω; settings Zoss=155Ω + [465Ω//103nF]

7.5 Electro-acoustic coupling

The electro acoustic coupling between handsfree microphone and loudspeaker has been measured by applying a signal to the loudspeaker and measuring the gain to the base microphone (across R58).
Fig. 21 shows the electro acoustic coupling occurring at printed circuit board PR46311 mounted on its cabinet. Clearly, the coupling has its maximum around -40dB (see also paragraph 3.4).

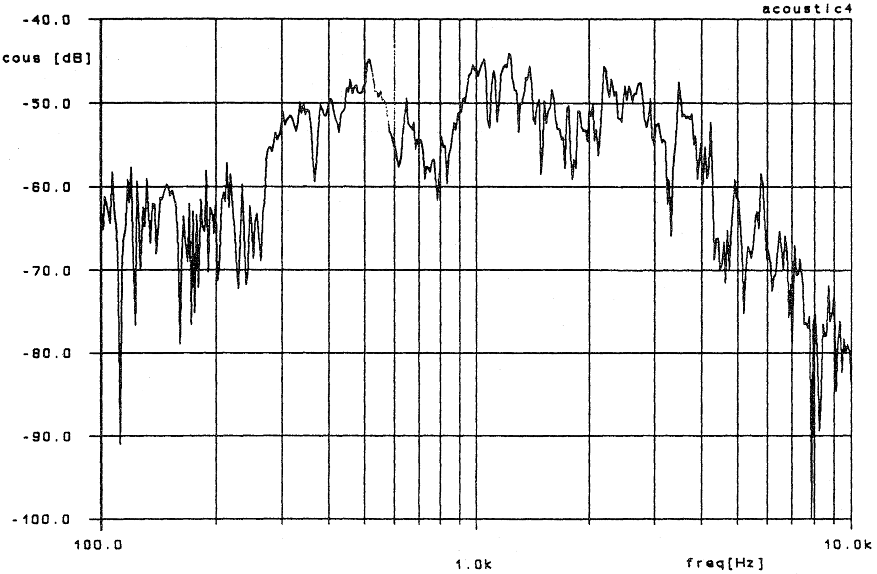


Figure 21 Electro acoustic coupling between loudspeaker and base microphone

7.6 Noise figures

Noise has been measured for both the transmit direction (at the line) and the receive direction (earpiece and loudspeaker). The handset microphone is replaced by a 150 ohm resistor and the handsfree microphone shorted to GND (R58=0). The loudspeaker is replaced by a 50Ω resistor.

7.6.1 Noise at the telephone line

The noise in HS mode is shown in Fig. 22.
Conditions: send channel of PCA1070 enabled for handset operation (SM=0)
MUTET = MUTER = high
VOL1 and DL/VOL2 don't care

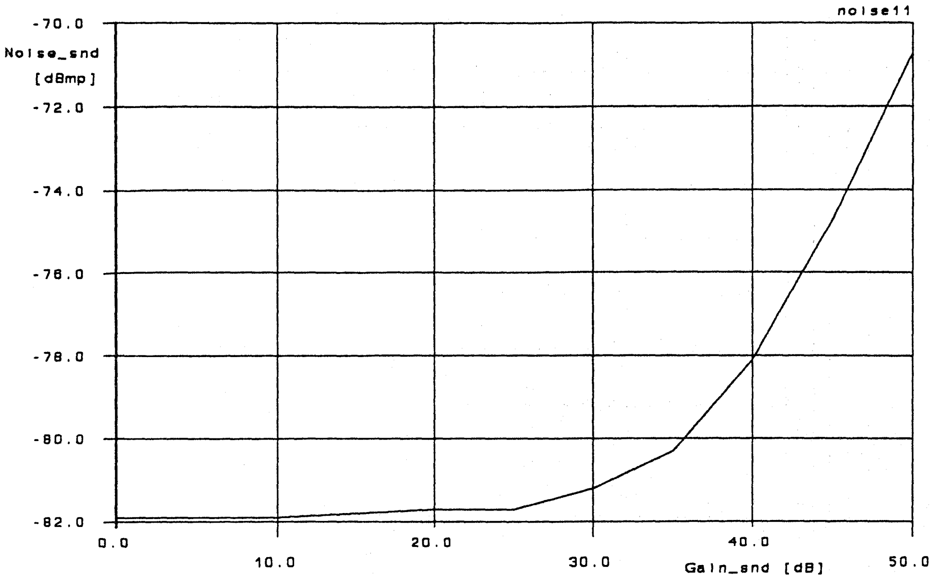


Figure 22 Psophometrically weighted noise at the line in HS mode versus gain setting of the HS microphone channel

Noise at the line in HF mode is given in Fig. 23.

Conditions:	send channel of PCA1070 enabled for DTMF (SM=1)	
	HF=high or low	
	TX mode	: MUTET = low, MUTER = high
	Idle_00	: Volume at -18dB, MUTET = low, MUTER = low
	Idle_11	: Volume at 0dB, MUTET = low, MUTER = low
	RX mode	: Volume at 0dB, MUTET = high, MUTER = low

Clearly, the noise at the line during handsfree is dependent on the mode of the TEA1093. In transmit mode (speech entering handsfree microphone) the noise level is at its highest value. However, during noise measurements the TEA1093 will automatically switch to idle mode if the RX channel is not muted. Fig. 23 represents noise figures in two idle modes with different volume setting. If volume is set to maximum, the idle mode (idle_11) represents a gain reduction of the TX and RX channel of 23dB (1/2 the switching range of 46dB). With volume set to minimum the RX gain is reduced with 18dB and thus the switching range is adapted to 46 - 18 = 28dB. Idle mode (idle_00) now represents a gain reduction of only 14dB.

In voice switched listening-in mode (LI1) mode the noise at the line will be the same as in Fig. 23 but the gain of the HS microphone channel is 9 dB less than the values indicated on the x-axis of the graph (See paragraph 3.3).

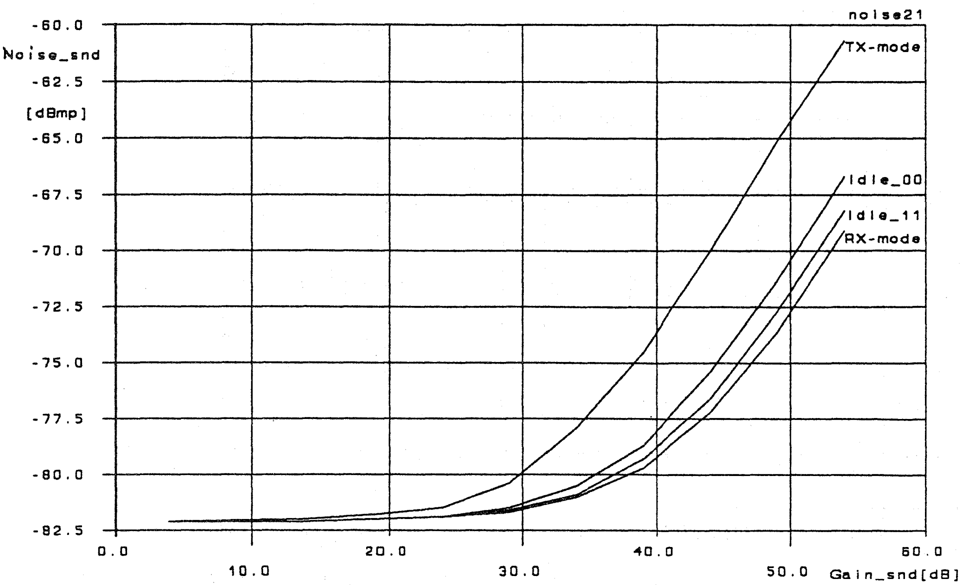


Figure 23 Psophometrically weighted noise at the line in HF mode versus gain setting of the HF microphone channel

7.6.2 Noise at the earpiece and the loudspeaker

Noise at the earpiece in basic listening-in mode (LI2) and handset mode (HS) is shown in Fig. 24. Conditions during measurement:

RT0/HF = high
MUTET = low
MUTER = high
SM = 1

Noise at the loudspeaker in HF and LI1 modes (Fig. 25) were measured as a function of volume setting with the following conditions:

Atx_1070 set to 25dB
MUTET = high
MUTER = low
SM=0

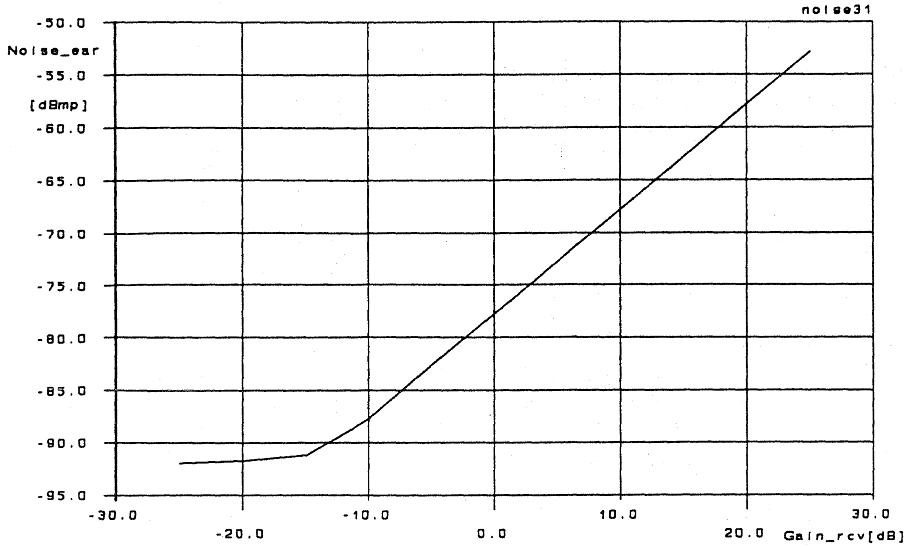


Figure 24 Psophometrically weighted noise at the earpiece in HS and LI2 modes versus gain setting of the HS earpiece channel

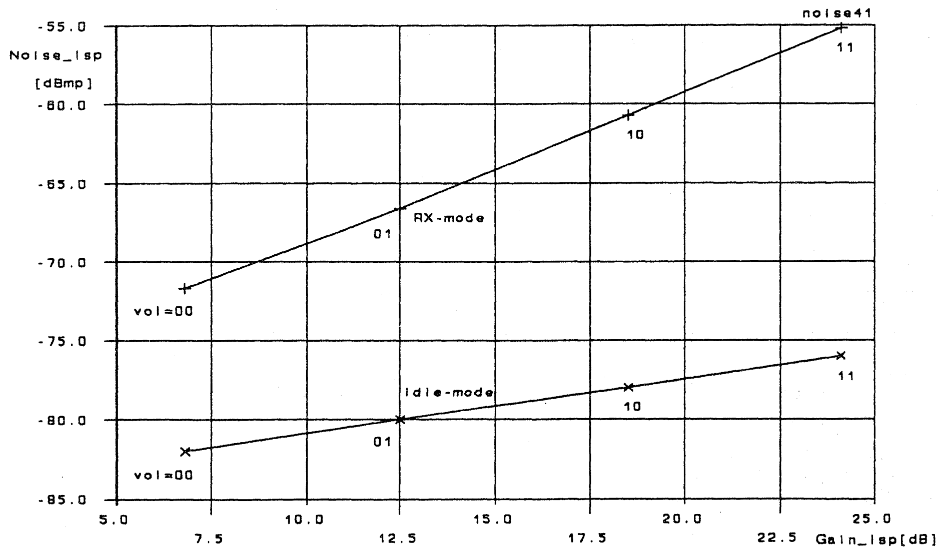


Figure 25 Psophometrically weighted noise at the loudspeaker in HF and LI1 modes versus volume setting

7.7 Transmit output swing on line

Fig. 26 shows for the HS mode the maximum possible peak to peak output swing on the line in sending direction as a function of line current. Fig.27 shows this for the HF or LI1 modes. Note that in HF and LI1 modes the dynamic limiter of the PCA1070 is not active because the sending signals are entered into PCA1070 via the DTMF input. Therefore the maximum swing on the line is somewhat higher than in HS mode and the clipping behaviour is different.

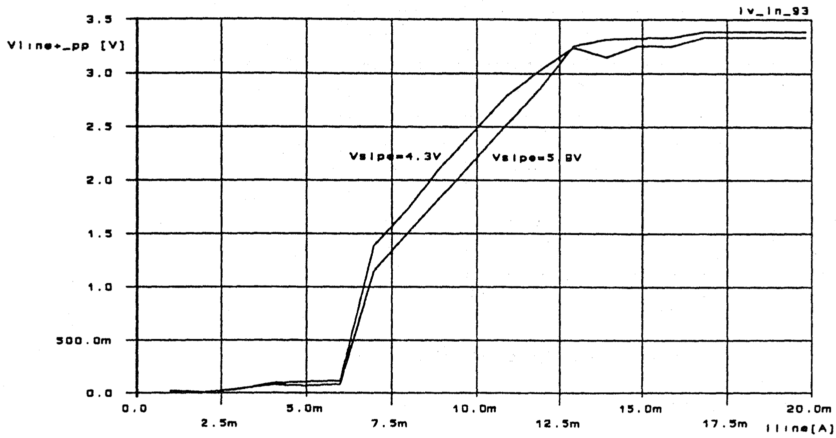


Figure 26 Maximum AC sending level on the line as a function of Iline in HS and LI2 modes; parameter is the DC setting V_{SLPE}=4.3V, 5.9V

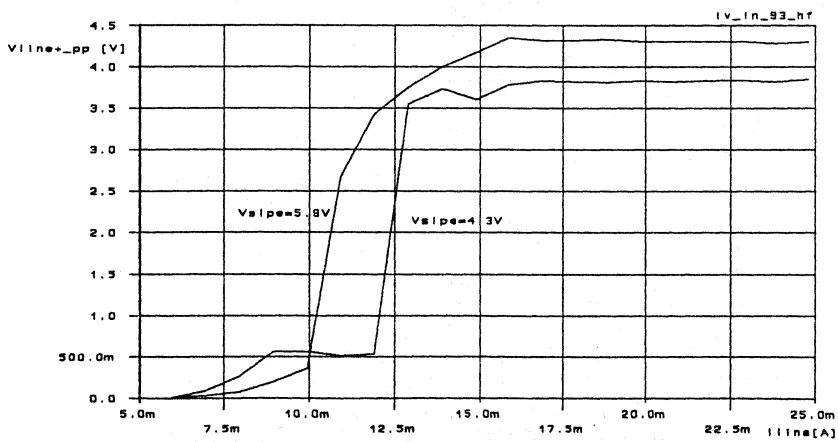


Figure 27 Maximum AC sending level in HF and LI1 modes as a function of Iline; V_{SLPE}=4.3V, 5.9V

7.8 Receive output swing on earpiece

Fig. 28 shows for the HS mode the output swing capabilities of the earpiece amplifier into a 150Ω symmetrical load.

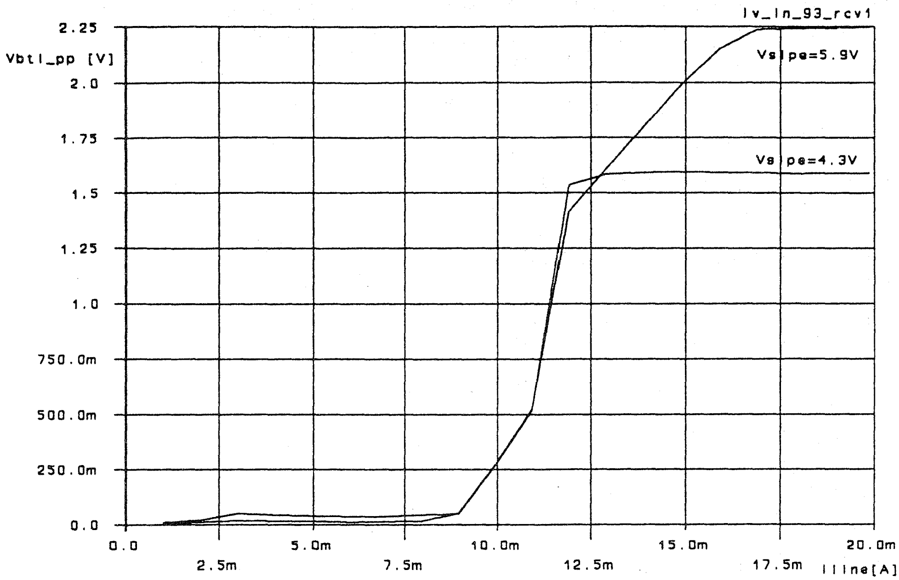


Figure 28 Maximum output swing of the HS receive amplifier into a 150Ω earpiece (BTL) versus Iline in HS and LI1,2 modes

7.9 Receive output swing on loudspeaker

The output power into a 50Ω load at the loudspeaker output of the TEA1093 has been measured versus line current. For all measurements shown in this paragraph, the AC signal at the telephone line has been limited to about 150mVpeak. Together with the 0.6V gap between SUP and VBB this level did not give rise to an efficiency reduction of the TEA1093 supply structure (see also Chapter 3.1). The 150mVpeak signal at the line is sufficient to guarantee that the loudspeaker output signal is always in overdrive and thus maximum swing occurs (THD < 2%).

Fig. 29 shows the output swing in case a 50Ω single ended load is used for two different DC settings of PCA1070. Clearly, the curves start at a different line current (10mA-25mA range). This is due to the DC power consumption of the total application. At higher DC settings more DC current is needed to supply the PCA1070 and TEA1093.
At the part were the output swing is increasing with line current, the output swing is limited due to lack of current. The VBB and shunt limiters of the TEA1093 are active in this area and guarantee that VBB will not drop below its preset value (see also Ref.[3]). The curves are limited to a maximum due to the DC setting of the TEA1093 which is fixed to 3.6V in this case. In this part the peak limiters of the TEA1093 prevent distortion of the output signal

(see Ref. [4]). A higher output swing is possible when the VBB setting of TEA1093 is increased (see also Chapter 3).

Fig. 30 shows the continuous output power into a 50Ω load. These curves are calculated out of Fig. 29. In the region were the output power is increasing with line current (10mA-25mA), the peak output power is much higher than the continuous output power shown. This is due to the charge in the VBB capacitor. During speech, which consists of bursts, the acoustic pressure from the speaker will be just as large in this region than in the region were the supply voltage determines the output power.

Fig. 31 gives an impression of the maximum output swing which can be reached during speech signals. The curve has been measured without any load at LSP1 or LSP2.

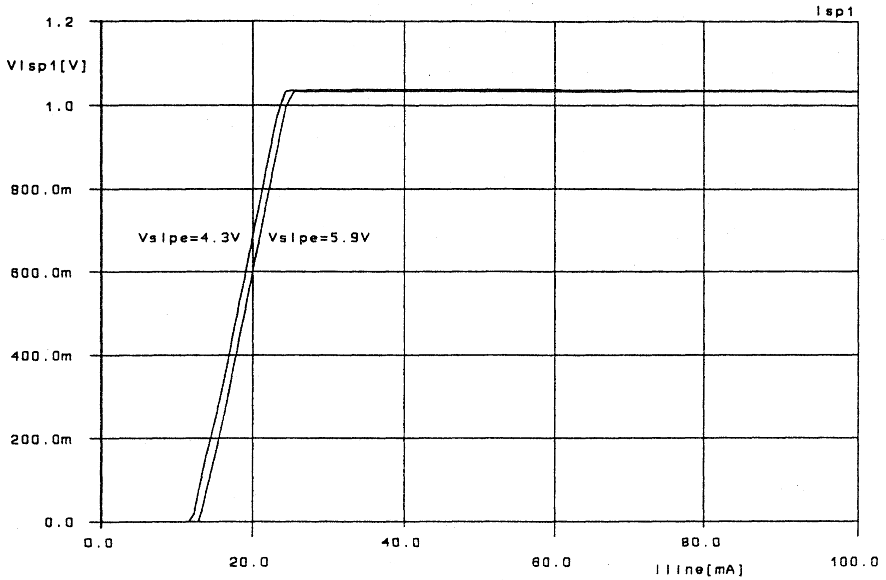


Figure 29 Output swing into a 50Ω single ended load at loudspeaker output LSP1; V_{SLPE}=4.3V, 5.9V

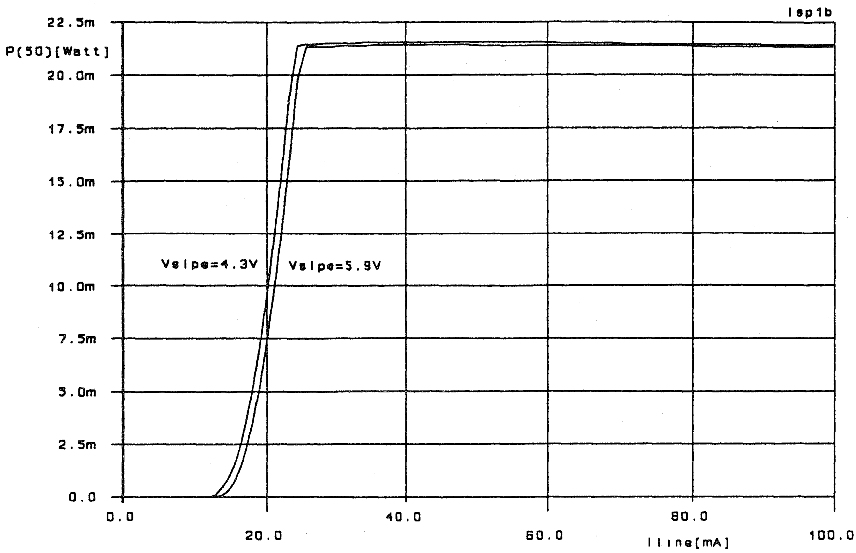


Figure 30 Output power into a 50Ω single ended load at Vslpe=4.3V, 5.9V

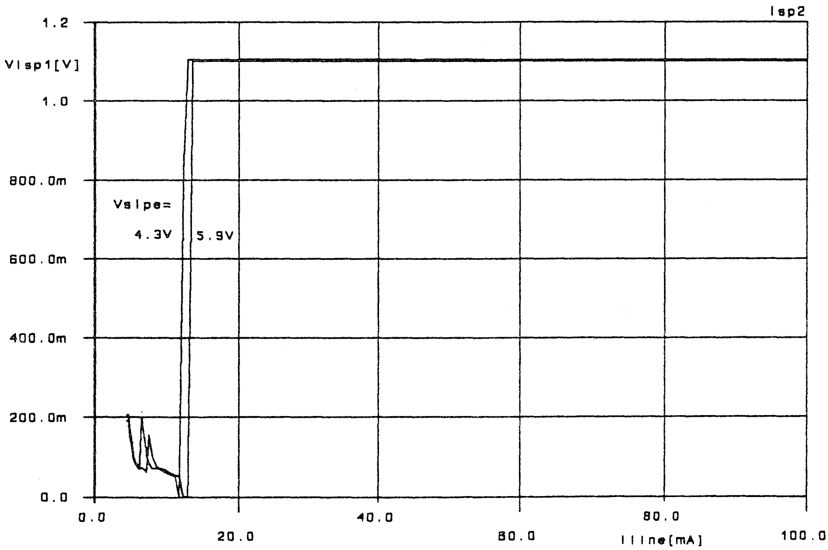


Figure 31 Output swing at LSP1 without load

7.10 Start-up behaviour

Start-up behaviour of the application has been tested with a 60V power supply with a series resistance of 2520Ω at the telephone line. This results in a line current of about 20mA. The trigger for the figures shown in this paragraph are derived from LN of the PCA1070.

Fig. 32 shows the DC start-up behaviour when the set is plugged in first time and all capacitors are fully discharged and the handset is lifted. The start-up times for Vab and VDD of the PCA1070 and the supply voltage VBB of the TEA1093 with Vslpe set to 4.3V are shown. Also the activity on the I^2C pin SCL is shown. At roughly 100ms the I^2C-bis becomes active and the µC loads all programmable parameters into PCA1070. This is done with the DST bit set to "1" to activate the DC start circuitry in the PCA1070 to minimize settling time. Then after a waiting time of about 150ms (t≈260ms) the µC loads all programmable parameters once more into PCA1070 with DST="0" and loads the display. The waiting time is necessary to ensure that the capacitors Clsi, Creg and the VDD, VBB supply capacitors are sufficiently charged. Then the line current is read via I^2C-bus and the gain of the microphone channel and the receive channel are adapted according to the values programmed in the EEPROM of the µC to match the connected line.

Fig. 33 shows the AC and DC start-up time for the handset microphone when the set is plugged in first time. This curve has been measured by applying an acoustic signal to the handset microphone. Clearly, the handset microphone is active within 150ms and reaches its final value at about 350ms.

Fig. 34 shows the AC and DC start-up behaviour when the 'HOOK' key is pressed. The set starts-up in HF mode. Assumption is that the set has been plugged into the wall socket long enough to charge the µC supply capacitor Cvmc=C5. The µC loads the transmission parameters into PCA1070 (with DST="1") after a waiting time of ≈50ms to ensure that the supply capacitor Cvdd=C7 of PCA1070 is charged sufficiently. Then after a delay of 150ms (supply capacitor C27 of TEA1093 must be charged) the PCA1070 is loaded again with DST="0" and the display is loaded. The start-up behaviour in HS mode (via the cradle contact S37) is about the same as shown in Fig. 34.

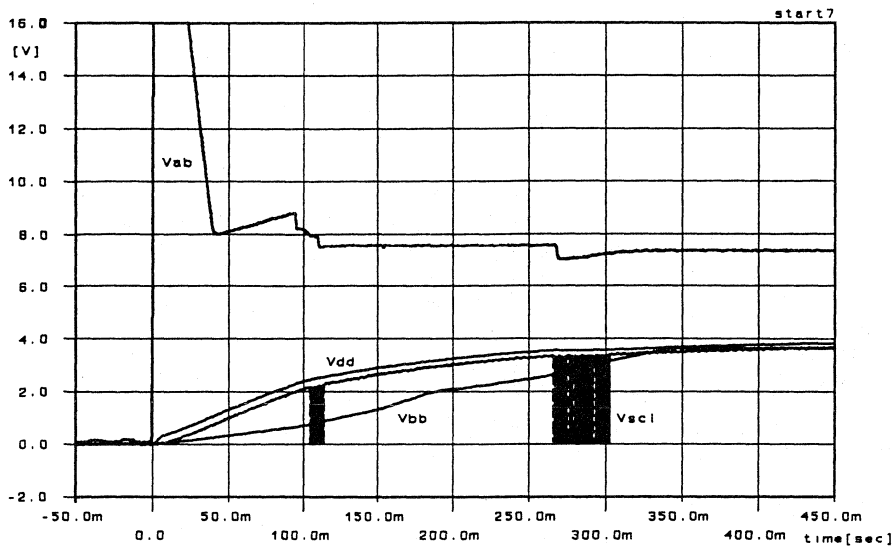


Figure 32 Start-up behaviour in HS mode (set plugged in first time); Vexch=60V, Rline=1520Ω, Rexch=1000Ω

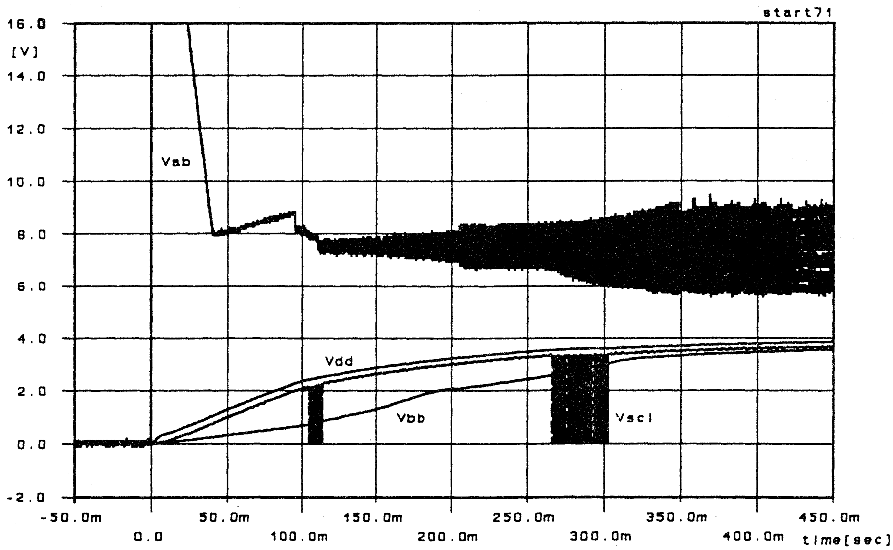


Figure 33 AC + DC start-up behaviour in HS mode (set plugged in first time); Vexch=60V, Rline=1520Ω, Rexch=1000Ω

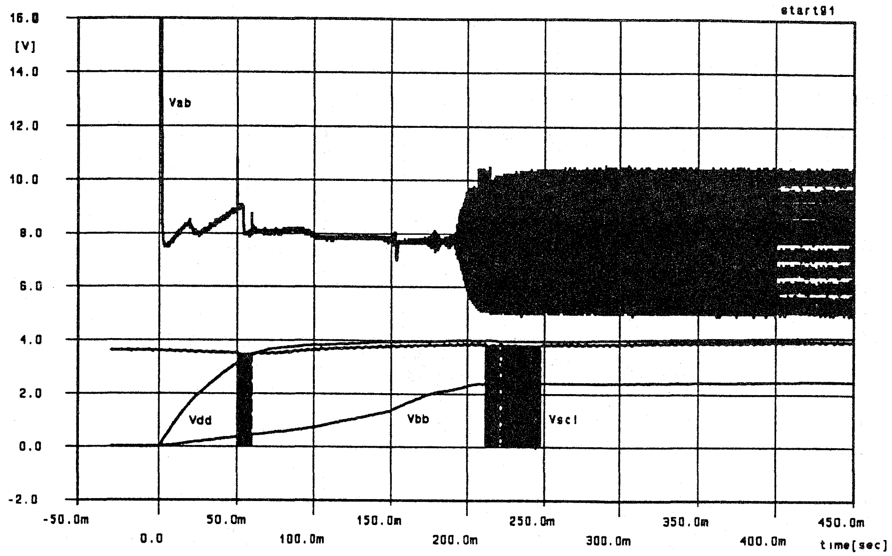


Figure 34 DC+ AC start-up time when key "HOOK" is pressed; Vexch=60V, Rline=1520Ω, Vexch=1000Ω

7.11 Hook-on

Fig. 35 shows the switch-off behaviour in HS mode via cradle contact S37. Fig. 36 gives the time diagram when key "HOOK" is pressed when the set is in HF mode.

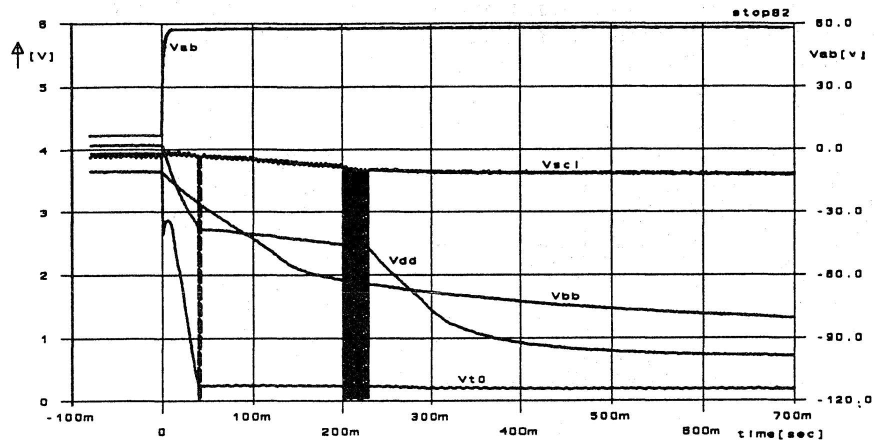


Figure 35 Hook-on (cradle switch S37), Vexch=60V, Rexch=1000Ω, Rline=1520Ω

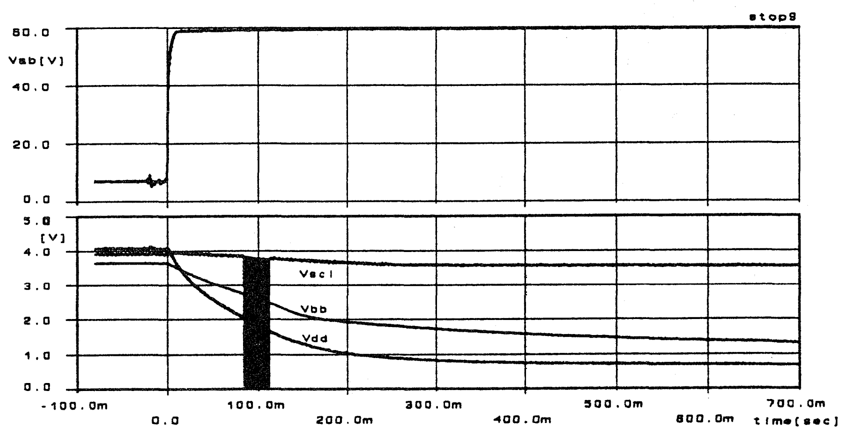


Figure 36 Hook-on in HF mode ("HOOK" key pressed)

7.12 Pulse dialling

Fig. 37 shows the line voltage Vab, Vslpe, VBB and I²C-bus activity at pin SCL during pulse dialling. The NSA (or DMO) function is used and it can be clearly seen that the DC voltage on pin SLPE of PCA1070 switches to a lower value (Vslpe_dm0=3.1V) at the start of pulse dialling (t=0). As a consequence of course the supply voltage of the PCA1070 and the TEA1093 will drop to a rather low value.

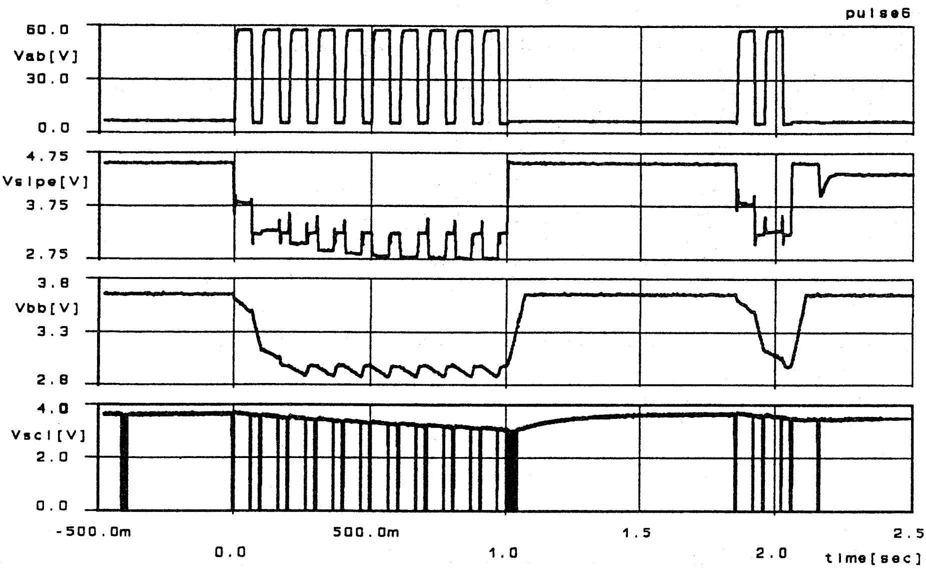


Figure 37 Pulse dialling in HS mode; Vexch=60V, Rexch=1000Ω, Rline=1520Ω; Vslpe=4.3V and Vslpe_dmo=3.1V.

7.13 Immunity to RF signals (EMC)

PCB layout: Double layer; virtually homogeneous ground-plane on components side.
EMC components: 6 capacitors of 2.2nF at all in and outgoing lines (handset cord, base cord) to the common ground plane.

Conditions for the set:
- Programmed settings as in Appendix A.
- HS mode
- Dummy microphone: 470µF
- Earpiece: 150Ω transducer

Test set-up:
- The German current injection method (common mode signal on a/b lines) is used (VDE 0878 part 200).
  This method is described in Ref. [16]
- Frequency range: 150kHz-150MHz.
- RF levels:
  Normal requirement: 3Vrms 150kHz-30MHz, 0.5Vrms 30-150Mhz.
- Amplitude modulation: 80%, 1kHz.
- Line current: 27mA.

Requirements:
- The allowed demodulation level towards the A/B lines or earpiece can be calculated from the required 40dB signal to interference ratio and the nominal line signal level. With a nominal signal of 100mV (send or receive, 1kHz) across the telephone line, this corresponds to a demodulation level of 1mV (-60dBV) at the A/B line and to -66dBV (receive gain PCA1070 -6dB) for the earpiece.

Test results:
- Are given as demodulated signal levels, with respect to 0dBV, on the A/B line terminals of the PCB (Fig. 38), across the earpiece in the handset (Fig. 39).

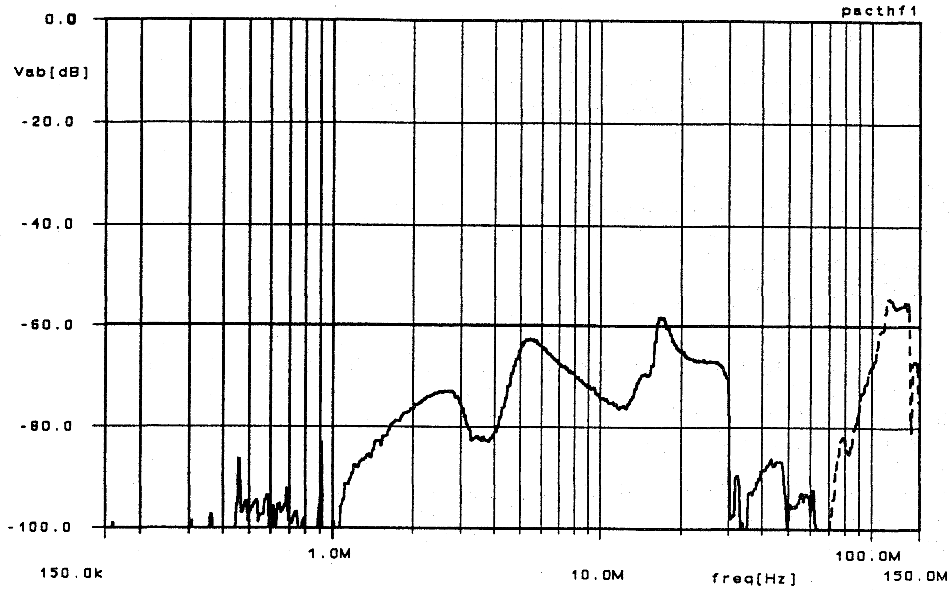


Figure 38 EMC characteristics: demodulated signal at the a/b lines

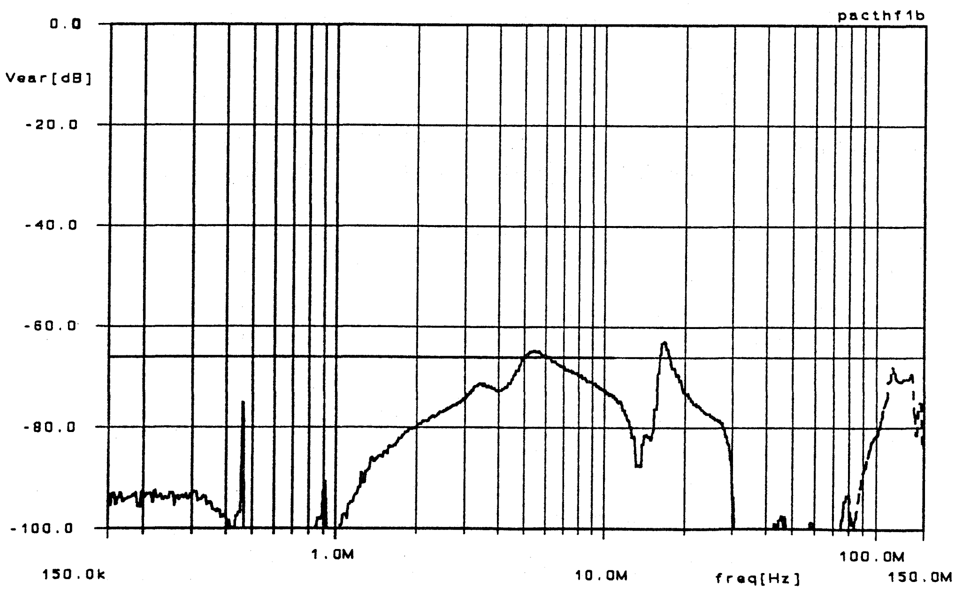


Figure 39 E.M.C. characteristics: demodulated signal at the earpiece

8 References

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9 Acknowledgement

The authors thank the people from the service group of PCALE for their excellent work on the evaluation board which has been used as the basis for this report.

Appendices

APPENDIX A Settings of programmable parameters.

Detailed explanation can be found in the SW description (Appendix D).

PRESS KEYS	DISPLAY DEFAULT	SETTING RANGE (with +/- keys)	STEP RESOLUTION
PROG, DC VOLT	dc_u=4.7	4.3 to 5.9 V **1)**	0.4 V
DC VOLT	dc_u_dmo=4.3	3.1 to 5.9 V	0.4 V
DC VOLT	dc_i_0=28.75	11.25 to 92.5 mA	≈2.5 to 3 mA
DC VOLT	dc_i_1=31.25	11.25 to 92.5 mA	
DC VOLT	dc_i_2=36.25	11.25 to 92.5 mA	
DC VOLT	dc_i_3=41.25	11.25 to 92.5 mA	
DC VOLT	dc_i_4=48.75	11.25 to 92.5 mA	
DC VOLT	dc_i_5=59.5	11.25 to 92.5 mA	
DC VOLT	dc_i_6=68.75	11.25 to 92.5 mA	
DC VOLT	dc_i_d_G=0dB	0 to 3 dB	1 dB
PROG			
PROG, GAIN MIC	G_mic=15dB **2)**	4 to 25 dB	1 dB
GAIN MIC	dlt=off	off=3.5Vp-p on=2.6Vp-p	
GAIN MIC	G_dtmf=1dB **3)**	-5 to +15 dB	1 dB
GAIN MIC	G_HF=20dB **4)**	2 to 25 dB	1 dB
GAIN MIC	G_Li1=20dB **5)**	4 to 25 dB	1 dB
GAIN MIC	G_Li2=15dB **6)**	2 to 25 dB	1 dB
GAIN MIC	Li=1	1 = LI1 mode 2 = LI2 mode	
PROG			
PROG, GAIN REC	G_rEc= -6dB **7)**	-19 to +11 dB	1 dB
GAIN REC	hPl=high	high=5.9Vp-p low=2.3Vp-p	
GAIN REC	rFc=off	on=C-load off=R-load	
GAIN REC	G_conF=-25dB **8)**	-25 to 0 dB	1 dB

PRESS KEYS	DISPLAY DEFAULT	SETTING RANGE (with +/- keys)	STEP RESOLUTION
PROG			
PROG, SET IMP	rA=200	0 to 600 Ohm	100 Ohm
SET IMP	rb=800	0, 600 to 1000 Ohm	100 Ohm
SET IMP	Fp=1915	828 to 5859 Hz, 12kHz	See Ref. 1
PROG			
PROG, SIDE TONE	rSA=492	See Ref. [1]	
SIDE TONE	rSB=1259	See Ref. [1]	
SIDE TONE	CS=134	See Ref. [1]	
PROG			
PROG, FLASH	F=2 S=4 m=4 u=4	1: Volume: u=1 to 4 2: Melody: m=1 to 4 3: Speed: S=1 to 4 4: Frequency: F=1,2	4=max, 1=min See Appendix D See Appendix D 1=20 to 60 Hz (1f) 2=40 to 120 Hz (2f)
PROG			
PROG, LNR	3-2 Pin7=vol2	1: pulse 2: DTMF 3: mark/space=3/2 4: mark/space=2/1 5: pin 7 is dynamic limiter input DL 6: pin 7 is VOL2 output	
PROG			

Notes:

1) The values 3.9, 3.5 and 3.1 are not to be used.
2) Gain of the send prog-amp is indicated on LCD. Total gain is 26 dB higher.
3) Gain of the send prog-amp is indicated on LCD. Total gain is 6 dB higher.
4) Gain of the send prog-amp is indicated on LCD. Total gain is 30 dB higher.
5) Gain of the send prog-amp is indicated on LCD. Total gain is 21 dB higher.
6) Gain of the send prog-amp is indicated on LCD. Total gain is 26 dB higher.
7) Gain of the rec prog-amp is indicated on LCD. This is the overall gain with symmetrical drive of the earpiece. In case of single ended drive, the overall gain is 6 dB lower.
8) Gain of the rec prog-amp is indicated on LCD. Total gain is about 19 dB less.

APPENDIX B Overview PCD335x(A) µC family

Type	ROM	RAM	OTP	I/O	EEPROM	Features	Package
PCD3350A	8k	256	-	34	256	clock	QFP44
PCD3351A	2k	64	-	20	128	-	DIL/SO28
PCD3352A	4k	128	-	20	128	-	DIL/SO28
PCD3353A	6k	128	-	20	128	-	DIL/SO28
PCD3354A	8k	256	-	36	256	-	QFP44
PCD3755A	-	128	8k	20	128	-	DIL/SO28

PCD335x(A) family of telecom and low-voltage µC
All have DTMF, Melody output, 8048 instruction set based, V_{DD} = 1.8V to 6.0V
(For DTMF and EEPROM programming V_{DD} = 2.5V to 6.0V).

APPENDIX C List of abbreviations

a/b	Line terminals (tip-ring)
A/D	Analog to digital
AGC	Automatic Gain Control (= automatic line loss compensation)
BRL	Balance Return Loss
BTL	Bridge Tied Load (symmetrical drive of load impedance)
CCITT	The International Telegraph and Telephone Consultative Committee
CLK	Clock signal input
C_{LSI}	Capacitor between LN and LSI
Clsp1	Capacitor at pin LSP1
Cmout	Capacitor at pin MOUT
CNET	Centre National d'Etudes des Télécommunications
C_{REG}	Capacitor between LN and REG
Crin1/2	Capacitor at pin RIN1 or RIN2
Cs	Capacitor of sidetone balance impedance Zoss
C_{VDD}	Capacitor between VDD and VSS
C_{VMC}	Capacitor between VMC and VSS
D	Germany
dB	Decibel
dBmp	Decibel with reference to 1mW into 600Ω psophometrically weighted (O41 curve of CCITT recommendations)
dBspl	Decibel with reference to $20\mu Pa=20\bullet 10^{-6}Pa$
DC	Direct Current
DLC	Dynamic limiter timing adjustment
DLT	Dynamic Limiter Threshold (bit)
DMO	Dial Mode Output
DOC	Dial Output Connection
DPI	Dial Pulse Input (bit)
DST	DC Start Time (bit)
DTMF	Dual Tone Multi Frequency
EEPROM	Electrical Erasable Programmable Read Only Memory
EMC	Electro Magnetic Compatibility
f	Frequency
f_{CLK}	Clock Frequency
fp	Pole Frequency of set impedance Zs
f_{SCL}	Serial Clock Frequency
FTZ	Fernmelde Technisches Zentralamt
Gatt1	Gain of attenuator 1
Gatt2	Gain of attenuator 2
G_{CTs}	Confidence Tone gain for Symmetrical drive (BTL)
G_{DTMF}	DTMF Gain
G_{M_HS}	Gain handset Microphone channel
Gma	Gain send prog-amp
GND	Ground
Gra	Gain receive prog-amp
G_{RA}	Gain Receive channel for Asymmetrical drive (SEL)
G_{RS}	Gain Receive channel for Symmetrical drive (BTL)
Gst	Sidetone Gain between microphone inputs and earpiece outputs
Gsupp	Sidetone suppression

Gswitch	Gain of switch
Gtx_1093	Transmit gain TEA1093 bewteen MIC and OMIC
HF	Handsfree
HPL	Hearing Protection Level (bit)
HS	Handset
I/O	Input and/or Output
I^2C	Inter IC connection
IC	Integrated Circuit
I_{DD}	Internal supply current
I_{LINE}	Line Current
I_{LN}	Current into pin LN
Ilsp	Current available to power loudspeaker amplifier
I_P	Current to external Peripheral circuitry
I_{PNP}	Current in PNP from collector to emitter
I_{SCR}	Current out of pin SCR
I_{SLPE}	Current into pin SLPE
I_{SUP0}	Bias current into pin SUP
I_{VMC}	Current into pin VMC
I_{VP}	Current out of pin VP
LCD	Liquid Crystal Display
LI1	Voice switched Listening-In
LI2	Basic Listening-In
LN	Positive Line terminal
LSI	Line Signal Input
LSP1/2	Loudspeaker amplifier output 1/2
µC	Micro Controller
MIC-	Inverting input of Microphone preamp
MIC+	Non-inverting input of Microphone preamp
MOUT	Microphone amplifier output
MUTE2	Logic signal used to decrease set voltage during pulse dialling (also known as NSA or DMO)
MUTER	Receive channel mute input
MUTET	Transmit channel mute input
NSA	Nummern Schalter Arbeitskontakt (also known as MUTE2 or DMO)
OMIC	Output Microphone preamplifier
OREC	Output Receive preamplifier
OTP	One Time Programmable
Pa	Pascal (= 1Newton/m^2)
PCB	Printed Circuit Board
PD	Power Down
PDx	Power Down bits
PNP	PNP transistor
prog-amp	programmable amplifier
PTT	Telephone company
QR-	Inverting output of BTL receive output amplifier
QR+	Non-inverting output of BTL receive output amplifier
Ra	Resistor Ra of Zs
Rb	Resistor Rb of Zs
RC	Combination of resistor and capacitor
REG	Voltage Regulator decoupling

Rexch	Feeding bridge Resistor in exchange
RF	Radio frequency
RFC	Resistive-not / Capacitive load select (bit)
RFI	Radio Frequency Interference
Rgar	Resistor at pin GAR to adjust receive gain
Rgat	Resistor at pin GAT to adjust microphone gain
RIN1/2	Receiver amplifier input 1 or 2
Rline	Resistance of telephone line
RLR	Receive Loudness Rating
R_{LSI}	Internal resistor between LSI and VSS
RM	Receive Mute
RMC	Reset output for Micro Controller
Rmout	Resistor at pin MOUT
Romic	Resitor at pin omic
Rsa	Resistor Rsa of Zoss
Rsb	Resistor Rsb of Zoss
R_{SCR}	Resistor between SCR and VSS
R_{SLPE}	Resistor between LN and SLPE
Rsref	Resistor between SREF and SUP
Rstab	Resistor at pin STAB
R_{SUP}	Resistor between SLPE and VDD
RTO	Ringer Tone Output
Rtx	Resistor between TX and VSS
Rva_vbb	Resistor between VA and VBB
Rva_vss	Resistor between VA and VSS
Rvdd	Resistor between SLPE and VDD
SCL	Serial Clock Line of I^2C-bus
SCR	Sending Current Resistor
SDA	Serial Data Line of I^2C-bus
SEL	Single Ended Load (asymmetrical drive of load impedance)
SLPE	Slope (DC resistance)
SLR	Send Loudness Rating
SM	Send Mute
SW	Software
Tamb	Ambient Temperature
THD	Total Harmonic Distortion
TONE	Output pin of DTMF generator on μC
TR	Technical Requirement
TST	Testpin
TX	Drive output of DC voltage stabilizer
Typ	Typically
VA	VBB voltage adjustment
Vab	Voltage at line terminals a/b
Vbridge	Voltage drop across diode bridge
Vdcslope	Voltage drop across Rslpe
VDD	Positive supply decoupling
$V_{DD}/2$	Voltage at VDD divided by 2
VDE	Verband Deutscher Elektrotechniker
Vexch	Supply Voltage in exchange
Vinter	Voltage drop across interrupter

V_{LN}	Voltage at pin LN
$V_{LN}p$	AC peak voltage at pin LN
$V_{LN}p$-p	AC peak to peak voltage at pin LN
V_{LSI}	Voltage at pin LSI
VMC	Input to sense supply voltage Micro Controller
V_{MIC}	AC Voltage between MIC+ and MIC-
VOL1/2	Volume setting pins
Vp-p	Peak to peak voltage
VP	Supply for electret microphone
VREF	Voltage Reference decoupling
V_{RING}	Voltage at point RING
Vschottky	Voltage drop across Schottky diode
V_{SLPE}	Voltage at pin SLPE
$V_{SLPE}p$-p	AC peak to peak voltage at pin SLPE
VSS	Negative line terminal
V_{VMC}	Voltage at pin VMC
Zbal	Balance impedance of anti sidetone bridge
Z_D	German reference impedance (220Ω+[820Ω//115nF])
Z_{LINE}	Telephone line impedance
Z_{LN}	Impedance between LN and VSS
Zoss	Balance impedance for Optimum Sidetone Suppression
Zref	Reference impedance of telephone line
Zs	Set impedance (programmable)

APPENDIX D Functional description of demo SW for μC PCD3755A to control PCA1070 and TEA1093

APPENDIX D

I C s f o r t e l e c o m

FUNCTIONAL DESCRIPTION OF DEMO SW for μC PCD3755A TO CONTROL PCA1070 AND TEA1093

J. C. F. van Loon

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Keywords
PCD3755A, micro-computer
PACT, CMOS-transmission
TEA1093 handsfree circuit
Pulse/DTMF/repertory dialler
Programmable parameters

Date : 94-09-08
Pages: D42

Summary:

Appendix D gives a description of a demonstration and test software to control the PCA1070, which is a Programmable Analog CMOS Transmission (PACT) circuit, and TEA1093, which is a handsfree circuit, with the PCD3755A micro-computer.

The PCD3755A is a member of the PCD335xA family of micro-computer special developed for telephone applications. This PCD3755A micro-controller is a OTP (one time programmable) member, and has on board DTMF-generator, 8K ROM, 128 bytes RAM, 128 bytes EEPROM, 20 I/O and a special ringer output and is fully able to control the PCA1070 via its software I^2C bus.

The PCA1070/TEA1093/PCD3755A circuits combination performs the following functions:

- Pulse and DTMF dialling
- Redial and repertory dial stored in EEPROM
- PCA1070 control
- PCA1070 variables stored in EEPROM
- PCA1070 variables programmable via I^2C-bus and Keyboard
- TEA1093 handsfree control
- Dialling options programmable via keyboard stored in EEPROM
- Ringer functions incorporated 3-tone, 4-speeds and 4 volumes
- Display

Purchase of Philips' I^2C components conveys a license under the Philips' I^2C patent to use the components in the I^2C-system provided the system conforms to the I^2C specifications defined by Philips

Table of Contents:

Figures and Tables

D1. GENERAL DESCRIPTION

This report decribes the software to control the PCA1070 (a Programmable Analog CMOS Transmission IC abbreviated to PACT) and the TEA1093 handsfree circuit. The PCD3755A and the PCA1070 are both fabricated in a low threshold voltage C-MOS technology and are members of the Philips family of telecom ICs.
The software can be used as a ROM-code for the PCD3353A, instead of the PCD3755A.

The PCD3755A micro-computer can together with PCA1070 and TEA1093 be used to demonstrate the possibilities of this PACT-IC in a handsfree application.
The programmable parameters of the PACT are stored in the internal EEPROM of the PCD3755A micro-computer, and will be transmitted to the PACT at off-hook, pulse dialling, DTMF dialling and AGC (automatic gain control).

The (maximum 10) repertory numbers, redial and some country specifications are also stored in EEPROM so that memory retention is guaranteed for 10 years without using a battery back-up.

The two on-chip tone generators are used for Dual Tone Multi-Frequency (DTMF) dialling, and for generating a melody during ringing, which is activated when a proper incoming ringer frequency is detected.

As an output transducer for the ringer a PXE, ringer out via the special ringer output which generates square wave ringer tones with a peak-to-peak voltage of V_{DD}-V_{SS}, can be used.

The operating supply voltage is 1.8 V (2.5 V in DTMF and ringer mode and EEPROM erase/write mode) to 6.0 V with a low current consumption in all operating modes: standby, conversation, dialling, programming and ringer.

D2. Features

D2.1. Pulse / DTMF dialling features

- Pulse dialling
- DTMF dialling
- Mixed mode dialling
- Number of digits per call is infinite (FIFO register)
- Flash or register recall
- Standard 3x4 keyboard for: 0 to 9 and * and #
- Function keys for: Flash, Mute, and Access Pause
- Selectable mark to space ratio (3:2 or 2:1) stored in EEPROM
- On-chip voltage reference for stabilized supply and temperature independent tone output
- On-chip filtering for low output distortion (CEPT CS 203 compatible)

D2.2. Number storage features

- Redial (maximum 24 digits) stored in internal EEPROM
- Storage for 10 repertory dial numbers (18 digits each) in internal EEPROM
- Access pause generation and termination
- Function keys for: LNRedial, Store, Access Pause, and 1 Key repertory dialling (10 keys)

D2.3. Ringer features

- Ringer input frequency detection
- Function key for: Program Ringer
- Three-tone ringer with 4 different ringer sequences
- Ringer melody generation, speed and output volume, keypad controlled
- Specifications which can be stored in EEPROM are:
    - Ringer input frequency detection selection
    - 4 Possible ringer melodies
    - 4 Possible ringer repetition rates
    - 4 Possible ringer volumes

D2.4. PCA1070 (PACT) control features

- PCA1070 control via the I^2C-bus
- Function key for: DC-values, Microphone gain, Receive gain, Set impedance and Side tone
- PCA1070 variables can be stored via the I^2C-bus or via keyboard
- Direct programming of the PCA1070 transmission parameters via keyboard
- Programmable AGC characteristics of microphone and receiving amplifiers

D2.5. TEA1093 handsfree control features

- Handsfree mode control
- Voice switched listening-in mode control (LI1)
- Basic listening-in mode control (LI2)

D2.6. General features

- Display control to simplify the dialling and program actions
- On-chip oscillator uses low-cost 3.58 MHz (tv colour burst) crystal or PXE resonator
- Reset is generated via PCA1070
- Supply voltage range 1.8 to 6.0 V (2.5 to 6.0 V in EEPROM erase/write and DTMF and ringer mode)
- Privacy switch (mute function)
- Possibility to copy data from an external memory into the internal EEPROM

D3. PCD3755A pinning and pin description

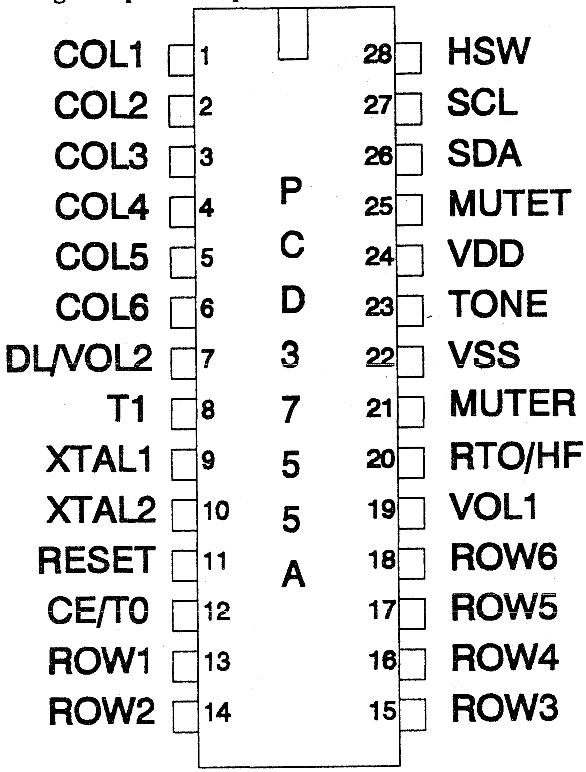


Figure D1. Pinning of the PCD3755A

Pin number	Pin name	Function
1	COL1	sense column keyboard input
2	COL2	sense column keyboard input
3	COL3	sense column keyboard input
4	COL4	sense column keyboard input
5	COL5	sense column keyboard input
6	COL6	sense column keyboard input
7	DL/VOL2	dynamic limiter input/volume control 2 output
8	T1	current loop detection
9	XTAL1	crystal/PXE oscillator input
10	XTAL2	crystal/PXE oscillator output
11	RESET	reset input
12	CE/T0	chip enable input/zero crossing of ringer signal
13	ROW1	scanning row keyboard output
14	ROW2	scanning row keyboard output
15	ROW3	scanning row keyboard output
16	ROW4	scanning row keyboard output
17	ROW5	scanning row keyboard output
18	ROW6	scanning row keyboard output
19	VOL1	volume control 1 output
20	RTO/HF	output of the melody ringer
21	MUTER	handsfree receive mute
22	VSS	negative supply
23	TONE	single or dual tone frequency output
24	VDD	positive supply
25	MUTET	handsfree transmit mute
26	SDA	serial data input/output of I^2C-bus
27	SCL	serial clock output of I^2C-bus
28	HSW	hook switch output

D3.1. COL1 to COL6, keyboard inputs

The sense column inputs COL1 to COL6 and the scanning row outputs ROW1 to ROW6 can directly be connected to a 6 x 6 single contact keyboard matrix.

D3.2. DL/VOL2, dynamic limiter input or volume output

Depending on the selection made this pin is used as input for the software controlled dynamic limiter or as extra volume control output pin so that the volume can be controlled in four steps. The selection can be made via keyboard and will be stored in EEPROM.

D3.3. T1, handset cradle input

If the handset is lifted, input T1 becomes HIGH, which indicate that conversation, programming or dialling mode is selected.

If input T1 is LOW the handsfree or the ringer mode is selected.

D3.4. XTAL1 and XTAL2, oscillator input/output

Time base for the PCD3755A is a crystal-controlled on-chip oscillator which is completed by connecting a 3.579545 MHz crystal or ceramic resonator (PXE) between XTAL1 and XTAL2.
The XTAL2 is the oscillator output and can be used as driver for the PCA1070 (PACT) oscillator input.

A low-cost quartz crystal from Philips (code nr. 4322 143 04401) is available, specially for telephony applications.

The oscillator starts when V_{DD} reaches the operating voltage level.

D3.5. RESET, reset input

In this application, the on-chip power-on-reset voltage of the PCD3755A has been chosen so low, that the PCA1070 controls the reset procedure. When the reset input becomes HIGH it initializes the IC.

D3.6. CE/T0, chip enable and zero crossing of ringer signal input

When the chip enable input becomes HIGH, then the PCD3755A leaves the standby mode, initialises the system, and switches to the handset, handsfree or ringer mode depending on if the cradle input is active (T1) or the HOOK-key is pressed input.

In ringer mode, via this input the time between two LOW-to-HIGH transients is measured, which is the actual ringer frequency.

D3.7. ROW1 to ROW6, keyboard outputs

The scanning row outputs ROW1 to ROW6 and the sense column inputs COL1 to COL6 can directly be connected to a 6 x 6 single contact keyboard matrix.

D3.8. VOL1, volume output

The VOL1 output is used to control the output volume, if pin 7 is used as VOL2 output a for steps volume control is possible, if pin 7 is used as DL only a two steps volume control is possible. The volume can be changed via keyboard, different values can be stored for ringing and handsfree/listening-in mode. The selected output levels are stored in EEPROM.

D3.9. RTO/HF, ringer tone/handsfree output

During ringing the output of the internal tone generator is connected to this RINGER output. This RINGER output signal has a peak-to-peak square output voltage of V_{DD}-V_{SS} (this is an optimal output for a PXE transducer).
At handsfree (HF), it selects between the HF basic microphone and HS microphone for voice switched listening-in.

D3.10. MUTER and MUTET, TEA1093 mode control outputs

The MUTER and MUTET outputs control the status of the TEA1093 handsfree circuit as follows:

MODES	MUTER	MUTET	RM	SM
Handset	HIGH	HIGH	0	0
Listening-in	LOW	HIGH	0	0
Voice switched listening-in	LOW	LOW	0	1
Handsfree	LOW	LOW	0	1

D3.11. V_{DD} and V_{SS}

V_{DD} and V_{SS} are the supply terminals.

D3.12. TONE, DTMF tone output

In DTMF dialling mode the dual tones which are provided at the output TONE are filtered by an on-chip switched capacitor filter, followed by an active RC low-pass filter.
As result, the total harmonic distortion of the DTMF tones fulfils the CEPT CS 203 recommendations.
An on-chip reference voltage ensures output tone levels independent of supply voltages.
The impedance is 100 ohm typically.
In case no tones are generated, the tone output is in tri-state mode.
In active state, the DTMF output is biased at $V_{DD}/2$.

D3.13. SDA, serial data input/output

This is the data input/output of a software I^2C-bus. The clock speed is about 10 KHz, fast enough to send and receive data to/from the PCA1070 and other I^2C-bus devices.

D3.14. SCL, serial clock output

This output is used as master clock for all I^2C-bus devices used in this application.

D3.15. HSW, hook-switch output

In listening-in and handsfree mode this output becomes active (HIGH), its function is to switch-on the electronic hook-switch.

D4. Keyboard

The sense column inputs COL1 to COL6 and the scanning row outputs ROW1 to ROW6 are directly connected to 6 x 6 single contact keyboard matrix. The keyboard organisation is shown in figure D2.

Keyboard entries are valid 20 ms (debounce time) after the leading edge of a keyboard entry.

If more than one key is pressed simultaneously, no operation is carried out.

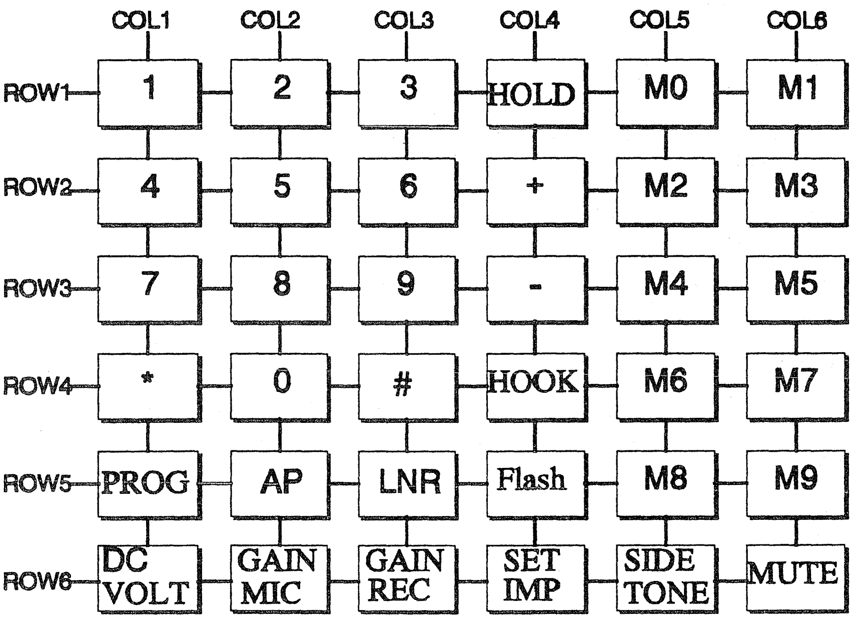


Figure D2. The keyboard organisation.

Function of the keys is:

0 to 9,* and #	Standard keyboard; in pulse dialling mode the valid keys are the 10 numeric keys (0 to 9), the 2 non-numeric keys (* and #) switch from pulse to DTMF dialling (mixed mode dialling). In DTMF dialling mode the 10 numeric keys and the 2 non-numeric keys are valid.
MUTE	Mute key (can only be used if dialling not active), this is a toggle function each time this key is pressed the send and receive mute of the PCA1070, via the I^2C-bus, are switched.

+	Increment key, in handsfree or listening-in mode and at ringer it increases volume, in the programming modes it change values.
-	Decrement key, in handsfree or listening-in mode and at ringer it decreases volume, in the programming modes it change values.
M0 to M9	One key abbreviated dialling, the 10 repertory numbers are directly accessible via push-buttons M0 to M9.
FLASH	FLASH key, this key starts a FLASH procedure. During programming it is used to select the ringer options.
LNR	Last number redial. During programming it is used to select dialling and an I/O option (pin 7).
AP	Access pause key, results in inserting an access pause of 3 seconds in the telephone number.
HOOK	Hook switch key, is used to switch from or to listening-in or handsfree mode.
HOLD	Hold switch key, is used to switch to the hold mode where the line current is measured.
PROG	Program or store key.
DC-VOLT	PACT DC program key, is used during programming it select the DC voltage, DC voltage during pulse dial, DC currents for AGC and the AGC delta gain.
GAIN-MIC	PACT microphone gain program key, is used during programming to select the microphone gain (in HS, HF, LI1 and LI2 modes), DTMF gain, dynamic limiter threshold and listening-in mode (LI1 or LI2).
GAIN-REC	PACT receiver gain program key, is used during programming to select the receive gain, confidence gain, protection level and load select bit.
SET-IMP	PACT impedance program key, is used during programming to select the set-impedance (set-resistor Ra, set resistor Rb and the set impedance pole frequency).
SIDE-TONE	PACT side tone program key, is used during programming to select the side tone balance impedance (resistor RSa, resistor RSb and capacitance CS).

D5. EEPROM organisation and programming procedure

D5.1. EEPROM organisation

The dialling options, PACT variables and options, ringer options and all the telephone numbers are all stored in EEPROM.
By using EEPROM no special backup provisions are needed such as battery, current from the line, or very big capacitors.
Table D1 describes the meaning of each EEPROM byte.
Table D2 describes the meaning of each bit of all the bytes that do not contain telephone numbers.

Table D1. EEPROM organisation table.

Function	Length	EEPROM byte places
Redial	13 bytes	00H till 0CH
M1	9 bytes	0DH till 15H
M2	9 bytes	16H till 1EH
M3	9 bytes	1FH till 27H
M4	9 bytes	28H till 30H
M5	9 bytes	31H till 39H
M6	9 bytes	3AH till 42H
M7	9 bytes	43H till 4BH
M8	9 bytes	4CH till 54H
M9	9 bytes	55H till 5DH
M10	9 bytes	5EH till 66H
PACT variables	23 bytes	67H till 7DH
dial options	1 byte	7EH
Ringer options	1 byte	7FH

Table D2. Exact bit arrangement in the EEPROM for the PACT and other variables

Function	EEPROM address	Bit 7	Bit 6	Bit 5	Data Bit 4	Bit 3	Bit 2	Bit 1	Bit 0
DC voltage	67H		VDC2	VDC1	VDC0				DST
Side-tone	68H	ZOSB3	ZOSB2	ZOSB1	ZOSB0	ZOSA3	ZOSA2	ZOSA1	ZOSA0
and set	69H	ZOSP3	ZOSP2	ZOSP1	ZOSP0		ZSA2	ZSA1	ZSA0
impedance	6AH		ZSB2	ZSB1	ZSB0	ZSP3	ZSP2	ZSP1	ZSP0
Gain mic. HS	6BH	DLT		GMA5	GMA4	GMA3	GMA2	GMA1	GMA0
Gain rec. HS	6CH	RFC	HPL	GRA5	GRA4	GRA3	GRA2	GRA1	GRA0
Control	6DH	PD1	PD0		RRG	RM	SM		DPI
DC-current-0	6EH				LC4-0	LC3-0	LC2-0	LC1-0	LC0-0
DC-current-1	6FH				LC4-1	LC3-1	LC2-1	LC1-1	LC0-1
DC-current-2	70H				LC4-2	LC3-2	LC2-2	LC1-2	LC0-2
DC-current-3	71H				LC4-3	LC3-3	LC2-3	LC1-3	LC0-3
DC-current-4	72H				LC4-4	LC3-4	LC2-4	LC1-4	LC0-4
DC-current-5	73H				LC4-5	LC3-5	LC2-5	LC1-5	LC0-5
DC-current-6	74H				LC4-6	LC3-6	LC2-6	LC1-6	LC0-6
Delta gain	75H							DG-1	DG-0
DTMF gain	76H			GDA5	GDA4	GDA3	GDA2	GDA1	GDA0
Conf. gain	77H			GCA5	GCA4	GCA3	GCA2	GCA1	GCA0
Gain mic. HF	78H			GHA5	GHA4	GHA3	GHA2	GHA1	GHA0
Gain mic. LI1	79H			GVA5	GVA4	GVA3	GVA2	GVA1	GVA0
Gain mic. LI2	7AH			GLA5	GLA4	GLA3	GLA2	GLA1	GLA0
DC_at_dial	7BH		VAD2	VAD1	VAD0				
Free	7CH								
"	7DH								
Dial-options	7EH			V-H1	V-H0	V/H	DYL	P/T	M/S
Ringer-opt.	7FH	Freq		SP-1	SP-0	M-1	M-0	V-R1	V-R0

D5.2. EEPROM programming procedure

The PCD3755A supports six EEPROM programming procedures:

1) LNR is described in chapter 7.4.1. "Last number redial".
2) Repertory numbers is described in chapter 7.4.5. "Storing repertory numbers".
3) PCA1070 variables described in chapter 7.5. "Programming PACT variables"
4) Telephone options described in chapter 7.5.4. "Programming dialling options"
5) Ringer options described in chapter 7.7.6. "Programming ringer options"
6) Via the SDA and SCL lines chapter 5.2.1.

Methods 1 to 5 can be used during demonstration and evaluation.
Method 6 can be used for factory programming of the internal EEPROM.

D5.2.1. EEPROM programming procedure via the SDA and SCL inputs

If the SDA and SCL pins are connected with the SDA and SCL pins of an external EEPROM (PCD8581/82) with series-address A0H, then after a hook-on/hook-off, or reset action the contents of the external EEPROM is read and stored in the internal EEPROM.

This EEPROM programming can only function when an external device is connected, because otherwise no acknowledge on the I^2C will be generated.

D6. Liquid Crystal Display

In principle every action on the keyboard will give an indication on a Liquid Crystal Display (LCD).
This software program uses the LTR700R-12 which is a 16-character, 7-segment with digital point module with additional indicators in Twisted Nematic technology.
The module is driven by an integrated driver PCF8576 in Chip-on-glass technology.
The complete layout of the display is given in figure D3.

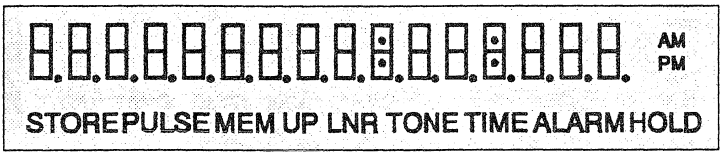


Figure D3. Complete layout of the LTR700R-12 LCD.

Outside a 7-segment display with digital point eleven labels are available of which six will be used in this program for the following functions:

- STORE	On	To indicate a program function is active (PROG-key).
- PULSE	On	The set is in pulse dial mode.
- MEM	On	One of the 10 repertory numbers is programmed or dialled.
- LNR	On	The last dialled number is redialled.
- TONE	On	The set is in or switched to DTMF dialling mode.
- HOLD	On	Then the set is in hold mode (with the HOLD-key).

The dialled telephone numbers are shown is the normal way, while the digital point indicates which number is dialled.

What exactly is shown on the LCD during the various dialling and programming modes is given the chapters where the functions are also explained.

The data to be displayed are transferred via the I^2C-bus to this LC-Display.

D7. Functional description and operating procedures

The PCD3755A has in total seven operation modes; standby, handsfree, listening_in, handset, ringer, dialling and programming.
The state diagram is given in figure D4.

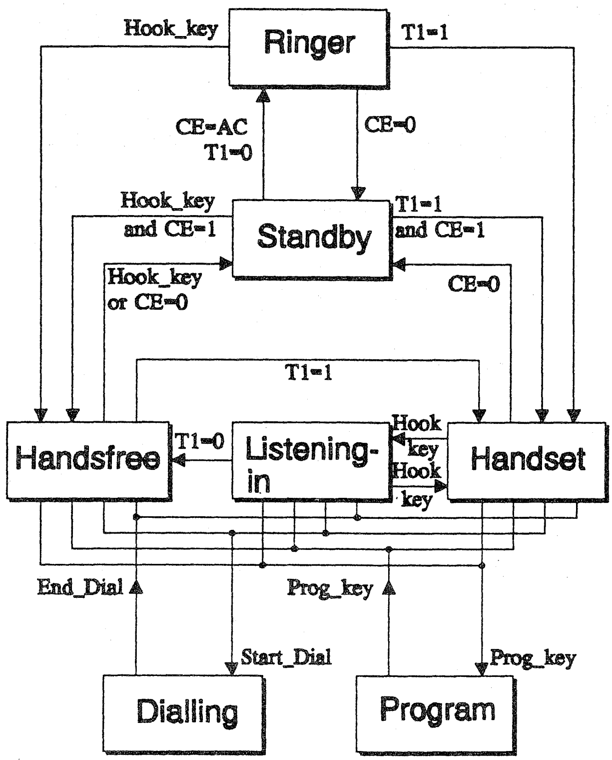


Figure D4. State diagram of the PCD3755A dialler/ringer

D7.1. Initialization

The initialisation starts when CE become high and ends when all the functions of the micro-computer and the PCA1070 (PACT) are set to the conversation state. This results in the following actions:

- The oscillator starts and the computer waits 1866 clock-pulses.
- All the outputs are put in the off-hook state, the internal RAM is cleared and the PCA1070 transmission parameters are copied from the EEPROM to the RAM.

When the micro-computer is activated from the stop mode an extra delay of 50 ms is added before the PACT parameters are sent to the PCA1070 to ensure that the supply voltage at the PCA1070 is high enough. However if the activation is done via the reset no delay is necessary, because then the supply voltage on the PCA1070 is ok.

- The cradle input (T1) and HOOK-key are tested, when T1="0" and no HOOK-key pressed the ringer mode is selected, if T1=1 the handset mode is selected (HSW = 0), if HOOK-key is pressed the handsfree mode is selected.
- The transmission parameters for handsfree or handset mode (stored in the EEPROM of the PCD3755A) are loaded into the PCA1070 with DST bit = HIGH to ensure short start-up time.
- After 150 ms this DST bit is resetted, by sending these transmission parameters again with DST bit = "0".
- Finally in both handset and handsfree mode the automatic gain control function is carried out. The DC line current is read via the I^2C bus from the PCA1070. This is compared with the values that are stored in the EEPROM for current start, stop and step resolution in dB and if necessary recalculates new receive and microphone gain. The new values will be send to the PCA1070.

Depending on the selected dialling mode the PULSE or the TONE label will be shown, as given in figure D5.

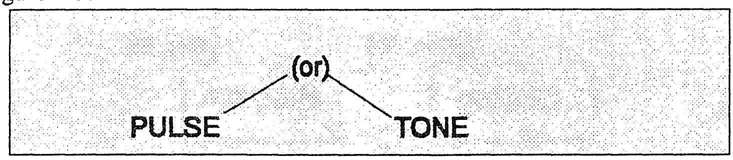


Figure D5. Display status during initialization (labels PULSE or TONE are on).

D7.2. On-hook

If in handsfree mode the HOOK-key is pressed, or in handset mode the craddle switch is opened, or the telephone line is disconnected then the micro-computer has to leave conversation mode and has to go to the stop mode to decrease current consumption.

When the handsfree mode is left by pressing the HOOK-key again then the telephone is disconnected by making output HSW LOW, the TEA1093 is disabled by making output MUTER and MUTET HIGH and the micro-computer is put direct in the STOP mode.

When the handset is layed-down or the telephone line is disconnected input CE/T0 of the micro-computer will become LOW, which results in the following actions:

1) The PCA1070 is put in power-down mode by sending the control byte with bits (PDx=01) to the PCA1070.

2) The CE/T0 input is tested during the reset delay time (t_{rds}), if it becomes HIGH again then the power-down mode is switched-off again by sending the control byte in which the PDx bits LOW to the PCA1070 and the micro-computer continues in the conversation mode.

3) After the reset delay time is ended (and CE/T0 is still low) the power-down mode is switched-off by sending the control byte in which the PDx bits LOW to the PCA1070 and the micro-computer is put in the STOP-mode.

D7.3. Control modes of the TEA1093 handsfree circuit

D7.3.1. Handsfree mode

Handsfree mode will be entered when one of the following events occurs:

- HOOK-key pressed in standby mode, during the above given initialization mode the here written actions also occur.
- HOOK-key pressed in ringer mode, the ringer mode is left and during the above given initialization mode the here written actions also occur.
- Handset layed down in listening-in mode, the here given actions are done direct after detection that the handset is layed down.

When entering this handsfree mode the following actions are to be taken by the micro-computer:

1) Activate the hook-switch output (HSW) by making it HIGH.

2) Enable the TEA1093 by making the two mute outputs MUTER and MUTET LOW.

3) Select the handsfree microphone by making RTO/HF HIGH.

4) Get the handsfree volume parameters from EEPROM and set the volume output pins VOL1 and when at pin 7 the VOL2 output is selected pin VOL2 to the correct value.

5) Put the PCA1070 in handsfree operation:
   a) Send the Gain microphone G_HF and Confidence gain G_conf (advised value -25dB) to the PCA1070.
   b) Put the PCA1070 in mute by sending the control byte with the receive and send mute bits HIGH (RM = 0, SM = 1).

D7.3.2. Voice switched listening-in mode LI1

Voice switched listening-in mode can only be entered from the handset mode by pressing the HOOK-key. The handset microphone signal is now routed through the TEA1093. The voice switching in the TEA1093 prevents howling.

Following actions are to be taken by the micro-computer for voice switching listening-in:

1) Activate the hook-switch output (HSW) by making it HIGH.

2) Enable the TEA1093 by making the two mute outputs MUTER and MUTET LOW.

3) Select the handset microphone path via TEA1093 by making RTO/HF LOW.

4) Get the handsfree volume parameters from EEPROM and set the volume output pins VOL1 and when at pin 7 the VOL2 output is selected pin VOL2 to the correct value.

5) Put the PCA1070 in voice switched listening-in operation:
   a) Send the Gain microphone G_LI1 and Gain receive handset G_rec to the PCA1070.
   b) Put the PCA1070 transmit part in mute by sending the control byte with the send mute bit HIGH (SM = 1).

D7.3.3. Basic listening-in mode LI2

Also this basic listening-in mode can only be entered from the handset mode by pressing the HOOK-key. The handset microphone signal is routed through the PCA1070 (no howling limitation).

Following actions are to be taken by the micro-computer for normal listening-in:

1) Activate the hook-switch output (HSW) by making it HIGH.

2) Enable the TEA1093 loudspeaker amplifier only by making the MUTER LOW and MUTET HIGH.

3) Get the handsfree volume parameters from EEPROM and set the volume output pins VOL1 and when at pin 7 the VOL2 output is selected pin VOL2 to the correct value.

4) Put the PCA1070 in basic listening-in operation by sending the Gain microphone G_LI2 and Gain receive G_rec to the PCA1070.

D7.4. Pulse / DTMF dialling function

The PCD3755A has two dialling modes, pulse dialling and Dual Tone Multi Frequency (DTMF) dialling. These modes can be programmed via keyboard (see chapter 8.3.4. "Programming of the dialling options") and will then be stored in the corresponding bit in the EEPROM of the micro-computer.

D7.4.1. Pulse dialling (the dialling mode option bit = LOW)

The keyboard entry initiates a recall of a previously stored number or is a simultaneous keying-in and pulsing-out activity, with storing for possible later recall.
If at keying-in the keys * or # are used this will result in a switch to the tempory DTMF dialling mode, all keys after this digits pressed will be send out in DTMF dialling mode.
Figure D6 shows the display status for a manual dialled number in pulse dial mode with digits 1-2-3-4-5-6-7-8-9-0, here the decimal point indicates which of the digits is dialled at that moment.

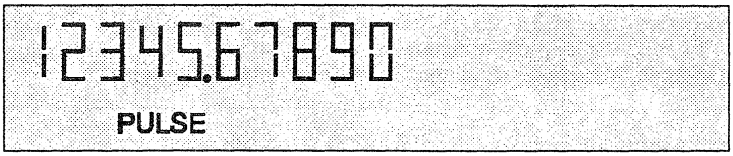


Figure D6. Display status at pulse dialling with number 1-2-3-4-5-6-7-8-9-0.

In this application everything necessary for correct line current interruption and muting is done in or via the PACT. The micro-computer makes the timing sequence and sends the correct commands to the PACT via the I^2C-bus as follows:

1) First mute the TEA1093 transmit path by making MUTET HIGH.

2) Set the volume at minimum, outputs VOL1 and if selected VOL2 LOW.

3) Put the PACT in the mute mode:
   a) Send the Gain confidence G_conf and Gain DTMF G_DTMF belonging by the mute status to the PCA1070.
   b) Switch-on the DST bit (for fast switching behaviour) and send the DC voltage belonging with pulse dialling (DC_u_dmo) to the PCA1070.
   c) Activate the actual mute by sending the control byte with the receive and send mute bits HIGH (RM = SM = 1).

4) Wait an inter-digit pause (840 mS).

5) Break: the line current interruption is done by sending a control byte in which the dial pulse input bit (DPI=1) and the power-down bits (PDx=01) are HIGH, of-course the mute bits are also still HIGH.

6) Make: the line current is switched-on again by sending the control byte in which the DPI and the PDx bits are LOW again, also here the mute bits stay HIGH.

Points 5 and 6 are repeated depending on the number of interruptions belonging to the corresponding digit.

7) At the end of the dialling string the PCA1070 is switched back to the conversation mode by sending the control byte with the receive and send mute bits LOW (RM = SM = 0).

8) After an extra delay of 100 ms the DST-bit and the gains are restored into the PCA1070 as follows:
   a) Switch-off the DST bit.
   b) Send the original receive and microphone gain to the PCA1070.

9) Finally return to the state in which the set was before pulse dialling was started, e.g. handset, handsfree or listening-in, by output MUTET.

The total pulse period time is 100 ms (10 Hz) with a mark-to-space ratio of 3:2 or 2:1, selected by an EEPROM bit (programmable via the keyboard). After transmission of a digit, the next digit is processed, again starting with an inter-digit pause.

D7.4.2. DTMF dialling (the dialling mode option bit = HIGH)

The PCD3755A converts keyboard inputs into data for the on-chip DTMF generator. Tones are transmitted via output TONE with a minimum tone burst/pause duration of 100/100 mS.
The maximum tone burst duration is equal to the key depression time.
With redial and repertory dialling tones are automatically fed at the programmed rate.
Figure D7 shows the display status for a manual dialled number in DTMF dial mode with digits 1-2-3-4-5-6-7-8-9-0-*-#, here the decimal point indicates which of the digits is dialled at that moment.

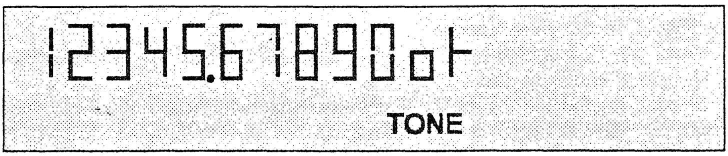


Figure D7. Display status at DTMF dialling with number 1-2-3-4-5-6-7-8-9-0-*-#.

For good DTMF operation the sending and receiver mutes as well as the send and receive gain of the PACT have to be adapted to the DTMF status via the I^2C-bus as follows:

1) First mute the TEA1093 transmit path by making MUTET HIGH.

2) Set the volume at minimum, outputs VOL1 and if selected VOL2 LOW.

3) Prepare the PCA1070 amplifiers by sending the new gain value for the send progamp (G_DTMF) and receive progamp (G_conf).

4) Activate the actual mute by sending the control byte with the receive and send mute bits HIGH (RM = SM = 1).

5) Generate a series of DTMF tones / pauses which are transmitted to the PACT via the tone output.

6) At the end of the dialling string the PCA1070 is switched back to the conversation mode by sending the control byte with the receive and send mute bits LOW (RM = SM = 0).

7) Re-install the normal amplifier values by sending the original in the PCD3755A stored microphone and receiver gain to the PCA1070.

8) Finally return to the state in which the set was before DTMF dialling was started, e.g. handset, handsfree or listening-in, by output MUTET.

D7.4.3. DTMF dialling in pulse dialling mode (mixed mode dialling)

If the controller is set to the pulse dial mode (the dialling mode option bit = LOW), activation of push-button * or # changes the dialling mode to DTMF.
Its entry is stored and all digits following are transmitted in the DTMF mode.
The digits entered after key TONE are not transmitted in the redial mode.
The * or # key which causes the switch-over is never transmitted.
The * and # keys pressed after a switch over to DTMF dialling are all transmitted.
If the controller is initially set to the DTMF mode, activation of push-button * or # are stored in the redial register and transmitted in DTMF mode.

D7.4.4. Flash function

Flash function is activated by the Flash-key which results in a timed line current break (100 ms) done with the PCA1070 controlled via I^2C-bus.
During the Flash function the keyboard is not scanned.

If the Flash is the second or later key pressed then all the digits added before the Flash are stored in the redial register (internal EEPROM) and the Flash action functions as a hook-on / hook-off cycles.

Thus:

- After dialling 1-2-3-"Fl"-on-hook/off-hook Redial is 1-2-3.
- After dialling 1-2-3-"Fl"-4-5-6-on-hook/off-hook Redial is 4-5-6

The timing sequence for Flash is as follows:

1) First mute the TEA1093 transmit part by making MUTET HIGH.

2) Set the volume at minimum, outputs VOL1 and if selected VOL2 LOW.

3) Mute the PCA1070 and interrupt the line current:
   a) Send the confidence and DTMF gain belonging by the mute status to the PCA1070.
   b) Switch-on the DST bit, to improve the switching behaviour.
   c) Activate the actual mute by sending the control byte with the receive and send mute bits HIGH (RM = SM = 1).
   d) Interrupt the line current by sending the control byte in which the DPI and PDx-bits are also HIGH, of course the mute bits are also still HIGH.

4) Wait for the FLASH-time (100 ms).

5) Put the PCA1070 back in conversation mode as follows:
   a) The line current is switched-on again by sending the control byte in which the DPI and the PDx bits are LOW again, also here the mute bits stay HIGH.
   b) Switch back to the conversation mode by sending the control byte with the receive and send mute bits LOW (RM = SM = 0).
   c) Switch-off the DST bit.
   d) Send the original receive and microphone gain to the PCA1070.

6) Finally return to the state in which the set was before this FLASH action was started, e.g. handset, handsfree or listening-in, by output MUTET.

After a FLASH the display is cleared and the micro-computer status is identical to that after start-up.

D7.4.5. Mute function (MUTE-key)

This mute function is used as privacy switch and is a toggle function. When no dialling is active, every time that the Mute key is pressed the receive mute (RM) and send mute (SM) bits are inverted and send to the PCA1070.

When another key is pressed and the mute bits where set then this function is switched off.

D7.4.6. Hold function (HOLD-key)

The hold function is used to switch the call to a parallel telephone set without disconnecting the line. A press on the hold key will cause the following actions:

1) First mute the TEA1093 by making MUTER and MUTET HIGH.

2) Mute the PCA1070 by sending the control byte with the receive and send mute bits HIGH (RM = SM = 1).

3) Read the line current via the I^2C-bus from the PCA1070.

4) Switch-on the HOLD label on the display.

5) If the handset was lifted this can be layed down, but the micro-computer stays in the hold mode (off-hook) till one of the following actions occur:
   a) The handset is lifted again, then the program returns to the handset mode.
   b) The HOOK-key is pressed, then the program returns to the handsfree mode.

6) In the hold mode, the line current which flows through the PCA1070, is continuously read by the PCD3755A via the I^2C-bus, if the latest measured line current is 5 mA lower than the line current measured at point 3 then the PCD3755A will go to the stop mode.

D7.4.7. Volume control (+/- keys)

During the conversation mode it is possible to change the output volume of the loudspeaker by pressing the + or - key.
A press on the + key will increase the volume by one step, but the contents of the volume buffer will never be higher then value 4 (see table below).
However, a press on the - key will result in a decrease of the volume by one step, but here the contents volume buffer will never be lower then value 1 (see table below).

If pin 7 is configurated as VOL2 output, then there are four volume steps possible, however if this pin 7 is used as dynamic limiter input DL, then only two steps are available (values 1 and 2).

The loudspeaker volume can be controlled by the port pins VOL1 and VOL2 and the last programmed value is stored in EEPROM.
The relation between the value stored in EEPROM and the outputs is given below:

Value	µC port pins		Volume
	VOL1	VOL2	
1	0	0	minimum
2	1	0	
3	0	1	
4	1	1	maximum

D7.5. Number storage function

D7.5.1. Last number redial (1 to 24 digits)

Recalling the last number which has been dialled and is stored in memory is done by pressing the LNR-key as the first key after hook-off.

If the first push-button pressed is 0 to 9 in pulse dialling or 0 to 9, * and # in DTMF dialling mode, digits are entered into the work register.

During the data entry the circuit starts immediately with transmission of the digit(s) and the minimum transmission time is unaffected by the speed of entry.
Transmission continues as long as further data input has to be processed.

Figure D8 gives the display for a REDIAL dialled number in pulse dial mode with digits 0-9-8-7-6-5-4-3-2-1, also here the decimal point indicates which of the digits is dialled at that moment.

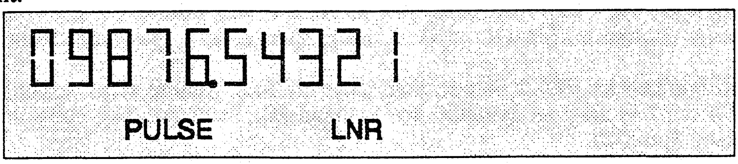


Figure D8. Display status at REDIAL with number 0-9-8-7-6-5-4-3-2-1.

Up to 24 digits can be stored in the redial register.
After the work register overflows, a 10 digits First-In-First-Out register (FIFO) takes over as buffer and the contents of the work register is not copied to the redial register.
After transmitting the first digit of the FIFO register this position is automatically cleared to provide space for the storage of new data. In this way, the total number that can be transmitted is unlimited, provided the key-in rate is not excessive.
However, if the FIFO register overflows (more than 10 digits in store) further input is ignored.

D7.5.2. 10-number repertory dialling

The PCD3755A includes a 10-number repertory dialler, 18 digits each, which is accessible with a one-key procedure.
The 10 repertory numbers can be recalled with the M0 to M9 keys.
Figure D9 gives the display for repertory number M0 in pulse dial mode with digits 0-7-0-AP-3-2-2-5-6-7-4, again the decimal point indicates which of the digits is dialled at that moment.

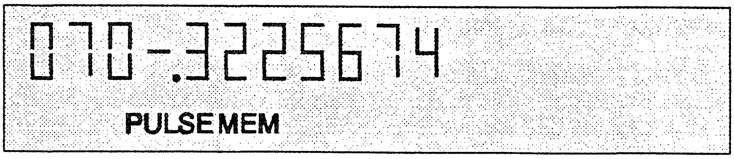


Figure D9. Display status at repertory M0 with number 0-7-0-AP-3-2-2-5-6-7-4.

The maximum length of these repertory numbers is 18 digits including the manually stored access pauses.

Chain dialling: Repertory numbers can be dialled-out after or before entering manual dialling, last number redial and by entering the memory locations in successive order.
During transmission of a number called from the memory location, the controller does not accept keyboard entries.
Dialling can be continued as soon as the number under transmission is completed.

The display will always show the last dialled repertory number.

Note that the last memory location which is transmitted is stored in the redial register.

D7.5.3. Access pause storage and activation

If during entering a telephone number via keyboard for normal dialling or during repertory number programming the AP-key (access pause key) is pressed, then an access pause is stored in the redial or repertory dial register. This will be indicated at the LCD by a horizontal bar see figure D9.

In this program one fixed wait (access pause) time is used in both pulse and DTMF dialling mode namely 3 seconds.

D7.5.4. Programming of the dialling and I/O options

In the dialling option part the pulse/DTMF dialling selection, mark-to-space ratio at pulse dialling and the Dynamic Limiter input/VOL2 output selection can be changed and stored. All these three options can be changed with this programming procedure.

Programming is only possible in off-hook mode, because supply is necessary.

Changing these options is done as follows:

- depress Prog (universal program key, LCD-label STORE is on)
- press the LNR-key (options mode, LCD shows stored modes see Fig. D10/D11)
- press key 1,2,3,4,5 or 6 (the LCD gives direct the new options see Fig. D10/11)
  in which key 1 = pulse dialling
  key 2 = DTMF dialling
  key 3 = 3:2 (60/40 ms)
  key 4 = 2:1 (66/33 ms)
  key 5 = Pin 7 is Dynamic limiter input (DL)
  key 6 = Pin 7 is VOL2 output
- depress Prog again (new modes are stored into EEPROM and LCD is cleared).

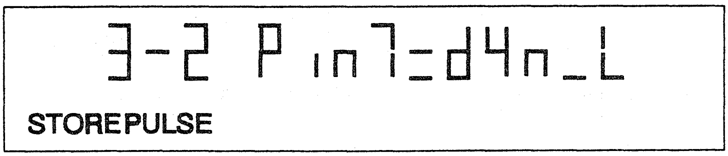


Figure D10. Display status during option programming when keys 1, 3 and 5 are pressed.

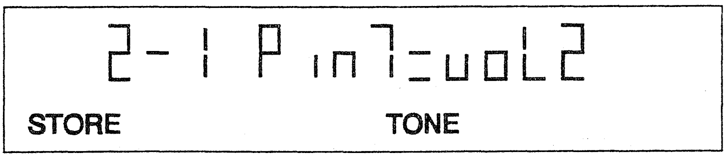


Figure D11. Display status during option programming when keys 2, 4 and 6 are pressed.

During programming the mute toggle function is active.

At the begin of the programming procedure (by pressing the Prog key) the dial functions are blocked, and after programming the dial functions are active again.

When during the program cycles a mistake has been made in the programming procedure then the program mode can always be left by pressing the Prog key (but then the already changed options will be stored in EEPROM) or by going on-hook (then nothing will be stored into EEPROM).

D7.5.5. Storing repertory numbers

The store mode starts after going off-hook and depressing the Prog-key.

In total 10 numbers can be stored.

Repertory numbers can be stored into EEPROM via the following one key access method:

- depress Prog (universal program key, LCD label STORE is on),
- depress one of the ten location keys (M1 to M10),
- enter the telephone number (maximum 18 digits inclusive AP, info on LCD)
- depress Prog key again (the new number will be stored into EEPROM, clear LCD).

Figure D12, gives the LC-Display when a repertory number 0-AP-0-9-AP-4-1-1-6-7-8-8-7-6 is pressed.

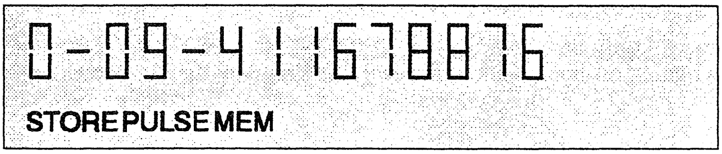


Figure D12. Display status during programming repertory number 0-AP-0-9-AP-4-1-1-6-7-8-8-7-6

The next number can now be stored by repeating the above procedure.

Memory clear: Cleaning of the memory location is possible via the same procedure as for storing a number, but now no telephone number is entered.

D7.6. PCA1070 (PACT) programming and evaluation

The PCA1070 variables, dialling options and ringer parameters which are stored in the EEPROM of the µC can be changed via the keyboard.
For programming of the PCA1070, five special programming keys (<DCVOLT>, <GAIN MIC>, <GAIN REC>, <SET IMP>, <SIDE TONE> are available.
The keyboard programming modes can be activated by pressing key <PROG> and this is indicated at the LCD by the label <STORE>. Subsequently one the programming modes can be chosen.

For demonstration purposes all in EEPROM stored PCA1070 variables can be changed via keyboard.

The values can be observed via the LC-Display.

The five special PACT programming and evaluation keys make the following corrections possible:

- DC-settings (normal and at pulse dialling) and AGC.
- Gain-microphone HS, gain-microphone HF, gain-microphone LI1, gain-microphone LI2, selection of LI1/LI2, gain-DTMF, and dynamic-limiter threshold.
- Gain-receive, gain-confidence, hearing-protection level and load select bit.
- Set-impedance (Ra, Rb and fpole).
- Sidetone balance impedance (RSa, RSb and Cs).

During all program modes the set must be in handset or handsfree mode, because supply is necessary, also the display shows clearly which mode is selected and what the new programmed value is.
The new programmed values will be active after an on-hook off-hook action.

The PCA1070 values that are changed by keyboard are send directly to the PCA1070. The EEPROM contents, however, will only be changed after ending the program procedure by pressing the program key <PROG>.

The new values will only be stored into the EEPROM after ending the procedure with the Prog-key, when going on-hook without pressing the Prog key the old values will be in the EEPROM.

D7.6.1. Programming the DC-settings and automatic gain control via keyboard

These in total ten programmable values can be changed and stored via one programming procedure.
Direct evaluation is only active for the new DC-voltage value, the changed DC-voltage value during dialling and the changed AGC values are active after ending the program procedures.

The programming procedure is as follows:

- Depress Prog (universal program key, LCD label STORE is on).

- Depress DC-key.
  a) First press, the DC-voltage program part is selected, the display status is given in figure D13 for a programmed voltage of 5.9 volt.

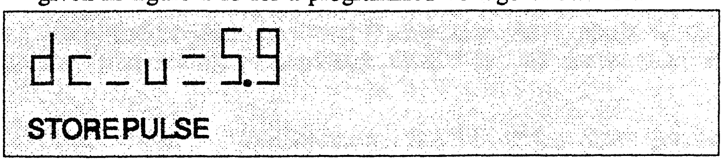


Figure D13. Display status at DC-programming showing the DC-voltage of 5.9 volt

  b) Second press, the DC-voltage at pulse dialling (DMO) program part is selected, the display status is given in figure D14 for a DMO voltage of 4.3 volt.

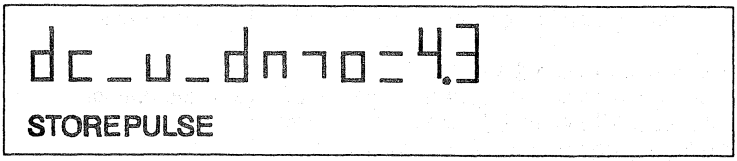


Figure D14. Display status at DC-programming showing the DMO-voltage of 4.3 volt

c) Third press, the DC-current START (dc-i-0) program part is selected, this is on the LCD given as DC-current_0 point see figure D15.

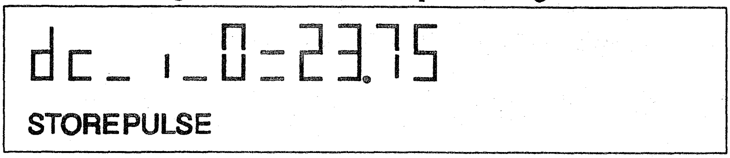


Figure D15. Display status at DC-programming showing the DC-current-0 of 23.75 mA

d) Fourth press, the DC-current first value program part is selected, on the display the 0 in figure D15 is replaced by a 1.
e) Fifth press, the DC-current second value program part is selected, on the display the 0 in figure D15 is replaces by a 2.
f) Sixth press, the DC-current third value program part is selected, on the display the 0 in figure D15 is replaced by a 3.
g) Seventh press, the DC-current fourth value program part is selected, on the display the 0 in figure D15 is replaced by a 4.
h) Eighth press, the DC-current fifth value program part is selected, on the display the 0 in figure D15 is replaced by a 5.
i) Ninth press, the DC-current STOP (dc-i-6) program part is selected, on the display the 0 in figure D15 is replaced by a 6.
j) Tenth press, the step resolution of gain control (dc-i-d-G) is selected, on the display this is shown as given in figure D16, actual value 1dB/step).

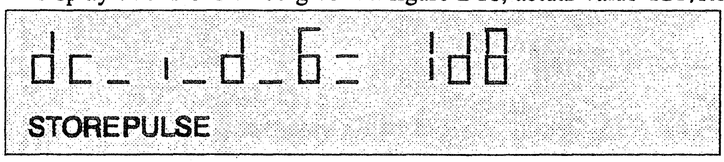


Figure D16. Display status at DC-programming showing a delta gain of 1 dB.

k) Tenth press, is identical to the first press.......goes to a)

- Depress the + or - key — When depressing the + or - key automatically the one higher or one lower value for the selected item will be selected. If the maximum or minimum value is reached the program stop.

- Depress Prog key This ends the DC programming part the corrected DC-voltage, DC-current and delta-gain values will be stored in EEPROM and the LCD is cleared.

D7.6.2. Programming the sending channel characteristics via keyboard

These in total seven programmable values microphone gain, handsfree gain, voice switched listening-in gain, basic listening-in gain, selection of the listening-in mode, threshold of dynamic limiter and the DTMF gain can be changed and stored via one programming procedure.
The pact evaluation mode is active for the microphone gain, handsfree gain, and threshold of dynamic limiter, the new DTMF gain value, voice switched listening-in gain, basic listening-in gain and the selection of listening-in mode are active after ending the program procedures.
The programming procedure is as follows:

- Depress Prog (universal program key, the LCD label STORE is on).

- Depress GAIN-MIC key
  - a) First press, the gain-microphone program part is selected, the display status for a microphone gain of 13 dB is given in figure D17.

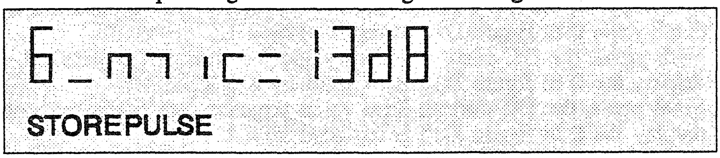


Figure D17. Display status at microphone gain showing a value of 13 dB.

  - b) Second press, the dynamic-limiter threshold program part is selected, the display status for DLT is off is given in figure D18.

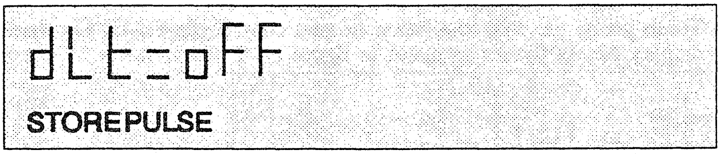


Figure D18. Display status at dlt showing that the function is off.

  - c) Third press, the gain-DTMF program part is selected, the display status for a DTMF gain of 1 dB is given in figure D19.

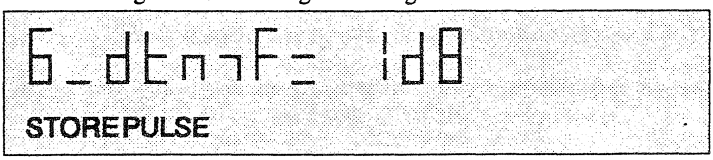


Figure D19. Display status at DTMF gain showing the value of 1 dB.

d) Fourth press, the gain-handsfree program part is selected, the display status for a handsfree gain of 3 dB is given in figure D20.

```
G_HF= 3dB
STOREPULSE
```

Figure D20. Display status at handsfree gain showing the value of 3 dB.

e) Fifth press, the gain-voice switched listening-in program part is selected, the display status for a voice switched listening-in gain of 4 dB is given in figure D21.

```
G_Li1= 4dB
STOREPULSE
```

Figure D21. Display status at voice switched listening-in gain showing the value of 4 dB.

f) Sixth press, the gain-basic listening-in program part is selected, the display status for a basic listening-in gain of 5 dB is given in figure D22.

```
G_Li2= 5dB
STOREPULSE
```

Figure D22. Display status at basic listening-in gain showing the value of 5 dB.

f) Seventh press, the listening-in mode selection program part is selected, the display status for the voice switched listening-in mode is given in figure D23, in case that basic listening-in is selected the "1" on display is changed to a "2".

```
Li= 1
STOREPULSE
```

Figure D23. Display status at voice switched listening-in mode.

g) Eighth press, is identical to the first press........goes to a).

- Depress the + or - key — When depressing the + or - key automatically the one higher or one lower value for the selected item will be selected. If the maximum or minimum value is reached the program stop.

- Depress Prog key — This will end the GAIN-MIC programming part all the corrected gain-microphone, dynamic-limiter and gain-DTMF values will be stored in EEPROM and the LCD is cleared.

D7.6.3. Programming the receive channel via keyboard

These in total four programmable values receive gain, hearing protection level, load select bit and confidence gain can be changed and stored via one programming procedure.
The pact evaluation mode is active for the receive gain, hearing protection level and load select bit, the new confidence gain value is active after ending the program procedures.

The programming procedure is as follows:

- Depress Prog (universal program key, LCD label STORE is on).

- Depress GAIN-REC key.
  a) First press, the gain-receiver program part is selected, the display status for a receiver gain of 3 dB is given in figure D24.

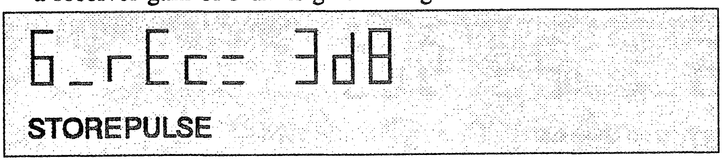


Figure D24. Display status at receiver gain showing the value of 3 dB.

  b) Second press, the hearing-protection level program part is selected, the display status for hpl is on is given in figure D25.

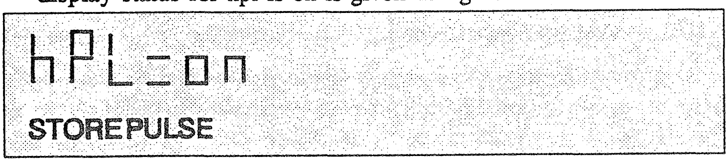


Figure D25. Display status for hpl function is on.

  c) Third press, the load select bit (capacitive/resistive) program part is selected, the display status for RFC is off (resistive load) is given in figure D26.

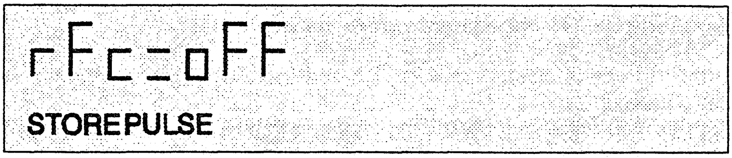


Figure D26. Display status for RFC = 0 (resistive load).

d) Fourth press, the gain-confidence program part is selected, the display status for a confidence tone gain of -25 dB is given in figure D27.

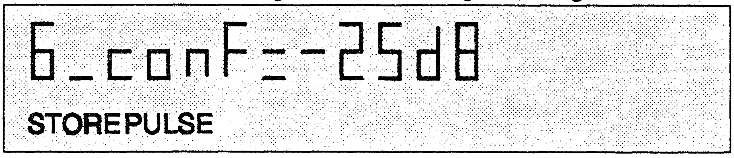


Figure D27. Display status at confidence gain showing the value of -25 dB.

e) Fifth press, is identical to the first press........goes to a).

- Depress the + or - key When depressing the + or - key automatically the one higher or one lower value for the selected item will be selected. If the maximum or minimum value is reached the program stop.

- Depress Prog key This will end the GAIN-REC programming part all the corrected gain-receiver, gain-confidence,receiver-feedback and hearing-protection values will be stored in EEPROM and the LCD is cleared.

D7.6.4. Programming the set-impedance values via keyboard

These in total three programmable values Ra, Rb and fpole can be changed and stored via one programming procedure.
The pact evaluation mode is active for all the changed values (Ra, Rb and fpole).

The programming procedure is as follows:

- Depress Prog (universal program key, LCD label STORE is on).

- Depress SET-IMP key.
  a) First press, the set-impedance Ra program part is selected, the display status for a set impedance Ra = 200 ohm is given in figure D28.

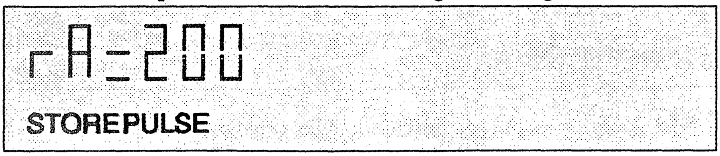


Figure D28. Display status at set impedance Ra, showing the Ra value of 200 ohm.

  b) Second press, the set-impedance Rb program part is selected, the display status for a set impedance Rb = 800 ohm is given in figure D29.

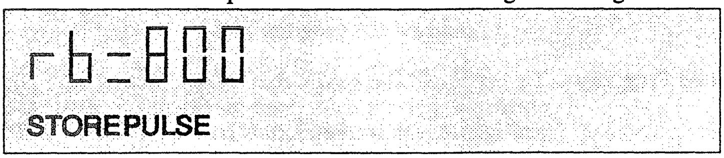


Figure D29. Display status at set impedance Rb, showing the Rb value of 800 ohm.

  c) Third press, the set-impedance fpole program part is selected, the display status for a set impedance Fpole = 1915 Hz is given in figure D30.

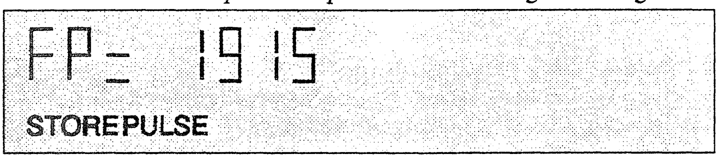


Figure D30. Display status at set impedance Fpole, showing the Fpole value of 1915 Hz.

  d) Fourth press is identical to the first press........goes to a).

- Depress the + or - key — When depressing the + or - key automatically the one higher or one lower value for the selected item will be selected. If the maximum or minimum value is reached the program stop.

- Depress Prog key — This will end the SET-IMP programming part all the corrected set-impedance values Ra, Rb and fpole will be stored in EEPROM and the LCD is cleared.

D7.6.5. Programming the sidetone balance impedance via keyboard

These in total three programmable values RSa, RSb and Cs can be changed and stored via one programming procedure.
The pact evaluation mode is active for all the changed values (RSa, RSb and Cs(.

The programming procedure is as follows:

- Depress Prog (universal program key, LCD label STORE is on).

- Depress SIDE-TONE key.
  a) First press, the sidetone RSa program part is selected, the display status for a sidetone RSa = 492 ohm is given in figure D31.

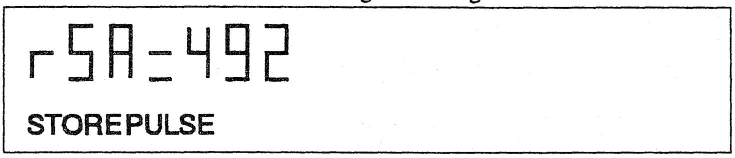


Figure D31. Display status at sidetone RSa, showing the RSa value of 492 ohm.

  b) Second press, the sidetone RSb program part is selected, the display status for a sidetone RSb = 1259 ohm is given in figure D32.

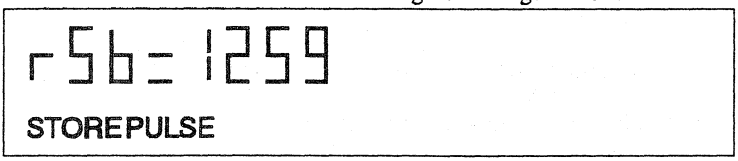


Figure D32. Display status at sidetone RSb, showing the RSb value of 1259 ohm.

  c) Third press, the sidetone Cs program part is selected, the display status for a sidetone Cs = 134 nF is given in figure D33.

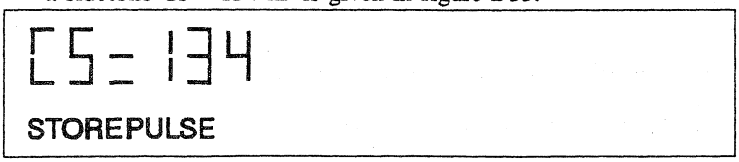


Figure D33. Display status at sidetone Cs, showing the Cs value of 134 nF.

d) Fourth press, is identical to the first press........goes to a).

- Depress the + or - key — When depressing the + or - key automatically the one higher or one lower value for the selected item will be selected. If the maximum or minimum value is reached the program stop.

- Depress Prog key — This will end the SIDE-TONE programming part all the corrected sidetone values RSa, RSb and Cs will be stored in EEPROM and the LCD is cleared.

D7.7. AGC control

The Automatic Gain Control (AGC) function allows correction of the receiver and microphone gain depending on the actual line current, this mode is active during handset and handsfree mode.

The AGC routine can be switched-off by making the AGC delta gain value equal to 0 dB. If active then the current is read from the PCA1070 and new gains are set every 200 ms, which makes testing and measurements easy.

The AGC-function works as follows:

- First the actual current is read from the PCA1070.
- Secondly the computer calculates the new value for receive and microphone gain by comparing the value read from the PCA1070 with the programmed start current value (dc_i_0), is the current higher then the gain is decreased with the programmed delta gain value. When the read current is bigger the current is compared with the next value (dc_i_1) and if necessary decreased again. This is repeated till or the current is lower or the last (dc_i_5) value is compared.
- During all these calculations it is tested if the function comes above the stop current point (dc_i_6), because then the gain decreasing will stop.
- Final the new values of receive and microphone gain are send to the PCA1070.

D7.8. Ringer function

The PCD3755A has a three-tone melody ringer with the following characteristics.

- Ringer volume change during conversation and ringer mode.
- Ringer melodies selection.
- Ringer repetition rate selection.
- Ringer input frequency measurement.

In figure D34 the timing diagram of the ringer function is given.

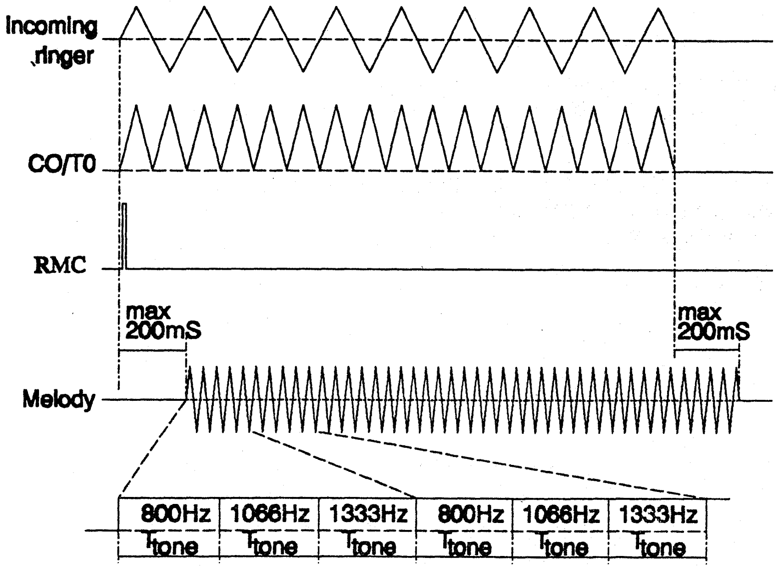


Figure D34. The timing diagram of the ringer function.

The ringer signal is sent via the special ringer output pin (RTO). At this output a block with a peak-to-peak output voltage of V_{DD}-V_{SS} is generated.

D7.8.1. Ringer option change during ringer mode (volume only)

If during the ringer output is active keys 1 to 4 are pressed then the port pins VOL1 and if pin 7 is configurated as VOL2 will go to the corresponding level and its value is stored in EEPROM.

D7.8.2. Ringer option change during conversation mode (all)

The ringer output volume, melody and speed and also the ringer input frequency detection can be changed by the following program procedure:

- Depress Prog-key (universal program key, LCD label STORE is on).
- Press Flash-key (ringer program key, LCD shows the ringer status see Fig. D35).
- Depress key 1 to 4

Where:	Key-1	Volume	(u=..)
	Key-2	Melody	(m=..)
	Key-3	Speed	(S=..)
	Key-4	Input frequency	(F=..)

- Depress key 1 to 4 (with input frequency selection 1 to 2)
  This will automatically change the selected mode (volumes, melodies, speeds and input frequencies), to the value belonging to the pressed key. The LCD gives direct the new ringer options see figure D35.
  Now the routine can be repeated by select volume, melody, speed or input-frequency, as done above.
- Depress PROG-key (the program cycles is ended and new value will be stored in EEPROM and the LCD is cleared.

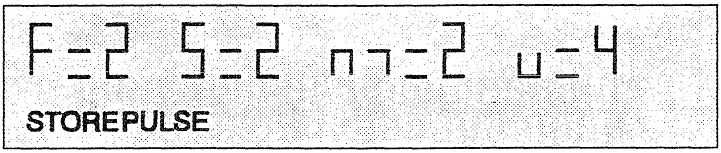


Figure D35. Display status in the ringer programming mode.

If the above routine is interrupted by an on-hook/off-hook action the procedure is ended and the old programmed values stay in the EEPROM.

D7.8.3. Ringer volume selection during conversation mode

The ringer volume can be controlled by the port pins VOL1 and VOL2 and its value is stored in EEPROM.
The status of the output pins, and thus also the ringer output volume, can be programmed via keyboard during conversation and ringer mode. This in conversation or ringer mode changed values will be stored in the EEPROM for use later. The relation between the keys and selected output status is given below:

Value	μC port pins		Volume
	VOL1	VOL2	
1	0	0	minimum
2	1	0	
3	0	1	
4	1	1	maximum

D7.8.4. Ringer melodies selection

The ringer melody generator can select out of four melodies. The selection can be done via keyboard, the new selected melody option will the be stored in the EEPROM. The four possible melodies are given below:

Value	Melody		
	f1 (Hz)	f2 (Hz)	f3 (Hz)
1	738	826	925
2	800	1067	1333
3	1455	1621	1810
4	1955	2223	2510

D7.8.5. Ringer repetition rate selection (speed)

The generated melody is built up out of three frequencies. These frequencies are generated successively in a selected repeat frequency. There are four steps and they can be programmed via keyboard after which the new value is stored in the EEPROM. The relation between the pressed keys and the repetition rate is given below:

Value	Speed	
	Frequency (Hz)	Tone time (ms)
1	7	47.6
2	11	30.3
3	15	22.2
4	20	16.6

D7.8.6. Ringer input frequency measurement

The melody ringer becomes active for all incoming ringer frequencies higher then the ringer detection LOW frequency and lower then the ringer detection HIGH frequency supplied to the CE/T0 input of the PCD3755A. The ringer detection LOW and ringer detection HIGH frequencies are selected such that it is possible to use this program for both single or double phase rectifier applications. Therefore this program has two ringer detection LOW and two ringer detection HIGH frequencies options. The selection can be done via keyboard and will be stored in the EEPROM for later use, the selectable values are given below:

Value	Input frequency (at pin T0)		
	Lower limit (Hz)	Higher limit (Hz)	Method
1	20	60	1f
2	40	120	2f

D8. PCD3755A Hardware characteristics

Here the hardware port options the oscillator type and the power-on-reset voltage of the actual device are given.

Pin number	Pin name	Port option	Reset option	OEF(1)notation
28	HSW	Standard I/O	HIGH	1S
1	COL1	Standard I/O	HIGH	1S
2	COL2	Standard I/O	HIGH	1S
3	COL3	Standard I/O	HIGH	1S
4	COL4	Standard I/O	HIGH	1S
5	COL5	Standard I/O	HIGH	1S
6	COL6	Standard I/O	HIGH	1S
7	DL/VOL2	Standard I/O	HIGH	1S
13	ROW1	Standard I/O	HIGH	1S
14	ROW2	Standard I/O	HIGH	1S
15	ROW3	Standard I/O	HIGH	1S
16	ROW4	Standard I/O	HIGH	1S
17	ROW5	Standard I/O	HIGH	1S
18	ROW6	Standard I/O	HIGH	1S
19	VOL1	Standard I/O	LOW	1R
20	RTO/HF	Standard I/O	LOW	1R
21	MUTER	Open-drain	HIGH	2S
25	MUTET	Open-drain	HIGH	2S
26	SDA	Open_drain	HIGH	2S
27	SCL	Open_drain	HIGH	2S

Other options	Type
Oscillator option	gml
Power-on-reset voltage	1.3 volt

Note (1): The abbreviation OEF stands for Order Entry Form and is the standard Faselec Zürich start form for mask making. On this OEF the here given notation are used.

D9. TIMING CHARACTERISTICS

V_{DD}=3V; V_{SS}=0V; T_{amb}=-25 to 70°C; f_{XTAL}= 3.579545 MHz (Gml).

parameter	symbol	min.	typ.	max.	unit
General					
Reset delay time	t_{rds}	155	160	165	ms
Keyboard debounce time	t_{db}	15	20	25	ms
Flash time	t_{fl}	90	100	110	ms
Access pause time	t_{pause}	2.9	3.0	3.1	s
Pulse dialling					
Dial frequency	f_d	9.5	10	10.5	Hz
Break time (M/S=3:2)	t_b	57	60	63	ms
Make time (M/S=3:2)	t_m	38	40	42	ms
Break time (M/S=2:1)	t_b	63.6	66.6	69.6	ms
Make time (M/S=2:1)	t_m	31.3	33.3	35.3	ms
Inter-digit pause	t_{idp}	800	840	900	ms
DTMF dialling					
Tone on/off time (automatic dial)	t_{on}	95	100	105	ms
	t_{off}	95	100	105	ms
Tone on/off time (manual dial)	t_{on}	95	-	key-on time	ms
	t_{off}	95	-	key-off time	ms

APPLICATION NOTE Nr	ETT/UM96006.0
TITLE	OM4784 System board of TEA1069N, TEA1093 and UBA1702/A
AUTHOR	E. M. Bosma
DATE	March 1996

CONTENTS

LIST OF FIGURES

LIST OF TABLES

1. INTRODUCTION

1.1 Purpose

This document is the user manual for the OM4784 system board. This kit has been designed to demonstrate the Speech/Dialler/Ringer IC TEA1069N in combination with the TEA1093 and the UBA1702/A. This document describes the functionality, the component layout, the electrical schematic and the bill of materials of the basic application board with these ICs.

1.2 Scope

This document describes all the information of the board, necessary to work with the board and to change the settings of the board. The test of functionality and performance is outside the scope of this report. This document is not intended as an Application Note.

1.3 Definitions, Acronyms and Abbreviations

A-B/B-A	Line terminals of application
AGC	Automatic Gain Control; line loss compensation facility
AP	Access Pause
BOM	Bill Of Materials
BRL	Balance Return Loss
DMO	Dial Mode Operation
DTMF	Dual Tone Multi Frequency
EMC	Electro Magnetic Compatibility
HF	Handsfree
LED	Light Emitting Diode
LNR	Last Number Redial
MIC	Microphone input
MOH	Music On Hold
PCB	Printed Circuit Board
PTS	Pulse / Tone Selection.
RCL	Recall
Speaker-phone button	Button "HOOK" on keypad.
TEL	Earpiece output
UBA1702	Line interrupter driver and ringer for PMOST
UBA1702A	Line interrupter driver and ringer for PNP
UBA1702/A	General indication of the UBA1702 as well as the UBA1702A

1.4 References

[1] Philips Semiconductors device specification TEA1069N

[2] Philips Semiconductors Application Note ETT/AN95023. "Application of the UBA1702/A line interrupter driver and ringer circuit", by H. Derks 01-Nov-95

[3] Philips Semiconductors Application Note ETT/AN93015. "Application of the TEA1093 handsfree circuit" by R. v. Leeuwen 30-Nov-95

2. GENERAL DESCRIPTION

The OM4784 system kit exists of a printed circuit board PR46932, a handset with handset cord and a userman-ual. The system kit is suited for the TEA1069N Speech/Dialler/Ringer. This IC integrates three parts:

1. Speech/Transmission
2. Dialler
3. Ringer

The purpose of the system board is to evaluate and demonstrate the features of the TEA1069N one-chip tele-phone IC. The board is a feature telephone. Using the handset and the keyboard, telephone calls can be made, and incoming calls can be accepted. Beside this it is also possible to make calls and accept incoming calls by means of a speaker-phone button (=electronic hook), a loudspeaker and a microphone. In order to accommodate easy testing, several testpins are available.

On the board some jumpers and dip-switches are implemented. Fig.1 gives the mechanical outline of the board. The circuit diagram of the board is given in Appendix [1] and Appendix [2]. The bill of materials is given in Appen-dix [3]. Components view and copper view are given in Appendix [4] and Appendix [5].

3. BASIC FUNCTIONALITY

As stated the board is a featurephone.

The following list contains the minimum features of the board:

- On hook/ off hook by means of the cradle-switch **or** the HOOK button in the matrix
- Off hook indication by means of LED
- Pulse dialling
- DTMF dialling
- On hook dialling
- Last number redial
- Flash or earth function
- Pause
- 10 number memory (by means of trickle current) and recall
- DMO option
- Handset operation
- Handsfree operation
- Four level keyboard controlled HF volume
- Hold/secret mode with flashing LED indication
- Automatic on hook in hold mode in case parallel connected set goes off hook
- Keytone via loudspeaker (jumper selection)
- Four level keyboard controlled ringer volume
- Four different keyboard controlled ringer melodies

The following transmission parameters are adjustable by means different component values:

- Set impedance
- DC line voltage
- Sidetone
- AGC
- Send gain (HF and handset mode)
- Receive gain (HF and handset mode)
- switching range (HF)
- switch over timing (HF)
- receive sensitivity (HF)
- transmit sensitivity (HF)

4. MECHANICAL OUTLINE

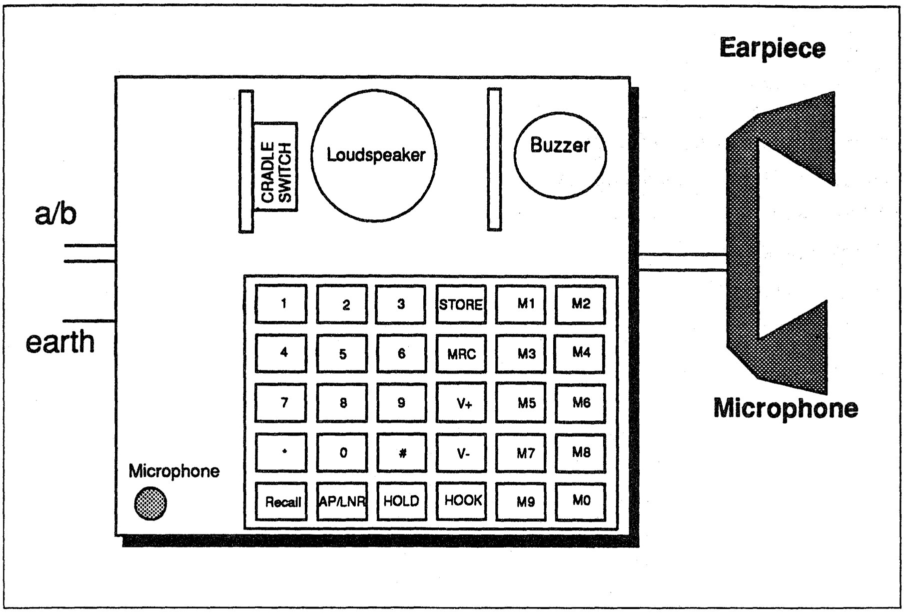


Fig.1 Outline of the board

The board contains the following controls:

Switch:	Meaning:
Cradle	Handset/ringer selection

Keyboard:	Meaning:
0 - 9, *,#	Digits
Recall	Flash/earth function
AP/LNR	Access Pause, Last number Redial
HOLD	Hold/secret mode
STORE	Memory enable
MRC	Memory recall
V+	Increase ringer/HF volume
V-	Decrease ringer/HF volume
HOOK	HF/handset selection
M0 - 9	Memories

Positioning of base microphone and loudspeaker has been chosen for minimum acoustic coupling. To increase sound pressure level the loudspeaker is mounted within a fitting enclosure as well as the PCB. On board, handset supports with cradle are provided. The handset and linecord can be connected by means of modular jacks. The proper connection for the linecord jack can be selected by means of jumpers.

4.1 Pin-names of the OM4784

On the board are several pins for inputsignals and measurement purposes:

A-B	=	line-input of the board
B-A	=	line-input of the board
Earth	=	earth connection input of the board
VEE	=	ground reference of UBA1702/A and TEA1069
Vline	=	line output voltage of UBA1702
SLPE	=	supply reference of the TEA1093
MIC +	=	non inverting microphone input (handset)
MIC -	=	inverting microphone input (handset)
QR	=	earpiece output

4.2 Telephone line connection/hookswitch

The board can be connected to the telephone line by a 6-pins modular jack plug (P1) via a 6x3 jumper matrix for selecting the proper A/B wires or by means of the A-B/B-A terminals.

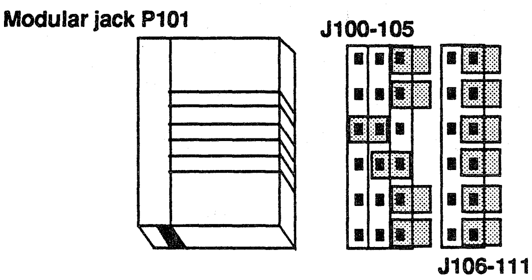


Fig.2 Default jumper settings

There is protection against high voltage peaks. Protection against high currents surges is provided by an internal (UBA1702/A) current limiter (max. current= 120 mA).

The selection between on-hook/off-hook is done by the single pole cradle switch S101 or by the speaker-phone button ("HOOK" on the key-pad).

4.3 Handset connection

A handset can be connected by the 4-pins modular jack plug (P2). The handset specifications are listed below:

Handset	Ericsson RLGN40201/8B6
Microphone	Sensitivity -44.5 dBV/Pa at 1 kHz, 2 kΩ load
Earpiece	150 Ω, sensitivity 49 dBPa/V

5. HARDWARE DESCRIPTION OF THE OM4784 BOARD

For numbering of components see Appendix [1]

5.1 Polarity guard and protection

The diode bridge D113, D114, D117, D118 ensures that in the off-hook situation the set functions properly independent of the polarity of the line voltage applied to it. In the on-hook situation the other diode bridge D104, D105, D109 and D110 rectifies the AC-ringer signal so a 2f signal is available.

Unprotected ICs might be destroyed by excessive current surges on the telephone lines if preventive measures are not taken. Protection is achieved by Break-Over-Diode (BOD) D111 (BR211-220) between the a/b lines and BOD's D112 and D115 for the earth connection. The LN-input of TEA1069N is voltage limited by the UBA1702/A.

During ringing the supply voltages of the ringer output stage and the V_{DD} of the TEA1069N are limited by the UBA1702/A.

Line Information

By means of the UBA1702/A (RFO), the incoming line signal (AC ringer signal or DC) is applied to CE_FDI input of the TEA1069N to provide information about the ringer frequency and the line status (on-hook/off-hook). The cradle-switch information [CSI] is applied directly to the TEA1069N and the UBA1702/A. The speaker-phone button information is applied to the UBA1702 by means of EHI and is applied to the TEA1069N by means of the [MATRIX]-signal.

5.2 UBA1702/A: line interrupter driver and ringer circuit.

This IC contains the interrupter switch driver -which is required for pulsedialling and flash operation-, line current detector and limiter, pin controlled line voltage limiter, ringer interface / outputstage and stabilized microcontroller supply.

Two versions are available:

- UBA1702, intended to drive a PMOST interrupter (BSP254A mounted on the board)
- UBA1702A, for a bipolar pnp interrupter (e.g. MPSA92).

Both versions and the belonging interrupter device can be used in the systemboard (board locations IC100 and TR101), however care must be taken for the correct connections of the interrupter device.

For extra information, please refer to Ref. [2] (the resistor names are also derived from Ref. [2])

5.2.1 DC settings

The voltage limit between SPO and VEE to protect the TEA1069N can be set by a resistor between ZPA (pin 9) and SPO (pin 6) (=RZPA1) of the UBA1702/A and a resistor between ZPA (pin9) and VEE (=RZPA2). When both are omitted (default) the voltage is limited to about 12 V. This voltage setting is recommended for the TEA106X family. When RZPA1 is replaced by a short (RZPA2 still open) the voltage is decreased to 9 V and in case RZPA2 is shorted (RZPA1 open) the voltage is increased to 18 V. Intermediate values can be achieved by applying appropriate resistors. The way these intermediate values can be achieved (by applying appropriate resistors) is the same for the various other adjustments of the UBA1702/A.

The current limit threshold can be influenced by a resistor between CLA (pin 25) and VEE (= RCLA). An open results in the lowest limit (about 45 mA), the highest limit can be achieved by a short (default setting), which results in a limit of approx. 120 mA.

5.2.2 Dial mode operation

The set resistance during the make period of pulse dialling can be decreased by making input pin MSI 'high' (DMO-mode), in this way the voltage between SPO and VEE is limited to a very low value. This value is default about 2.7 V and can be decreased by placing a short between SPO (pin 6) and MSA (pin 10).

5.2.3 Ringer mode operation

The input impedance in ringer mode for large ringer signals is mainly determined by R108, C111 and C112. Almost at the same time an AC ringersignal is applied a square wave signal with twice the ringer frequency is available on pin RFO. The ringer melody generated by the microcontroller is made audible by the piezo transducer H1, if the voltage on pin VRR crosses a certain upper ringer threshold of about 11 V (default setting). The ringer output stage is switched off again if a lower threshold has been reached (hysteresis). These thresholds can be influenced by changing resistors at pin RTA (pin 11). Decreasing the resistance (=RRTA1) between RTA and VRR (pin 21), results in decreasing the ringer threshold level, decreasing the resistance (=RRTA2) between RTA and VEE (leaving RRTA1 open) results in an increased ringer threshold level.

The ringer melody generated by the microcontroller (available on TONE output) is AC coupled by C102 to the RMI pin of the UBA1702/A.

5.3 TEA1069N: transmission part

5.3.1 DC-setting

For DC the TEA1069N (between pin LN and VEE) is a voltage stabilizer with a slope of 20 Ω. The DC voltage can be increased by means of R109 (between pins REG and SLPE). The OM4784 board is adjusted to $V_{A/B}$ = 7V at I_{LINE} = 20mA.

5.3.2 Set impedance

For AC (300 - 3400 Hz) the set impedance is mainly determined by R104 and is set to 600 Ω.

5.3.3 Power Control

Reset conditions

The TEA1069N has an internal reset circuit POR (Power On Reset) that monitors the supply voltage V_{DD}. If V_{DD} is below 2.5 V the circuit is in reset-mode. In this mode the dialler functions are disabled, but the transmission still operates although with less performance see Ref. [1].

The external reset overrules the internal reset at first start-up. This is done for proper initialization of the dialler registers in case V_{DD} rises the first time from 0 V to 2.5 V. The reset is delayed until V_{DD} is above 2.5V.

Start-up and switch-off behaviour

Off-hook procedure:

In handset mode:

After switching the cradleswitch to the off-hook position, line current will be applied to the line input LN. The supply capacitors C124, C108 and C104 will be charged. Above 2.5 V the TEA1069N RESET is disabled and the dialler circuit becomes active. As both CSI and CE_FDI of the TEA1069N are HIGH, TEA1069N will select the speech mode.

In handsfree mode:

After pressing the speaker-phone button 'HOOK', line current will be applied to the line input LN. The supply capacitors C124, C108 and C104 will be charged. Above 2.5 V the TEA1069N RESET is disabled and the dialler circuit becomes active. As both CE_FDI and an input via the matrix (when speaker-phone button is applied, COL4 and ROW5 are connected) are HIGH, TEA1069 will select the speech mode.

On-hook procedure:

After switching the cradle switch or the speaker-phone button to the on-hook position (see also 5.6.1), CSI and CE_FDI are LOW and the TEA1069N goes in stand-by mode. The V_{CC} capacitor C104 will be discharged, but the V_{DD} capacitor C124 will remain above 1.4 V. This is achieved by a small trickle current from the telephone line (R100 = 5.6 MΩ). So memory retention is possible this way.

Ringer procedure:

In case of an incoming call, the supply capacitors C119 and C124 will be charged by the ringer signal. So the ringer part of the TEA1069N can analyse the applied ringer signal. In this case CSI = LOW and CE_FDI = HIGH (square wave) and the TEA1069N will select the ringer mode.

5.3.4 Microphone inputs

The TEA1069N has symmetrical microphone inputs. Its typ. specified voltage gain is 52 dB. Dynamic, magnetic and piezoelectric microphones can be used. Use of electret requires supply from VCC. Some microphone arrangements are shown in Fig.3

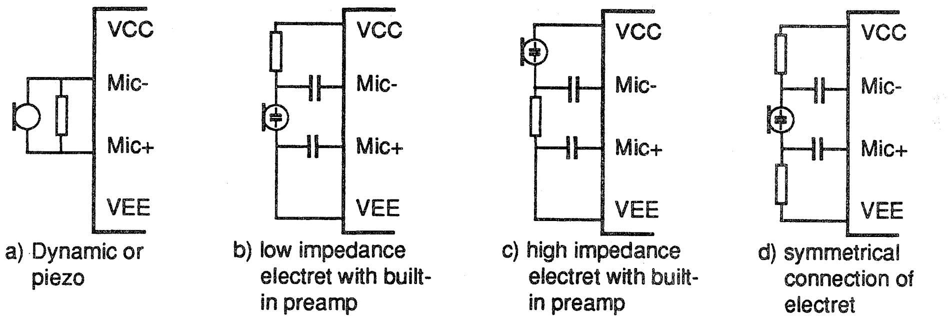
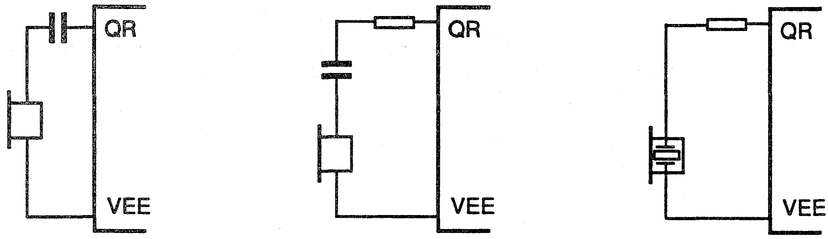


Fig.3 Microphone arrangements

Option d) is used on the OM4784. The gain-loss due to R125, R129 and R120 (Gain reduction due to sensitive electret microphone and EMC optimization) is 4 dB. Together with C118 and C120 they form a first-order high-pass filter with a cut-off frequency of 260 Hz.
The voltage gain of the TEA1069N is set to 44 dB, so the overall voltage gain (MIC <-> LN) is 40 dB. The voltage gain can be changed by adjusting **resistor R111**. Stability is ensured by two external capacitors, C107 connected between GAS1 and SLPE, C110 connected between GAS1 and VEE. With R111 and C107 a first-order low-pass filter is obtained with a cut-off frequency of 5.9 kHz. For more information see Ref. [1]

5.3.5 Earpiece outputs

The TEA1069N receiving amplifier has one input (IR) and a non-inverting output (QR). The typ. IR to QR specified voltage gain is 31 dB (from LN to QR is -1 dB, because the anti-sidetone network attenuates -32 dB) see Ref. [1]. Some earpiece arrangements are shown in Fig.4.



a) Dynamic earpiece

b) Magnetic earpiece (resistor is to prevent distortion, inductive load)

c) Piezo-electric earpiece (resistor is required to ensure stability, capacitive load)

Fig.4 Earpiece arrangements

The outputs (QR - VEE) may be used to connect dynamic, magnetic or piezoelectric earpieces.

Option a) is used on the OM4784. R_L (earpiece impedance) forms together with C128 a first-order high-pass filter with a cut-off frequency of 30Hz.
The voltage gain of the TEA1069N is set to 27.7 dB. The overall voltage gain (LN <-> TEL+) is -6 dB. The voltage gain can be changed by adjusting resistor R119. Stability is ensured by two external capacitors, C117 connected between QR and GAR, C121 connected between GAR and VEE. With R119 and C117 a first-order low-pass filter is obtained with a cut-off frequency of 4.2 kHz. For more information see Ref. [1]

5.3.6 Sidetone adjustment

In the OM4784 the anti-sidetone network consists of R104, R113, R117, R118, R128, R123+R127+C125 (= Zbal) and is optimized for 5 km length (0.5 mm diameter twisted pair copper cable with a DC resistance of 176 Ω/km, 38 nF/km and an average attenuation of 1.2 dB/km) see Ref. [1].

When following conditions are fulfilled, maximum compensation is obtained:

$$R_{128} \times R_{113} = R_{104} \times (R_{117} + R_{118})$$

$$Z_{bal} = \frac{R_{117} \times (R_{118} + R_{128})}{(R_{113} \times R_{128})} \times Z_{line}$$

5.3.7 DTMF-level

When the DTMF input is enabled ($\overline{\text{MUTE}}$ is LOW) dialling tones from output TONE are sent on the line. The DTMF level is -6 dBm (line load = 600 Ω)at $V_{A/B}$ and depends on resistor R111 in the same way as the microphone gain. R101 and R102 must be changed for only adjusting the DTMF level. The tones can be heard in the earpiece at low level (confidence tone).

5.3.8 MOH-level

When pressing the MUTE/HOLD button the DTMF input is enabled ($\overline{\text{MUTE}}$ is LOW) and melody tones from output TONE are sent on the line. The MOH level is -26 dBm at $V_{A/B}$ and depends on resistor R111 in the same way as the microphone gain. R103 must be changed for only adjusting the MOH level. The tones can be heard in the earpiece at low level (confidence tone).

5.3.9 AGC control

The Automatic Gain Control varies the gain of the microphone amplifier and the gain of the receiving amplifier in accordance with the DC-line current. The control range is 5.8 dB (which corresponds to a line length of 5 km for a

0.5 mm diameter twisted pair copper cable with a DC resistance of 176 Ω/km and an average attenuation of 1.2 dB/km). The AGC can be adapted by means of resistor R114 (OM4784 default 110 kΩ). R114 should be chosen in accordance with the exchange supply voltage and its feeding bridge resistance see Ref. [1]. When the resistor is removed, the AGC-function is disabled.

5.4 Dialler part

For numbering of components see Appendix [1]

5.4.1 Operation modes of the OM4784

The OM4784 has three different modes:

1. STAND-BY mode
2. RINGER mode
3. SPEECH mode

Actions to be taken to enter or to leave each mode, including the status or changes of the dialler signals, are described below. Extended dialling functions and programming operations are described in detail in Ref. [1]

STAND-BY mode

- Entering the STAND-BY mode:

  STAND-BY mode (ON-hook) is only entered if CE_FDI and CSI are low for a specific time.

- Leaving the STAND-BY mode:

  STAND-BY mode is cancelled if CE_FDI goes high. The following actions are to be taken by the dialler:

  - Start-up from TEA1069N
  - Scan all I/Os for their status
  - If S101 activated or speaker-phone button: SPEECH mode CSI = high or HF is high, CE_FDI = high
  - If S101 and speaker-phone button deactivated: RINGER mode CSI = low, CE_FDI = high

RINGER mode

- Entering the RINGER mode being ON-hook:
  Ringer mode can only be entered from the STAND-BY mode after detection of an incoming ringer signal CSI = low, CE_FDI = high (square signal)
- Leaving the RINGER mode:
  RINGER mode is left:
  - If CE_FDI remains low for time out: STAND-BY mode
  - If S101 or speaker-phone button is switched to OFF-hook =>: SPEECH mode. CSI = high, CE_FDI = high.

SPEECH mode

- Entering the SPEECH mode:
  SPEECH mode can be entered by switching S101 or speaker-phone button OFF-hook: CSI = high or HF = high, CE_FDI = high.
- Leaving the SPEECH mode:
  - S1 and speaker-phone button switched-over to ON-hook: STAND-BY mode. CSI = low, CE_FDI = low.

In speech mode several dialling functions can be made.

5.4.2 Dialling functions

Dialling functions in the SPEECH mode, CSI = high, CE_FDI = high
--> See chapter 5.6 'Diode options dip-switches S2' to select the parameters of dialling and ringer functions.

- **Dialling functions**
  - Keys 0 up to 9, */T and #: manual dialling (Dip switch PTS selects Pulse or DTMF dialling).
  - */T: In Pulse mode pressing this button switches from Pulse to DTMF mode until going off-hook or pressing RECALL
  - AP/LNR: last number redial if pressed directly after switching S1 =>, otherwise Access Pause
  - RECALL: activation of FLASH function (Time selectable by diode switches FES A and FES B)
  - STORE / MRC: to store/recall repertory numbers:
- **Memory store**:
  To store numbers into the memory locations:
  - Depress STORE
  - Enter telephone number or depress LNR (in case LNR must be stored)

  Now two options are available:
  1. *Direct* memory access store: depress M0-9 (only one of them).
  2. *Indirect* memory access store: depress MRC and then depress memory location button, corresponding numerical key (0-9)

- **Memory recall:**
  To recall repertory numbers:
  1. *Direct:* depress button M0-9
  2. *Indirect:* depress button MRC and then depress one of the numerical keys (0-9)
- **Memory clear:**
  To clear memory content:
  - Depress STORE
  1. *Direct:* depress M0-9
  2. *Indirect:* depress MRC + the assigned numerical button

5.4.3 Ringer functions

The melody ringer of the OM4784 can be programmed to become active for incoming signals at the CE_FDI input with frequencies between 40-120 Hz or 29-146Hz (both rectified AC-ringer signal = 2f detection method).
--> See chapter 5.6 'Diode options dip-switches S2' to select the parameters of dialling and ringer functions.

To change ringer melodies:
- Depress buttons 1, 2, 3 or 4 during ringer burst

5.5 Ringer output stage

The output-stage is incorporated in the UBA1702. The volume level can be adjusted during ringing by pressing the V+ or V- button of the pad.The buzzer is:

PXE Buzzer	Murata PKM34EW-1224

5.6 TEA1093: handsfree

This chapter describes the parameter setting of the TEA1093. The circuit diagram is shown in Appendix [1].

5.6.1 Choosing between handsfree and handset operation

There are several possibilities to choose between handset and handsfree operation:

1. The set is ON-HOOK. Handsfree operation after pressing the "HOOK"-key,
2. The set is in handset-mode. Handsfree operation after pressing the "HOOK"-key,
3. The set is in handsfree-mode. Handset operation after taking the handset (going OFF-HOOK)
4. The set is in handset-mode. Handsfree operation after pressing the "HOOK"-key, After pressing the "HOOK"-key again, handset mode is achieved again.

5.6.2 DC setting

The TEA1093 splits up the line current into a constant current for the TEA1069 (approximately 3 mA) and handles the excess current for generating its own stabilized supply voltage VBB (3.6V nominal). In case (due to AC line signals) the line voltage drops below the VBB voltage plus 0.4V, the excess current is sunk to GND to prevent distortion.

For stability reasons, capacitor C105 of 4.7 nF between pins SREF and VEE and capacitor C216 of 4.7 nF between pins SUB and VBB are needed. For correct functioning of the TEA1093 a capacitor of 470 μF is mounted at VBB (C215).

5.6.3 Microphone amplifier

To be able to connect a handsfree as well as a handset microphone, an analogue switch (IC102) itransfers either the signals coming from the handset or the handsfree microphone to input MIC of the TEA1069N. It is controlled by signal HF of the TEA1069N.

The handsfree microphone is supplied from VBB.The sensitivity for the handsfree microphone is set with R207.

The overall gain needed for the handsfree microphone can be adjusted by varying R210. An appropriate line level has been achieved with R210 set to 9.5 kΩ in combination with R207 set to 820 Ω. Capacitor C214 of 3.9 nF forms a low pass filter together with R210 (4.3 kHz). The gain from MIC to MOUT is proportional to the resistor between GAT and MOUT. R210 may be varied between 9.53 kΩ up to 95.3 kΩ providing a gain setting range of 5 dB up to 25 dB.

The microphone signal is coupled to MIC via capacitor C211 of 56 nF. With this capacitor and the input impedance at MIC (20 kΩ) a high pass filter is realized (150 Hz).

The microphone amplifier of the TEA1093 can be muted by means of applying a voltage >1.5V between pins MUTET and GND.

5.6.4 Loudspeaker amplifier

The input RIN1 of the loudspeaker amplifier is connected to QR+ via C202 of 100nF. Input RIN2 is connected to VEE of the TEA1069 via C204 of 33nF. With these DC blocking capacitors and the input impedance at RIN1 and RIN2 (2 *x* 20 kΩ) a high pass filter is realized (80 Hz).

The gain of the loudspeaker amplifier from the inputs RIN1 and RIN2 to LSP1 is proportional to the resistor between GAR and LSP1. On the board R204 is set to 220 kΩ resulting in a loudspeaker gain of 28.4dB. With the TEA1069 receiver gain being set at -6 dB, the overall gain from telephone line to LSP1 is 22.9 dB. If another gain setting is required for the TEA1093 loudspeaker amplifier, resistor R204 may be varied between 11.5 kΩ and 365kΩ providing a gain setting range of 3 dB up to 33 dB.

At pin LSP2, in case of the TEA1093, the inverted signal from LSP1 is present and may be used to obtain a bridge tied load drive configuration. In this configuration, the gain is increased 6 dB. On the board, the loudspeaker is connected as a single ended load.

Capacitor C207 of 180pF is mounted to form a low pass filter with resistor R204 (4 kHz).

The loudspeaker amplifier can be muted by making pin DLC/$\overline{\text{MUTER}}$ lower than 0.2V.

5.6.5 Volume control

The volume is controlled with the pins VOL1 and VOL2 of the TEA1069. These pins control IC200. The functionality is explained by means of Fig.5. With the two digital volume bits levels can be achieved via steps of approximately 6 dB. The control of the volumebits is shown in TABLE 1.

TABLE 1 Volume control

Vol2	Vol1	Applied level	Attenuation
0	0	0 (soft)	-18 dB
0	1	1 (middle1)	-12 dB
1	0	2 (middle2)	- 6 dB
1	1	3 (loud)	0 dB

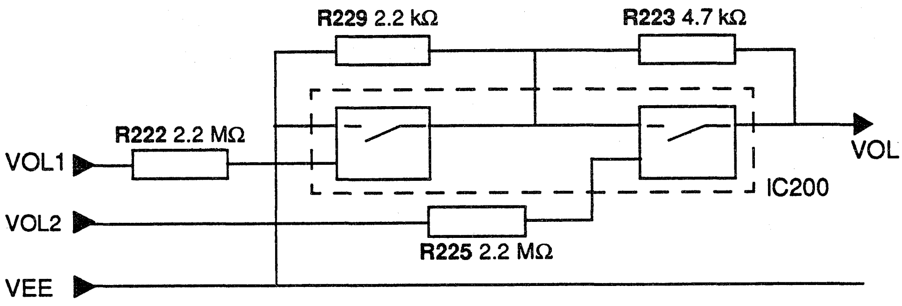


Fig.5 The volume control of the board

5.6.6 Dynamic limiter

The dynamic limiter (TEA1093) will reduce the gain of the loudspeaker amplifier in the following cases:

a) VBB drops below 2.8 V (VBB-limiter).

b) VBB starts dropping because supply current is insufficient (current limiter).

c) The loudspeaker signal starts to cause saturation (peak-limiter).

The attack and release times of these limiters are all proportional to the value of capacitor C200 connected to DLC. With C200 = 470 nF, attack times are in the order of a few milliseconds for limiters a) and c). Limiter b) is much slower and has an attack time in the order of seconds.

5.6.7 Switching range

The switching range is proportional to the ratio between R212 and R213. The resistor R213 is fixed to 3.65 kΩ. The resistor R212 can be varied between 3.65 kΩ and 1.45 MΩ resulting in a switching range of 0 dB up to 52 dB.

The volume setting affects the switching range such that the sum of the microphone and loudspeaker amplifier gain is kept constant. Therefore, volume control has a maximum range equal to the switching range.

For more details about how to adjust the switching range, refer to Ref. [3]

On the board the switching range is set to 40 dB via the 365 kΩ resistor R212.

If AGC is used, the gain from MIC+/- to QR of the TEA1069 is lowered with approximately 12 dB for very short line lengths below 5km.

5.6.8 Envelope detectors

The sensitivity of the detectors can be adjusted with R201 and R203. They can best be adjusted such that currents in the order of 10 μA_{rms} are flowing through them at nominal signals. The DC blocking capacitors C201 and C206 form a high pass filter together with R201 and R203.

The dial tone detector threshold is proportional to the value of R203. In formula:

$$R203 = V_{dial} * 78.7\,k\Omega$$

On the board R203 is set to 4.7 kΩ giving a dial tone detector threshold of about 60mV between RIN1 and RIN2. At the telephone line this means the dial tone detector level is set to 100mV (receive gain TEA1069 is -6 dB).

Capacitor C201 (180 nF) in series with R201 blocks DC and forms a high pass filter (225 Hz).

As explained in more detail in the appropriate application reports (Ref. [3]), the value of R201 can be determined with the following formula:

$$20\log(R201) = 20\log(R203) - 3\ dB$$

Thus R201 = 3.3 kΩ.

With AGC active:

$$20\log(R201) = 20\log(R203) - 3\ dB$$

thus R201 = 6.7 kΩ

As a compromise between the two situations, R201 is set to 3.9 kΩ on the board.

The charge and discharge times of both signal envelope detectors are proportional to the value of the capacitors C203 and C208 giving a maximum rise slope in the order of 85 dB/ms maximum, and a maximum fall slope of 0.7 dB/ms.

The noise envelope timing is proportional to the value of the capacitors C205 and C210. On the board 4.7µF is mounted at TNOI providing a maximum rise slope of 0.07 dB/ms maximum, and a maximum fall slope in the order of 0.7 dB/ms. It is also possible to apply a resistor of 10 kΩ at TNOI. With this resistor the influence background noise is being reduced.

5.6.9 Switch over timing

The switch over time from Tx to Rx or vice versa is proportional to the value of capacitor C217 connected to pin SWT. On the board 220nF is mounted for C217, resulting in switch over times of around 13ms.

The switch over time from Idle mode to Tx or Rx is also proportional to the value of C217. With the value chosen, switch over will take 4ms.

The timing from Tx or Rx to idle mode is determined by resistor R217 in conjunction with C217. On the board 2.2MΩ is mounted for R217 resulting in an idle mode timing of about 2s.

5.7 Extra features

In this paragraph extra features of the OM4784 like keytone and parallel set detection will be discussed

Keytone means that every time a key is pressed a tone through the speaker is heard. This option can be turned ON or OFF by means of jumper J200 (see also TABLE 3).

There is also an option called "parallel set detection". This means that when the set is in the MOH-mode (there is a tune "always is little shortcoat ill" (typical Dutch song) audible), the set can be put ON-HOOK. The tune continues playing. During MOH, the LED will flash. When another parallel set is put OFF-HOOK. An additional circuit consisting of R215, R226, R216, R227, R221, R224, C219, C220, C221, TR204, TR205, TR206 detects the decrease of the line current and makes the input HOLD of the TEA1069N low for a very short time. Now the set goes ON-hook.

NOTE: when this option is desired, diode-option HMS must be ON! (see also TABLE 2)!

Another function is the Earth-function.

5.8 Diode options dip-switches S2

The OM4784 has ten dip-switches which have the following functions. See Appendix [4] for board layout.

TABLE 2 Diode option

Number	DIODE	Condition	Function	On	Off
1	M/B		Make/Break ratio	3:2	2:1
2	RFS		Ringer Frequency Select	29 - 146 Hz	40 - 120 Hz
3	HMS		Hold/Mute Select	Hold mode	Mute mode
4	APT		Access Pause Time	4 seconds	2 seconds
5	PTS		Pulse/Tone selection	Pulse mode	DTMF mode
6	KBS		KeyBoard select	10 direct memories	3 direct and 10 indirect memories
7	GOS		German Output Select	Pin 8= Earth / Pin 27 is DMO	Pin 8= Keytone / Pin 27 is MOH output
8	TBT		Tone Burst/pause Time	85/85 ms	100/100 ms
9	FES A	FES B = On	Flash/Earth time Select	Earth of 400 ms	Flash of 600 ms
10	FES A	FES B = Off	Flash/Earth time Select	Flash of 270 ms	Flash of 100 ms

5.9 Jumper settings

The OM4784 has two jumpers which have the following functions. See Appendix [4] for board layout

TABLE 3 Jumper settings

Jumper 106 --> closed	DMO-function enabled
Jumper 200 --> closed	keytone enabled

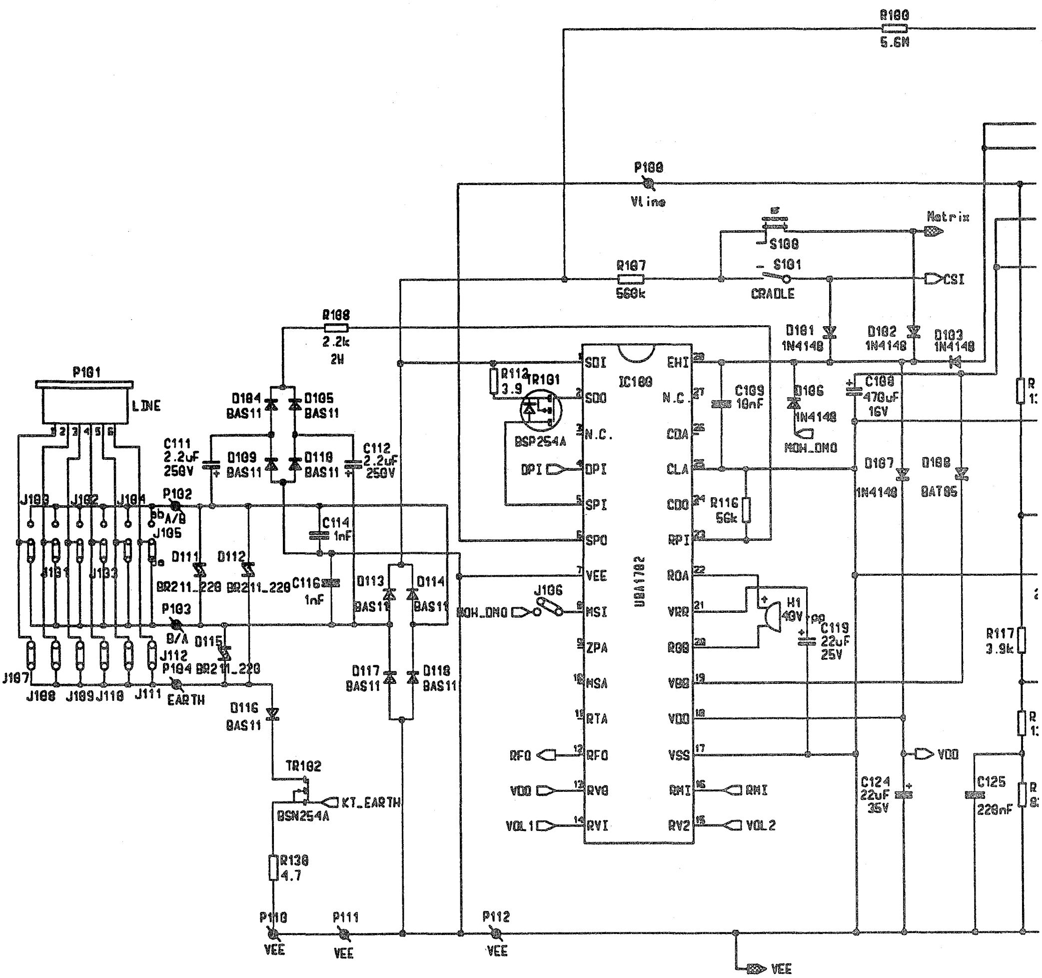
6. ELECTROMAGNETIC COMPATIBILITY

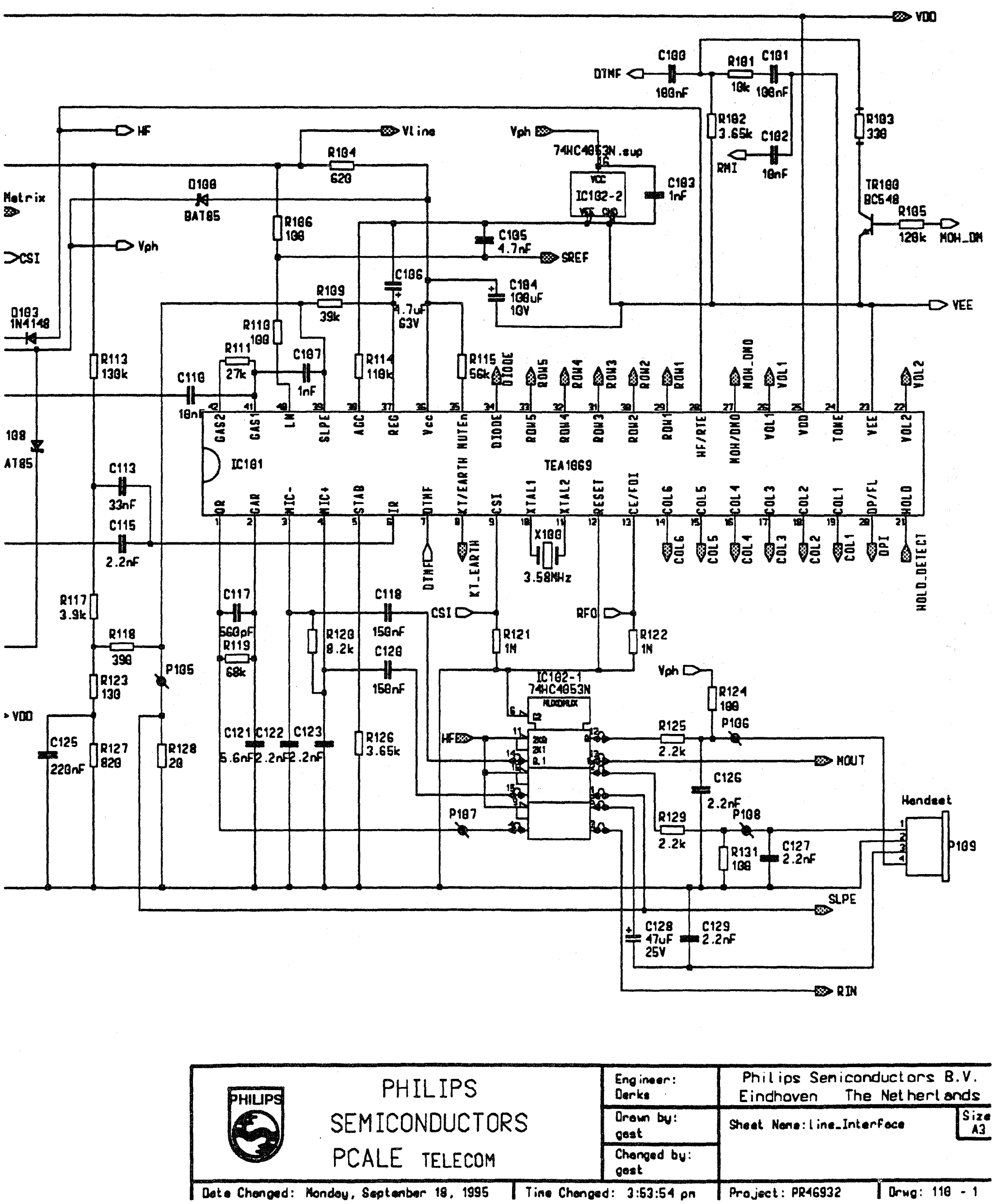
In the presence of high-intensity electromagnetic fields, common-mode amplitude modulated R.F. signals can be introduced in the a/b lines. These common-mode signals can become differential-mode signals as a result of asymmetrical parasitic impedances to ground (for example through the hand of the subscriber holding the handset). Preventive measures have to be taken to avoid these signals being detected and the low-frequency modulation appearing as unwanted signals at the earpiece or on the line.

The layout of the OM4784 has been designed with respect to electromagnetic compatibility. Basic protection is provided by means of discrete capacitors to suppress the unwanted R.F. signals before they can enter the circuit. Capacitors types suitable for high frequencies are used. Those are at A/B-line connection (C114, C116), MIC/TEL connections (C122, C123, C129) and one between IR and VEE (C115). Besides this, there are extra filters at microphone inputs and earpiece output (R125, C126, R129, C127).

The EMC capacitors at the A/B terminals (C114, C116) slightly influence the balance return loss and the sidetone characteristics at higher frequencies.

APPENDIX 1 ELECTRICAL SCHEMATIC (part 1)




IC101
TEA1069
IC102-2
IC102-1
74HC4053N
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PCALE TELECOM
Engineer: Derks
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Date Changed: Monday, September 18, 1995
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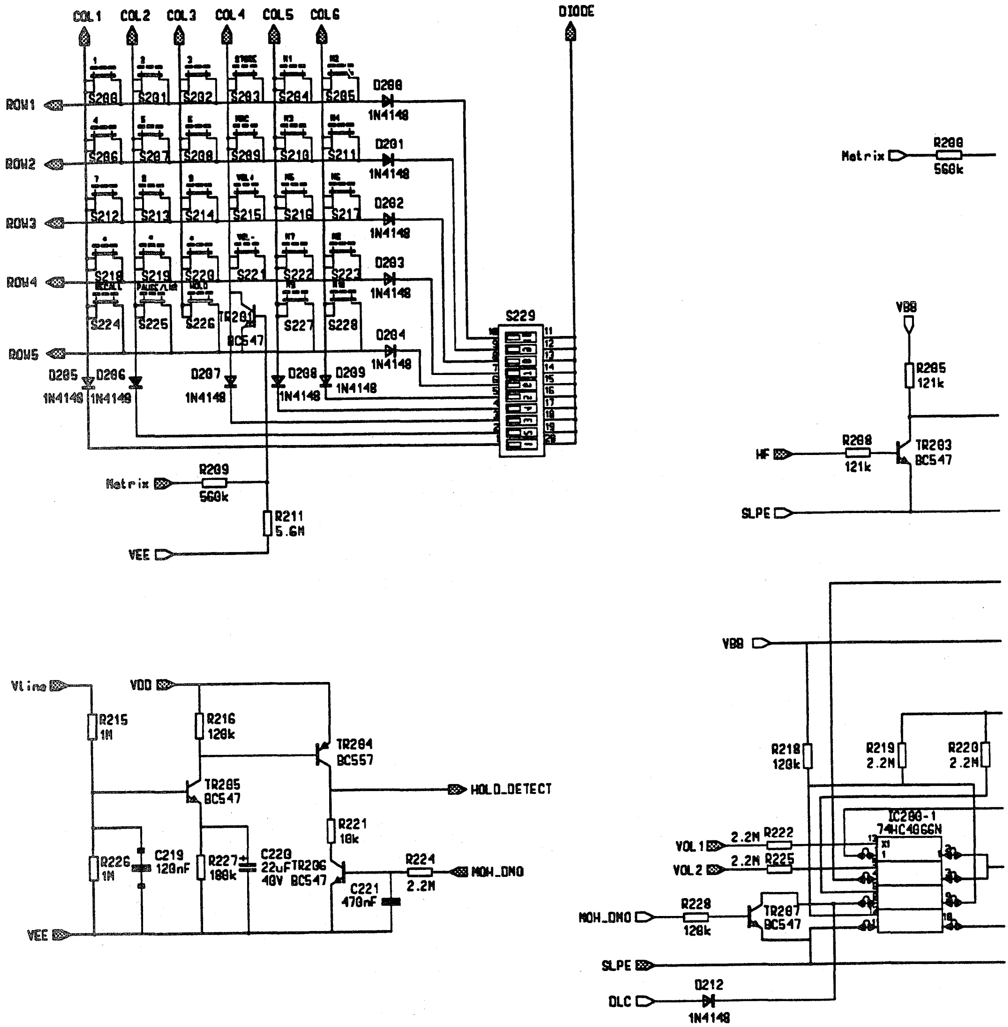
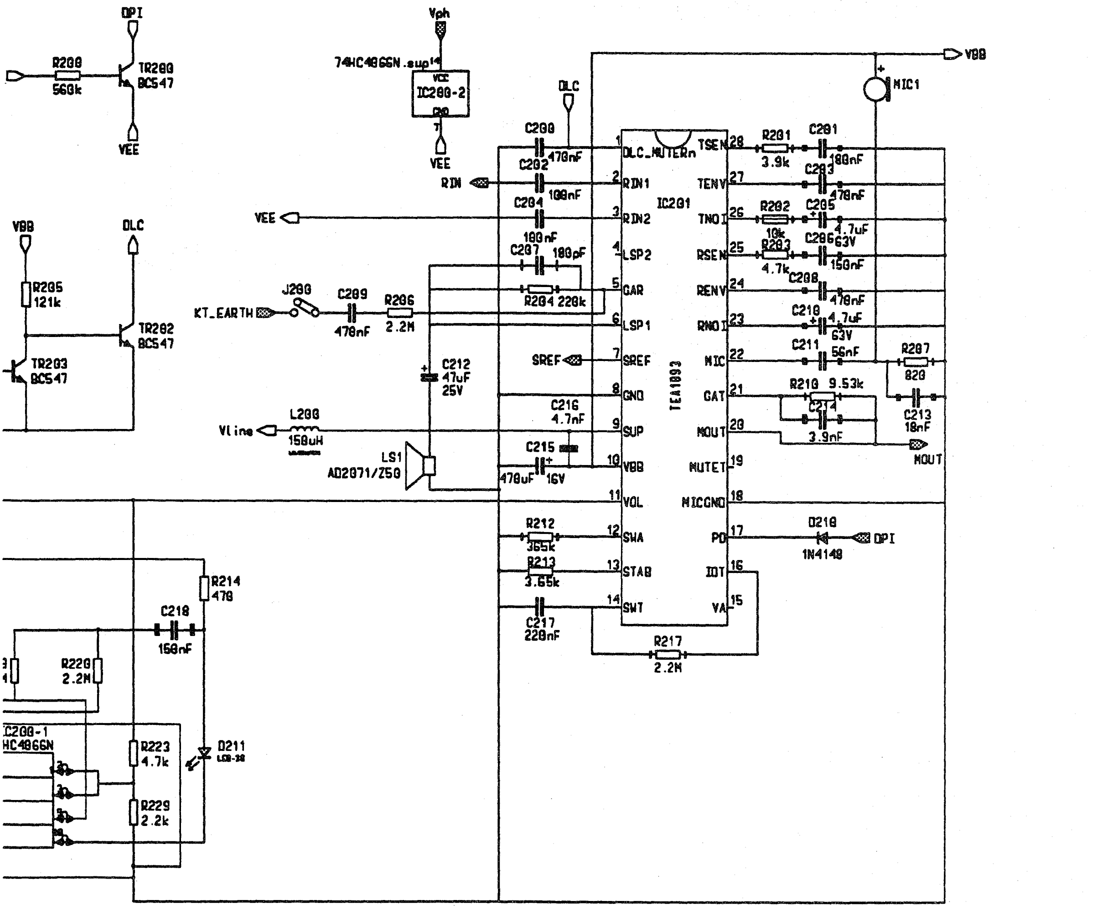
APPENDIX 2 ELECTRICAL SCHEMATIC (part 2)





PHILIPS SEMICONDUCTORS PCALE Telecom	Engineer: derks	Philips Semiconductors B.V. Eindhoven The Netherlands
	Drawn by: gest	Sheet Name: HF+INTERFACE — Size A3
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APPENDIX 3 BILL OF MATERIALS

TABLE 4 Bill Of Materials

Reference	Count	Component	Tolerance	Rating
	25	BOARD 160x100_bi		
TR102	1	BSN254A		
D111, D112, D115	3	BR211_220		
MIC1	1	MCE100		
P101	1	MOD_JACK_6p		
P109	1	MOD_JACK_4p		
D211	1	LED-red		
L200	1	150uH	10%	
S101	1	SPDT		
S100, S200, S201, S202, S203, S204, S205, S206, S207, S208, S209, S210, S211, S212, S213, S214, S215, S216, S217, S218, S219, S220, S221, S222, S223, S224, S225, S226, S227, S228	30	DE703-001		
TR101	1	BSP254A		
IC102	1	74HC4053N		
IC200	1	74HC4066N		
D100, D108	2	BAT85		
D104, D105, D109, D110, D113, D114, D116, D117, D118	9	BAS11		
TR204	1	BC557		
TR100	1	BC548		
TR200, TR207, TR206, TR205, TR203, TR202, TR201	7	BC547		
D101, D102, D103, D106, D107, D200, D201, D202, D203, D204, D205, D206, D207, D208, D209, D210, D212	17	1N4148		
P100, P102, P103, P104, P105, P106, P107, P108, P110, P111, P112	11	SOLDER-PIN_large		
S229	1	DIP-SWITCH_10		
J100, J101, J102, J103, J104, J105	6	JUMPER_3p		
J106, J107, J108, J109, J110, J111, J112, J200	8	JUMPER_2p		
LS1	1	AD2071/Z50		
R108	1	2.2k	5%	2W
R120	1	8.2k	5%	0.5W
R207	1	820	5%	0.5W
R127	1	820	5%	0.5W
R119	1	68k	5%	0.5W
R104	1	620	5%	0.5W
R100, R211	2	5.6M	5%	0.5W
R107, R200, R209	3	560k	5%	0.5W
R115, R116	2	56k	5%	0.5W

TABLE 4 Bill Of Materials

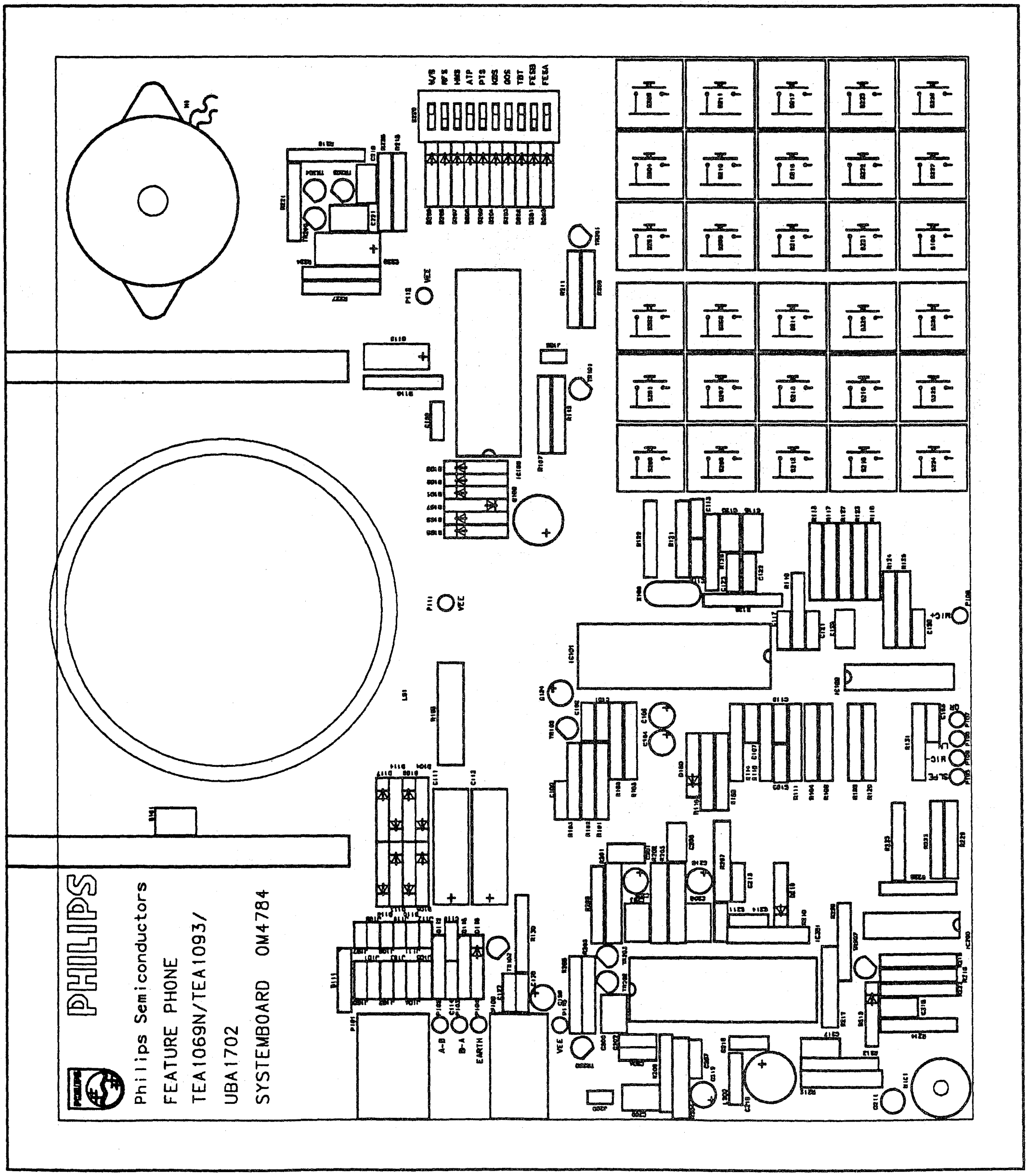
Reference	Count	Component	Tolerance	Rating
R130	1	4.7	5%	0.5W
R203	1	4.7k	5%	0.5W
R223	1	4.7k	5%	0.5W
R214	1	470	5%	0.5W
R112	1	3.9	5%	0.5W
R109	1	39k	5%	0.5W
R201	1	3.9k	5%	0.5W
R117	1	3.9k	5%	0.5W
R118	1	390	5%	0.5W
R103	1	330	5%	0.5W
R111	1	27k	5%	0.5W
R217	1	2.2M	5%	0.5W
R206, R219, R220, R222, R224, R225	6	2.2M	5%	0.5W
R204	1	220k	5%	0.5W
R125, R129, R229	3	2.2k	5%	0.5W
R128	1	20	5%	0.5W
R227	1	180k	5%	0.5W
R113	1	130k	5%	0.5W
R123	1	130	5%	0.5W
R105, R216, R218, R228	4	120k	5%	0.5W
R114	1	110k	5%	0.5W
R121, R122, R215, R226	4	1M	5%	0.5W
R101, R221	2	10k	5%	0.5W
R106, R110, R124, R131	4	100	5%	0.5W
R210	1	9.53k	1%	0.6W
R212	1	365k	1%	0.6W
R102, R126, R213	3	3.65k	1%	0.6W
R205, R208	2	121k	1%	0.6W
C117	1	560pF	10%	100V
C105	1	4.7nF	10%	100V
C214	1	3.9nF	10%	100V
C207	1	180pF	10%	100V
C216	1	4.7nF	-20+80%	63V
C115, C122, C123, C126, C127, C129	6	2.2nF	-20+80%	63V
C109, C110	2	10nF	-20+80%	63V
C103, C107, C114, C116	4	1nF	-20+80%	63V
C102	1	10nF	10%	250V
C113	1	33nF	10%	100V
C213	1	18nF	10%	100V
C211	1	56nF	10%	63V
C203, C208	2	470nF	10%	63V
C200, C209, C221	3	470nF	10%	63V
C125 C217	2	220nF	10%	63V

TABLE 4 Bill Of Materials

Reference	Count	Component	Tolerance	Rating
C201	1	180nF	10%	63V
C206, C218	2	150nF	10%	63V
C118, C120	2	150nF	10%	63V
C219	1	120nF	10%	63V
C100, C101, C202, C204	4	100nF	10%	63V
C111, C112	2	2.2uF	-10/+50%	250V
C128, C212	2	47uF	20%	25V
C108, C215	2	470uF	20%	16V
C106, C205, C210	3	4.7uF	20%	63V
C104	1	100uF	20%	10V
C124	1	22uF	20%	35V
C220	1	22uF	-10/+50%	40V
C119	1	22uF	-10/+50%	25V
IC100	1	UBA1702		
IC101	1	TEA1069		
H1	1	40V-pp		
X100	1	3.58MHz		
IC201	1	TEA1093_C1		
C121	1	5.6nF	10%	100V
	1	DIL_SOCKET_SHR_ 42p		
	16	SOLDER_PIN_large		
	2	SCREW_TAP_TORX_T6x6.5_St		
	20	SOLDER_STRIP_BUS_0.1_part		
	2	DIL_SOCKET_28p		
	1	DIL_SOCKET_16p		
	1	DIL_SOCKET_14p		
	14	JUMPER_CAP		

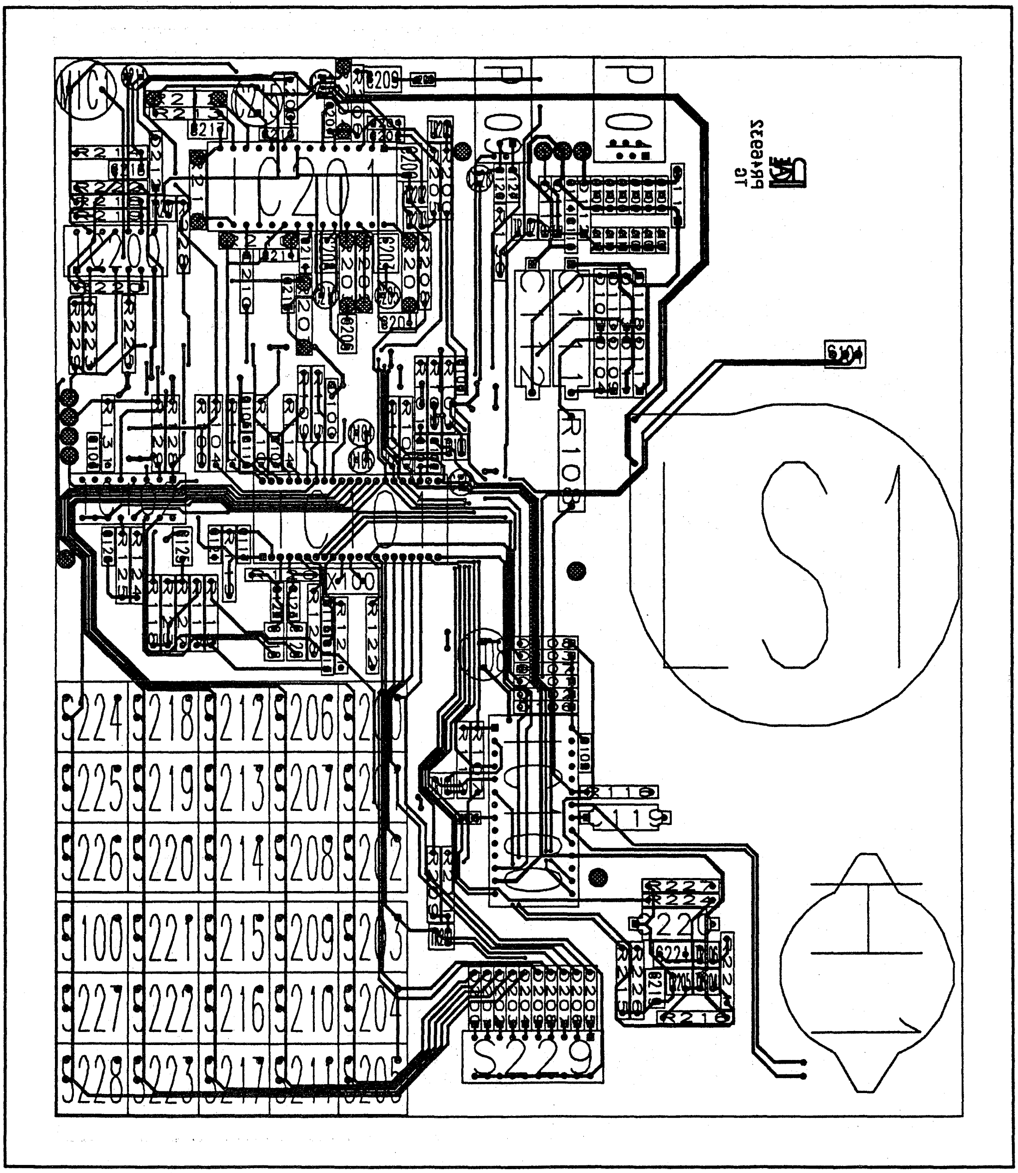
APPENDIX 4 COMPONENT LAYOUT

This picture shows the component layout with components laying inside the blocks and circles.

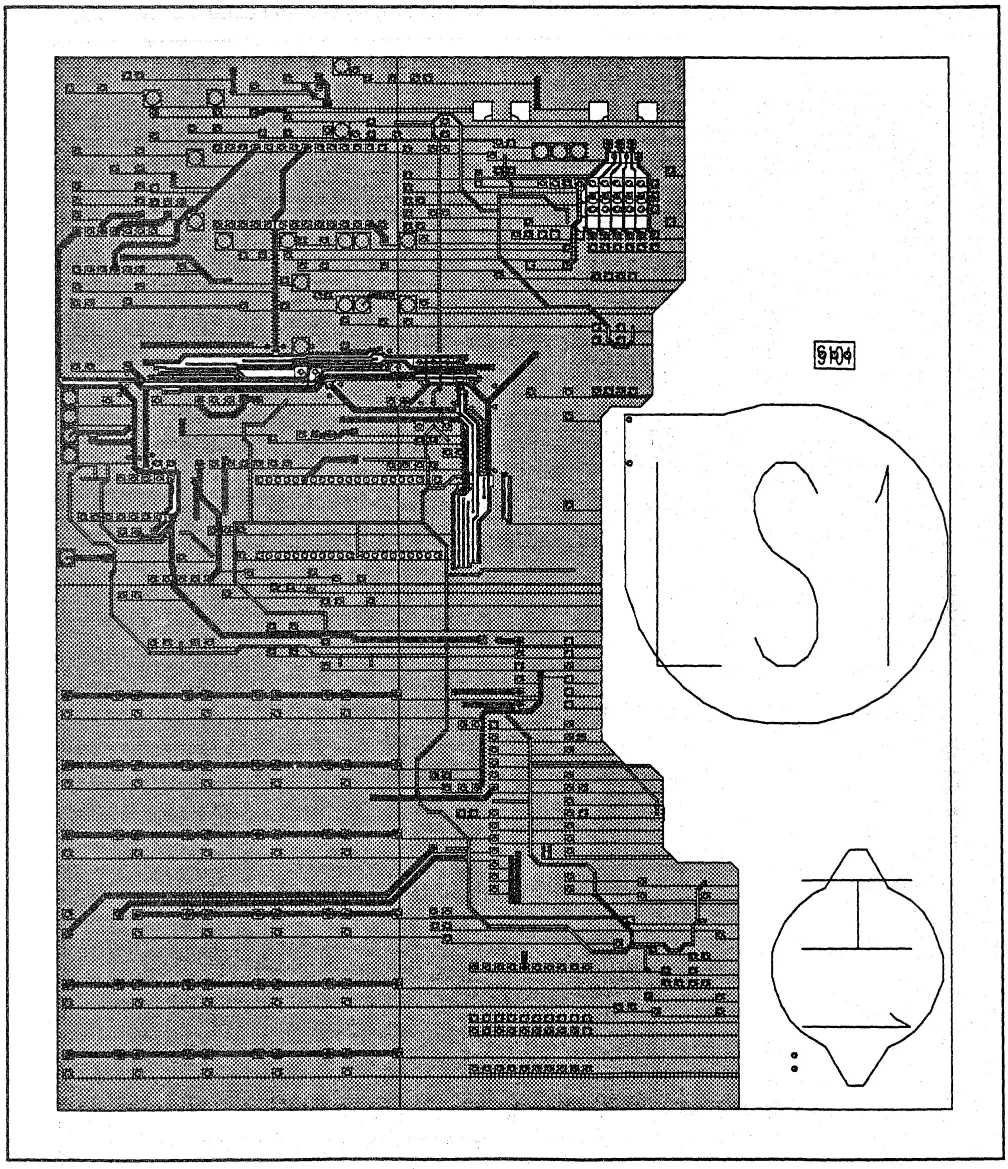


APPENDIX 5 PRINTED CIRCUIT BOARD LAYOUT

This picture shows the printed board layout of the OM4784. This picture represents the bottom of the PCB.



This picture shows the top of the PCB



8 REFERENCE MATERIAL

CONTENTS

APPLICATION NOTE Nr	IEEE Journal of Solid-State Circuits, Vol SC-19, Nr 3, June 1984 pp 379 - 388
TITLE	Dual tone and Modem Frequency Generator with On-Chip Filters and Voltage Reference
AUTHOR	J. Mulder, W. Oswald
DATE	June 1984

Abstract—A single-chip CMOS circuit is described that contains a dual-tone multifrequency (DTMF) and modem frequency generator. For optimum performance and economy, switched-capacitor techniques are used for the on-chip bandgap reference voltage, D–A converters, and filter. CEPT recommendations on output level stability and distortion are met without recourse to external filtering and without a stabilized supply or external reference voltage. A self-aligned contact (SAC) CMOS process with 4 μm design rules and with a 500 Å thick gate oxide is used to manufacture the circuit.

I. Introduction

IN modern telephony, dual-tone multifrequency dialing (DTMF) is replacing the old pulse dialing system. It is much faster and more reliable than the old rotary dial system. DTMF frequencies were standardized by the CCITT in 1968 [1] and later supported by CEPT recommendations on distortion and output level stability. The standard tones are illustrated in Fig. 1.

A problem facing subscriber set circuits is that their supply comes from the exchange via the telephone line. The voltage developed depends on the country, locality, and the length and diameter of the lines. Modern telephone sets use a transmission circuit to interface speech and DTMF circuits to the line, as well as providing a line termination that regulates the supply voltage for other circuits to between 2.5 and 6 V.

Bipolar [2] circuits have been described for DTMF tone generation. As well as requiring an external filter to meet CEPT CS203 recommendations, some have required a stabilized supply or an external reference voltage.

The circuit described here has been designed for modern subscriber set architecture such as that shown in Fig. 2. The input is designed for microcomputer control, either parallel or serial (for an I^2C bus), and the circuit can accommodate the line voltage variations mentioned. It requires only an external crystal. Filters and a reference voltage source are included on the chip. It is interfaced to the line by a transmission circuit.

Other features are that, in addition to generating DTMF tones, it can provide standard modem signaling frequencies, as well as two octaves (in semitone steps) of the musical scale for such purposes as electronic ringing.

Manuscript received October 27, 1983; revised January 16, 1984.
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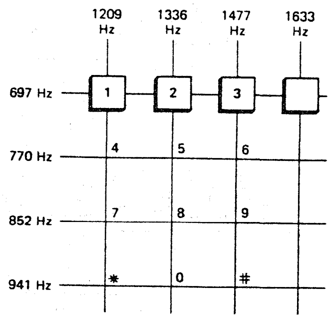


Fig. 1. CCITT and CEPT standardized dialing tone combinations.

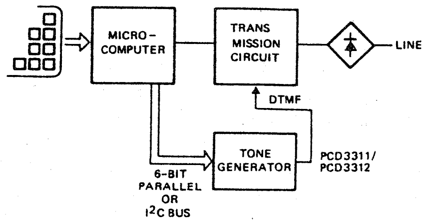


Fig. 2. Typical architecture for a microcomputer-controlled subscriber set.

The circuit is manufactured in a self-aligned contact (SAC) CMOS process using 4 μm design rules. Switched-capacitor techniques are used for sine wave synthesis, digital-to-analog conversion (DAC), on-chip filtering, and for an on-chip voltage reference source.

II. General Description

Tone frequencies are derived from a crystal oscillator whose frequency is divided by two parallel variable dividers (to provide the dual tones), each followed by a fixed divider (see Fig. 3). Either of the variable dividers can be disabled to allow the generation of single tones for the modem signals and for the musical tones.

The variable dividers are programmed via a ROM that is controlled by the data input (6 bits). The input may be parallel (PCD3311) or serial (PCD3311 or PCD3312) for connection to an I^2C bus [3]. The variable dividers drive logic circuits that control a sine wave synthesizer circuit.

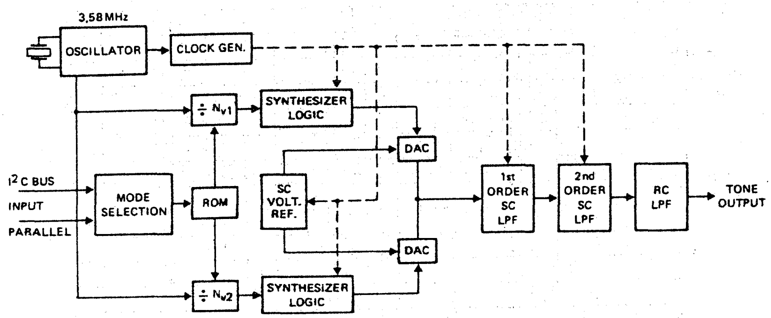


Fig. 3. Block diagram of the IC. Switched-capacitor techniques are used for the voltage reference, for the DAC's, and first-order and second-order filters.

The fixed dividers are combined with sine wave synthesizers comprising two DAC arrays and a summation stage, together also forming the first stage of a switched-capacitor (SC) low-pass filter. The capacitance ratio between the two DAC arrays provides the required (2 dB) preemphasis for the upper group of frequencies. This stage is followed by a second-order low-pass SC filter, the final filtering being performed by an active *RC* low-pass filter.

III. Circuit Description

For convenience, the various circuit functions are described in the sequence in which they were considered in the design phase.

A. The Tone Generators

The tone frequencies are derived by frequency division from a 3.579545 MHz crystal.[1] As mentioned, there are two channels, each with a variable divider (with divisor N_r) followed by fixed frequency dividers (divisor N_f). As will be seen later, N_f needs to be as high as possible to ease filter requirements. A computer program selected 23 as the highest common factor that allows integers for all values of N_r while yielding tones that are within 0.5 percent of CCITT frequencies and also meet the requirements for modem frequencies.[2] Although CEPT CS203 allows 1.5 percent tolerance, we chose 0.5 percent for quality reasons. Table I shows the results of these two divisions.

TABLE I
Accuracy of Generated Frequencies

	f_{tone} (Hz)	N_r	f_{obt} (Hz)	T_{acc} (%)	T_{obt} (%)	
Dual-Tone Frequencies	697	223	697.9	0.50	0.13	
	770	202	770.5	0.50	0.06	
	852	183	850.5	0.50	−0.18	
	941	165	943.2	0.50	0.24	
	1209	129	1206.5	0.50	−0.21	
	1336	116	1341.7	0.50	0.42	
	1447	105	1482.2	0.50	0.35	
	1633	95	1638.2	0.50	0.32	
Modem Frequencies	980	159	978.8	0.61	−0.12	
	1180	132	1179.0	0.51	−0.08	V.21
	1650	94	1655.7	0.36	0.34	
	1850	84	1852.8	0.32	0.15	V.21
	1300	120	1296.9	0.77	−0.24	
	2100	74	2103.1	0.48	0.15	V.23
	1070	145	1073.3	0.50	0.31	
	1270	123	1265.3	0.50	−0.37	Bell 103
	2025	77	2021.2	0.50	−0.19	
	2225	70	2223.3	0.50	−0.08	Bell 103
	1200	130	1197.2	1.00	−0.24	
	2200	71	2192.0	1.00	−0.36	Bell 203

f_{obt} = frequency obtained, T_{acc} = acceptable tolerance, T_{obt} = obtained tolerance.

B. Filter Requirements

Fig. 4 shows CEPT CS203 recommendations regarding unwanted frequencies. In this case, the components that determine the filter requirements will be those occurring at 22 times ($= N_f - 1$) the fundamentals. These will be 26.9 dB below the fundamentals due to a zero-order sample-and-hold effect in the SC filter, and they will occur in the frequency range 15–50 kHz (arrow-tipped horizontal lines on the right of Fig. 4). Harmonics due to quantization in the sine wave synthesis are assumed to have no influence on the filter requirement. Table II shows the required minimum attenuation for three orders of filter.

Table III shows the maximum cutoff frequency for Butterworth and Chebyshev (0.2 dB ripple) filters. The minimum cutoff frequency (Table IV) is dictated by the highest fundamental frequency (2225 Hz) and the acceptable ripple in the passband, which we took as 0.2 dB.

It will be seen that a second-order Chebyshev with 0.2 dB ripple is just theoretically acceptable. However, we decided on a third-order Chebyshev because the higher ratio between maximum and minimum cutoff frequencies allowed us greater latitude for a tradeoff between perfor-

[1] This crystal is used as color burst generator in CTV receivers and is, therefore, in large-scale production.

[2] It would have been convenient if we could have taken 24 so that only 6 steps need be generated in the sine wave synthesis. However, this would not have enabled us to meet the recommendations for both DTMF and modem frequencies.

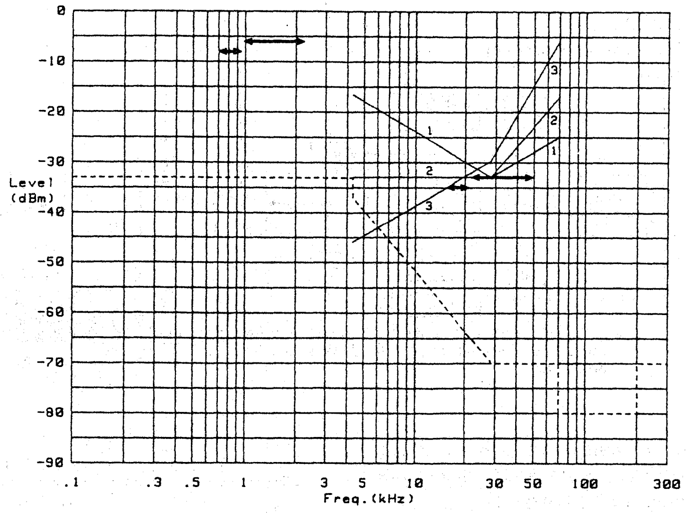


Fig. 4. CEPT recommendations on unwanted frequencies (broken line). Horizontal lines tipped with arrows show (upper left) the fundamental tones and (center right) the major unwanted harmonics. The lines numbered 1, 2, and 3 illustrate the situation as it would appear if the CEPT recommendations were relaxed by the roll off of (respectively) first-, second-, and third-order filters.

TABLE II
REQUIRED FILTER ATTENUATION

Order	Freq. (Hz)	Attenuation (dB)
1	27 940	36.6
2	27 940	36.6
3	15 334	24.2

TABLE III
MAXIMUM CUTOFF FREQUENCY

Order	Butterworth Freq. (Hz)	Chebyshev (0.2 dB) Freq. (Hz)
1	413	90
2	3398	2236
3	6062	5585

TABLE IV
MINIMUM CUTOFF FREQUENCY

Order	Butterworth Freq. (Hz)	Chebyshev (0.2 dB) Freq. (Hz)
1	10 249	2225
2	4775	2225
3	3702	2225

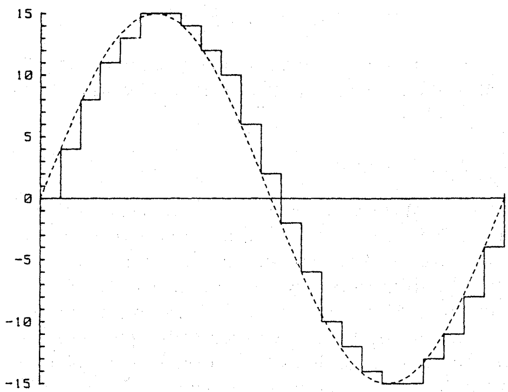


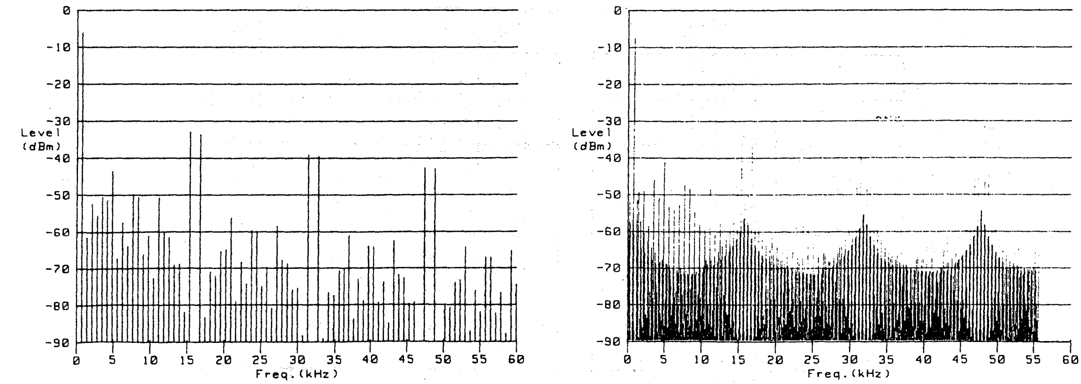
Fig. 5. The synthesized sine wave. The number of time slots is the same (23) for all tones. The amplitudes are quantized to 5 bit accuracy.

mance and chip area. We also elected for a switched-capacitor filter.

C. The Choice of Filter Clock Frequency

The filter clock frequency of 111.86 kHz is derived by dividing the crystal oscillator frequency by 32. This frequency is a fair compromise between filter op-amp power consumption and the requirements of the analog *RC* filter, but it is not without influence on the spectrum of the synthesized waves. These are sampled at a frequency $f_{crystal}/N_r$, so with a filter clock frequency of $f_{crystal}/32$, the filter will oversample by a factor of $N_r/32$. As in no case is N_r exactly divisible by 32, a number of new components will always be introduced.

The effects of this are illustrated in Figs. 5 and 6(a) and (b). Fig. 5 shows the synthesized waveform. Fig. 6(a) shows the spectrum of a 697 Hz tone (the lowest DTMF tone). The components on either side of the synthesizer sampling frequency are prominent. Fig. 6(b) shows the same spectrum, but with the effect of the oversampling added. N_r for



(a) (b)

Fig. 6. (a) The spectrum of a synthesized 697 Hz tone. Note the strong components on both sides of multiples of the sampling frequency. (b) The same spectrum as Fig. 6(a), but now showing the effects of oversampling, and without the sample-and-hold effect.

this tone is 223, a prime number, so there are 32 periods of synthesized wave before repetition occurs. The result is 31 new components between the original ones, plus some alteration in the amplitudes of the original components. This effect is most prominent for the smaller components, but that at 22 times the fundamental frequency is also slightly increased.

We calculated the effects of this on the filter requirement, and found it rather small. A reduction in the maximum filter cutoff frequency from 5585 to 5505 Hz is sufficient. Because an SC filter is time discrete, an analog *RC* postfilter is used to remove components around multiples of the filter clock frequency. The strongest of these is at 111.86–2.225 kHz (i.e., 109.635 kHz). Extra attenuation needed to meet CEPT CS203 recommendations is 40.1 dB.

D. The SC Filter

By making a third-order filter from a first- and a second-order filter in cascode, we could neatly combine the first-order filter with the DAC's and the summation stage.

The transfer characteristics of the first and second section are

$$H_1(z) = -0.398\frac{z}{z-0.602}$$

$$H_2(z) = -0.062\frac{z^2}{z^2-1.653z+0.715}.$$

These filter characteristics for a clock frequency of 111.86 kHz are shown in Fig. 7. No hold effect is taken into account. Fig. 8 shows the circuit diagram (the synthesis method used is that due to Fleischer and Laker [4]) and Fig. 9 shows the clock-phase diagram. By sampling the output of op-amp 3 on clock-phase 2 when it has stabilized, direct feedthrough from input to output is prevented.

E. The DAC's

The circuit diagram of both DAC's and the filter section are shown in Fig. 10. This replaces the first-order section of Fig. 8. The DAC's are two binary weighted capacitor arrays, one for the lower tone group, with $1C$ (unit capacitance) as base, and one for the upper. The capacitance ratio of 1.27:1 between arrays fixes the preemphasis for the higher group, making it slightly more than the recommended 2 dB to compensate for the low-pass characteristics of the line interface circuit.

A number of considerations play a role in determining the output voltage swing of op-amp 3.

1) The output voltage swing should not be more than about 1 V peak-to-peak to allow the circuit to work properly down to the minimum supply voltage of 2.5 V.

2) If the reference voltage is too low, clock feedthrough on it can become significant.

3) The ratio of switched to unswitched capacitance is fixed by the filter requirement (1:1.51), but the absolute value must be kept low to conserve chip area.

4) The minimum DAC capacitance is $1C$ and the capacitance ratio between DAC's is 1:1.27, so the total DAC capacitance is fixed at $(1+1.27)\times 15C = 34.1C$.

With a reference voltage of 200 mV (see Section III-G) and a switched capacitance over the op-amp of $13.9C$, the peak-to-peak output voltage becomes an acceptable 0.98 V. Similar scaling of the other sections of the filter gives about the same output voltage swing for the other op-amp's.

The negative half-period of the DAC waveform is generated by reversing the logic switching sequence and the switching sequence of the DAC array (see Figs. 5 and 11).

F. Analog Postfilter

As mentioned earlier, an analog postfilter is needed because of the time-discrete nature of the preceding filter.

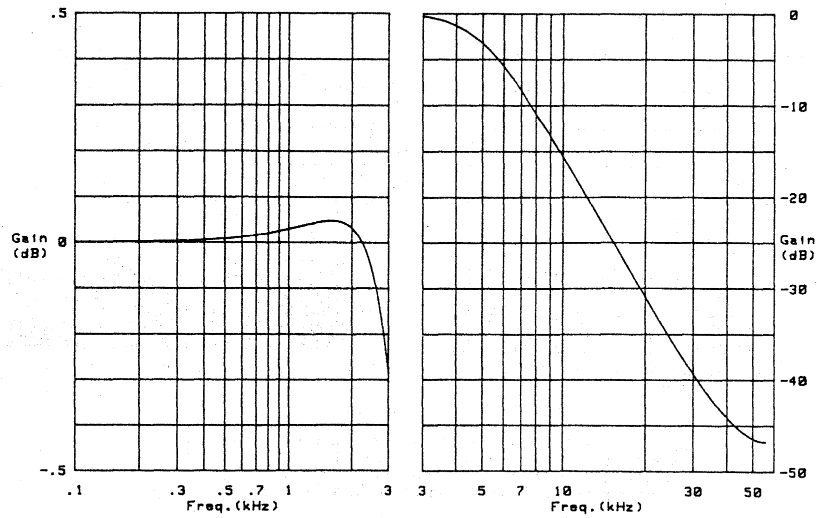


Fig. 7. Frequency characteristic of the SC filter. On the left, the characteristic below 3 kHz in expanded form is shown.

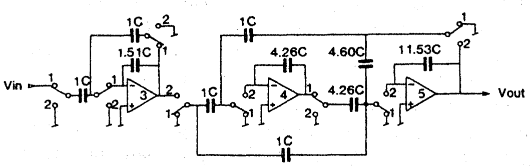


Fig. 8. Circuit diagram of the SC filter. Numbers adjacent to the switches refer to their positions during clock phases 1 and 2. *C* represents unit capacitance.

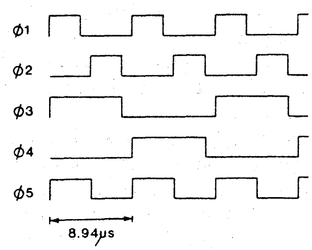


Fig. 9. Clock-phase diagram for SC filter (1 and 2), offset compensation (3 and 4), and voltage reference (5).

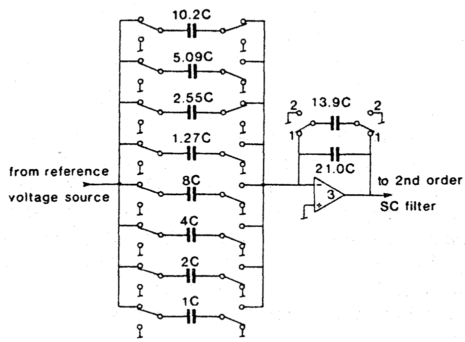


Fig. 10. Dual DAC arrays with first-order filter.

It was mentioned that a minimum of 40.1 dB attenuation at 111.86–2.2225 kHz is needed.

Here we chose a third-order Butterworth to allow for component tolerances, but we raised the attenuation requirement to 50 dB to compensate for clock feedthrough via the switch at the output of the SC filter.

Fig. 12 shows the circuit used. The resistors are p-well resistors with distributed capacitance both to the substrate (thus to V_{DD}) and to a polysilicon layer above them. The filter capacitors are actually situated on top of the resistors, and are comprised of a polysilicon layer covering the resistors, a silicon nitride insulator, and an aluminum top plate. The polysilicon layer is connected to V_{SS} or the op-amp output.

The frequency response was determined by computer analysis. This is shown in Fig. 13 for minimum acceptable component values. The circles indicate measurements made on a typical sample and show them to be well within the requirement.

G. The Reference Voltage Source

Variations in the tone level output to the line can have two causes: variations in the gain of the transmission circuit, and variations in the voltage reference source. CEPT CS203 allows a total tolerance of ± 2 dB under all supply and ambient temperature conditions. 1 dB for each of the above causes seems a reasonable allowance.

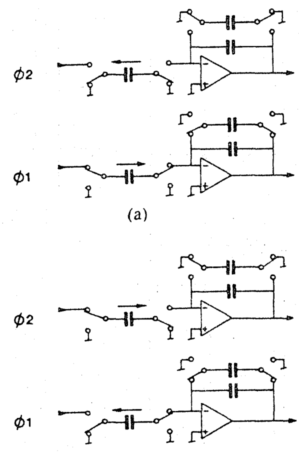


Fig. 11. Switch sequences for negative (a) and positive (b) op-amp output voltages.

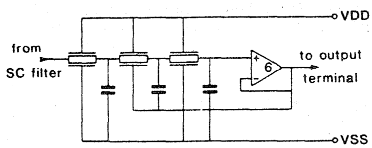


Fig. 12. Circuit of analog *RC* filter. The lines on either side of the resistors represent distributed capacitances.

The reference voltage source is derived by the well-known bandgap principle, not in the least because the process used allows us to implement bipolar transistors and diodes. In Fig. 14, *TR*1 and *TR*2 are connected as diodes with their collectors returned to V_{DD}. *TR*1 switches between a current of I and $9I$; *TR*2 remains constant at I.

The temperature coefficient of delta V_{BE} of *TR*1 is positive. That of the V_{BE} of *TR*2 is negative. We used switched-capacitor techniques to compensate both temperature coefficients, to scale the output voltage, and to shift the output voltage by $(V_{DD} - V_{SS})/2$.[3]

If we assume that the switched capacitor across op-amp 1 has a value of 1C, then the minimum temperature dependence is obtained when the output voltage of the op-amp (with respect to $(V_{DD} - V_{SS})/2$) equals [5]

$$V_{\text{out}} = V_{GO} + A \times V_T$$

where V_{GO} is the bandgap voltage (1205 mV), A is a constant dependent on the semiconductor material and the temperature dependence of the diode current source, and V_T is kT/q.

Preliminary measurements indicated 3.2 as the value for A, resulting in V_{out} being 1280 mV. To reduce this to 200 mV (the required reference voltage), the actual value of the switched capacitance across op-amp 1 must be $(1280/200)C = 6.4C$. Tests and simulations showed, however, that the output voltage was being slightly reduced by clock feedthrough; we compensated for this by reducing the capacitance to 6.25C.

A computer simulation of the reference voltage is given in Fig. 15(a). It was done for a supply voltage of 3 V and with the circuit loaded for both clock phases with the fully discharged capacitor array of both DAC's.

The DAC's sample the reference voltage for both positive and negative half cycles at the end of clock phases 1 and 2, respectively, as indicated by the arrows in Fig. 15(a). Inequalities in the positive and negative amplitudes of a sine wave cause even harmonic distortion. With the inequalities of Fig. 15(a), we are, however, a safe 46 dB below the fundamental.

Fig. 15(b) is an oscillogram showing (top) the reference voltage source operating in single tone mode (the double trace is the result of superpositioning). The lower trace is clock phase 3 (for reference).

H. The Op-Amps

The basic circuit of the op-amps is shown in Fig. 16. It is a well-known structure with an active p-channel output stage and a passive n-channel current sink. The differential stage, with a tail current of 2.4 μA, is identical for all op-amps; only the output stages and current are adapted to the specific requirement. The load on op-amps 1–5 is almost purely capacitive; that of op-amp 6 (the output) is not wholly specified.

[3]This technique was described by O. Leuthold in an unpublished lecture at the IEEE MOS Analog Circuits Meeting, Berne, Switzerland, Oct. 22, 1981.

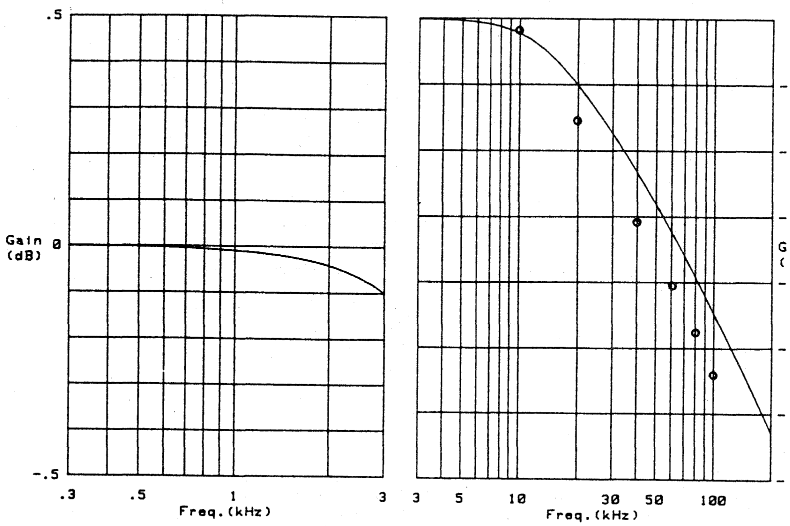


Fig. 13. Calculated response of the analog filter for minimum component values. The circles show measured response, which is well within the requirement.

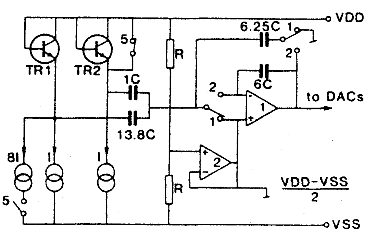


Fig. 14. Circuit diagram of voltage reference circuit.

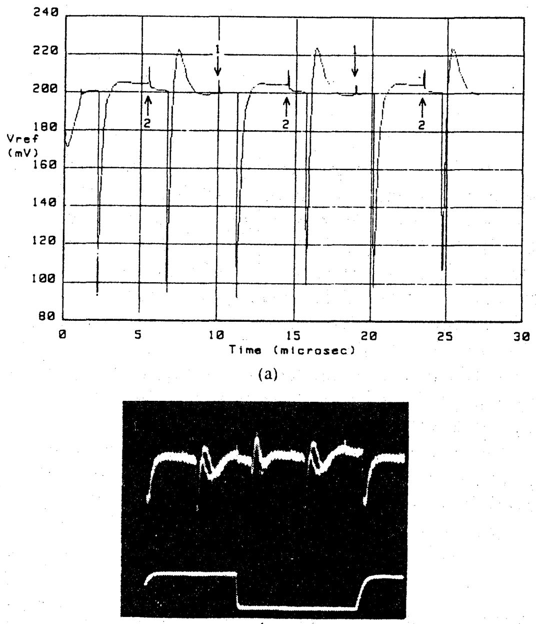
A first approximation in the frequency domain, based on the feedback and load situation shown in Fig. 17(a), gave us f_1 [see broken line in Fig. 17(b)]. The low-frequency gain is

$$G_1 = \frac{C_a + C_b}{C_b}.$$

With settling error being a first-order effect (f_2 being much higher than f_1), and accepting an error of 0.1 percent, we can take it as equal to $\exp(-2\pi f_2 t)$. With t being at maximum $3/8 \times 8.94$ μs (see Fig. 9), the minimum frequency of f_1 is 328 kHz. The gain-bandwidth product is, therefore, $G_1 \times 328$ kHz and, as only the output stage is adapted, it is mainly dependent on the Miller capacitance.

To a first approximation, f_2 is determined by the load capacitance $(C_aC_b/(C_a + C_b) + C_c)$ and by the width-to-length ratio of the output p-channel FET and the current through it. With this, we designed the output stage for an f_2 of 1.2 f_1, while adding an extra 5 pF to the load to allow on-chip measurements with a test probe.

Frequency domain analysis takes, of course, no account of such factors as slew rate. So, after calculating the maximum output voltage step, which occurs when the zero crossings of two tone signals of the highest frequencies coincide, we made a final analysis in the time domain. Time domain analysis must be done for both positive and negative output steps and for both clock phases. As an



(a)

(b)

Fig. 15. (a) Simulated output of voltage reference circuit. Performed for a supply voltage of 3 V and with the circuit loaded on both clock phases with the completely discharged capacitor array of both DAC's. The numbered arrows indicate the end of clock-phases 1 and 2. (b) Measured output of voltage reference circuit (upper trace: 20 mV/div) with clock-phase 3 (lower trace: 2 V/div). Time base: 1 μs/div.

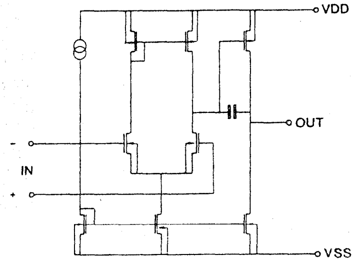


Fig. 16. Circuit of standard op-amp's.

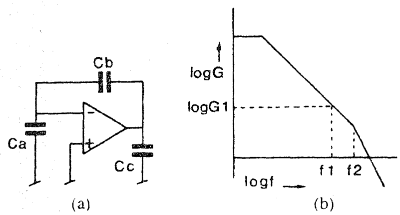


Fig. 17. Feedback and load capacitances of an op-amp (a) and calculation of pole frequencies (f_1 and f_2) and gain–bandwidth product (b).

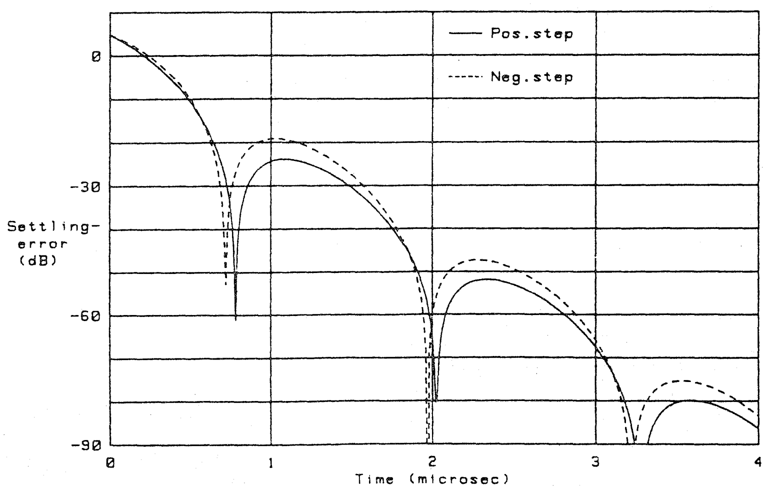


Fig. 18. Settling error (signless) of op-amp 3 during clock-phase 1 versus time.

example, Fig. 18 shows the settling error in the output of op-amp 3 during clock-phase 1. It can be seen that it is 60 dB down within the required 3.35 μs. The second-order behavior is quite obvious.

The offset voltage of the op-amps on either side of the DAC's determines in large measure the amplitudes and distortion in the output. By offset compensating (chopped input stage [6]) op-amps 1 and 3 (see Fig. 19), the parasitic effects are modulated to half the clock frequency where subsequent filters will attenuate them sufficiently.

I. The Switches

All switches are combinations of identically dimensioned p- and n-channel devices. The switch signals driving them are in antiphase so that switch signal feedthrough is largely compensated. The dimensions, and thereby the series resistance, of each switch are adapted to the capacitance connected to it. The settling error of the voltage on the capacitance is $\exp(-t/RC)$. Because a switch sometimes has more than one capacitance connected to it, and

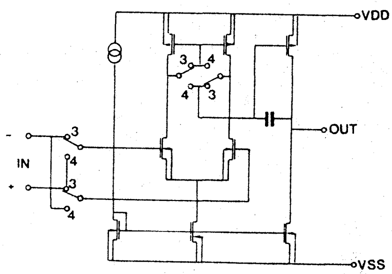


Fig. 19. Circuit of offset compensated op-amps.

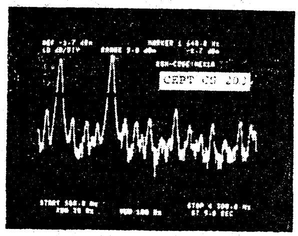


Fig. 21. Frequency spectrum of the output from 0.3 to 4.3 kHz with CEPT CS203 recommendations.

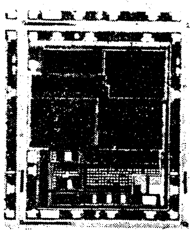


Fig. 20. Photomicrograph of chip.

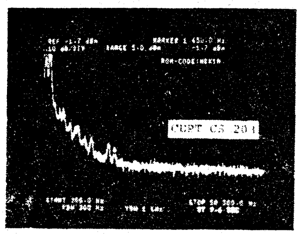


Fig. 22. As Fig. 21, but extending to 50.3 kHz.

because, in general, capacitors are connected to more than one switch, careful examination is needed to determine the value of C in the above equation. Settling error of the SC switches was held to 0.1 percent.

IV. Circuit Implementation

A. Technology

The circuit is manufactured in a CMOS process using 4 μm design rules. Boron is implanted to form a p-well in $\langle 100 \rangle$ oriented 4.5 $\Omega \cdot$cm n-type silicon. Polysilicon is deposited on 500 Å gate oxide which is then covered by silicon nitride (Si_3N_4). Both layers are plasma etched and N_2 implanted. This prevents the growth of silicon oxide in the source and drain regions during a subsequent oxidation process, allowing it only on the "open" vertical edges of the polysilicon. The polysilicon is thereby completely insulated by SiO_2 or Si_3N_4, which allows the contacts to be placed with zero gap to the gate (design rule). For this reason, it is called the self-aligned contact (SAC) process [7]. It makes possible the formation of floating capacitors without additional process stages. The plates are metal and polysilicon with nitride as the dielectric. The specific capacity is 620 pF/mm^2.

B. Layout

Fig. 20 shows a micrograph of the device, which measures 2.72×3.34 mm^2. The input stage for the serial and parallel modes is in the upper right-hand corner. In the upper left-hand corner is the ROM which translates the data input, with the latches to store it right below. The two identical blocks of logic in the middle of the chip contain the counters and the data concerning the sine wave synthesis. Below, on the left, is the clock generator and, towards the right, the analog circuits in the following sequence: SC voltage reference, SC filter and DAC's, and finally, the RC filter.

C. Packaging

The chip is assembled in DIL-8 or SO-8 package for the serial I^{2}C bus (PCD3312) or in a DIL-14 package for serial and parallel (6 bit) input (PCD3311).

V. Circuit Performance and Results

The complete performance can be seen simply from the output signal. Figs. 21 and 22 show the spectrum of the output, together with the CEPT recommendations. The preemphasis is 2.15 dB$\pm$0.15 dB over the full supply and temperature range. The behavior of the output under the same conditions is shown in Fig. 23. The output voltage stability is largely due to the stability of the reference source.

VI. Conclusions

A tone generator has been described which includes an on-chip voltage reference for output stability and filters for reducing unwanted frequency components to meet CEPT recommendations on distortion and output level. The cir-

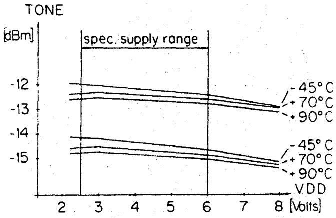


Fig. 23. Output signals versus supply voltage and ambient temperature.

cuit provides all DTMF frequency combinations as well as their single tones, modem frequencies, and two octaves of the musical scale in semitone steps. The standby current is typically below 1.5 μA over the full range of voltage (2.5–6.0 V) and temperature (-25–$+70$°C).

Acknowledgment

The authors would like to thank U. W. Brugger for the initial filter design concept, A. H. M. van Roermond for his assistance in the filter and op-amp design, J. J. van der Kam for spectrum calculations, and J. Fasser for general assistance in design and layout questions.

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APPLICATION NOTE Nr	No number
TITLE	Quick reference data of Speech/transmission IC's: TEA106X, TEA111X and PCA1070
AUTHOR	F. van Dongen
DATE	March 1996

DC - CHARACTERISTICS

IC - type	V_{LN}(V) @ 15mA	V_{LN}(V) @ R_{REG-LN} (Ω)	V_{LN}(V) @ $R_{REG-SLPE}$ (Ω)
TEA1060	4.45 ± 0.20	3.80 ± 0.25 68k	5.0 ± 0.30 39k
TEA1061	4.45 ± 0.20	3.80 ± 0.25 68k	5.0 ± 0.30 39k
TEA1062/A	4.0 + 0.25/ − 0.45	3.50 68k	4.5 39k
TEA1064A 1)	3.60 + 0.4/ − 0.2		4.6 20k
TEA1064B	3.50 ± 0.25		4.4 ± 0.35 20k
TEA1065	4.45 ± 0.30	3.90 ± 0.25 68k	5.0 ± 0.30 39k
TEA1066T	4.45 ± 0.20	3.80 0.25 68k	5.0 ± 0.30 39k
TEA1067	3.90 ± 0.25	3.40 ± 0.30 68k	4.5 ± 0.30 39k
TEA1068	4.45 ± 0.25	3.80 ± 0.30 68k	5.0 ± 0.35 39k
TEA1112/A	3.65 ± 0.25		4.4 27k
TEA1113	4.0 ± 0.30	3.6 82k	4.75 27k
TEA1118/A	3.65 ± 0.25		4.4 27k
PCA1070 2)	5.0 ± 0.40 @ 20mA	2)	2)

MICROPHONE GAIN 3)

IC - type	Setting range (dB)	Gain (dB) @ R_{gas} = 68kΩ	Default setting (dB)
TEA1060	44 - 60	52 ± 1.0	
TEA1061	30 - 46	38 ± 1.0	
TEA1062/A	44 - 52	52 ± 1.5	
TEA1064A	44 - 52	52 ± 1.0	
TEA1064B	44 - 52	52 ± 1.0	
TEA1065	30 - 46	38 ± 1.0	
TEA1066T	30 - 46	38 ± 1.0	
	44 - 60	52 ± 1.0	
TEA1067	44 - 52	52 ± 1.0	
TEA1068	44 - 60	52 ± 1.0	
TEA1112/A	39 - 52		52 ± 1.0 4)
TEA1113	39 - 52		52 ± 1.0 4)
TEA1118	0 - 11		11 4)
TEA1118A	Fixed gain: 11		
PCA1070	30 - 51	2)	41 ± 1.5

RECEIVE GAIN 3)

IC - type	Setting range (dB)	Gain (dB) @ R_{gar} = 100kΩ		Default setting (dB) SEL
		SEL	BTL	
TEA1060	17 - 39	25 ± 1.0	31 ± 1.0	
TEA1061	17 - 39	25 ± 1.0	31 ± 1.0	
TEA1062/A	20 - 31	31 ± 1.5		
TEA1064A	20 - 45	31 ± 1.0	37 ± 1.0	
TEA1064B	20 - 45	31 ± 1.0	37 ± 1.0	
TEA1065	20 - 45	31 ± 1.0	37 ± 1.0	
TEA1066T	17 - 39	25 ± 1.0	31 ± 1.0	
TEA1067	20 - 45	31 ± 1.0	37 ± 1.0	
TEA1068	17 - 39	25 ± 1.0	31 ± 1.0	
TEA1112/A	19 - 31			31 ± 1.0 4)
TEA1113	19 - 31			31 ± 1.0 4)
TEA1118/A	19 - 31			31 ± 1.0 4)
PCA1070	−25 - +11 5)	2)	2)	−12 ± 1.5 5)

DTMF GAIN

IC - type	DTMF gain (dB) / general	Gain (dB) @ R_{gas} = 68kΩ	Default setting (dB)
TEA1060	Microphone gain – 26.5	25.5 ± 1.0	
TEA1061	Microphone gain – 12.5	25.5 ± 1.0	
TEA1062/A	Microphone gain – 26.5	25.5 ± 1.5	
TEA1064A	Microphone gain – 26.0	26.0 ± 1.0	
TEA1064B	Microphone gain – 26.0	26.0 ± 1.0	
TEA1065	Microphone gain – 12.5	25.5 ± 1.0	
TEA1066T	Microph. gain Low - 12.5	25.5 ± 1.0	
	Microph. gain High - 26.5	25.5 ± 1.0	
TEA1067	Microphone gain – 26.5	25.5 ± 1.0	
TEA1068	Microphone gain – 26.5	25.5 ± 1.0	
TEA1112/A	Microphone gain – 26.5		25.5 ± 1.0
TEA1113	Microphone gain – 26.5		25.5 ± 1.0
TEA1118	No DTMF amplifier		
TEA1118A	Fixed gain: 17		
PCA1070	Default: microph.gain – 20.0 Range: 1 - 21	2)	21.0 ± 1.0

Notes:
1) Stabilized supply-voltage. Option for regulated line-voltage
2) Programmable transmission IC / parameters programmable via I^2C-bus
3) AGC range: 6 dB
4) On-chip default setting; gain reduction by external resistor
5) Total receive gain from LN to QR

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