Signetics

Linear Data Manual Volume 3 Video

Signetics

Linear Data Manual Video 3

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Linear Products

1989 Linear Data Manual Volume 3: Video

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Preface

Linear Products

The Linear Division, one of four Signetics product divisions, is a major supplier of a broad line of linear integrated circuits ranging from high performance application specific designs to many of the more popular industry standard devices.

A fifth Signetics division, the Military Division, provides military-grade integrated circuits, including Linear. Please consult the Signetics Military data book for information on such devices.

Employing Signetics' high quality processing and screening standards, the Linear Division is dedicated to providing high-quality linear products to our customers worldwide.

The three 1989 Linear Data and Applications Manuals provide extensive technical data and application information for a broad range of products serving the needs of a wide variety of markets.

Volume 1 — Communications:
Contains data and application information concerning our radio and audio circuits, compandors, phase-locked loops, compact disk circuits, and ICs for RF communication, fiber optic communication, telephony and modem applications.

Volume 2 — Industrial:
Contains data and application information concerning our data conversion products (analog-to-digital and digital-to-analog), sample-and-hold circuits, comparators, driver/receiver ICs, amplifiers, position measurement devices, power conversion and control ICs and music/speech synthesizers.

Volume 3 — Video:
Contains data and application information concerning our video products. This includes tuning, video IF and audio IF circuits, sync processors/generators, color decoders and encoders, video processing ICs, vertical deflection circuits, and power supply controllers for video applications.

Each volume contains extensive product-specific application information. In addition there are selector guides and product-specific symbols and definitions to facilitate the selection and understanding of Linear products. A functional Table of Contents for each of the three volumes and a complete product and application note listing is also included.

Although every effort has been made to ensure the accuracy of information in these manuals, Signetics assumes no liability for inadvertent errors.

Your suggestions for improvement in future editions are welcome.

Product Status

DEFINITIONS		
Data Sheet Identification	**Product Status**	**Definition**
Objective Specification	**Formative or In Design**	This data sheet contains the design target or goal specifications for product development. Specifications may change in any manner without notice.
Preliminary Specification	**Preproduction Product**	This data sheet contains preliminary data and supplementary data will be published at a later date. Signetics reserves the right to make changes at any time without notice in order to improve design and supply the best possible product.
Product Specification	**Full Production**	This data sheet contains Final Specifications. Signetics reserves the right to make changes at any time without notice in order to improve design and supply the best possible product.

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Application Notes by Part Numbers

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Application Notes by Part Numbers

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Volume 2
Industrial

Cross Reference Guide by Manufacturer

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Pin-for-Pin Functionally-Compatible* Cross Reference by Manufacturer

Manufacturer	Manufacturer Part Number	Signetics Part Number	Temperature Range (°C)	Package
AMD	AM26LS30PC	AM26LS30CN	0 to +70	Plastic
	AM26LS31PC	AM26LS31CN	0 to +70	Plastic
	AM26LS32PC	AM26LS32CN	0 to +70	Plastic
	AM25LS33PC	AM26LS33CN	0 to +70	Plastic
	AM6012DC	AM6012F	0 to +70	Ceramic
	DAC-08AQ	DAC-08AF	-55 to +125	Ceramic
	DAC-08CN	DAC-08CN	0 to +70	Plastic
	DAC-08CQ	DAC-08CF	0 to +70	Ceramic
	DAC-08EN	DAC-08EN	0 to +70	Plastic
	DAC-08EQ	DAC-08EF	0 to +70	Ceramic
	DAC-08HN	DAC-08HN	0 to +70	Plastic
	DAC-08HQ	DAC-08HF	0 to +70	Ceramic
	DAC-08Q	DAC-08F	-55 to +125	Ceramic
	LF198H	LF198H	-55 to +125	Metal Can
	LF198H	SE5537H	-55 to +125	Metal Can
	LF398H	LF398H	0 to +70	Metal Can
	LF398H	NE5537H	0 to +70	Metal Can
	LF398L	LF398D	0 to +70	Plastic
	LF398L	NE5537D	0 to +70	Plastic
	LF398N	LF398N	0 to +70	Plastic
	LF398N	NE5537N	0 to +70	Plastic
Datel	AM-453-2	NE5534/AF	0 to +70	Ceramic
	AM-453-2C	NE5534/AF	0 to +70	Ceramic
	AM-453-2M	SE5534/AF	-55 to +125	Ceramic
	DAC-UP10BC	NE5020N	0 to +70	Plastic
	DAC-UP8BC	NE5018N	0 to +70	Plastic
	DAC-UP8BM	SE5019F	-55 to +125	Ceramic
	DAC-UP8BQ	SE5018F	-55 to 125	Ceramic
Exar	XR-558CN	NE558F	0 to +70	Ceramic
	XR-558CP	NE558N	0 to +70	Plastic
	XR-558M	SE558F	-55 to +125	Ceramic
	XR-L567CN	NE567F	0 to +70	Ceramic
	XR-L567CP	NE567N	0 to +70	Plastic
	XR-1488CP	MC1488N	0 to +70	Plastic
	XR-1489/ACP	MC1489/AN	0 to +70	Plastic
	XR-1524N	SG3524F	0 to +70	Ceramic
	XR-1524P	SG3524N	0 to +70	Plastic
	XR-2524P	SG3524N	0 to +70	Plastic
	XR-3524N	SG3524F	0 to +70	Ceramic
	XR-3524P	SG3524N	0 to +70	Plastic
	XR-4558CP	NE4558N	0 to +70	Plastic
	XR-5532/A N	NE5532/AF	0 to +70	Ceramic
	XR-5532/A P	NE5532/AN	0 to +70	Plastic
	XR-5534/A CN	NE5534/AF	0 to +70	Ceramic
	XR-5534/A CP	NE5534/AN	0 to +70	Plastic
	XR-5534/A M	SE5534/AF	-55 to +125	Ceramic
	XR-6118CP	NE594N	0 to +70	Plastic
	XR-13600CP	NE5517N	0 to +70	Plastic
Harris	HA-2539N	NE5539N	0 to +70	Plastic
	HA-2420-2/8B	SE5060F	-55 to +125	Ceramic
	HA-2425N	NE5060N	0 to +70	Plastic
	HA-2425B	NE5060F	0 to +70	Ceramic
	HA-5320B	NE5060F	0 to +70	Ceramic

Manufacturer	Manufacturer Part Number	Signetics Part Number	Temperature Range (°C)	Package
	HA1-5102-2	SE5532/AF	-55 to +125	Ceramic
	HA1-5135-2	SE5534/AF	-55 to +125	Ceramic
	HA1-5135-5	NE5534/AF	0 to +70	Ceramic
	HA1-5202-5	NE5532/AF	0 to +70	Ceramic
	HA3-5102-5	NE5532/AN	0 to +70	Plastic
Intersil	ADC0803LCD	ADC0803-1 LCF	-40 to +85	Ceramic
	ADC0804	ADC0804-1 CN	0 to +70	Plastic
	ADC0805	ADC0805-1 LCN	-40 to +85	Plastic
	ICM7555CBA	ICM7555CD	0 to +70	Plastic
	ICM7555IPA	ICM7555IN	-40 to +85	Plastic
	ACM7555CP	ICM7555CN	0 to +70	Plastic
Motorola	AM26LS31PCD	AM26LS31CD	0 to +70	Plastic
	AM26LS31PC	AM26LS31CN	0 to +70	Plastic
	AM26LS32PC	AM26LS32CN	0 to +70	Plastic
	AM26LS32PCD	AM26LS32CD	0 to +70	Plastic
	DAC-08CD	DAC-08CN	0 to +70	Plastic
	DAC-08CQ	DAC-08CF	0 to +70	Ceramic
	DAC-08ED	DAC-08EN	0 to +70	Plastic
	DAC-08EF	DAC-08EF	0 to +70	Ceramic
	DAC-08HQ	DAC-08HF	0 to +70	Ceramic
	DAC-08Q	DAC-08F	-55 to +125	Ceramic
	LM2901N	LM2901N	-40 to +85	Plastic
	LM311J-8	LM311F	0 to +70	Ceramic
	LM311N	LM311N	0 to +70	Plastic
	LM324J	LM324F	0 to +70	Ceramic
	LM324N	LM324N	0 to +70	Plastic
	LM339/A J	LM339/AF	0 to +70	Ceramic
	LM339/A N	LM339/AN	0 to +70	Plastic
	LM358N	LM358N	0 to +70	Plastic
	LM393A/J	LM393/AF	0 to +70	Ceramic
	LM393A/N	LM393/AN	0 to +70	Plastic
	MC1408L	MC1408F	0 to +70	Ceramic
	MC1408P	MC1408N	0 to +70	Plastic
	MC1488L	MC1488F	0 to +70	Ceramic
	MC1488P	MC1488N	0 to +70	Plastic
	MC1489/A L	MC1489/AF	0 to +70	Ceramic
	MC1489/A P	MC1489/AN	0 to +70	Plastic
	MC1496L	MC1496F	0 to +70	Ceramic
	MC1496P	MC1496N	0 to +70	Plastic
	MC3302L	MC3302F	-40 to +85	Ceramic
	MC3302P	MC3302N	-40 to +85	Plastic
	MC3361D	MC3361D	0 to +70	Plastic
	MC3361P	MC3361N	0 to +70	Plastic
	MC3403L	MC3403F	0 to +70	Ceramic
	MC3403P	MC3403N	0 to +70	Plastic
	MC3410CL	MC3410CF	0 to +70	Ceramic
	MC3410L	MC3410F	0 to +70	Ceramic
		NE5410F	0 to +70	Ceramic
	MC3510L	MC5410F	-55 to +125	Ceramic
	NE565N	NE565N	0 to +70	Plastic
	NE592F	NE592F-8	0 to +70	Ceramic
	NE592F	NE592F-14	0 to +70	Ceramic
	NE592N	NE592N-14	0 to +70	Plastic

Cross Reference Guide by Manufacturer

Manufacturer	Manufacturer Part Number	Signetics Part Number	Temperature Range (°C)	Package
	SE592F	SE592F-8	-55 to +125	Ceramic
	SE592F	SE592F-14	-55 to +125	Ceramic
	SE592H	SE592H	-55 to +125	Metal Can
National	ADC0803F	ADC0803-1 LCF	-40 to +85	Ceramic
	ADC0803N	ADC0803-1 LCN	-40 to +85	Plastic
	ADC0805	ADC0805-1 LCN	-40 to +85	Plastic
	ADC0820CCN	ADC0820CNEN	0 to +70	Plastic
	ADC0820CCD	ADC0820CSAN	-40 to +85	Plastic
	ADC0820CD	ADC0820CSEF	-55 to +125	Ceramic
	DAC0800LCJ	DAC-08EF	0 to +70	Ceramic
	DAC0800LJ	DAC-08F	-55 to +125	Ceramic
	DAC0800LCN	DAC-08EN	0 to +70	Plastic
	DAC0801LCJ	DAC-08CF	0 to +70	Ceramic
	DAC0801LCN	DAC-08CN	0 to +70	Plastic
	DAC0802LJ	DAC-08AF	-55 to +125	Ceramic
	DAC0802LCJ	DAC-08HF	0 to +70	Ceramic
	DAC0802LCN	DAC-08HN	0 to +70	Plastic
	DAC0806LCJ	MC1408-6F	0 to +70	Ceramic
	DAC0806LCN	MC1408-6N	0 to +70	Plastic
	DAC0807LCJ	MC1408-7F	0 to +70	Ceramic
	DAC0807LCN	MC1408-7N	0 to +70	Plastic
	DAC0808LCJ	MC1408F	0 to +70	Ceramic
	DAC0808LCN	MC1408N	0 to +70	Plastic
	DAC0808LD	MC1408F	0 to +70	Ceramic
	DS3691N	AM26LS30CN	0 to +70	Plastic
	DS3691M	AM26LS30CD	0 to +70	Plastic
	LF198H	SE5537H	-55 to +125	Metal Can
	LF398H	NE5537H	0 to +70	Metal Can
	LF398N	NE5537N	0 to +70	Plastic
	LM13600AN	NE5517N	0 to +70	Plastic
	LM13600N	NE5517N	0 to +70	Plastic
	LM1458N	MC1458N	0 to +70	Plastic
	LM161H	SE529H	-55 to +125	Metal Can
	LM161J	SE529F	-55 to +125	Ceramic
	LM2524J	SG3524F	0 to +70	Ceramic
	LM2524N	SG3524N	0 to +70	Plastic
	LM2901N	LM2901N	-40 to +85	Plastic
	LM2903N	LM2903N	-40 to +85	Plastic
	LM3089	CA3089N	-55 to +125	Plastic
	LM319J	LM319F	0 to +70	Ceramic
	LM319N	LM319N	0 to +70	Plastic
	LM324J	LM324F	0 to +70	Ceramic
	LM324N	LM324N	0 to +70	Plastic
	LM324AD	LM324AD	0 to +70	Plastic
	LM324AN	LM324AN	0 to +70	Plastic
	LM339/AJ	LM339/AF	0 to +70	Ceramic
	LM339/AN	LM339/AN	0 to +70	Plastic
	LM3524J	SG3524F	0 to +70	Ceramic
	LM3524N	SG3524N	0 to +70	Plastic
	LM358H	LM358H	0 to +70	Metal Can
	LM358N	LM358N	0 to +70	Plastic
	LM361H	NE529H	0 to +70	Metal Can
	LM361J	NE529D	0 to +70	
	LM361N	NE529N	0 to +70	Plastic
	LM393/AN	LM393/AN	0 to +70	Plastic
	LM555J	NE555F	0 to +70	Ceramic
	LM555N	NE555N	0 to +70	Plastic
	LM556J	SE556-1F	-55 to +125	Ceramic
	LM556N	SE556-1N	-55 to +125	Plastic
	LM556CJ	NE556-1F	0 to +70	Ceramic
	LM556CN	NE556-1N	0 to +70	Plastic

Manufacturer	Manufacturer Part Number	Signetics Part Number	Temperature Range (°C)	Package
	LM565CN	NE565N	0 to +70	Plastic
	LM566N	SE566N	-55 to +125	Plastic
	LM566CN	NE566N	0 to +70	Plastic
	LM567CN	NE567N	0 to +70	Plastic
	LM733CN	μA733CN	0 to +70	Plastic
	LM741CJ	μA741CF	0 to +70	Ceramic
	LM741CN	μA741CN	0 to +70	Plastic
	LM741J	μA741F	-55 to +125	Ceramic
	LM741N	μA741N	-55 to +125	Plastic
	LM747CJ	μA747CF	0 to +70	Ceramic
	LM747CN	μA747CN	0 to +70	Plastic
	LM747J	μ747F	-55 to +125	Ceramic
	LM747N	μA747N	-55 to +125	Plastic
	LMC555CN	ICM7555CN	0 to +70	Plastic
	LMC555CM	ICM7555CD	0 to +70	Plastic
	μA080/DA	DAC-08F	0 to +70	Ceramic
	μA0801CDC	MC1408F	0 to +70	Ceramic
	μA0801CPC	MC1408N	0 to +70	Plastic
	μA0801EDC	DAC-08EF	0 to +70	Ceramic
	μA0801EPC	DAC-08AF	0 to +70	Ceramic
	μA124J	LM124F	-55 to +125	Ceramic
	μA1458TC	MC1458N	0 to +70	Plastic
	μA1488DC	MC1488F	0 to +70	Ceramic
	μA1488PC	MC1488N	0 to +70	Plastic
	μA1489/A PC	MC1489/AF	0 to +70	Ceramic
	μA1489/A PC	MC1489/AN	0 to +70	Plastic
	μA198HM	NE5537H	0 to +70	Metal Can
	μA198RM	NE5537N	0 to +70	Plastic
	μA2901DC	LM2901F	-40 to +85	Ceramic
	μA2901PC	LM2901N	-40 to +85	Plastic
	μA311RC	LM311F	0 to +70	Ceramic
	μA324DC	LM324F	0 to +70	Ceramic
	μA324PC	LM324N	0 to +70	Plastic
	μA3302DC	MC3302F	-40 to +85	Ceramic
	μA3302PC	MC3302N	-40 to +85	Plastic
	μA339/ADC	LM339/AF	0 to +70	Ceramic
	μA339/APC	LM339/AN	0 to +70	Plastic
	μA3403DC	MC3403F	0 to +70	Ceramic
	μA3403PC	MC3403N	0 to +70	Plastic
	μA398HC	SE5537H	-55 to +125	Metal Can
	μA398RC	SE5537N	-55 to +125	Plastic
	μA555TC	NE555N	0 to +70	Plastic
	μA556PC	NE556-1N, NE556N	0 to +70	Plastic
	μA723DC	μA723CF	0 to +70	Ceramic
	μA723DM	μA723F	-55 to +125	Ceramic
	μA723PC	μA723CN	0 to +70	Plastic
	μA733DC	μA733F	0 to +70	Ceramic
	μA733DM	μA733F	-55 to +125	Ceramic
	μA733PC	μA733N	0 to +70	Plastic
	μA741NM	μA741N	-55 to +125	Plastic
	μA741RC	μA741CF	0 to +70	Ceramic
	μA741TC	μA741CN	0 to +70	Plastic
	μA747DC	μA747CF	0 to +70	Ceramic
	μA747PC	μA747CN	0 to +70	Plastic
	UC3842D	UC3842D	0 to +70	Plastic
	UC3842J	UC3842FE	0 to +70	Ceramic
	UC3842N	UC3842N	0 to +70	Plastic
	UC2842D	UC2842D	0 to +70	Plastic
	UC2842J	UC2842FE	0 to +70	Ceramic
	UC2842N	UC2842N	0 to +70	Plastic

Cross Reference Guide

Manufacturer	Manufacturer Part Number	Signetics Part Number	Temperature Range (°C)	Package
	UC1842J	UC1842FE	-55 to +125	Ceramic
	UC1842N	UC1842N	-55 to +125	Plastic
NEC	μPC1571C	NE571N	0 to +70	Plastic
PMI	CMP-05GP	NE5105N	0 to +70	Plastic
	CMP-05CZ	SE5105F	-55 to +125	Ceramic
	CMP-05BZ	SE5105F	-55 to +125	Ceramic
	CMP-05GZ	SA5105N	-40 to +85	Plastic
	CMP-05FZ	SA5105N	-40 to +85	Plastic
	DAC1408A-6P	MC1408-6N	0 to +70	Plastic
	DAC1408A-6Q	MC1408-6F	0 to +70	Ceramic
	DAC1408A-7N	MC1408-7N	0 to +70	Plastic
	DAC1408A-7Q	MC1408-7F	0 to +70	Ceramic
	DAC1408A-8N	MC1408-8N	0 to +70	Plastic
	DAC1408A-8Q	MC1408-8F	0 to +70	Ceramic
	DAC1508A-8Q	MC1408-8F	-55 to +125	Ceramic
	DAC312FR	AM6012F	0 to +70	Ceramic
	OP27BZ	SE5534AFE	-55 to +125	Ceramic
	OP27CZ	SE5534FE	-55 to +125	Ceramic
	PM747Y	μA747N	-55 to +125	Plastic
	SMP-10AY	SE5060F	-55 to +125	Ceramic
	SMP-10EY	NE5060N	0 to +70	Plastic
	SMP-11AY	SE5060F	-55 to +125	Ceramic
	SMP-11EY	NE5060N	0 to +70	Plastic
Raytheon	RC4805DE	NE5105N	0 to +70	Plastic
	RC4805EDE	NE5105AN	0 to +70	Plastic
	RM4805DE	SE5105F	-55 to +125	Ceramic
	RM4805ADE	SE5105AF	-55 to +125	Ceramic
	RC5532/A DE	NE5532/AF	0 to +70	Ceramic
	RC5532/A NB	NE5532/AN	0 to +70	Plastic
	RC5534/A DE	NE5534/AF	0 to +70	Ceramic
	RC5534/A NB	NE5534/AN	0 to +70	Plastic
	RM5532/A DE	SE5532/AF	-55 to +125	Ceramic
	RM5534/A DE	SE5534/AF	-55 to +125	Ceramic
Silicon General	SG3524J	SG3524F	0 to +70	Ceramic
	SG3526N	SG3526N	0 to +70	Plastic
Sprague	UDN6118A	SA594N	-40 to +85	Plastic
	UDN6118R	SA594F	-40 to +85	Ceramic
	ULN3524A	SG3524	0 to +70	Plastic
	ULN8142M	UC3842N	0 to +70	Plastic
	ULN8160A	NE5560N	0 to +70	Plastic
	ULN8160R	NE5560F	0 to +70	Ceramic
	ULN8161M	NE5561N	0 to +70	Plastic
	ULN8168M	NE5568N	0 to +70	Plastic
	ULN8564A	NE564N	0 to +70	Plastic
	ULN8564R	NE564F	0 to +70	Ceramic
	ULS8564R	SE564F	-55 to +125	Ceramic
TI	ADC0803N	ADC0803-1 LCN	-40 to +85	Plastic
	ADC0804CN	ADC0804-1 CN	0 to +70	Plastic
	ADC0805N	ADC0805-1 LCN	-40 to +85	Plastic
	LM111J	LM111F	-55 to +125	Ceramic

Manufacturer	Manufacturer Part Number	Signetics Part Number	Temperature Range (°C)	Package
	LM311D	LM311D	0 to +70	Plastic
	LM311J	LM311F	0 to +70	Ceramic
	LM311JG	LM311FE	0 to +70	Ceramic
	LM324D	LM324N	0 to +70	Plastic
	LM324J	LM324F	0 to +70	Ceramic
	LM339/AJ	LM339/AF	0 to +70	Ceramic
	LM339/AN	LM339/AN	0 to +70	Plastic
	LM358P	LM358N	0 to +70	Plastic
	LM393/A P	LM393/AN	0 to +70	Plastic
	MC1458P	MC1458N	0 to +70	Plastic
	NE5532/A JG	NE5532/AF	0 to +70	Ceramic
	NE5532/A P	NE5532/AN	0 to +70	Plastic
	NE5534/A JG	NE5534/AF	0 to +70	Ceramic
	NE5534/A P	NE5534/AN	0 to +70	Plastic
	NE555JG	NE555N	0 to +70	Plastic
	NE555P	NE555N	0 to +70	Plastic
	NE556P	NE556N	0 to +70	Plastic
	NE556J	NE556-1F	0 to +70	Ceramic
	NE556N	NE556-1N	0 to +70	Plastic
	NE592	NE592N14	0 to +70	Plastic
	NE592A	NE592F14	0 to +70	Ceramic
	NE592J	NE592F	0 to +70	Ceramic
	NE592N	NE592N-14	0 to +70	Plastic
	SA556P	SA556N	-40 to +85	Plastic
	SE5534/A JG	SE5534/AF	-55 to +125	Ceramic
	SE555JG	SE555N	-55 to +125	Plastic
	SE556J	SE556-1F	-55 to +125	Ceramic
	SE556N	SE556-1N	-55 to +125	Plastic
	SE592	SE592N14	-55 to +125	Plastic
	SE592J	SE592F-14	-55 to +125	Ceramic
	SE592N	SE592N-14	-55 to +125	Plastic
	SN55107AJ	NE521F	0 to +70	Plastic
	SN55108AJ	SE522F	-55 to +125	Ceramic
	SN75107AJ	NE521F	0 to +70	Plastic
	SN75107AN	NE521N	0 to +70	Plastic
	SN75108AJ	NE522F	0 to +70	Ceramic
	SN75108AN	NE522N	0 to +70	Plastic
	SN75188J	MC1488F	0 to +70	Ceramic
	SN75188N	MC1488N	0 to +70	Plastic
	SN75189AJ	MC1489AF	0 to +70	Ceramic
	SN75189AN	MC1489AN	0 to +70	Plastic
	SN75189J	MC1489F	0 to +70	Ceramic
	SN75189N	MC1489A	0 to +70	Plastic
	TL592A	NE592F14	0 to +70	Ceramic
	TL592P	NE592NB	0 to +70	Plastic
	μA723CJ	μA723CF	0 to +70	Ceramic
	μA723CN	μA723CN	0 to +70	Plastic
	μA723MJ	μA723F	-55 to +125	Ceramic
Unitrode	UC3524J	SG3524F	0 to +70	Ceramic
	UC3524N	SG3524N	0 to +70	Plastic

*THERE MAY BE PARAMETRIC DIFFERENCES BETWEEN SIGNETICS' PARTS AND THOSE OF THE COMPETITION.

Cross Reference Guide by Numeric Listing

NUMERIC	DESCRIPTION	SIGNETICS	ANALOG DEVICES	EXAR	FAIRCHILD	HITACHI	LINEAR TECH	MOTOROLA	NATIONAL	NEC	PMI	RAY-THEON	RCA	SGS/ THOMSON	SILICON GENERAL	SPRAGUE	TI	OTHERS
DAC-08	8-Bit D/A Converter	DAC-08F DAC-08AF DAC-08CF, CN NE5007F, N DAC-08ED, EN NE5008D, F, N SE5008F DAC-08HF, HN NE5009F, N SE5009F	ADDAC-08		μA080/DA μA0801E	HA17008		DAC-08	DAC-0800 DAC-0801 DAC-0802	μPC624	DAC-08							DATEL DAC-08 AMD DAC-08 Harris-HI5618
08031 0804/ 0805	8-Bit A/D Converter	ADC0803LCF, LCN ADC0804CN, LCD, LCF, LCN, ADC0805 LCN							ADC0803 ADC0804 ADC0805								ADC0803 ADC0804 ADC0805	Intersil ADC0803 0840 0805
0820	8-Bit CMOS A/D Converter	ADC0820 CNED ADC0820CNEN	AD7820						ADC0820									Maxim Max150
111	Voltage Comparator	LM111FE	AD111		μA111		LM111	LM111	LM111		PM111	LM111			SG111		LM111	
119	Dual Comparator	LM119F					LT119 LM119		LM119		PM119							
124	Quad OP Amp	LM124F, N			LM124		LT1014	LM124	LM124				CA124		SG124		LM124	
13600	High Performance Dual Transcon Amp	NE5517AN NE5517D, N		XR13600					LM13600/A									
139	Quad Comparator	LM139AF LM139F, N			μA139			LM139	LM139		PM139 CMP-04	LM139		CA139			LM139	
1408/ 1508	8-Bit D/A Converter	MC1408-6F, N MC1408-7F, N MC1408-8D, F, N MC1508-8F	AD1408		μA0801C	HA17408		MC1408/ 1508	DAC0806 0807 0808		DAC-1408	DAC-1408						Harris HI5618
1458/ 1558	Dual Op Amp	MC1458D, N MC1558N SA1458N			μA1458			MC1458 MC1558	LM1458 LM1558	μPC251	OP-14		CA1458	MC1458			MC1458	Harris CM1458 Samsung MC1458 Micro Power MP OP-14
1488	Quad Line Driver	MC1488D, F, N		XR1488	μA1488			MC1488	DS1488					MC1488			SN75188 MC1488	
1489	Quad Line Receiver	MC1489A, D, F, N MC1489D, F, N		XR1489/ A	μA1489/A			MC1489/A	DS1489/A					MC1489	SG1489/A		SN75189/A MC1489/A	
1496/ 1596	Balanced Modulator/ Demodulator	MC1496F, N MC1596F, N			μA796			MC1496 MC1596	LM1496 LM1596						SG1496			Plessey SL1496
1524	Improved SMPS Control Circuit	SG1524CF, CN		XR1524			LT1524						CA1524	SG1524	SG1524	ULN8124	SG1524	Cherry CS1524 Unitrode UC1524
158	Dual Op Amp	LM158FE, N NE532FE, N						LM158	LM158					LM158			LM158	Intersil CA158
193	Dual Comparator	LM193AFE LM193FE			μA193			LM193/A	LM193/A								LM193/A	

Cross Reference Guide by Numeric Listing (Continued)

NUMERIC	DESCRIPTION	SIGNETICS	ANALOG DEVICES	EXAR	FAIRCHILD	HITACHI	LINEAR TECH	MOTOROLA	NATIONAL	NEC	PMI	RAYTHEON	RCA	SGS/ THOMSON	SILICON GENERAL	SPRAGUE	TI	OTHERS
198	Sample-and-Hold Amp	LF198FE, H SE5537FE, H			µA198		LF198		LF198									AMD LF198 Harris HA2430
211	Voltage Comparator	LM211D, FE, N	AD211					LM211	LM211		PM211				SG211		LM211	
219	Dual Comparator	LM219D, F, N							LM219					TDE0119				
224	Quad Op Amp	LM224D, F, N SA534D, F, N			µA224	HA17224		LM224	LM224					LM224			LM224	
239	Quad Voltage Comparator	LM239AN LM239F, N			µA239			LM239	LM239		PM239 CMP-04	LM239	CA239				LM239	
2524	Improved SMPS Control IC	SG2524CN													SG2524			Cherry CS2524 Unitrode UC2524
258	Dual Op Amp	LM258N SA532D, N			µA258	HA17258		LM258	LM258	µPC258			CA258	LM258			LM258	
2577	Sync with Vert Osc and Driver	TDA2577A												TDA2577				
2593	Horizontal Combination	TDA2593												TDA2593				Plessey TA2593
26LS31	Quad Hi-Speed Line Driver	AM26LS31 CD, CN, IN, MN			AM26LS31			AM26LS31	DS26LS31								AM26LS31	AMD AM26LS31
2901	Quad Voltage Comparator	LM2901D, F, N			µA2901			LM2901	LM2901								LM2901	
2902	Quad Op Amp	LM2902D, N SA534D, F, N			µA2902			LM2902	LM2902								LM2902	
2903	Dual Voltage Comparator	LM2903D, FE, N			µA2903			LM2903	LM2903								LM2903	
2904	Dual Op Amp	LM2904D, N			µA2904			LM2904	LM2904								LM2904	
293	Dual Comparator	LM293AFE, AN LM293FE, N						LM293/A	LM293/A								LM293/A	
3089	FM IF System	CA3089N							LM3089				CA3089					
311	Voltage Comparator	LM311D, FE, N			µA311			LM311	LM311								LM311	
319	High-Speed Dual Comparator	LM319D, F, N							LM319	µPC319				LM319				
324	Quad Op Amp	LM324AD, AN LM324D, F, N			µA324	HA17324		LM324/A	LM324/A					LM324			LM324	Samsung LM324
3302	Quad Voltage Comparator	MC3302D, F, N			µA3303			MC3302										
3303	Quad Op Amp	MC3303F, N			µA3303			MC3303						MC3303			M3303	
3361	Low Power FM IF	MC3361D, N						MC3361										Samsung MC3361
339	Quad Voltage Comparator	LM339AF, AN LM339D, F, N			µA339			LM339/A	LM339/A	µPC339	PM339	LM339	CA339	LM339			LM339	
3403/ 3503	Quad Op Amp	MC3403D, F, N MC3505, F, N		µA3403				MC3403 MC3503				RM4137		MC3403 MC3503			MC3403 MC3503	

Cross Reference Guide by Numeric Listing (Continued)

NUMERIC	DESCRIPTION	SIGNETICS	ANALOG DEVICES	EXAR	FAIRCHILD	HITACHI	LINEAR TECH	MOTOROLA	NATIONAL	NEC	PMI	RAYTHEON	RCA	SGS/ THOMSON	SILICON GENERAL	SPRAGUE	TI	OTHERS
3410/ 3510	10-Bit D/A Converter	MC3410F MC3410CF MC3510F						MC3410/C MC3510										Harris HI-5610
3524	SMPS Control Circuit	SG3524D, F, N		XR3524			LT3524		LM3524				CA3524	SG3524	SG3524	ULN3524	SG3524	Cherry CS3524 Unitrode UC3524
3524C	Improved SMPS Control Circuit	SG3524C, D, N														SG3524B		Unitrode UC3524A
3526	SMPS	SG3526F, N						SG3526							SG3526	ULN8126		Unitrode UC3526
358	Dual Op Amp	LM358AD, AN LM358D, N NE532D, N				HA17358		LM358/A	LM358/A	µPC358	OP-221		CA358/A	LM358			LM358/A	Sanyo LA6358
361	See 529																	
3842	SMPS IC	UC3842N, D						UC3842AN							SG3842M			Unitrode UC3842N/D Cherry CS3842AN
387	See 542																	
393	Dual Comparator	LM393AFE, AN LM393D, N LM393FE-Sole Source				HA17393		LM393/A	LM393/A					LM393			LM393/A	Sanyo LA6393
398	Sample-and-Hold Amp	LF398D, FE, H, N NE5537D, FE, H, N			µA398		LF398		LF398		SMP-10							AMD LF398 Harris HA2425
4558	Dual General Purpose Op Amp	NE4558D, FE, N SA4558FE, N SE4558FE, N		XR4588				MC4558					RC4558					
5007	See DAC-08C																	
5008	See DAC-08E																	
5009	See DAC-08H																	
5018	8-Bit Converter Voltage Out	NE5018D, F, N SE5018F																AMD AM6081 Datel DAC µP8B
5019	8-Bit D/A Converter Voltage Out	NE5019F, N SE5019F																Datel DAC µP8BM
5020	10-Bit D/A Converter Voltage Out	NE5020F, N																Datel DAC µP10
5060	High-Speed Precision Sample-and-Hold Amp	NE5060F	AD583								SMP-10 SMP-11							Harris HA2420 HA2425 HA5320
5105	High-Speed Precision Comparator	NE5105D, N SA5105AN (NE5105AD, AN-sole source)									CMP-05	RCA805						

Cross Reference Guide by Numeric Listing (Continued)

NUMERIC	DESCRIPTION	SIGNETICS	ANALOG DEVICES	EXAR	FAIRCHILD	HITACHI	LINEAR TECH	MOTOROLA	NATIONAL	NEC	PMI	RAYTHEON	RCA	SGS/ THOMSON	SILICON GENERAL	SPRAGUE	TI	OTHERS
5118	8-Bit D/A Converter Current Out	NE5118F, N SE5118F																Datel DAC-UP
5170	Octal Line Driver	NE5170A, N																Unitrode UC5170
5180	Octal Line Receiver	NE5180A, N																Unitrode UC5180
529	High-Speed Comparator	NE529D, F, H, N SE529F, H							LM161 LM361									
531	High Slew Rate Op Amp	NE531FE, H, N										RC4531						Harris HA2515
532	See 358																	
542	Low Noise Dual PreAmp	NE542N							LM387									
5517	See 13600																	
5532	Dual Low Noise Op Amp	NE5532AFE, AN NE5532D, FE, N SE5532AFE, FE		XR5532/A								RC5532/A					NE5532/A	Harris HA35102-5
5533	Dual Low Noise Op Amp	NE5533AN NE5533D, N		XR5533													NE5533/A	
5534	Low Noise Op Amp	NE5534AD, AN (NE5534AFE-sole source) NE5534D, FE, N SA5534AD, AN SA5534N SE5534AFE, AN SE5534FE, N		XR5534							OP-27	RC5534/A					NE5534/A	Analog Systems MA332 Datel AM453-2C Harris HA5101/11
5537	See 398																	
5539	Fast Op Amp	NE5539D, F, N SE5539, F, H	AD5539															Harris HA2539
555	Timer	NE555D, FE, N SA555D, N SE555CN, FE, N		XR555	μA555	HA17555		NE555 MC1455	LM555	μPC555		RC555	CA555	NE555			NE555	Intersil NE555
556	Dual Timer	NE556D, F, N SA556N SE556CN, F, N			μA556			NE556 MC1456	LM556					NE556			NE556	Samsung NE556
5560	SMPS Control Circuit	NE5560D, F, N SE5560F, N														ULN8160 *disc		Cherry CS5560C IPS *disc IP5560C
5561	SMPS Control Circuit	NE5561D, FE, N SE5561FE, N														ULN8161 *disc		Cherry CS5561 IPS *disc IP5561C
5568	SMPS Control Circuit	NE5568D, N														ULN8168 *disc		Cherry CS5568 IPS *disc IP5568C

Cross Reference Guide by Numeric Listing (Continued)

NUMERIC	DESCRIPTION	SIGNETICS	ANALOG DEVICES	EXAR	FAIRCHILD	HITACHI	LINEAR TECH	MOTOROLA	NATIONAL	NEC	PMI	RAYTHEON	RCA	SGS/ THOMSON	SILICON GENERAL	SPRAGUE	TI	OTHERS
558	Quad Timer	NE558D, F, N SA558N SE558F, N		XR558														
564	High Frequency Phase-Locked Loop	NE564N (NE564D, F-sole source)														ULN8564		
565	Phase-Locked Loop	NE565D, F, N SE565F, N						NE565	LM565									
566	Function Generator	NE566D, F, N SE566F, N							LM566									
567	Tone Decoder Phase-Locked Loop	NE567D, F, FE, N SE567FE, F, N (SE567D-sole source)		XR567 XR2567					LM567									MCE MCE-567 Samsung LM567
571	Compandor	NE571D, F, N (SA571D, F, N-sole source)								µPC1571C								
583	See 5060																	
592	Video Amplifier	NE592: D14, D8, F14, F8, H, HD14, HD8, HN14, HN8, N14, N8 SA592D8, N8 SE592: F14, F8, H, N14, N8			µA592C			NE592	LM592								NE592 TL592	Intersil NE592
594	Vacuum Fluorescent, Display Driver	NE594D, F, N SA594D, F, N SE594F, N		XR6118												ULN6188		Sanyo LB1290 Toshiba TD62781
6012	12-Bit D/A Converter	AM6012F (AM6012D-sole source)		XR3464					NS8464		DAC312							AMD AM6012 Harris HI562A
6081	See 5018																	
6456	1GHz Prescaler	SAB6456PN, TD																Siemens SD4211
723	Precision Voltage Regulator	µA723CD, CF, CN µA723F, N SA723CN			µA723	HA17723		MC1723	LM723			RC723 LM723	CA723 LM723	LM723	SG723		µA723	Intersil LM723
733	Differential Video Amp	µA733CF, CN µA733F, N			µA733	HA17733		MC1733	LM733								µA733	Intersil µA733
741	General Purpose Op Amp	µA741CD, CFE, CN µA741FE, N SA741CFE, CN			µA741	HA17741		MC1741	LM741		OP-02			LM741	SG741		µA741	Micropower MPOP-02 Plessey SL562 Samsung LM741

Cross Reference Guide by Numeric Listing (Continued)

NUMERIC	DESCRIPTION	SIGNETICS	ANALOG DEVICES	EXAR	FAIRCHILD	HITACHI	LINEAR TECH	MOTOROLA	NATIONAL	NEC	PMI	RAYTHEON	RCA	SGS/ THOMSON	SILICON GENERAL	SPRAGUE	TI	OTHERS
747	Dual Op Amp	μA747CD, CF, CN μA747F, N SA747CN			μA747	HA17747		MC1747	LM747	μPC1418	OP-04 PM747	RC747	CA747				μA747	Micropower MPOP-04
75188	See 1488																	
75189	See 1489																	
7555	CMOS TIMER	ICM7555CN, CD ICM7555IN, ID ICM7555MN							LMC555								TLC555	Intersil- ICM7555
7820	See 0820																	
8126	See 3526																	
8160	See 5560																	
8161	See 5561																	
8168	See 5568																	
8464	See 6012																	
8564	See 564																	

Signetics

SO Availability List

Linear Products

PART NUMBER	SMD PACKAGE	DESCRIPTION
ADC0820D	SOL-20	8-Bit CMOS A/D
*DAC08ED	SO-16	8-Bit D/A Converter
*LF398D	SO-14	Sample-and-Hold Amp
LM1870D	SOL-20	Stereo Demodulator
LM2901D	SO-14	Quad Volt Comparator
LM2903D	SO-8	Dual Volt Comparator
LM311D	SO-8	Voltage Comparator
LM319D	SO-14	High-Speed Dual Comparator
LM324AD	SO-14	Quad Op Amp
LM324D	SO-14	Quad Op Amp
LM339D	SO-14	Quad Volt Comparator
LM358AD	SO-8	Dual Op Amp
LM358D	SO-8	Dual Op Amp
LM393D	SO-8	Dual Comparator
*MC1408-8D	SO-16	8-Bit D/A Converter
MC1458D	SO-8	Dual Op Amp
MC1488D	SO-14	Quad Line Driver
MC1489D	SO-14	Quad Line Receiver
MC1489AD	SO-14	Quad Line Receiver
MC3302D	SO-14	Quad Volt Comparator
MC3361D	SOL-16	Low Power FM IF
MC3403D	SO-14	Quad Low Power Op Amp
NE4558D	SO-8	Dual Op Amp
*NE5018D	SOL-24	8-Bit D/A Converter
*NE5019D	SOL-24	8-Bit D/A Converter
*NE5036D	SO-14	6-Bit A/D Converter
NE5037D	SO-16	6-Bit A/D Converter
NE5044D	SO-16	Prog 7-Channel Encoder
NE5045D	SO-16	7-Channel Decoder
NE5090D	SOL-16	Address Relay Driver
NE5105/AD	SO-8	High-Speed Comparator
NE5170A	PLCC-28	Octal Line Driver
NE5180A	PLCC-28	Octal Line Receiver
NE5204D	SO-8	High-Frequency Amp
NE5205D	SO-8	High-Frequency Amp
NE521D	SO-14	High-Speed Dual Comparator
NE5212D8	SO-8	Transimedance Amplifier
NE522D	SO-14	High-Speed Dual Comparator
NE5230D	SO-8	Low Voltage Op Amp
NE527D	SO-14	High-Speed Comparator
NE529D	SO-14	High-Speed Comparator

PART NUMBER	SMD PACKAGE	DESCRIPTION
NE532D	SO-8	Dual Op Amp
*NE544D	SOL-16	Servo Amp
*NE5512D	SO-8	Dual Hi-Perf Op Amp
*NE5514D	SOL-16	Quad Hi-Perf Op Amp
NE5517D	SO-16	Dual Hi-Perf Amp
NE5520D	SOL-16	LVDT Signal Cond Ckt
*NE5532D	SOL-16	Dual Low-Noise Op Amp
*NE5533D	SOL-16	Low-Noise Op Amp
NE5534AD	SO-8	Low-Noise Op Amp
NE5534D	SO-8	Low-Noise Op Amp
NE5537D	SO-14	Sample-and-Hold Amp
NE5539D	SO-14	Hi-Freq Amp Wideband
NE555D	SO-8	Single Timer
NE556D	SO-14	Dual Timer
NE5560D	SO-16	SMPS Control Ckt
NE5561D	SO-8	SMPS Control Ckt
NE5562D	SOL-20	SMPS Control Ckt
NE5568D	SO-8	SMPS Control Ckt
NE558D	SOL-16	Quad Timer
NE5592D	SO-14	Dual Video Amp
NE564D	SO-16	Hi-Frequency PLL
*NE565D	SO-14	Phase Locked Loop
NE566D	SO-8	Function Generator
NE567D	SO-8	Tone Decoder PLL
NE568D	SOL-20	PLL
NE571D	SOL-16	Compandor
NE572D	SOL-16	Prog Compandor
*NE587D	SOL-20	7 Seq LED Driver (Anode)
*NE589D	SOL-20	7 Seq LED Driver (Cath)
NE5900D	SOL-16	Call Progress Decoder
NE592D14	SO-14	Video Amp
NE592D8	SO-8	Video Amp
NE592HD14	SO-14	Hi-Gain Video Amp
NE592HD8	SO-8	Hi-Gain Video Amp
*NE594D	SOL-20	Vac Fluor Disp Driver
NE602D	SO-8	Double Bal Mixer/ Oscillator
NE604D	SO-16	Low Power FM IF System
NE605	SOL-20	FM IF System
NE612D	SO-8	Double Balanced Mixer/Oscillator
NE614D	SO-16	Low Power FM IF System
*PCD3311TD	SO-16	DTMF/Melody Generator

SO Availability List

PART NUMBER	SMD PACKAGE	DESCRIPTION
PCD3312TD	SO-8	DTMF/Melody Generator With ICC
PCD3315TD	SOL-28	Repertory Pulse Dial
PCD3360TD	SO-16	Progress Tone Ringer
PCF2100TD	SOL-28	LCD Duplex Driver (40)
PCF2111TD	VSO-40	LCD Duplex Driver (64)
PCF2112TD	VSO-40	LCD Duplex Driver (32)
PCF8570TD	SO-8	Static RAM (256 × 8)
PCF8571TD	SO-8	1K Serial RAM
PCF8573TD	SO-16	Clock/Timer
PCF8574TD	SO-16	Remote I/O Expander
PCF8576TD	VSO-56	MUX/Static Driver
PCF8577TD	VSO-40	32-/64-Segment LCD Driver
SA5105/AD	SO-8	High-Speed Comparator
SA5230D	SO-8	Low Voltage Op Amp
SA5212D8	SO-8	Transimpedance Amp
SA532D	SO-8	Dual Op Amp
SA534D	SO-14	Dual Op Amp
SA555D	SO-8	Single Timer
SA571D	SOL-16	Compandor
SA572D	SOL-16	Compandor
*SA594D	SOL-20	Vac Fluor Disp Driver
SA602D	SO-8	Double Bal Mixer/ Oscillator
SA604D	SO-16	Lower Power FM IF System

PART NUMBER	SMD PACKAGE	DESCRIPTION
SAA3004TD	SOL-20	R/C Transmitter
SG3524D	SO-16	SMPS Control Circuit
TDA1001BTD	SO-16	Noise Suppressor
TDA1005ATD	SO-16	Stereo Decoder
TDA3047TD	SO-16	IR Preamp
TDA3048TD	SO-16	IR Preamp
TDA5040TD	SO-8	Brushless DC Motor Driver
TDA7010TD	SO-16	FM Radio Circuit
TDA7050TD	SO-8	Mono/Stereo Amp
TDD1742TD	SOL-28	Frequency Synthesizer
ULN2003D	SO-16	Transistor Array
ULN2004D	SO-16	Transistor Array
μA723CD	SO-14	Voltage Regulator
μA741CD	SO-8	Single Op Amp
μA747CD	SO-14	Dual Op Amp

NOTE:
*Non-standard pinout.

NOTE:
For information regarding additional SO products released since the publication of this document, contact your local Signetics Sales Office.

Signetics

Ordering Information for Prefixes ADC, AM, AU, CA, DAC, ICM, LF, LM, MC, NE, SA, SE, SG, μA, UC

Linear Products

Signetics' Linear integrated circuit products may be ordered by contacting either the local Signetics sales office, Signetics representatives and/or Signetics authorized distributors. A complete listing is located in the back of this manual.

Minimum Factory Order:

Commercial Product:

$1000 per order
$250 per line item per order

Military Product:

$250 per line item per order

Table 1 provides part number information concerning Signetics originated products.

Table 2 is a cross reference of both the old and new package suffixes for all presently existing types, while Tables 3 and 4 provide appropriate explanations on the various prefixes employed in the part number descriptions.

As noted in Table 3, Signetics defines device operating temperature range by the appropriate prefix. It should be noted, however, that an SE prefix (−55°C to +125°C) indicates only the operating temperature range of a device and *not* its military qualification status. The military qualification status of any Linear product can be determined by either looking in the Military Data Manual and/or contacting your local sales office.

Table 1. Part Number Description

PART NUMBER	CROSS REF PART NO.	PRODUCT FAMILY	PRODUCT DESCRIPTION
N E 5 5 3 7 N	LF398	LIN	Sample-and-Hold Amp

- Sample-and-Hold Amp → Description of Product Function
- LIN → Linear Product Family
- N (suffix) → Package Descriptions — See Table 2
- 5537 → Device Number
- NE → Device Family and Temperature Range Prefix — See Tables 3 & 4

Ordering Information

Table 2. Package Descriptions

OLD	NEW	PACKAGE DESCRIPTION
A, AA	N	14-lead plastic DIP
A	N-14	14-lead plastic DIP (selected analog products only)
B, BA	N	16-lead plastic DIP
	D	Microminiature package (SO)
F	F	14-, 16-, 18-, 22-, and 24-lead ceramic DIP (Cerdip)
I, IK	I	14-, 16-, 18-, 22-, 28-, and 4-lead ceramic DIP
K	H	10-lead TO-100
L	H	10-lead high-profile TO-100 can
NA, NX	N	24-lead plastic DIP
Q, R	Q	10-, 14-, 16-, and 24-lead ceramic flat
T, TA	H	8-lead TO-99
U	U	SIP plastic power
V	N	8-lead plastic DIP
XA	N	18-lead plastic DIP
XC	N	20-lead plastic DIP
XC	N	22-lead plastic DIP
XL, XF	N	28-lead plastic DIP
	A	PLCC
	EC	TO-46 header
	FE	8-lead ceramic DIP

Table 3. Signetics Prefix and Device Temperature

PREFIX	DEVICE TEMPERATURE RANGE
NE	0 to +70°C
SE	−55°C to +125°C
SA	−40°C to +85°C

Table 4. Industry Standard Prefix

PREFIX	DEVICE FAMILY
ADC	Linear Industry Standard
AM	Linear Industry Standard
CA	Linear Industry Standard
DAC	Linear Industry Standard
ICM	Linear Industry Standard
LF	Linear Industry Standard
LM	Linear Industry Standard
MC	Linear Industry Standard
NE	Linear Industry Standard
SA	Linear Industry Standard
SE	Linear Industry Standard
SG	Linear Industry Standard
μA	Linear Industry Standard
UC	Linear Industry Standard

Signetics

Ordering Information for Prefixes HE, OM, PC, PN, SA, TD, TE

Linear Products

Signetics' integrated circuit products may be ordered by contacting either the local Signetics sales office, Signetics representatives and/or Signetics authorized distributors.

Minimum Factory Order:

Commercial Product:

$ 1000 per order

$ 250 per line item per order

Table 1 provides part number information concerning Signetics/Philips integrated circuits.

Table 2 provides package suffixes and descriptions for all presently existing types. Letters following the device number <u>not</u> <u>used in Table 2</u> are considered to be part of the device number.

Table 3 provides explanations on the various prefixes employed in the part number descriptions. As noted in Table 3, Signetics/Philips device operating temperature is defined by the appropriate prefix.

OPERATING TEMPERATURE:

The third letter of the prefix, in a three-letter prefix, is the temperature designator.

The letters A to F give information about the operating temperature:

A: Temperature range not specified. See data sheet.
 e.g. TDA2541N

B: 0 to +70°C
 e.g. PCB8573PN

C: −55°C to +125°C
 e.g. PCC2111PN

D: −25°C to +70°C
 e.g. PCD8571PN

E: −25°C to +85°C
 e.g. PCE2111PN

F: −40°C to +85°C
 e.g. PCF2111PN

Table 1. Part Number Description

PART NUMBER	PRODUCT FAMILY	PRODUCT DESCRIPTION
TDA2541N	LIN	Video IF Amplifier

Video IF Amplifier → Description of Product Function

LIN → Product Family Linear

N → Package Description — See Table 2A

2541 → Device Number

TDA → Device Family and Temperature Range Prefix — See Table 3A

Table 2. Package Description

SUFFIX	PACKAGE DESCRIPTION
PN	8-, 14-, 16-, 18-, 20-, 24-, 28-, 40-lead plastic DIP
TD	Microminiature Package (SO)
DF	14-, 16-, 18-, 22-, 24-lead ceramic DIP
U	Single in-line plastic (SIP) and SIP power packages

Table 3. Device Prefix

PREFIX	DEVICE FAMILY
HEx	CMOS circuit
OM	Linear circuit
PCx	CMOS circuit
PNx	NMOS circuit
SAx	Digital circuit
TDx	Linear circuit
TEx	Linear circuit

Signetics

Section 2
Quality and Reliability

2

Linear Products

INDEX

SIGNETICS' ZERO DEFECTS PROGRAM

In recent years, American industry has demanded increased product quality of its IC suppliers in order to meet growing international competitive pressures. As a result of this quality focus, it is becoming clear that what was once thought to be unattainable — zero defects — is, in fact, achievable.

The IC supplier committed to a standard of zero defects provides a competitive advantage to today's electronics OEM. That advantage can be summed up in four words: *reduced cost of ownership*. As IC customers look beyond purchase price to the total cost of doing business with a vendor, it is apparent that the quality-conscious supplier represents a viable cost reduction resource. Consistently high quality circuits reduce requirements for expensive test equipment and personnel, and allow for smaller inventories, less rework, and fewer field failures.

REDUCING THE COST OF OWNERSHIP THROUGH TOTAL QUALITY PERFORMANCE

Quality involves more than just IC's that work. It also includes cost-saving advantages that come with error-free service — on-time delivery of the right quantity of the right product at the agreed-upon price. Beyond the product, you want to know you can place an order and feel confident that no administrative problems will arise to tie up your time and personnel.

Today, as a result of Signetics' growing appreciation of the concern with cost of ownership, our quality improvement efforts extend out from the traditional areas of product conformance into every administrative function, including order entry, scheduling, delivery, shipping, and invoicing. Driving this process is a Corporate Quality Improvement Team, comprised of the president and his staff, which oversees the activities of 30 other Quality Improvement Teams throughout the company.

LINEAR PRODUCT QUALITY

Signetics has put together a winning process for the manufacturing of Linear Integrated Circuits. The circuits produced by our Linear Division must meet rigid criteria as defined in our design rules and as evaluated through product characterization over the device operating temperature range.

Product conformance to specification is measured throughout the manufacturing cycle. Signetics calls the first submittal to a Product or Quality Assurance gate our Estimated Process Quality or EPQ. It is an internal measure used to drive our Quality Improvement Programs toward our goal of Zero Defects. All product acceptance sampling plans have zero as their acceptance criteria. Only shipments that demonstrate zero defects during these acceptance tests may be shipped to our customers. This is in accordance with our commitment to our Zero Defect policy.

Our standard is Zero Defects and our customers' statistics and awards for outstanding product quality demonstrate our advance toward this goal. Nowhere is this more evident than at our Electrical and Visual-Mechanical Outgoing Product Assurance inspection gates. Over the past eight years, the measured defect level at the first submission to Electrical Product Assurance for Linear products has dropped from over 4000PPM (0.4%) to under 50PPM (0.005%) (See Figure 1a). Similarly our Visual-Mechanical (body defects, lead bend, etc.) defect level has improved remarkably (see Figure 1b). The results from our Quality Improvement Program have allowed Signetics to take the industry leadership position with its Zero Defects Limited Warranty policy. No longer is it necessary to negotiate a mutually acceptable AQL between buyer and Signetics. Signetics will replace any lot in which a customer finds one verified defective part.

QUALITY DATABASE REPORTING SYSTEM — QA05

The capabilities of our manufacturing process are measured and the results are recorded through our corporate-wide QA05 database system. The QA05 system collects the results on all finished lots and feeds this data back to concerned organizations where appropriate corrective actions can be taken. The QA05 reports Estimated Process Quality (EPQ) data which are the sample inspection results for first submittal lots to Quality Assurance inspection for electrical, visual/mechanical, hermeticity, and documentation. Data from this system is available upon request and is distributed routinely to our customers who have formally adopted our Ship-to-Stock program.

CUSTOMER/VENDOR COOPERATION IS AT THE HEART OF ZERO DEFECTS AND REDUCED COSTS

Working to a zero defects standard requires that emphasis be consistently placed, not on "catching" defects, but on preventing them from ever occurring. This strong preventive focus, which demands that quality be "built-in" rather than "inspected in," includes a much greater attention to ongoing communication on quality-related issues. At Signetics, a focus on this cooperative approach has resulted in better service to all customers and the development of two innovative customer/vendor programs: Ship-to-Stock and Self-Qual.

Signetics' Ship-to-Stock Program

Ship-to-Stock is a joint program between Signetics and a customer which formally certifies specific parts to go directly into inventory or to the assembly line from the customer's receiving dock without incoming inspection. This program was developed at the request of several major customers after they had worked with us and had a chance to experience the data exchange and joint corrective action that occurs as part of our quality improvement program.

The key elements of the Ship-to-Stock program are:

- Signetics and customer agree on a list of products to be certified, complete device correlation, and sign a specification.
- The product Estimated Product Quality (EPQ) must be 300ppm or less for the past 3 months.
- Signetics will share Quality (QA05) and Reliability data on a regular basis.
- Signetics will alert Ship-to-Stock customers of any changes in quality or reliability which could adversely impact their product.

Any customer interested in the benefits of the Ship-to-Stock program should contact his local Signetics sales office for a brochure and further details.

As a result of their participation in the Ship-to-Stock Program, many of our customers have eliminated costly incoming testing on selected ICs. We will work together with any customer interested to establish a Ship-to-Stock Program, and identify the products to be included in the program and finalize all neces-

Quality and Reliability

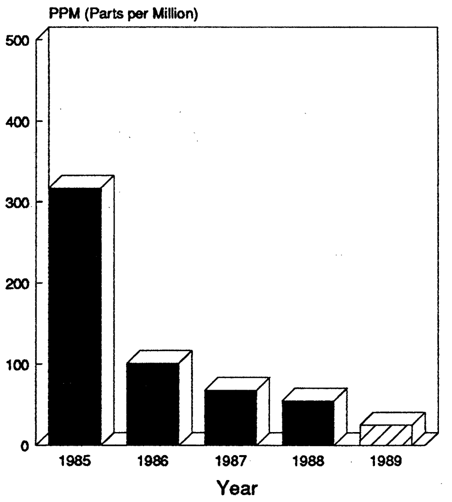


Figure 1a. Product Electrical Quality

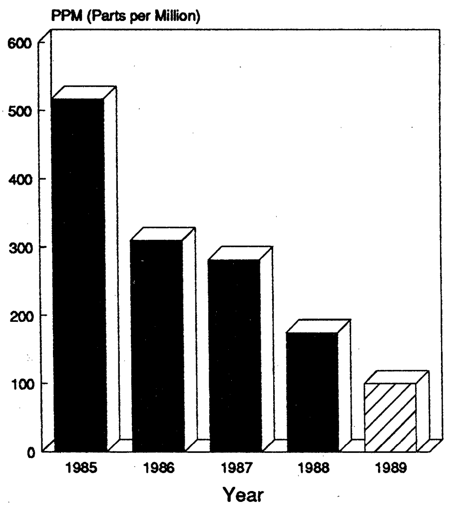


Figure 1b. Visual-Mechanical Quality

sary terms and conditions. From that point, the specified products can go directly from the receiving dock to the assembly line or into inventory. Signetics then provides, free of charge, monthly reports on those products.

In our efforts to continually reduce cost of ownership, we are now using the experience we have gained with Ship-to-Stock to begin developing a Just-in-Time Program. With Just-in-Time, products will be delivered to the receiving dock just as they are needed, permitting continuous-flow manufacturing and eliminating the need for expensive inventories.

Signetics Self-Qual Program

Like Ship-to-Stock, our Self-Qual Program employs a cooperative approach based on ongoing information exchange. At Signetics, formal qualification procedures are required for all new or changed materials, processes, products, and facilities. Prior to 1983, we created our qualification programs independently. Our major customers would then test samples to confirm our findings. Now, under the new Self-Qual Program, customers can be directly involved in the prequalification stage. When we feel we have a promising enhancement to offer, customers will be invited to participate in the development of the qualification plan. This eliminates the need to duplicate expensive qualification testing and also adds another dimension to our ongoing efforts to build in quality.

WE WANT TO WORK WITH YOU

At Signetics, we know that our success depends on our ability to support all our customers with the defect-free, higher density, higher performance products needed to compete effectively in today's demanding business environment. To achieve this goal, quality in another arena — that of communications — is vital. Here are some specific ways we can maintain an ongoing dialogue and information exchange between your company and ours on the quality issue:

- Periodical face-to-face exchanges of data and quality improvement ideas between the customer and Signetics can help prevent problems before they occur.
- Test correlation data is very useful. Line pull information and field failure reports also help us improve product performance.
- When a problem occurs, provide us as soon as possible with whatever specific data you have. This will assist us in taking prompt corrective action.

Quality products are, in large measure, the result of quality communication. By working together, by opening up channels through which we can talk openly to each other, we will insure the creation of the innovative, reliable, cost effective products that help insure a competitive edge.

QUALITY AND RELIABILITY ASSURANCE

Signetics' Linear Division Quality and Reliability Assurance Department is involved in all stages of the production of our Linear ICs:

Quality and Reliability

- Product Design and Process Development
- Wafer Fabrication
- Assembly
- Inspection and Test
- Product Reliability Monitoring
- Customer liaison

The result of this continual involvement at all stages of production enables us to provide feedback to refine present and future designs, manufacturing processes, and test methodology to enhance both the quality and reliability of the products delivered to our customers.

RELIABILITY BEGINS WITH THE DESIGN

Quality and reliability must begin with design. No amount of extra testing or inspection will produce reliable ICs from a design that is inherently unreliable. Signetics follows very strict design and layout practices with its circuits. To eliminate the possibility of metal migration, current density in any path cannot exceed 5×10^5 amps/cm^2. Layout rules are followed to minimize the possibility of shorts, circuit anomalies, and SCR type latch-up effects. All circuit designs are computer-checked using the latest CAD software for adherence to design rules. Simulations are performed for functionality and parametric performance over the full operating ranges of voltage and temperature before going to production. These steps allow us to meet device specifications not only the first time, but also every time thereafter.

PRODUCT CHARACTERIZATION

Before a new design is released, the characterization phase is completed to insure that the distribution of parameters resulting from lot-to-lot variations is well within specified limits. Such extensive characterization data also provides a basis for identifying unique application-related problems which are not part of normal data sheet guarantees.

RELIABILITY MEASUREMENT PROGRAMS

Signetics has developed comprehensive product and process qualification programs to assure that its customers are receiving highly reliable products for their critical applications. Additionally, ongoing reliability monitoring programs, SURE III and Product Monitor, sample standard production product on a regularly established basis (see Table I below).

DESCRIPTION OF STRESSES

SHTL — Static High Temperature Life: SHTL stressing applies static DC bias to the device. This has specific merit in detecting ionic contamination problems which require continuous uninterrupted bias to drive contaminants to the silicon surface. DHTL stressing is not as effective in detecting such problems because the bias continuously changes, intermittently generating and healing the problem.

HTSL — High Temperature Storage Life: This stress exposes the parts to elevated temperatures (150°C – 175°C) with no applied bias.

THBS — Biased Temperature-Humidity, Static: This accelerated temperature and humidity bias stress is performed at 85°C and 85% relative humidity (85°C/85% RH).

TMCL — Temperature Cycling, Air-to-Air: The device is cycled between the specified upper and lower temperature without power in an air or nitrogen environment. Normal temperature extremes are −65°C and +150°C with a minimum 10 minute dwell and 5 minute transition per Mil-STD-883C, Method 1010.5, Condition C. This is a good test to measure the overall package to die mechanical compatibility, because the thermal expansion coefficients of the plastic are normally very much higher than those of the die and leadframe.

PPOT — Pressure Pot: This stress exposes the devices to saturated steam at elevated temperature and pressure. The standard condition is 20 PSIG which occurs at a temperature of 127°C and 100% RH. The stress is used to test the moisture resistance of plastic encapsulated devices. Because the steam environment has an unlimited supply of moisture and ample temperature to catalyze thermally activated events, it is effective at detecting corrosion problems, contamination in-

Table I. RELIABILITY ASSURANCE PROGRAMS

RELIABILITY FUNCTION	TYPICAL STRESS	FREQUENCY
New Process Qualification	High Temperature Operating Life Biased Temperature-Humidity, Static High Temperature Storage Life Pressure Pot Temperature Cycle	Each new wafer fab process
New Product Qualification	High Temperature Operating Life Biased Temperature-Humidity, Static High Temperature Storage Life Pressure Pot Temperature Cycle Electrostatic Discharge Characterization	Each new product
SURE III	High Temperature Operating Life Biased Temperature-Humidity, Static High Temperature Storage Life Pressure Pot Temperature Cycle Thermal Shock	Each fab process family, every four weeks
Product Monitor	Pressure Pot Thermal Shock	Each package type and technology family at each assembly plant, every week

duced leakage problems, and general glassivation stability and integrity.

TMSK — Thermal Shock, Liquid-to-Liquid: Similar to TMCL, however, heating and cooling are done by immersing the units in hot and cold inert liquid. Temperature extremes are −65°C to +150°C with a minimum 5 minute dwell and less than 10 second transition per Mil-STD-883C, Method 1011.4, Condition C. Since heat transfer by conduction is generally much faster than by convection, the liquid-based thermal shock causes more rapid temperature changes in the part.

PRODUCT QUALIFICATION

Linear products are subjected to rigorous qualification procedures for all new products or redesigns to current products. Qualification testing consists of:

- High Temperature Operating Life: T_J = 150°C, 1000 hours, static bias
- High Temperature Storage Life: T_J = 175°C, 1000 hours, unbiased
- Temperature Humidity Biased Life: 85°C, 85% relative humidity, 1000 hours, static bias
- Pressure Cooker: 20 psig, 127°C, 168 hours, unbiased
- Temperature Cycle: −65°C to +150°C, 500 cycles, 10 minute dwell, air to air, unbiased

Formal qualification procedures are required for all new or changed products, processes, and facilities. These procedures ensure the high level of product reliability our customers expect. New facilities are qualified by corporate groups as well as by the quality organizations of specific units that will operate in the facility. After qualification, products manufactured by the new facility are subjected to highly accelerated environmental stresses to ensure that they can meet rigorous failure rate requirements. New or changed processes are similarly qualified.

ONGOING RELIABILITY ASSESSMENT PROGRAMS

The SURE Program

The SURE (Systematic and Uniform Reliability Evaluation) program audits products from each of Signetics Linear Division's process families: Bipolar Junction, Single Layer Metal, Dual Layer Metal, Gold-Doped and Schottky; Oxide Isolated and ACMOS, under a variety of accelerated stress conditions. This program, first introduced in 1964, has evolved to suit changing product complexities and performance requirements.

The Audit Program

Samples are selected from each process family every four weeks and are subjected to each of the following stresses:

- High Temperature Operating Life: T_J = 150°C, 1000 hours, static bias
- Temperature Humidity Biased Life: 85°C, 85% relative humidity, 1000 hours, static bias
- Pressure Cooker: 20 psig, 127°C, 72 hours, unbiased
- Thermal Shock: −65°C to +150°C, 300 cycles, 5 minute dwell, liquid-to-liquid, unbiased
- Temperature Cycling: −65°C to +150°C, 1000 cycles, 10 minute dwell, air-to-air, unbiased

The Product Monitor Program

In addition, each Signetics assembly plant performs Pressure Cooker and Thermal Shock SURE Product Monitor stresses on a weekly basis on each molded package by pin count per the same conditions as the SURE Program.

Product Reliability Reports

The data from these test matrices provides a basic understanding of product capability, an indication of major failure mechanisms, and an estimated failure rate resulting from each stress. This data is compiled periodically and is available to customers upon request. Many customers use this information in lieu of running their own qualification tests, thereby eliminating time-consuming and costly additional testing.

Reliability Engineering

In addition to the product performance monitors encompassed in the Linear SURE program, Signetics' Corporate and Division Reliability Engineering departments sustain a broad range of evaluation and qualification activities.

Included in the engineering process are:

- Evaluation and qualification of new or changed materials, assembly/wafer-fab processes and equipment, product designs, facilities, and subcontractors.
- Device or generic group failure rate studies.
- Advanced environmental stress development.
- Failure mechanism characterization and corrective action/prevention reporting.

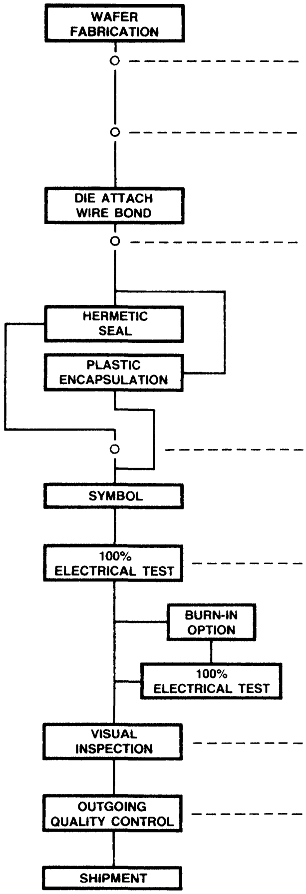
The environmental stresses utilized in the engineering programs are similar to those utilized for the SURE monitor; however, more highly-accelerated conditions and extended durations typify these engineering projects. Additional stress systems such as biased pressure pot, power-temperature cycling, and cycle-biased temperature-humidity, are also included in some evaluation programs.

Failure Analysis

The SURE Program and the Reliability Engineering Program both include failure analysis activities and are complemented by corporate, divisional, and plant failure analysis departments. These engineering units provide a service to our customers who desire detailed failure analysis support, who in turn provide Signetics with the technical understanding of the failure modes and mechanisms actually experienced in service. This information is essential in our ongoing effort to accelerate and improve our understanding of product failure mechanisms and their prevention.

Quality and Reliability

LINEAR DIVISION LINEAR PROCESS FLOW



SCANNING ELECTRON MICROSCOPE CONTROL
Wafers are sampled daily by the Quality Control Laboratory from each fabrication area and subjected to SEM analysis. This process control reveals manufacturing defects such as contact and oxide step coverage in the metalization process which may result in early failures.

DIE SORT VISUAL ACCEPTANCE
Product is inspected for defects caused during fabrication, wafer testing, or the mechanical scribe and break operation. Defects such as scratches, smears and glassivated bonding pads are included in the lot acceptance criteria.

DIE ATTACH AND WIRE BONDING
The latest automated equipment is used under statistical process control program.

PRE-SEAL VISUAL ACCEPTANCE
Product is inspected to detect any damage incurred at the die attach and wire bonding stations. Defects such as scratches, contamination and smeared ball bonds are included in the lot acceptance criteria.

SEAL TESTS
Hermetic package seal integrity is ensured by 100% and fine gross leak testing.

SYMBOL
Devices are marked with the Signetics logo, device number and period date code of assembly or custom symbol per individual specification requirements.

100% PRODUCTION ELECTRICAL TESTING
Every device is tested to all data sheet parameters guaranteeing temperature specifications.

BURN-IN (SUPR II LEVEL B OPTION)
Devices are burned in for 21 hours at 155°C maximum Junction Temperature.

100% PRODUCTION ELECTRICAL TESTING
Every device is tested to all data sheet parameters guaranteeing temperature specifications.

VISUAL
All products are visually inspected per the requirements specified in Signetics' or customer documents.

FINAL QUALITY ASSURANCE GATE
The final QA inspection step guarantees the specified mechanical and electrical AQL's. Every shipment is sealed and identified by QA personnel.

PF00850S

Quality and Reliability

SIGNETICS' MANUFACTURING FACILITIES

Signetics, as part of a multinational corporation, utilizes manufacturing facilities for wafer fabrication, package assembly, and test in three states and three overseas countries as shown in Table II. All wafer fabrication is performed in Signetics operated fabs which report to the Vice President of Die Manufacturing Operations (DMO) in Sunnyvale. Similarly, Signetics Assembly operations in Utah, Korea, and Thailand, report to the Vice President of Assembly Manufacturing Operations (AMO). Assembly subcontractors, Pebei and Anam, are scheduled and controlled through the AMO organization. Assembly subcontractors process all product to Signetics' specifications and materials. Signetics has on-site quality assurance personnel at each subcontractor to audit assembly processes and procedures.

All Signetics Linear products are electrically tested in Signetics operated facilities. These facilities report to the manufacturing organization (DMO or AMO) operating the facility at which they are located.

Table II. Signetics' Linear Product Manufacturing Facilities

WAFER FABRICATION FACILITIES		
Designation	**Location**	**Process Families**
Fab 01	Sunnyvale, California	Bipolar Junction Isolated
Fab 09	Orem, Utah	Bipolar Gold Doped
Fab 16	Sunnyvale, California	Oxide Isolated
Fab 21	Orem, Utah	Bipolar Schottky
Fab 22	Albuquerque, New Mexico	ACMOS
ASSEMBLY FACILITIES		
Designation	**Location**	**Package**
SigKor	Seoul, Korea	DIP, SO, and PLCC
SigThai	Bangkok, Thailand	DIP and CERDIP
Orem	Orem, Utah	Military "Jan" Hermetic
Pebei	Kaohsiung, Taiwan	SO
Anam	Seoul, Korea	SO and Metal Can
TEST FACILITIES		
Designation	**Location**	**Package**
TA03	Sunnyvale, California	Wafer Sort, Final Test and Quality Assurance
SigKor	Seoul, Korea	Final Test and Quality Assurance
SigThai	Bangkok, Thailand	Final Test and Quality Assurance
Sacto	Sacramento, California	Military Final Test and Quality Assurance

Quality and Reliability

SYMBOLIZATION INFORMATION

Signetics' Linear Division products are symboled with the following information on each package:

- Signetics' Logo
- Product Identification and Package Designator
- Traceability Code*
- Assembly Date and Plant Codes*
- Product Revision Level*
- SUPR II B Processing Code (if applicable)

* May appear on the backside of SO 8, 14 & 16 lead packages due to space limitations on topside symbol.

Example:

S NE5534N	line 1
FBW5491	line 2
8901VCB	line 3

Line 1:
S = Signetics' Logo
NE5534 = Product type designation
N = Package type:
N = Dual-in-Line Plastic
F = Dual-in-Line CerDip
D = Small Outline (SO) Surface Mount
A = Plastic Leaded Chip Carrier (PLCC)
E or H = Metal Header

Line 2:
FBW5491 = 7 character Traceability Code assigned to each Assembly Lot which maintains product traceability back to the Wafer Fabrication.
(May be truncated on SO-8 and metal headers.)
Line 3:
8901 = Assembly Date Code (YYWW) specifies the year (YY) (YYWW) and week number (WW) that begins the 4 week assembly period during which the product was manufactured. Thus, 8901 indicates that the product was packaged during the first four weeks of 1989. The first digit of the year may be omitted on some packages: 901.

V = Assembly Plant Code which indicates the assembly facility in which the finished product was packaged.
Assembly Plants Codes are:
V = Signetics Bangkok, Thailand
K = Signetics Seoul, Korea
B = Philips Kaohsiung, Taiwan
L = Anam Seoul, Korea
C = Product Revision Level
B = SUPR II B Burn-in Processing Code (if present)
indicates that the product was processed through 100% SUPR II B Burn-in for 21 hours under biased operation at a junction temperature (Tj) of 155°C

Section 3
Small Area Networks

INDEX

Introduction to I^2C

THE I^2C CONCEPT

The Inter-IC bus (I^2C) is a 2-wire serial bus designed to provide the facilities of a small area network, not only between the circuits of one system, but also between different systems; e.g., teletext and tuning.

Philips/Signetics manufactures many devices with built-in I^2C interface capability, any of which can be connected in a system by simply "clipping" it to the I^2C bus. Hence, any collection of these devices around the I^2C bus is known as "clips."

The I^2C bus consists of two bidirectional lines: the Serial Data (SDA) line and the Serial Clock (SCL) line. The output stages of devices connected to the bus (these devices could be NMOS, CMOS, I^2C, TTL, ...) must have an open-drain or open-collector in order to perform the wired-AND function. Data on the I^2C bus can be transferred at a rate up to 100kbits/sec. The physical bus length is limited to 13 feet and the number of devices connected to the bus is solely dependent on the limiting bus capacitance of 400pF.

The inherent synchronization process, built into the I^2C bus structure using the wired-AND technique, not only allows fast devices to communicate with slower ones, but also eliminates the "Carrier Sense Multiple Access/Collision Detect" (CSMA/CD) effect found in some local area networks, such as Ethernet.

Master-slave relationships exist on the I^2C bus; however, there is no central master. Therefore, a device addressed as a slave during one data transfer could possibly be the master for the next data transfer. Devices are also free to transmit or receive data during a transfer.

To summarize, the I^2C bus eliminates interfacing problems. Since any peripheral device can be added or taken away without affecting any other devices connected to the bus, the I^2C bus enables the system designer to build various configurations using the same basic architecture.

Application areas for the I^2C bus include:

- Video Equipment
- Audio Equipment
- Computer Terminals
- Home Appliances
- Telephony
- Automotive
- Instrumentation
- Industrial Control

Signetics

I^2C Bus Specification

Linear Products

INTRODUCTION

For 8-bit applications, such as those requiring single-chip microcomputers, certain design criteria can be established:

- **A complete system usually consists of at least one microcomputer and other peripheral devices, such as memories and I/O expanders.**
- **The cost of connecting the various devices within the system must be kept to a minimum.**
- **Such a system usually performs a control function and does not require high-speed data transfer.**
- **Overall efficiency depends on the devices chosen and the interconnecting bus structure.**

In order to produce a system to satisfy these criteria, a serial bus structure is needed. Although serial buses don't have the throughput capability of parallel buses, they do require less wiring and fewer connecting pins. However, a bus is not merely an interconnecting wire, it embodies all the formats and procedures for communication within the system.

Devices communicating with each other on a serial bus must have some form of protocol which avoids all possibilities of confusion, data loss and blockage of information. Fast devices must be able to communicate with slow devices. The system must not be dependent on the devices connected to it, otherwise modifications or improvements would be impossible. A procedure has also to be resolved to decide which device will be in control of the bus and when. And if different devices with different clock speeds are connected to the bus, the bus clock source must be defined.

All these criteria are involved in the specification of the I^2C bus.

THE I^2C BUS CONCEPT

Any manufacturing process (NMOS, CMOS, I^2L) can be supported by the I^2C bus. Two wires (SDA – serial data, SCL – serial clock) carry information between the devices connected to the bus. Each device is recognized by a unique address – whether it is a microcomputer, LCD driver, memory or keyboard interface – and can operate as either a transmitter or receiver, depending on the function of the device. Obviously an LCD driver is only a receiver, while a memory can both receive and transmit data. In addition to transmitters and receivers, devices can also be considered as masters or slaves when performing data transfers (see Table 1). A master is the device which initiates a data transfer on the bus and generates the clock signals to permit that transfer. At that time, any device addressed is considered a slave.

The I^2C bus is a multi-master bus. This means that more than one device capable of controlling the bus can be connected to it. As masters are usually microcomputers, let's consider the case of a data transfer between two microcomputers connected to the I^2C bus (Figure 1). This highlights the master-slave and receiver-transmitter relationships to be found on the I^2C bus. It should be noted that these relationships are not permanent, but only depend on the direction of data transfer at that time. The transfer of data would follow in this way:

1) Suppose microcomputer A wants to send information to microcomputer B
   - microcomputer A (master) addresses microcomputer B (slave)
   - microcomputer A (master transmitter) sends data to microcomputer B (slave receiver)
   - microcomputer A terminates the transfer.
2) If microcomputer A wants to receive information from microcomputer B
   - microcomputer A (master) addresses microcomputer B (slave)
   - microcomputer A (master receiver) receives data from microcomputer B (slave transmitter)
   - microcomputer A terminates the transfer.

Even in this case, the master (microcomputer A) generates the timing and terminates the transfer.

The possibility of more than one microcomputer being connected to the I^2C bus means that more than one master could try to initiate a data transfer at the same time. To avoid the chaos that might ensue from such an event, an arbitration procedure has been developed. This procedure relies on the wired-AND connection of all devices to the I^2C bus.

If two or more masters try to put information on to the bus, the first to produce a one when the other produces a zero will lose the arbitration. The clock signals during arbitration are a synchronized combination of the clocks generated by the masters using the wired-AND connection to the SCL line (for more detailed information concerning arbitration see Arbitration and Clock Generation).

Generation of clock signals on the I^2C bus is always the responsibility of master devices; each master generates its own clock signals when transferring data on the bus. Bus clock signals from a master can only be altered when they are stretched by a slow slave

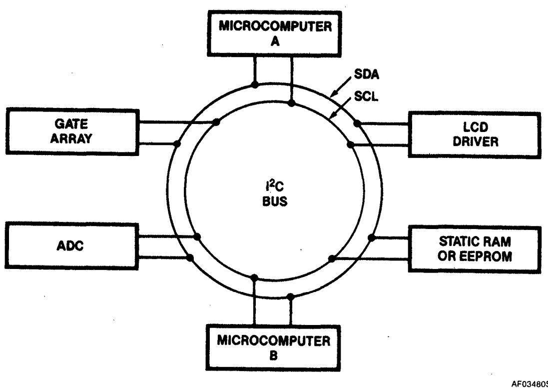


Figure 1. Typical I^2C Bus Configuration

I^2C Bus Specification

Table 1. Definition of I^2C Bus Terminology

TERM	DESCRIPTION
Transmitter	The device which sends data to the bus
Receiver	The device which receives data from the bus
Master	The device which initiates a transfer, generates clock signals and terminates a transfer
Slave	The device addressed by a master
Multi-master	More than one master can attempt to control the bus at the same time without corrupting the message
Arbitration	Procedure to ensure that if more than one master simultaneously tries to control the bus, only one is allowed to do so and the message is not corrupted
Synchronization	Procedure to synchronize the clock signals of two or more devices

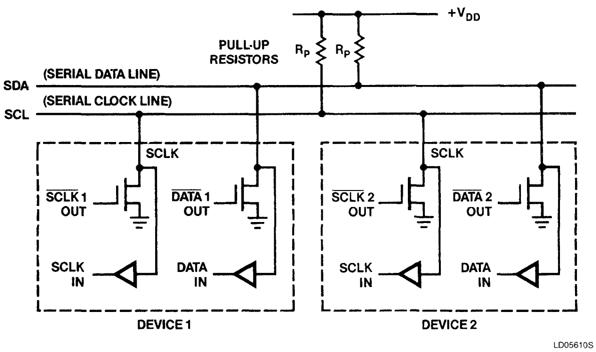


Figure 2. Connection of Devices to the I^2C Bus

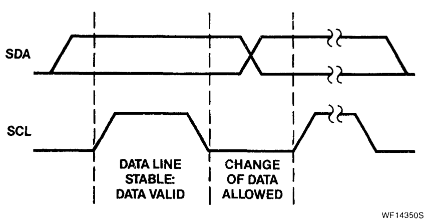


Figure 3. Bit Transfer on the I^2C Bus

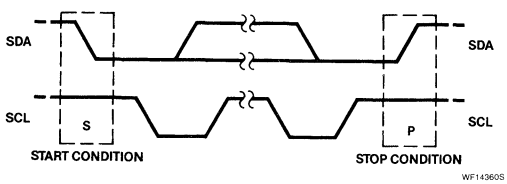


Figure 4. Start and Stop Conditions

device holding down the clock line or by another master when arbitration takes place.

GENERAL CHARACTERISTICS

Both SDA and SCL are bidirectional lines, connected to a positive supply voltage via a pull-up resistor (see Figure 2). When the bus is free, both lines are High. The output stages of devices connected to the bus must have an open-drain or open-collector in order to perform the wired-AND function. Data on the I^2C bus can be transferred at a rate up to 100kbit/s. The number of devices connected to the bus is solely dependent on the limiting bus capacitance of 400pF.

BIT TRANSFER

Due to the variety of different technology devices (CMOS, NMOS, I^2L) which can be connected to the I^2C bus, the levels of the logical 0 (Low) and 1 (High) are not fixed and depend on the appropriate level of V_{DD} (see Electrical Specifications). One clock pulse is generated for each data bit transferred.

Data Validity

The data on the SDA line must be stable during the High period of the clock. The High or Low state of the data line can only change when the clock signal on the SCL line is Low (Figure 3).

Start and Stop Conditions

Within the procedure of the I^2C bus, unique situations arise which are defined as start and stop conditions (see Figure 4).

A High-to-Low transition of the SDA line while SCL is High is one such unique case. This situation indicates a start condition.

A Low-to-High transition of the SDA line while SCL is High defines a stop condition.

Start and stop conditions are always generated by the master. The bus is considered to be busy after the start condition. The bus is considered to be free again a certain time after the stop condition. This bus free situation will be described later in detail.

Detection of start and stop conditions by devices connected to the bus is easy if they possess the necessary interfacing hardware. However, microcomputers with no such interface have to sample the SDA line at least twice per clock period in order to sense the transition.

TRANSFERRING DATA

Byte Format

Every byte put on the SDA line must be 8 bits long. The number of bytes that can be transmitted per transfer is unrestricted. Each byte must be followed by an acknowledge bit.

I²C Bus Specification

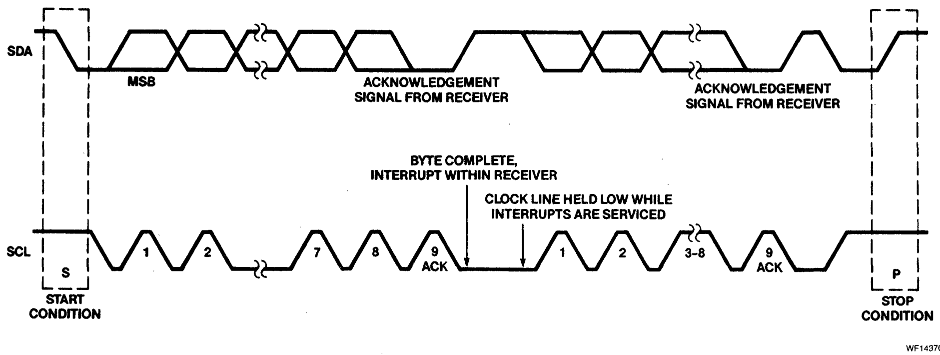


Figure 5. Data Transfer on the I²C Bus

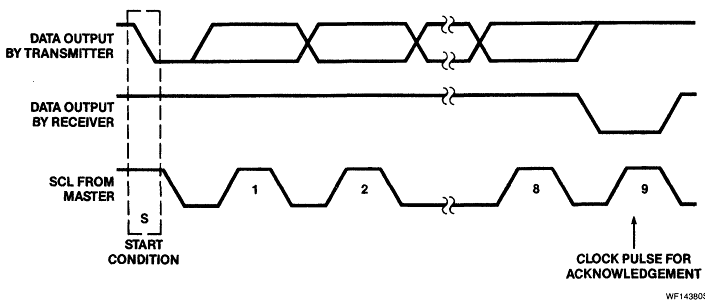


Figure 6. Acknowledge on the I²C Bus

Data is transferred with the most significant bit (MSB) first (Figure 5). If a receiving device cannot receive another complete byte of data until it has performed some other function, for example, to service an internal interrupt, it can hold the clock line SCL Low to force the transmitter into a wait state. Data transfer then continues when the receiver is ready for another byte of data and releases the clock line SCL.

In some cases, it is permitted to use a different format from the I²C bus format, such as CBUS compatible devices. A message which starts with such an address can be terminated by the generation of a stop condition, even during the transmission of a byte. In this case, no acknowledge is generated.

Acknowledge

Data transfer with acknowledge is obligatory. The acknowledge-related clock pulse is generated by the master. The transmitting device releases the SDA line (High) during the acknowledge clock pulse.

The receiving device has to pull down the SDA line during the acknowledge clock pulse so that the SDA line is stable Low during the high period of this clock pulse (Figure 6). Of course, setup and hold times must also be taken into account and these will be described in the Timing section.

Usually, a receiver which has been addressed is obliged to generate an acknowledge after each byte has been received (except when the message starts with a CBUS address.

When a slave receiver does not acknowledge on the slave address, for example, because it is unable to receive while it is performing some real-time function, the data line must be left High by the slave. The master can then generate a STOP condition to abort the transfer.

If a slave receiver does acknowledge the slave address, but some time later in the transfer cannot receive any more data bytes, the master must again abort the transfer. This is indicated by the slave not generating the acknowledge on the first byte following. The slave leaves the data line High and the master generates the STOP condition.

In the case of a master receiver involved in a transfer, it must signal an end of data to the slave transmitter by not generating an acknowledge on the last byte that was clocked out of the slave. The slave transmitter must release the data line to allow the master to generate the STOP condition.

ARBITRATION AND CLOCK GENERATION

Synchronization

All masters generate their own clock on the SCL line to transfer messages on the I²C bus. Data is only valid during the clock High period on the SCL line; therefore, a defined clock is needed if the bit-by-bit arbitration procedure is to take place.

Clock synchronization is performed using the wired-AND connection of devices to the SCL LINE. This means that a High-to-Low transi-

I²C Bus Specification

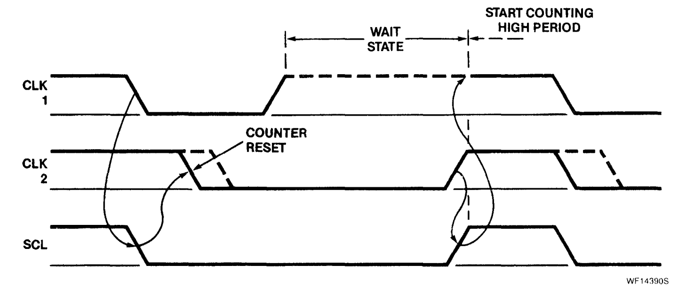


Figure 7. Clock Synchronization During the Arbitration Procedure

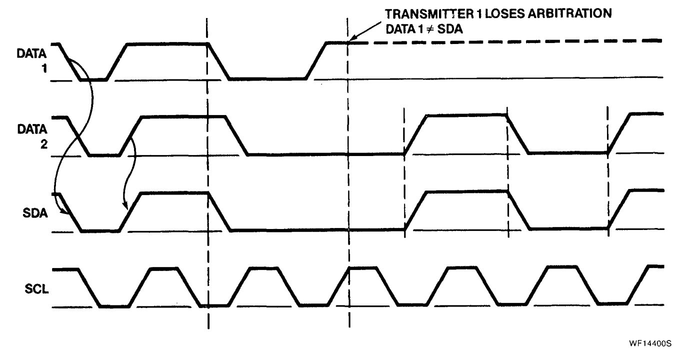


Figure 8. Arbitration Procedure of Two Masters

tion on the SCL line will affect the devices concerned, causing them to start counting off their Low period. Once a device clock has gone Low it will hold the SCL line in that state until the clock High state is reached (Figure 7). However, the Low-to-High change in this device clock may not change the state of the SCL line if another device clock is still within its Low period. Therefore, SCL will be held Low by the device with the longest Low period. Devices with shorter Low periods enter a High wait state during this time.

When all devices concerned have counted off their Low period, the clock line will be released and go High. There will then be no difference between the device clocks and the state of the SCL line and all of them will start counting their High periods. The first device to complete its High period will again pull the SCL line Low.

In this way, a synchronized SCL clock is generated for which the Low period is determined by the device with the longest clock Low period while the High period on SCL is determined by the device with the shortest clock High period.

Arbitration

Arbitration takes place on the SDA line in such a way that the master which transmits a High level, while another master transmits a Low level, will switch off its DATA output stage since the level on the bus does not correspond to its own level.

Arbitration can carry on through many bits. The first stage of arbitration is the comparison of the address bits. If the masters are each trying to address the same device, arbitration continues into a comparison of the data. Because address and data information is used on the I²C bus for the arbitration, no information is lost during this process.

A master which loses the arbitration can generate clock pulses until the end of the byte in which it loses the arbitration.

If a master does lose arbitration during the addressing stage, it is possible that the winning master is trying to address it. Therefore, the losing master must switch over immediately to its slave receiver mode.

Figure 8 shows the arbitration procedure for two masters. Of course more may be involved, depending on how many masters are connected to the bus. The moment there is a difference between the internal data level of the master generating DATA 1 and the actual level on the SDA line, its data output is switched off, which means that a High output level is then connected to the bus. This will not affect the data transfer initiated by the winning master. As control of the I²C bus is decided solely on the address and data sent by competing masters, there is no central master, nor any order of priority on the bus.

Use of the Clock Synchronizing Mechanism as a Handshake

In addition to being used during the arbitration procedure, the clock synchronization mechanism can be used to enable receiving devices to cope with fast data transfers, either on a byte or bit level.

On the byte level, a device may be able to receive bytes of data at a fast rate, but needs more time to store a received byte or prepare another byte to be transmitted. Slave devices can then hold the SCL line Low, after reception and acknowledge of a byte, to force the master into a wait state until the slave is ready for the next byte transfer in a type of handshake procedure.

On the bit level, a device such as a microcomputer without a hardware I²C interface on-chip can slow down the bus clock by extending each clock Low period. In this way, the speed of any master is adapted to the internal operating rate of this device.

I^2C Bus Specification

FORMATS

Data transfers follow the format shown in Figure 9. After the start condition, a slave address is sent. This address is 7 bits long; the eighth bit is a data direction bit (R/$\overline{W}$). A zero indicates a transmission (WRITE); a one indicates a request for data (READ). A data transfer is always terminated by a stop condition generated by the master. However, if a master still wishes to communicate on the bus, it can generate another start condition, and address another slave without first generating a stop condition. Various combinations of read/write formats are then possible within such a transfer.

At the moment of the first acknowledge, the master transmitter becomes a master receiver and the slave receiver becomes a slave transmitter. This acknowledge is still generated by the slave.

The stop condition is generated by the master.

During a change of direction within a transfer, the start condition and the slave address are both repeated, but with the R/$\overline{W}$ bit reversed.

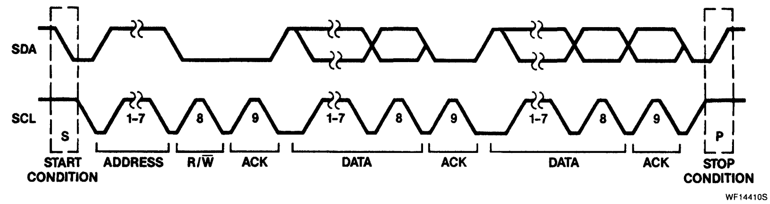
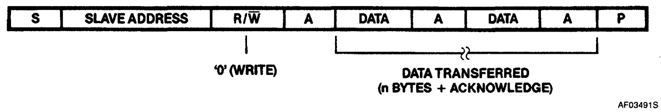


Figure 9. A Complete Data Transfer

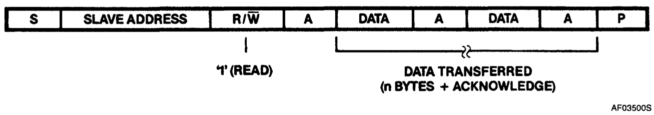
Possible Data Transfer Formats are:

a) Master transmitter transmits to slave receiver. Direction is not changed.

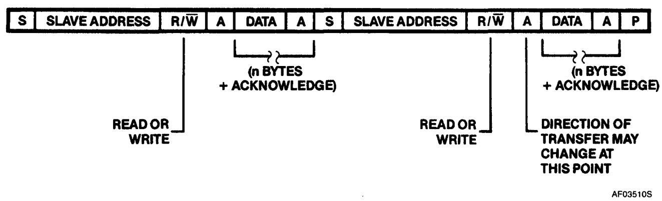
A = ACKNOWLEDGE
S = START
P = STOP



b) Master reads slave immediately after first byte.



c) Combined formats.



NOTES:

1. Combined formats can be used, for example, to control a serial memory. During the first data byte, the internal memory location has to be written. After the start condition is repeated, data can then be transferred.
2. All decisions on auto-increment or decrement of previously accessed memory locations, etc., are taken by the designer of the device.
3. Each byte is followed by an acknowledge as indicated by the A blocks in the sequence.
4. I^2C devices have to reset their bus logic on receipt of a start condition so that they all anticipate the sending of a slave address.

I^2C Bus Specification

ADDRESSING

The first byte after the start condition determines which slave will be selected by the master. Usually, this first byte follows that start procedure. The exception is the general call address which can address all devices. When this address is used, all devices should, in theory, respond with an acknowledge, although devices can be made to ignore this address. The second byte of the general call address then defines the action to be taken.

Definition of Bits in the First Byte

The first seven bits of this byte make up the slave address (Figure 10). The eighth bit (LSB – least significant bit) determines the direction of the message. A zero on the least significant position of the first byte means that the master will write information to a selected slave; a one in this position means that the master will read information from the slave.

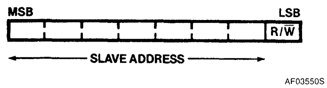


Figure 10. The First Byte After the Start Procedure

When an address is sent, each device in a system compares the first 7 bits after the start condition with its own address. If there is a match, the device will consider itself addressed by the master as a slave receiver or slave transmitter, depending on the R/$\overline{W}$ bit.

The slave address can be made up of a fixed and a programmable part. Since it is expected that identical ICs will be used more than once in a system, the programmable part of the slave address enables the maximum possible number of such devices to be connected to the I^2C bus. The number of programmable address bits of a device depends on the number of pins available. For example, if a device has 4 fixed and 3 programmable address bits, a total of eight identical devices can be connected to the same bus.

The I^2C bus committee is available to coordinate allocation of I^2C addresses.

The bit combination 1111XXX of the slave address is reserved for future extension purposes.

The address 1111111 is reserved as the extension address. This means that the addressing procedure will be continued in the next byte(s). Devices that do not use the extended addressing do not react at the reception of this byte. The seven other possibilities in group 1111 will also only be used for extension purposes but are not yet allocated.

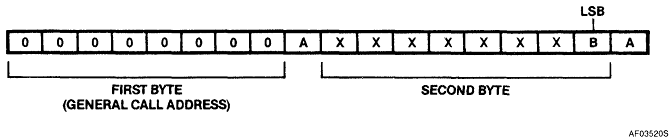


Figure 11. General Call Address Format

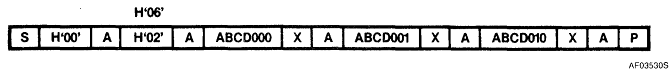


Figure 12. Sequence of a Programming Master

The combination 0000XXX has been defined as a special group. The following addresses have been allocated:

FIRST BYTE			
Slave Address		R/$\overline{W}$	
0000	000	0	General call address
0000	000	1	Start byte
0000	001	X	CBUS address
0000	010	X	Address reserved for different bus format
0000	011	X	To be defined
0000	100	X	To be defined
0000	101	X	To be defined
0000	110	X	To be defined
0000	111	X	To be defined

No device is allowed to acknowledge at the reception of the start byte.

The CBUS address has been reserved to enable the intermixing of CBUS and I^2C devices in one system. I^2C bus devices are not allowed to respond at the reception of this address.

The address reserved for a different bus format is included to enable the mixing of I^2C and other protocols. Only I^2C devices that are able to work with such formats and protocols are allowed to respond to this address.

General Call Address

The general call address should be used to address every device connected to the I^2C bus. However, if a device does not need any of the data supplied within the general call structure, it can ignore this address by not acknowledging. If a device does require data from a general call address, it will acknowledge this address and behave as a slave receiver. The second and following bytes will be acknowledged by every slave receiver capable of handling this data. A slave which cannot process one of these bytes must ignore it by not acknowledging.

The meaning of the general call address is always specified in the second byte (Figure 11).

There are two cases to consider:

1. When the least significant bit B is a zero.
2. When the least significant bit B is a one.

When B is a zero, the second byte has the following definition:

00000110 (H'06') Reset and write the programmable part of slave address by software and hardware. On receiving this two-byte sequence, all devices (designed to respond to the general call address) will reset and take in the programmable part of their address.
Precautions must be taken to ensure that a device is not pulling down the SDA or SCL line after applying the supply voltage, since these low levels would block the bus.

00000010 (H'02') Write slave address by software only. All devices which obtain the programmable part of their address by software (and which have been designed to respond to the general call address) will enter a mode in which they can be programmed. The device will not reset.

I²C Bus Specification

An example of a data transfer of a programming master is shown in Figure 12 (ABCD represents the fixed part of the address).

00000100 (H'04') Write slave address by hardware only. All devices which define the programmable part of their address by hardware (and which respond to the general call address) will latch this programmable part at the reception of this two-byte sequence. The device will not reset.

00000000 (H'00') This code is not allowed to be used as the second byte.

Sequences of programming procedure are published in the appropriate device data sheets.

The remaining codes have not been fixed and devices must ignore these codes.

When B is a one, the two-byte sequence is a hardware general call. This means that the sequence is transmitted by a hardware master device, such as a keyboard scanner, which cannot be programmed to transmit a desired slave address. Since a hardware master does not know in advance to which device the message must be transferred, it can only generate this hardware general call and its own address, thereby identifying itself to the system (Figure 13).

The seven bits remaining in the second byte contain the device address of the hardware master. This address is recognized by an intelligent device, such as a microcomputer, connected to the bus which will then direct the information coming from the hardware master. If the hardware master can also act as a slave, the slave address is identical to the master address.

In some systems an alternative could be that the hardware master transmitter is brought in the slave receiver mode after the system reset. In this way, a system configuring master can tell the hardware master transmitter (which is now in slave receiver mode) to which address data must be sent (Figure 14). After this programming procedure, the hardware master remains in the master transmitter mode.

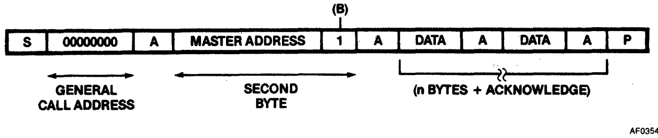
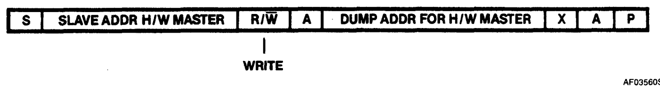
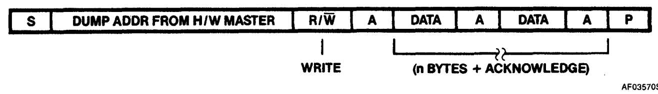


Figure 13. Data Transfer From Hardware Master Transmitter



a. Configuring master sends dump address to hardware master



b. Hardware master dumps data to selected slave device

Figure 14. Data Transfer of Hardware Master Transmitter Capable of Dumping Data Directly to Slave Devices

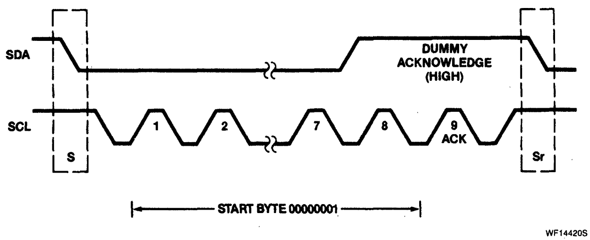


Figure 15. Start Byte Procedure

Start Byte

Microcomputers can be connected to the I²C bus in two ways. If an on-chip hardware I²C bus interface is present, the microcomputer can be programmed to be interrupted only by requests from the bus. When the device possesses no such interface, it must constantly monitor the bus via software. Obviously, the more times the microcomputer monitors, or polls, the bus, the less time it can spend carrying out its intended function.

Therefore, there is a difference in speed between fast hardware devices and the relatively slow microcomputer which relies on software polling.

In this case, data transfer can be preceded by a start procedure which is much longer than normal (Figure 15). The start procedure consists of:

a) A start condition, (S)
b) A start byte 00000001
c) An acknowledge clock pulse
d) A repeated start condition, (Sr)

After the start condition (S) has been transmitted by a master requiring bus access, the start byte (00000001) is transmitted. Another microcomputer can therefore sample the SDA line on a low sampling rate until one of the seven zeros in the start byte is detected. After detection of this Low level on the SDA line, the microcomputer is then able to switch to a higher sampling rate in order to find the second start condition (Sr) which is then used for synchronization.

A hardware receiver will reset at the reception of the second start condition (Sr) and will therefore ignore the start byte.

After the start byte, an acknowledge-related clock pulse is generated. This is present only to conform with the byte handling format used on the bus. No device is allowed to acknowledge the start byte.

I^2C Bus Specification

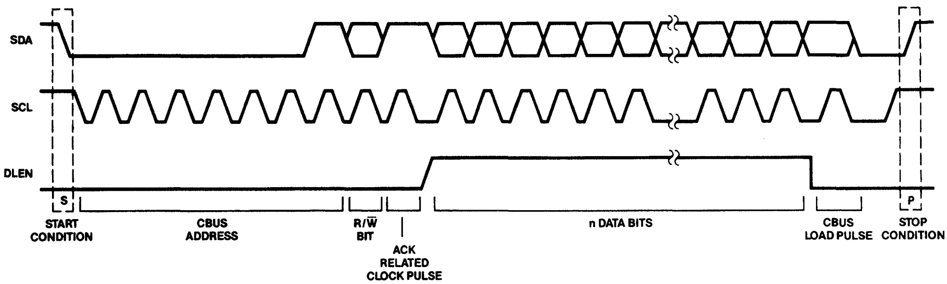


Figure 16. Data Format of Transmissions With CBUS Receiver/Transmitter

CBUS Compatibility

Existing CBUS receivers can be connected to the I^2C bus. In this case, a third line called DLEN has to be connected and the acknowledge bit omitted. Normally, I^2C transmissions are multiples of 8-bit bytes; however, CBUS devices have different formats.

In a mixed bus structure, I^2C devices are not allowed to respond on the CBUS message. For this reason, a special CBUS address (0000001X) has been reserved. No I^2C device will respond to this address. After the transmission of the CBUS address, the DLEN line can be made active and transmission, according to the CBUS format, can be performed (Figure 16).

After the stop condition, all devices are again ready to accept data.

Master transmitters are allowed to generate CBUS formats after having sent the CBUS address. Such a transmission is terminated by a stop condition, recognized by all devices. In the low speed mode, full 8-bit bytes must always be transmitted and the timing of the DLEN signal adapted.

If the CBUS configuration is known and no expansion with CBUS devices is foreseen, the user is allowed to adapt the hold time to the specific requirements of device(s) used.

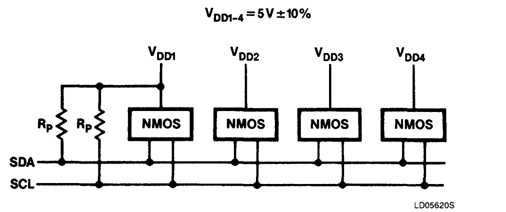


Figure 17. Fixed Input Level Devices Connected to the I^2C Bus

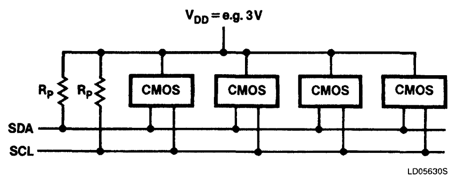


Figure 18. Devices With a Wide Range of Supply Voltages Connected to the I^2C Bus

ELECTRICAL SPECIFICATIONS OF INPUTS AND OUTPUTS OF I^2C DEVICES

The I^2C bus allows communication between devices made in different technologies which might also use different supply voltages.

For devices with fixed input levels, operating on a supply voltage of +5V ± 10%, the following levels have been defined:

V_{ILmax} = 1.5V (maximum input Low voltage)

V_{IHmin} = 3V (minimum input High voltage)

Devices operating on a fixed supply voltage different from +5V (e.g. I^2L), must also have these input levels of 1.5V and 3V for V_{IL} and V_{IH}, respectively.

For devices operating over a wide range of supply voltages (e.g. CMOS), the following levels have been defined:

V_{ILmax} = $0.3V_{DD}$ (maximum input Low voltage)

V_{IHmin} = $0.7V_{DD}$ (minimum input High voltage)

For both groups of devices, the maximum output Low value has been defined:

V_{OLmax} = 0.4V (max. output voltage Low) at 3mA sink current

The maximum low-level input current at V_{OLmax} of both the SDA pin and the SCL pin of an I^2C device is −10μA, including the leakage current of a possible output stage.

The maximum high-level input current at $0.9V_{DD}$ of both the SDA pin and SCL pin of an I^2C device is 10μA, including the leakage current of a possible output stage.

The maximum capacitance of both the SDA pin and the SCL pin of an I^2C device is 10pF.

Devices with fixed input levels can each have their own power supply of +5V ± 10%. Pull-up resistors can be connected to any supply (see Figure 17).

However, the devices with input levels related to V_{DD} must have one common supply line to which the pull-up resistor is also connected (see Figure 18).

I^2C Bus Specification

When devices with fixed input levels are mixed with devices with V_{DD}-related levels, the latter devices have to be connected to one common supply line of +5V ± 10% along with the pull-up resistors (Figure 19).

Input levels are defined in such a way that:

1. The noise margin on the Low level is 0.1 V_{DD}.
2. The noise margin on the High level is 0.2 V_{DD}.
3. Series resistors (R_S) up to 300Ω can be used for flash-over protection against high voltage spikes on the SDA and SCL line (due to flash-over of a TV picture tube, for example) (Figure 20).

The maximum bus capacitance per wire is 400pF. This includes the capacitance of the wire itself and the capacitance of the pins connected to it.

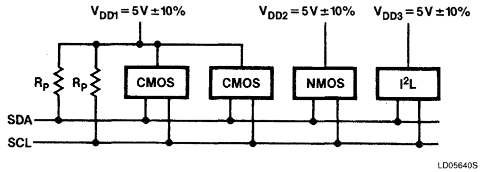


Figure 19. Devices With V_{DD} Related Levels Mixed With Fixed Input Level Devices on the I^2C Bus

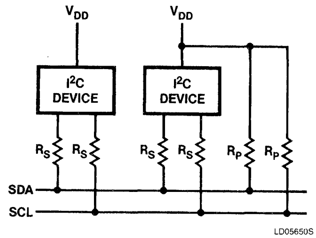


Figure 20. Serial Resistors (R_S) for Protection Against High Voltage

TIMING

The clock on the I^2C bus has a minimum Low period of 4.7μs and a minimum High period of 4μs. Masters in this mode can generate a bus clock with a frequency from 0 to 100kHz.

All devices connected to the bus must be able to follow transfers with frequencies up to 100kHz, either by being able to transmit or receive at that speed or by applying the clock synchronization procedure which will force the master into a wait state and stretch the Low periods. In the latter case the frequency is reduced.

Figure 21 shows the timing requirements in detail. A description of the abbreviations used is shown in Table 2. All timing references are at V_{ILmax} and V_{ILmin}.

LOW-SPEED MODE

As explained previously, there is a difference in speed on the I^2C bus between fast hardware devices and the relatively slow microcomputer which relies on software polling. For this reason a low speed mode is available on the I^2C bus to allow these microcomputers to poll the bus less often.

Start and Stop Conditions

In the low-speed mode, data transfer is preceded by the start procedure.

Data Format and Timing

The bus clock in this mode has a Low period of 130μs ± 25μs and a High period of 390μs ± 25μs, resulting in a clock frequency of approx. 2kHz. The duty cycle of the clock has this Low-to-High ratio to allow for more efficient use of microcomputers without an on-chip hardware I^2C bus interface. In this mode also, data transfer with acknowledge is obligatory. The maximum number of bytes transferred is not limited (Figure 22).

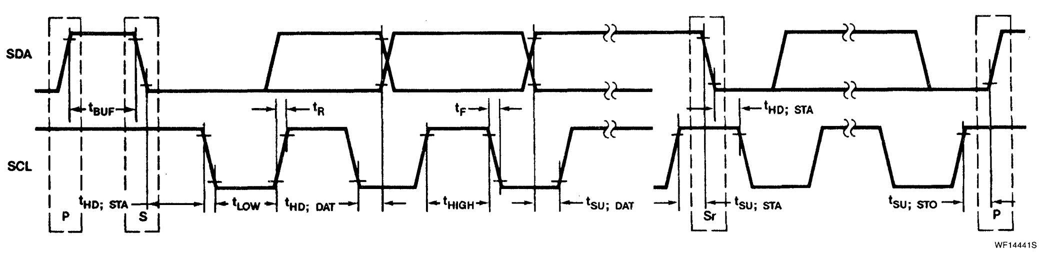


Figure 21. Timing Requirements for the I^2C Bus

I^2C Bus Specification

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Table 2. Timing Requirement for the I^2C Bus

SYMBOL	PARAMETER	LIMITS		UNIT
		Min	Max	
f_{SCL}	SCL clock frequency	0	100	kHz
t_{BUF}	Time the bus must be free before a new transmission can start	4.7		μs
$t_{HD;\,STA}$	Hold time start condition. After this period the first clock pulse is generated	4		μs
t_{LOW}	The Low period of the clock	4.7		μs
t_{HIGH}	The High period of the clock	4		μs
$t_{SU;\,STA}$	Setup time for start condition (Only relevant for a repeated start condition)	4.7		μs
$t_{HD;\,DAT}$	Hold time DATA for CBUS compatible masters for I^2C devices	 5 0*		 μs μs
$t_{SU;\,DAT}$	Setup time DATA	250		ns
t_R	Rise time of both SDA and SCL lines		1	μs
t_F	Fall time of both SDA and SCL lines		300	ns
$t_{SU;\,STO}$	Setup time for stop condition	4.7		μs

NOTES:
All values referenced to V_{IH} and V_{IL} levels.
* Note that a transmitter must internally provide a hold time to bridge the undefined region (300ns max.) of the falling edge of SCL.

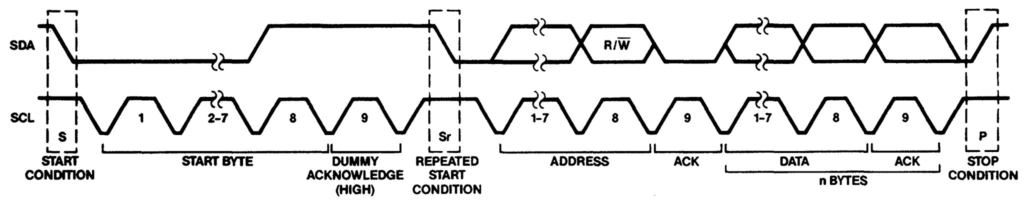


Figure 22. Data Transfer Low-Speed Mode

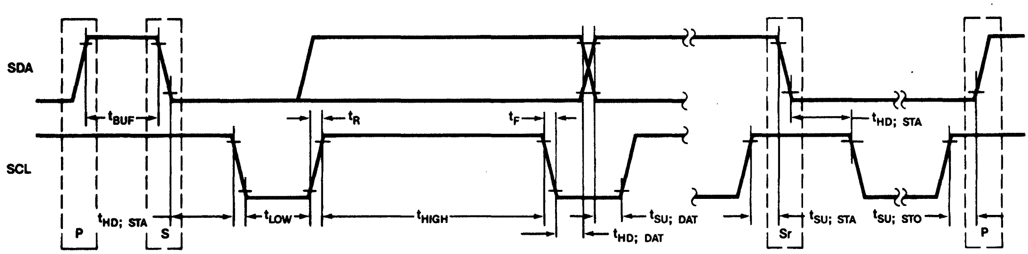


Figure 23. Timing Low-Speed Mode

I^2C Bus Specification

LOW SPEED MODE

CLOCK	: t_{LOW} = 130μs ±25μs
DUTY CYCLE	: t_{HIGH} = 390μs ±25μs
	: 1:3 Low-to-High (Duty cycle of clock generator)
START BYTE	: 0000 0001
MAX. NO. OF BYTES	: UNRESTRICTED
PREMATURE TERMINATION OF TRANSFER	: NOT ALLOWED
ACKNOWLEDGE CLOCK BIT	: ALWAYS PROVIDED
ACKNOWLEDGEMENT OF SLAVES	: OBLIGATORY

In this mode, a transfer cannot be terminated during the transmission of a byte.

The bus is considered busy after the first start condition. It is considered free again one minimum clock Low period, 105μs, after the detection of the stop condition. Figure 23 shows the timing requirements in detail, Table 3 explains the abbreviations.

Table 3. Timing Low Speed Mode

SYMBOL	PARAMETER	LIMITS		UNIT
		Min	Max	
t_{BUF}	Time the bus must be free before a new transmission can start	105		μs
$t_{HD;\ STA}$	Hold time start condition. After this period the first clock pulse is generated	365		μs
$t_{HD;\ STA}$	Hold time (repeated start condition only)	210		μs
t_{LOW}	The Low period of the clock	105	155	μs
t_{HIGH}	The High period of the clock	365	415	μs
$t_{SU;\ STA}$	Setup time for start condition (Only relevant for a repeated start condition)	105	155	μs
t_{HD}; t_{DAT}	Hold time DATA for CBUS compatible masters for I^2C devices	 5 0*		 μs μs
$t_{SU;\ DAT}$	Setup time DATA	250		ns
t_R	Rise time of both SDA and SCL lines		1	μs
t_F	Fall time of both SDA and SCL lines		300	ns
$t_{SU;\ STO}$	Setup time for stop condition	105	155	μs

NOTES:
All values referenced to V_{IH} and V_{IL} levels.
* Note that a transmitter must internally provide a hold time to bridge the undefined region (300ns max.) of the falling edge of SCL.

I^2C Bus Specification

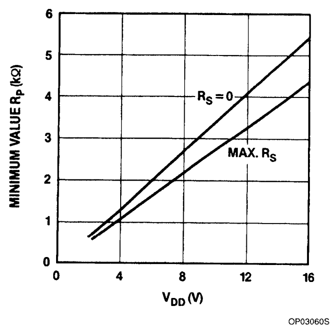
APPENDIX A

Maximum and minimum values of the pull-up resistors R_P and series resistors R_S (See Figure 20).

In a I^2C bus system these values depend on the following parameters:

- Supply voltage
- Bus capacitance
- Number of devices (input current + leakage current)

1) The supply voltage limits the minimum value of the R_P resistor due to the specified 3mA as minimum sink current of the output stages, at 0.4V as maximum low voltage. In Graph 1, V_{DD} against R_{Pmin} is shown.

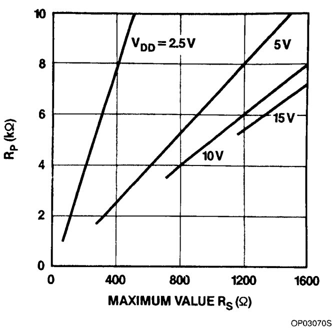


Graph 1

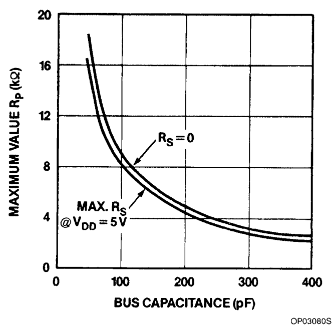
The desired noise margin of 0.1 V_{DD} for the low level limits the maximum value of R_S.

In Graph 2, R_{Smax} against R_P is shown.

2) The bus capacitance is the total capacitance of wire, connections, and pins. This capacitance limits the maximum value of R_P because of the specified rise time of 1μs.



Graph 2

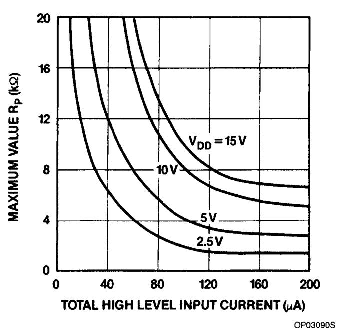


Graph 3

In Graph 3, the bus capacitance – R_{Pmax} relationship is shown.

3) The maximum high-level input current of each input/output connection has a specified value of 10μA max. Due to the desired noise margin of 0.2 V_{DD} for the high level, this input current limits the maximum value of R_P. This limit is dependent on V_{DD}.

In Graph 4 the total high-level input current – R_{Pmax} relationship is shown.



Graph 4

I^2C LICENSE

Purchase of Signetics or Philips I^2C components conveys a license under the Philips I^2C patent rights to use these components in an I^2C system, provided that the system conforms to the I^2C standard specification as defined by Philips.

Signetics

Linear Products

AN168
The Inter-Integrated Circuit (I^2C) Serial Bus: Theory and Practical Consideration

Application Note

Author: Carl Fenger

INTRODUCTION

The I^2C (Inter-IC) bus is becoming a popular concept which implements an innovative serial bus protocol that needs to be understood. On the hardware level I^2C is a collection of microcomputers (MAB8400, PCD3343, 83C351, 84CXX) and peripherals (LCD/LED drivers, RAM, ROM, clock/timer, A/D, D/A, IR transcoder, I/O, DTMF generator, and various tuning circuits) that communicate serially over a two-wire bus, serial data (SDA) and serial clock (SCL). The I^2C structure is optimized for hardware simplicity. Parallel address and data buses inherent in conventional systems are replaced by a serial protocol that transmits both address and bidirectional data over a 2-line bus. This means that interconnecting wires are reduced to a minimum; only V_{CC}, ground and the two-wire bus are required to link the controller(s) with the peripherals or other controllers. This results in reduced chip size, pin count, and interconnections. An I^2C system is therefore smaller, simpler, and cheaper to implement than its parallel counterpart.

The data rate of the I^2C bus makes it suited for systems that do not require high speed. An I^2C controller is well suited for use in systems such as television controllers, telephone sets, appliances, displays or applications involving human interface. Typically an I^2C system might be used in a control function where digitally-controllable elements are adjusted and monitored via a central processor.

The I^2C bus is an innovative hardware interface which provides the software designer the flexibility to create a truly multi-master environment. Built into the serial interface of the controllers are status registers which monitor all possible bus conditions: bus free/busy, bus contention, slave acknowledgement, and bus interference. Thus an I^2C system might include several controllers on the same bus each with the ability to asynchronously communicate with peripherals or each other. This provision also provides expandability for future add-on controllers. (The I^2C system is also ideal for use in environments where the bus is subject to noise. Distorted transmissions are immediately detected by the hardware and the information presented to the software.) A slave acknowledgement on every byte also facilitates data integrity.

An I^2C system can be as simple or sophisticated as the operating environment demands. Whether in a single master or multimaster system, noisy or 'safe', correct system operation can be insured under software control.

CONTROLLERS

Currently the family of I^2C controllers include the MAB8400, and the PCD 3343 (the PCD3343 is basically a CMOS version of the MAB8400). The MAB8400 is based on the 8048 architecture with the I^2C interface built-in. The instruction set for the MAB8400 is similar to the 8048, with a few instructions added and a few deleted. Tables 1 and 2 summarize the differences.

Programs for the MAB8400 and PCD 3343 may be assembled on an 8048-assembler using the macros listed in Appendix A. The serial I/O instructions involve moving data to and from the S0, S1, and S2 serial I/O control registers. The block diagram of the I^2C interface is shown in Figure 1.

SERIAL I/O INTERFACE

A block diagram of the Serial Input/Output (SIO) is shown in Figure 1. The clock line of the serial bus (SCL) has exclusive use of Pin 3, while the Serial Data (SDA) line shares Pin 2 with parallel I/O signal P23 of port 2. Consequently, only three I/O lines are available for port 2 when the I^2C interface is enabled.

Communication between the microcomputer and interface takes place via the internal bus of the microcomputer and the Serial Interrupt Request line. Four registers are used to store data and information controlling the operation of the interface:

- data shift register S0
- address register S0′
- status register S1
- clock control register S2.

THE I^2C BUS INTERFACE: SERIAL CONTROL REGISTERS S0, S1

All serial I^2C transfers occur between the accumulator and register S0. The I^2C hardware takes care of clocking out/in the data, and receiving/generating an acknowledge. In addition, the state of the I^2C bus is controlled and monitored via the bus control register S1. A definition of the registers is as follows:

Data Shift Register S0 — S0 is the data shift register used to perform the conversion between serial and parallel data format. All transmissions or receptions take place through register S0 MSB first. All I^2C bus receptions or transmissions involve moving data to/from the accumulator from/to S0.

Table 1. MAB8400 Family Instructions not in the MAB8048 Instruction Set

SERIAL I/O	REGISTER	CONTROL	CONDITIONAL BRANCH

MOV A,Sn
MOV Sn,A
MOV Sn,#data
EN SI
DIS SI
	DEC @Rr
	DJNZ @Rr,addr
		SEL MB2
		SEL MB3
			JNTF addr

Table 2. MAB8048 Instructions not in the MAB8400 Family Instruction Set

DATA MOVES	FLAGS	BRANCH	CONTROL

MOVX A,@R
MOVX @R,A
MOVP3 A,@A
MOVD A,P
MPVD P,A
ANLD P,A
ORLD P,A
	CLR F0
	CPL F0
	CLR F1
	CPL F1
		*JNI addr
		JF0 addr
		JF1 addr
		
		*replaced by
		JT0, JNT0
			ENTOCLK

3

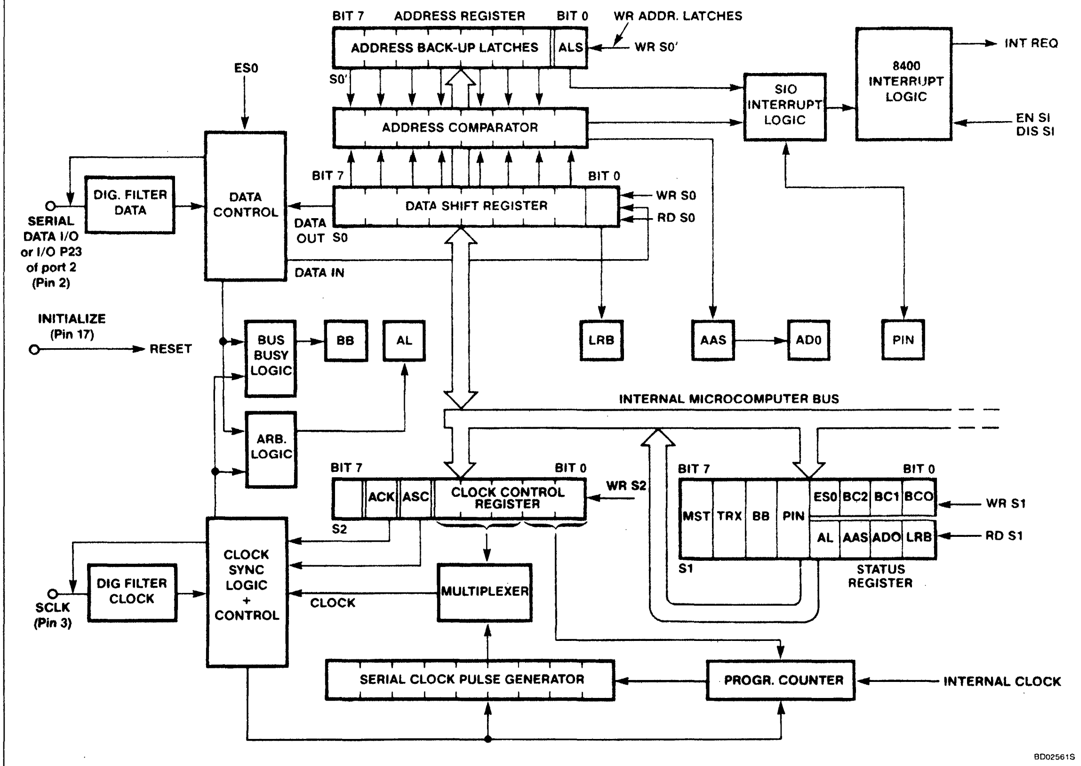


Figure 1. Block Diagram of the MAB8400 SIO Interface

Address Register S0' — In multi-master systems, this register is loaded with a controller's slave address. When activated, (ALS = 0), the hardware will recognize when it is being addressed by setting the AAS (Addressed As Slave) flag. This provision allows a master to be treated as a slave by other masters on the bus.

Status Register S1 — S1 is the bus status register. To control the SIO interface, information is written to the register. The lower 4 bits in S1 serve dual purposes; when written to, the control bits ES0, BC2, BC1, BC0 are programmed (Enable Serial Output and a 3-bit counter which indicates the current number of bits left in a serial transfer). When reading the lower four bits, we obtain the status information AL, AAS, AD0, LRB (Arbitration Lost, Addressed As Slave, Address Zero (the general call has been received), the Last Received Bit (usually the acknowledge bit)). The upper 4 bits are the MST, TRX, BB, and PIN control bits (Master, Transmitter, Bus Busy, and Pending Interrupt Not). These bits define what role the controller has at any particular time. The values of the master and transmitter bits define the controller as either a master or slave (a master initiates a transfer and generates the serial clock; a slave does not), and as a transmitter or receiver. Bus Busy keeps track of whether the bus is free or not, and is set and reset by the 'Start' and 'Stop' conditions which will be defined. Pending Interrupt Not is reset after the completion of a byte transfer + acknowledge, and can be polled to indicate when a serial transfer has been completed. An alternative to polling the PIN bit is to enable the serial interrupt; upon completion of a byte transfer, an interrupt will vector program control to location 07H.

SERIAL CLOCK/ACKNOWLEDGE CONTROL REGISTER S2

Register S2 contains the clock-control register and acknowledge mode bit. Bits S20 – S24 program the bus clock speed. Bit S26 programs the acknowledge or not-acknowledge mode (1/0). The various I^2C bus clock speed possibilities are shown in Table 3.

Table 3. Clock Pulse Frequency Control When Using a 4.43MHz Crystal

HEX S20 – S24 CODE	DIVISOR	APPROX. f_{CLOCK} (kHz)
0	Not Allowed	
1	39	114
2	45	98
3	51	87
4	63	70
5	75	59
6	87	51
7	99	45
8	123	36
9	147	30
A	171	26
B	195	23
C	243	18
D	291	15
E	339	13
F	387	11
10	483	9.2
11	579	7.7
12	675	6.6
13	771	5.8
14	963	4.6
15	1155	3.8
16	1347	3.3
17	1539	2.9
18*	1923	2.3
19*	2307	1.9
1A*	2691	1.7
1B*	3075	1.4
1C	3843	1.2
1D	4611	1.0
1E	5379	0.8
1F	6147	0.7

*only values that may be used in the low speed mode (ASC = 1).

These speeds represent the frequency of the serial clock bursts and do not reflect the speed of the processor's main clock (i.e. it controls the bus speed and has no effect on the CPU's execution speed).

BUS ARBITRATION

Due to the wire-AND configuration of the I^2C bus, and the self-synchronizing clock circuitry of I^2C masters, controllers with varying clock speeds can access the bus without clock contention. During arbitration, the resultant clock on the bus will have a low period equal to the longest of the low periods; the high period will equal the shortest of the high periods. Similarly, when two masters attempt to drive the data line simultaneously, the data is 'ANDed', the master generating a low while the other is driving a high will win arbitration. The resultant bus level will be low, and the loser will withdraw from the bus and set its 'Arbitration Lost' flag (S1 bit 3).

The losing Master is now configured as a slave which could be addressed during this very same cycle. These provisions allow for a number of microcomputers to exist on the same bus. With properly written subroutines, software for any one of the controllers may regard other masters as transparent.

I^2C PROTOCOL AND ASSEMBLY LANGUAGE EXAMPLES

I^2C data transfers follow a well-defined protocol. A transfer always takes place between a master and a slave. Currently a microcomputer can be master or slave, while the 'CLIPS' peripherals are always slaves. In a 'bus-free' condition, both SCL and SDA lines are kept logical high by external pull-up resistors. All bus transfers are bounded by a 'Start' and a 'Stop' condition. A 'Start' condition is defined as the SDA line making a high-to-low transition **while the SCL line is high**. At this point, the internal hardware on all slaves are activated and are prepared to clock-in the next 8 bits and interpret it as a 7-bit address and a $R/\overline{W}$ control bit (MSB first). All slaves have an internal address (most have 2 – 3 programmable address bits) which is then compared with the received address. The slave that recognized its address will respond by pulling the data line low during a ninth clock generated by the master (all I^2C byte transfers require the master to generate 8 clock pulses plus a ninth acknowledge-related clock pulse). The slave-acknowledge will be registered by the master as a '0' appearing in the LRB (Last Received Bit) position of the S1 serial I/O status register. If this bit is high after a transfer attempt, this indicates that a slave did not acknowledge, and that the transfer should be repeated.

After the desired slave has acknowledged its address, it is ready to either send or receive data in response to the master's driving clock. All other slaves have withdrawn from the bus. In addition, for multi-master systems, the start condition has set the 'Bus Busy' bit of the serial I/O register S1 on all masters on the bus. This gives a software indication to other masters that the bus is in use and to wait until the bus is free before attempting an access.

There are two types of I^2C peripherals that now must be defined: there are those with only a chip address such as the I/O expander, PCF8574, and those with a chip address plus an internal address such as the static RAM, PCF8570. Thus after sending a start condition, address, and $R/\overline{W}$ bit, we must take into account what type of slave is being addressed. In the case of a slave with only a chip address, we have already indicated its address and data direction ($R/\overline{W}$) and are therefore ready to send or receive data. This is performed by the master generating bursts of 9 clock pulses for each byte that is sent or received. The transaction for writing one byte to a slave with a chip address only is shown in Figure 3.

In this transfer, all bus activity is invoked by writing the appropriate control byte to the serial I/O control register S1, and by moving data to/from the serial bus buffer register S0. Coming from a known state (MOV S1,#18H-Slave, Receiver, Bus not Busy) we first load the serial I/O buffer S0 with the desired

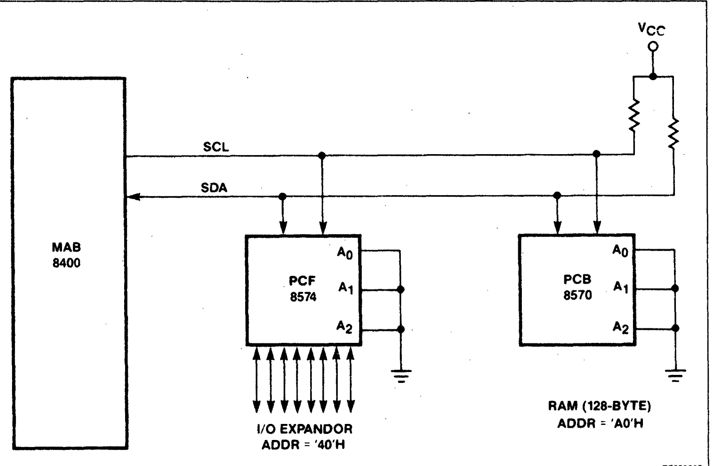
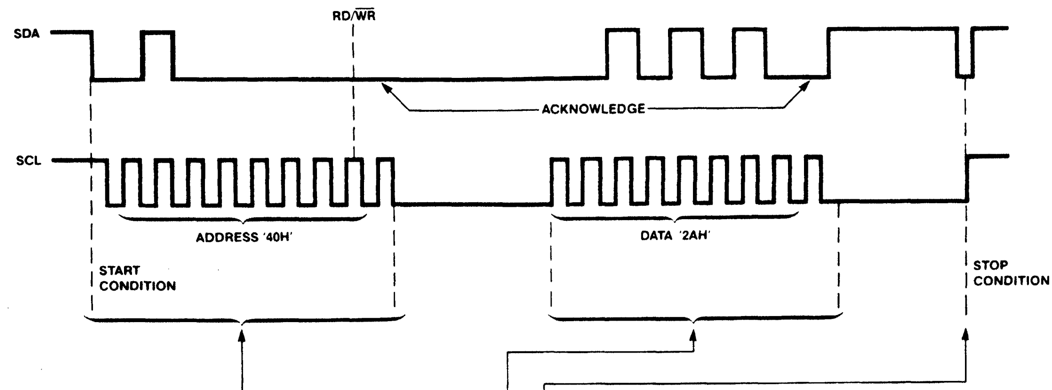


Figure 2. Schematic for Assembly Examples




```
MOV S1,#18H        ;Initialize S1-Slave, Receiver, Bus not
                   ;Busy, Enable Serial I/O.
MOV S0,#40H        ;Preload S0 with Slave's address &
                   ;R/W bit.
MOV S1,#0F8H       ;Invoke start condition & slave address
                   ;(Master, Transmitter, Bus Busy, Enable
                   ;Serial I/O, Bit Counter = 000).
CALL ACKWT:        ;Check for transmission complete, ack.
                   ;received, no arbitration, etc.
MOV A,#2AH         ;Get a data byte.
MOV S0,A           ;Transmit data byte.
CALL ACKWT:        ;Wait for transmission complete again.
MOV S1,#0D8H       ;Generate Stop condition
                   ;(Master, Transmitter, Bus not Busy).
```

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Figure 3

slave's address (MOV S0,#40H). To transmit this preceded by a start condition, we must first examine the control register S1, which, after initialization, looks like this:

MASTER	TRANS	BUS BUSY	PIN	ESO	BC2	BC1	BC0
0	0	0	1	1	0	0	0

To transmit to a slave, the Master, Transmitter, Bus Busy, PIN (Pending Interrupt Not), and ESO (Enable Serial Output) must be set to a 1. This results in an 'F8H' being written to S1. This word defines the controller as a Master Transmitter, invokes the transfer by setting the 'Bus Busy' bit, clears the Pending Interrupt Not (an inverted flag indicating the completion of a complete byte transfer), and activates the serial output logic by setting the Enable Serial Output (ESO) bit.

BIT COUNTER S12, S11, S10

BC2, BC1, and BC0 comprise a bit-counter which indicates to the logic how long the word is to be clocked out over the serial data line. By setting this to a 000H, we are telling it to produce 9 clocks (8 bits plus an acknowledge clock) for this transfer. The bit counter will then count off each bit as it is transmitted. The bit counter possibilities are shown in Table 4.

Thus the bit counter keeps track of the number of clock pulses remaining in a serial transfer. Additionally, there is a not-acknowledge mode (controlled through bit 6 of clock control register S2) which inhibits the acknowledge clock pulse, allowing the possibility of straight serial transfer. We may thus define the word size for a serial transfer (by preloading BC2, BC1, BC0 with the appropriate control number), with or without an acknowledge-related clock pulse being generated. This makes the controller able to transmit serial data to most any serial device regardless of its protocol (e.g., C-bus devices).

CHECKING FOR SLAVE ACKNOWLEDGE

After a 'Start' condition and address have been issued, the selected slave will have recognized and acknowledged its address by

Table 4. Binary Numbers in Bit-Count Locations BC2, BC1 and BC0

BC2	BC1	BC0	BITS/BYTE WITHOUT ACK	BITS/BYTE WITH ACK
0	0	1	1	2
0	1	0	2	3
0	1	1	3	4
1	0	0	4	5
1	0	1	5	6
1	1	0	6	7
1	1	1	7	8
0	0	0	8	9

pulling the data line low during the ninth clock pulse. During this period, the software (which runs on the processor's 4MHz clock) will have been either waiting for the transfer to be completed by polling the PIN bit in S1 which goes low on completion of a transfer/reception (whose length is defined by the pre-loaded Bit-counter value), or by the hardware in Serial Interrupt mode. The serial interrupt (vectored to 07H) is enabled via the EN SI (enable serial interrupt) instruction.

At the point when PIN goes low (or the serial interrupt is received) the 9-bit transfer has been completed. The acknowledgement bit will now be in the LRB position of register S1, and may be checked in the routine 'ACKWT' (Wait for Acknowledge) as shown in Figure 4.

This routing must go one step further in multi-master systems; the possibility of an Arbitration Lost situation may occur if other masters are present on the bus. This condition may be detected by checking the 'AL' bit (bit 3). If arbitration has been lost, provisions for re-attempting the transmission should be taken. If arbitration is lost, there is the possibility that the controller is being addressed as a Slave. If this condition is to be recognized, we must test on the 'AAS' bit (bit 2). A 'General Call' address (00H) has also been defined as an 'all-call' address for all slaves; bit 1, AD0, must be tested if this feature is to be recognized by a Master.

After a successful address transfer/acknowledge, the slave is ready to be sent its data. The instruction MOV SO,A will now automatically send the contents of the accumulator out on the bus. After calling the ACKWT routine once more, we are ready to terminate the transfer. The Stop condition is created by the instruction 'MOV S1, #0D8H'. This resets the bus-busy bit, which tells the hardware to generate a Stop — the data line makes a low-to-high transition while the clock remains high. All bus-busy flags on other masters on the bus are reset by this signal.

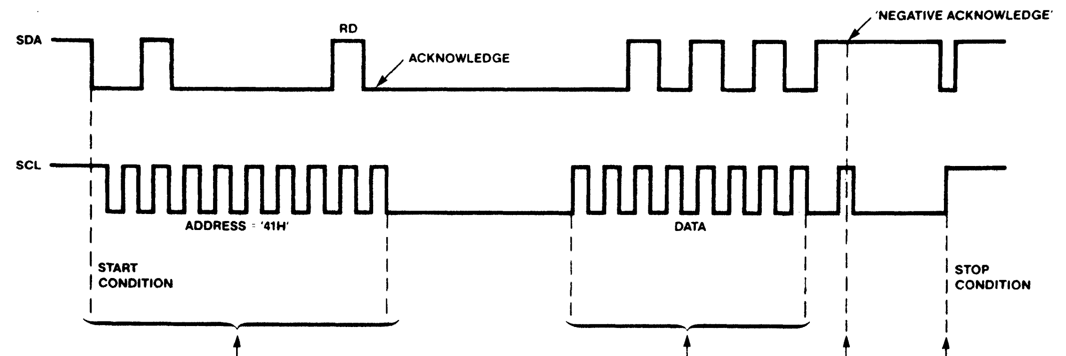
The transfer is now complete — PCF8574 I/O Expandor will transfer the serial data stream to its 8 output pins and latch them until further update.

```
ACKWT:   MOV A,S1        ;Get bus status word
                         ;from S1.
         JB4 ACKWT       ;Poll the PIN bit
                         ;until it goes low
                         ;indicating transfer
                         ;completed
         JBO BUSERR      ;Jump to BUSERR
                         ;routine if acknowledge
                         ;not received.
         RET             ;transfer complete,
                         ;acknowledge received - return.
```

Figure 4

MASTER READS ONE BYTE FROM SLAVE

A read operation is a similar process; the address, however, will be 41H, the LSB indicating to the I/O device that a read is to be performed. During the data portion of a read, the I/O port 8574 will transmit the contents of its latches in response to the clock generated by the master. The Master/Receiver in this case generates a low-level acknowledge on reception of each byte (a 'positive' acknowledge). Upon completion of a read, the master must generate a 'negative' acknowledge during the ninth clock to indicate to the slaves that the read operation is finished. This is necessary because an arbitrary number of bytes may be read within the same transfer. A negative acknowledge consists of a high signal on the data line during the ninth clock of the last byte to be read. To accomplish this, the master 8400 must leave the acknowledge mode just before the final byte, read the final byte (producing only 8 clock pulses), program the bit-counter with 001 (preparing for a one-bit negative acknowledge pulse), and simply move the contents of S0 to the accumulator. This final instruction accomplishes two things simultaneously: it transfers the final byte to the accumulator and produces one clock pulse on the SCL line. The structure of the serial I/O register S0 is such that a read from it causes a double-buffered transfer from the I^2C bus to S0, while the original contents of S0 are transferred to the accumulator. Because the number of clocks produced on the bus is determined by the control number in the Bit Counter, by presetting it to 001, only one clock is generated. At this point in time the slave is still waiting for an acknowledge; the bus is high due to the pull-up, as single clock pulse in this condition is interpreted as a 'negative' acknowledge. The slave has now been informed that reading is completed; a Stop condition is now generated as before. The read process (one byte from a slave with only a chip address) is shown in Figure 5.




```
        MOV S1,#18H          ;Initialize serial I/O control
                             ;register.
        MOV S0,#41H          ;Preload serial register S0
                             ;with slave address and RD
                             ;control bit.
        MOV S1,#0F8H         ;Send address to bus along with
                             ;start condition.
        CALL ACKWT           :Wait for acknowledge (as
                             ;before).
        MOV S2,#01H          ;Leave acknowledge mode.
        MOV S0,A             ;Read data from slave to S0.
WAIT:   MOV A,S1             ;Test for byte received by
                             ;testing S1 PIN bit.
        JB4 Wait             ;Wait until PIN received.
        MOV S1,#0A9H         ;Set Bit Counter to 1 and
                             ;become a receiver (A9 =
                             ;Mst,Rec,Bus Busy,Bit Coutner =
                             ;001).
        MOV A,S0             ;Move data to accumulator and
                             ;clock out a negative
        MOV S1,#0D8H         ;acknowledge.
                             ;Generate Stop Condition.
```

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Figure 5

The Inter-Integrated Circuit (I^2C) Serial Bus: Theory and Practical Consideration

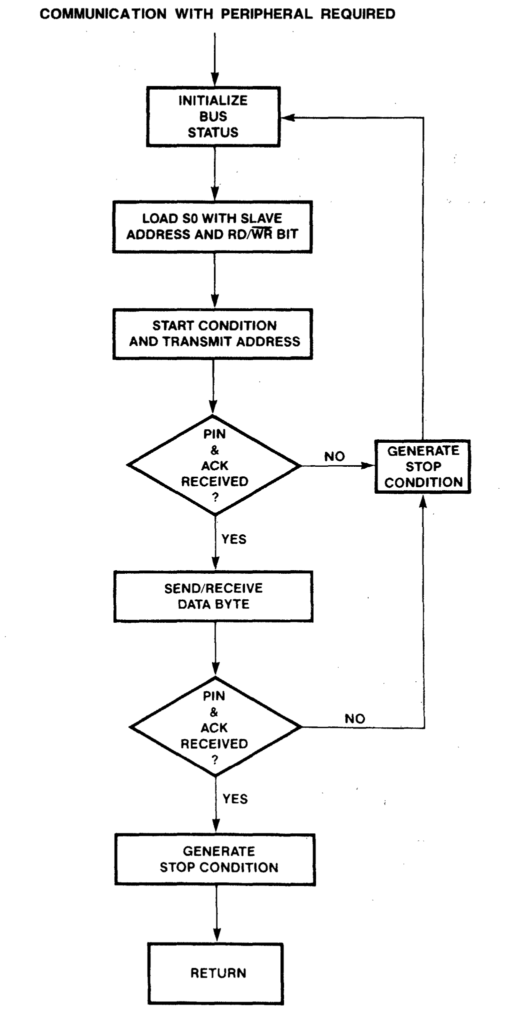


Figure 6. Flowchart for Reading/Writing One Byte to an I^2C Peripheral; Single-Master, Single-Address Slave

```
        MOV S1, #18H      ;Initialize bus-status register
                          ;Master, Transmitter,
                          ;Bus-not-Busy, Enable SIO.
        MOV S0, #0A0H     ;Load S0 with RAM's chip
                          ;address.
        MOV S1, #0F8H     ;Start cond. and transmit
                          ;address.
        CALL ACKWT        ;Wait until address received.
        MOV A,#00H        ;Set up for transmitting RAM
                          ;location address.
        MOV S0,A          ;Transmit first RAM address.
        CALL ACKWT        ;Wait.
        MOV S1, #18H      ;Set up for a repeated Start
                          ;condition.
        MOV A,#0A1H       ;Get RAM chip address & RD bit.
        MOV S0,A          ;Send out to bus
        MOV S1,#0F8H      ;preceded by repeated Start.

        CALL ACKWT        ;Wait.
        MOV A,S0          ;First data byte to S0.
        CALL ACKWT        ;Wait.
        MOV A,S0          ;Second data byte to S0.
                          ;And First data byte to Acc.
        CALL ACKWT        ;Wait.
        MOV R0,A          ;Save first byte in R0.
        MOV A,S0          ;Third data byte to S0
                          ;and second data byte to Acc.
        CALL ACKWT        ;Wait.
        MOV R1,A          ;Save second data byte
                          ;in R1.
        MOV S2,#01H       ;Leave ack. mode.
                          ;Bit Counter=001 for neg ack.
        MOV A,S0          ;Third data byte to acc
                          ;negative ack. generated.
        MOV R2,A          ;Save third data byte in R2.
WAIT1:  MOV A,S1          ;Get bus status.
        JB4 WAIT1         ;Wait until transfer complete.
        MOV S1,#0D8H      ;Stop condition.
        MOV S2,#41H       ;Restore acknowledge mode.
```

DF05700S

Figure 7

These examples apply to a slave with a chip address — more than one byte can be written/read within the same transfer; however, this option is more applicable to I^2C devices with sub-addresses such as the static RAMs or Clock/Calendar. In the case of these types of devices, a slightly different protocol is used. The RAM, for example, requires a chip address and an internal memory location before it can deliver or accept a byte of information. During a write operation, this is done by simply writing the secondary address right after the chip address — the peripheral is designed to interpret the second byte as an internal address. In the case of a Read operation, the slave peripheral must send data back to the Master after it has been addressed and sub-addressed. To accomplish this, first the Start, Address, and Sub-address is transmitted. Then we have a **repeated** start condition to reverse the direction of the data transfer, followed by the chip address and RD, then a data string (w/ acknowledges). This repeated Start does not affect other peripherals — they have been deactivated and will not reactivate until a Stop condition is detected. I^2C peripherals are equipped with auto-incrementing logic which will automatically transmit or receive data in consecutive (increasing) locations. For example, to read 3 consecutive bytes to PCB8571 RAM locations 00, 01 and 02, we use the following format as shown in Figure 7.

This routine reads the contents of location 00, 01 and 02 of the PCB8571 128-byte RAM and puts them in registers R0, R1, and R2. The auto-incrementing feature allows the programmer to indicate only a starting location, then read an arbitrary block of consecutive memory addresses. The WAIT 1 loop is required to poll for the completion of the final byte because the ACKWT routine will not recognize the negative acknowledge as a valid condition.

BUS ERROR CONDITIONS: ACKNOWLEDGE NOT RECEIVED

In the above routines, should a slave fail to acknowledge, the condition is detected during the 'ACKWT' routine. The occurrence may indicate one of two conditions: the slave has failed to operate, or a bus disturbance has occurred. The software response to either event is dependent on the system application. In either case, the 'BusErr' routine should reinitialize the bus by issuing a 'Stop' condition. Provision may then be taken to repeat the transfer an arbitrary number of times. Should the symptom persist, either an error condition will be entered, or a backup device can be activated.

These sample routines represent single-master systems. A more detailed analysis of multi-master/noisy environment systems will be treated in further application notes. Examples of more complex systems can be found in the 'Software Examples' manual; publication 9398 615 70011.

APPENDIX A

Only the 8048 assembler is capable of assembling MAB8400 source code when it has at least a "DATA" or "Define Byte" assembler directive, possibly in combination with a MACRO facility.

The new instructions can be simply defined by MACROs. The instructions which are not in the MAB8400 should not be in the MAB8400 source program.

An example of a macro definitions list is given here for the Intel Macro Assembler.

This list can be copied in front of a MAB8400 source program; the new instructions are added to the MAB8400 source program by calling the MACRO via its name in the opcode field and (if required) followed by an operand in the operand field.

MACRO DEFINITIONS

LINE SOURCE STATEMENT

```
$MACROFILE
;MACROS FOR 8048 ASSEMBLER RECOGNITION
;OF 8400 COMMANDS
                MOVSOA      MACRO       ;MOV SO,A
                DB 3CH
                ENDM
                MOVASO      MACRO       ;MOV A,SO
                DB 0CH
                ENDM
                MOVS1A      MACRO       ;MOV S1,A
                DB 3DH
                ENDM
                MOVAS1      MACRO       ;MOV A,S1
                DB 0DH
                ENDM
                MOVS2A      MACRO       ;MOV S2,A
                DB 3EH
                ENDM
                MOVSO       MACRO L     ;MOV SO,#DATA
                DB 9CH,L
                ENDM
                MOVS1       MACRO L     ;MOV S1,#DATA
                DB 9DH,L
                ENDM
                MOVS2       MACRO L     ;MOV S2,#DATA
                DB 9EH,L
                ENDM
                ENSI        MACRO       ;EN SI
                DB 85H
                ENDM
                DISSI       MACRO       ;DIS SI (Disable serial
                                         interrupt)
                DB          95H
                ENDM
;
; PORT 0 INSTRUCTIONS:
;               INAPO       MACRO       ;IN A,PO
                DB          08H
                ENDM
;
                OUTPOA      MACRO       ;OUTL PO,A
                DB          38H
                ENDM
;
                ORLPO       MACRO L     ;ORL PO,#DATA
                DB          88H,L
                ENDM
;
                ANLPO       MACRO L     ;ANL PO,#DATA
                DB          98H,L
                ENDM
;
```

MACRO DEFINITIONS (Continued)

LINE SOURCE STATEMENT

```
; DATA MEMORY INSTRUCTIONS:
        DECARO      MACRO           ;DEC @R0
        DB          0C0H
        ENDM
;
        DECAR1      MACRO           ;DEC @R1
        DB          OC1H
        ENDM
;
; SELECT MEMORY BANK INSTRUCTIONS:
        SELMB2      MACRO           ;SEL MB2
        DB          0A5H
        ENDM
;
        SELMB3      MACRO           ;SEL MB3
        DB          0B5H
        ENDM
;
; CONDITIONAL JUMP INSTRUCTIONS:
        DJNZA0      MACRO L         ;DJNZ @R0,ADDR
        DB          0E0H,L AND 0FFH
        ENDM
;
        DJNZA1      MACRO L         ;DJNZ @R1,ADDR
        DB          0E1H,L AND 0FFH
        ENDM
;
        JNTF        MACRO L         ;JUMP IF TIMERFLAG IS
                                     NON ZERO
        DB          06H,L AND 0FFH
        ENDM

; END OF MACRO DEFINITIONS
```

THE 8400 INSTRUCTIONS BUILT FROM THE MACRO LIST

LOC/OBJ	LINE	SOURCE STATEMENT		
0000	1	ORG 0		
	2	MOVAS0		;MACRO for MOV A,S0
0000 0C	3 +	DB	0CH	
	4	MOVAS1		;MACRO for MOV A,S1
0001 0D	5 +	DB	0DH	
	6	MOVS0A		;MACRO for MOV S0,A
0002 3C	7 +	DB	3CH	
	8	MOVS1A		;MACRO For MOV S1,A
0003 3D	9 +	DB	3DH	
	10	MOVS2A		;MACRO For MOV S2,A
0004 3E	11 +	DB	3EH	
	12	MOVS0	56H	;MACRO For MOV S0, #56H
0005 9C	13 +	DB	9CH,56H	
0006 56				
	14	MOVS1	9FH	;MACRO for MOV S1, #9FH
0007 9D	15 +	DB	9DH,9FH	
0008 9F				
	16	MOVS2	0E8H	;MACRO for MOV S2, #0E8H
0009 9E	17 +	DB	9EH,0E8H	
000A E8				
	18	ENS1		;MACRO for EN S1
000B 85	19 +	DB	85H	
	20	DISSI		;MACRO for DIS SI
000C 95	21 +	DB	95H	
	22	INAP0		;MACRO for IN A,P0
000D 08	23 +	DB	08H	
	24	OUTP0A		;MACRO for OUTL P0,A
000E 38	25 +	DB	38H	
	26	ORLP0	5AH	;MACRO for ORL P0,A
000F 88	27 +	DB	88H,5AH	
0010 5A				
	28	ANLP0	2FH	;MACRO for ANL P0,A
0011 98	29 +	DB	98H,2FH	
0012 2F				
	30	DECAR0		;MACRO for DEC @R0
0013 C0	31 +	DB	0C0H	
	32	DECAR1		;MACRO for DEC @R1
0014 C1	33 +	DB	0C1H	
	34	SELMB2		;MACRO for SEL MB2
0015 A5	35 +	DB	0A5H	
	36	SELMB3		;MACRO for SEL MB3
0016 B5	37 +	DB	0B5H	
	38	DJNZA0	567H	;MACRO for DJNZ @R0, 567H
0017 E0	39 +	DB	0E0H,567H AND 0FFH	
0019 67				
	40	DJNZA1	0EFEH	;MACRO for DJNZ @R1, 0EFEH
0019 E1	41 +	DB	0E1H,0EFEH AND 0FFH	
001A FE				
	42	JNTF	789H	;MACRO for JNTF 789H
001B 06	43 +	DB	06H, 789H AND 0FFH	
001C 89				
	44	END		

Signetics

Section 4
Tuning Systems

Linear Products

4

INDEX

Signetics

PCF8570
256 × 8 Static RAM

Product Specification

Linear Products

DESCRIPTION

The PCF8570 is a low power 2048-bit static CMOS RAM organized as 256 words by 8-bits. Addresses and data are transferred serially via a two-line bidirectional bus (I^2C). The built-in word address register is incremented automatically after each written or read data byte. Three address pins — A0, A1, and A2 — are used for programming the hardware address, allowing the use of up to eight devices connected to the bus without additional hardware.

FEATURES

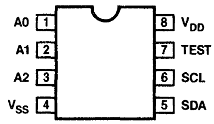
- **Operating supply voltage: 2.5V to 6V**
- **Low data retention voltage: min. 1.0V**
- **Low standby current: max. 5μA**
- **Power saving mode: typ. 50nA**
- **Serial input/output bus (I^2C)**
- **Address by 3 hardware address pins**
- **Automatic word address incrementing**
- **8-lead DIP package**

APPLICATIONS

- **Telephony RAM expansion for stored numbers in repertory dialing (e.g., PCD3343 applications)**
- **Radio and television channel presets**
- **Video cassette recorder**
- **General purpose RAM expansion for the microcomputer families MAB8400 and PCF84C00**

PIN CONFIGURATION

N, D Packages



TOP VIEW

CD11900S

PIN NO.	SYMBOL	DESCRIPTION
1	A0	Address inputs
2	A1	Address inputs
3	A2	Address inputs
4	V_{SS}	Negative supply
5	SDA	Serial data line (I^2C bus)
6	SCL	Serial clock line (I^2C bus)
7	TEST	Test input for test speed-up; must be connected to V_{SS} when not in use. (Power saving mode, see Figures 12 and 13)
8	V_{DD}	Positive supply

ORDERING INFORMATION

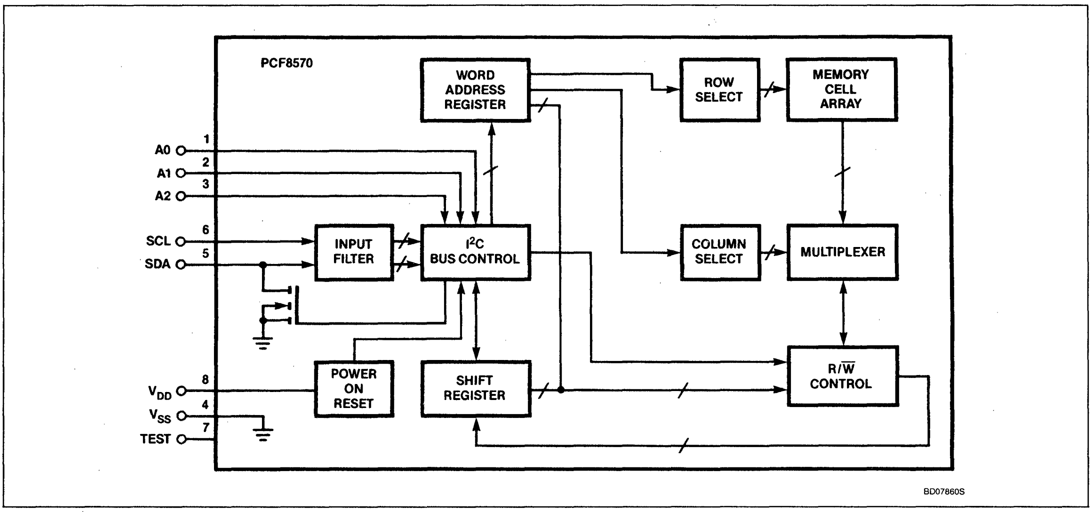
DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
8-Pin Plastic DIP (SOT-97A)	−40°C to +85°C	PCF8570PN
8-Pin Plastic SO (SO-8L; SOT-176)	−40°C to +85°C	PCF8570TD

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{DD}	Supply voltage range (Pin 8)	−0.8 to +8.0	V
V_I	Voltage range on any input	−0.8 to V_{DD} +0.8	V
$\pm I_I$	DC input current (any input)	10	mA
$\pm I_O$	DC output current (any output)	10	mA
$\pm I_{DD}$; I_{SS}	Supply current (Pin 4 or Pin 8)	50	mA
P_{TOT}	Power dissipation per package	300	mW
P_O	Power dissipation per output	50	mW
T_{STG}	Storage temperature range	−65 to +150	°C
T_A	Operating ambient temperature range	−40 to +85	°C

4

BLOCK DIAGRAM



DC ELECTRICAL CHARACTERISTICS V_{DD} = 2.5 to 6V; V_{SS} = 0V; T_A = −40°C to +85°C, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supply					
V_{DD}	Supply voltage	2.5		6	V
	Supply current at f_{SCL} = 100kHz; $V_I = V_{SS}$ or V_{DD}				
I_{DD}	operating			200	µA
I_{DDO}	standby			15	µA
I_{DDO}	standby at T_A = −25 to +70°C			5	µA
V_{POR}	Power-on reset voltage level[1]	1.5	1.9	2.3	V
Input SCL; input/output SDA					
V_{IL}	Input voltage LOW[2]	−0.8		0.3 × V_{DD}	V
V_{IH}	Input voltage HIGH[2]	0.7 × V_{DD}		V_{DD} + 0.8	V
I_{OL}	Output current LOW at V_{OL} = 0.4V	3			mA
I_{OH}	Output leakage current HIGH at $V_{OH} = V_{DD}$			250	nA
$\pm I_I$	Input leakage current (A0, A1, A2) at $V_I = V_{DD}$ or V_{SS}			250	nA
f_{SCL}	Clock frequency (Figure 5)	0		100	kHz
C_I	Input capacitance (SCL, SDA) at $V_I = V_{SS}$			7	pF
t_{SW}	Tolerable spike width on bus			100	ns
LOW V_{DD} data retention					
V_{DDR}	Supply voltage for data retention	1		6	V
I_{DDR}	Supply current at V_{DDR} = 1V			5	µA
I_{DDR}	Supply current at V_{DDR} = 1V; T_A = −25 to +70°C			2	µA
Power saving mode					
I_{DDR}	Supply current at T_A = 25°C; TEST = V_{DDR}		50	400	nA

NOTES:
1. The power-on reset circuit resets the I²C bus logic when $V_{DD} < V_{POR}$.
2. If the input voltages are a diode voltage above or below the supply voltage V_{DD} or V_{SS} an input current will flow; this current must not exceed ±0.5mA.

CHARACTERISTICS OF THE I²C BUS

The I²C bus is for 2-way, 2-line communication between different ICs or modules. The two lines are a serial data line (SDA) and a serial clock line (SCL). Both lines must be connected to a positive supply via a pull-up resistor when connected to the output stages of a device. Data transfer may be initiated only when the bus is not busy.

Bit Transfer

One data bit is transferred during each clock pulse. The data on the SDA line must remain stable during the HIGH period of the clock pulse, as changes in the data line at this time will be interpreted as control signals.

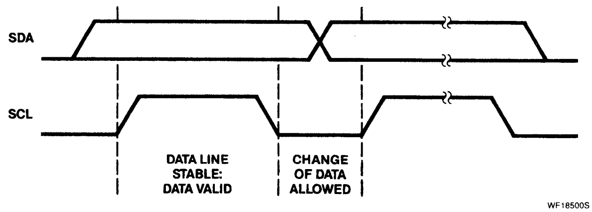


Figure 1. Bit Transfer

Start and Stop Conditions

Both data and clock lines remain HIGH when the bus is not busy. A HIGH-to-LOW transition of the data line while the clock is HIGH is defined as the start condition (S). A LOW-to-HIGH transition of the data line while the clock is HIGH is defined as the stop condition (P).

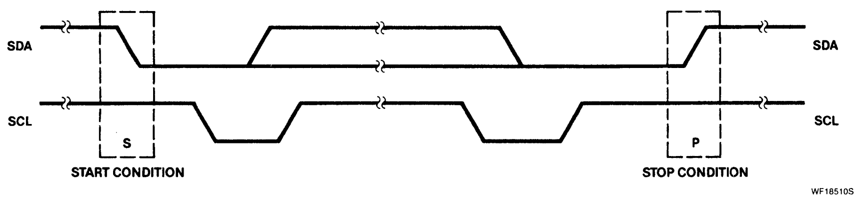


Figure 2. Definition of Start and Stop Conditions

System Configuration

A device generating a message is a "transmitter"; a device receiving a message is the "receiver". The device that controls the message is the "master" and the devices which are controlled by the master are the "slaves".

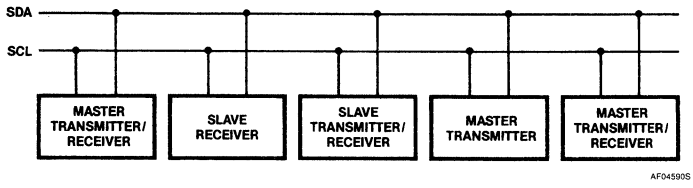


Figure 3. System Configuration

256 × 8 Static RAM | PCF8570

Acknowledge

The number of data bytes transferred between the start and stop conditions from transmitter to receiver is not limited. Each byte of eight bits is followed by one acknowledge bit. The acknowledge bit is a HIGH level put on the bus by the transmitter whereas the master generates an extra acknowledge related clock pulse. A slave receiver which is addressed must generate an acknowledge after the reception of each byte. Also a master must generate an acknowledge after the reception of each byte that has been clocked out of the slave transmitter. The device that acknowledges has to pull down the SDA line during the acknowledge clock pulse, so that the SDA line is stable LOW. During the HIGH period of the acknowledge related clock pulse, setup and hold times must be taken into account. A master receiver must signal an end of data to the transmitter by *not* generating an acknowledge on the last byte that has been clocked out of the slave. In this event the transmitter must leave the data line HIGH to enable the master to generate a stop condition.

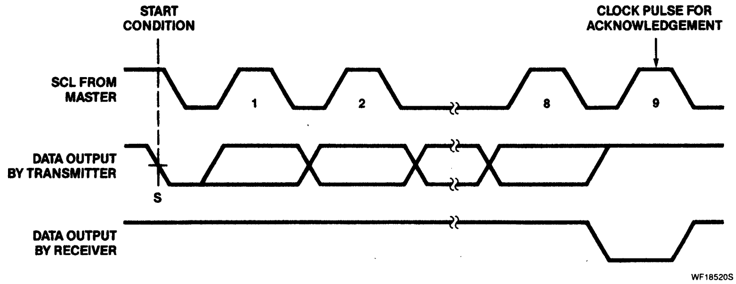
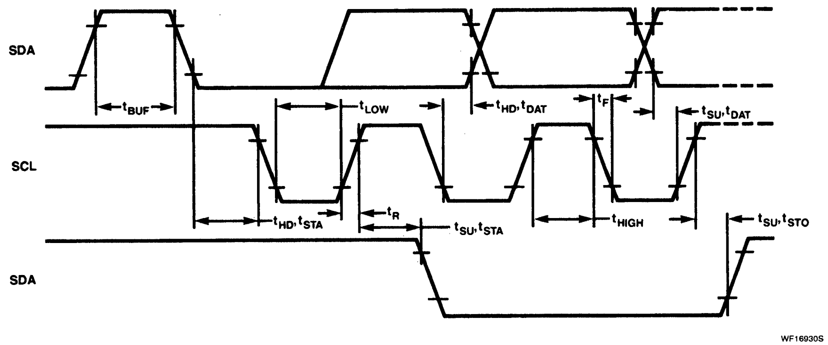


Figure 4. Acknowledge on the I²C Bus

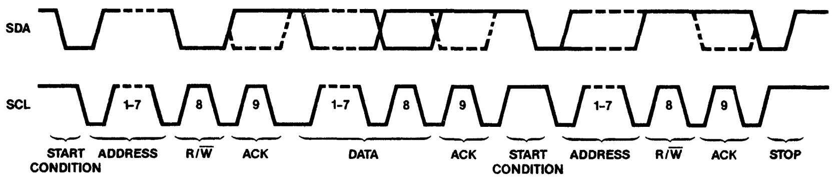


Where:

t_{BUF}	$t \geqslant t_{LOWmin}$	The minimum time the bus must be free before a new transmission can start
$t_{HD;}$ t_{STA}	$t \geqslant t_{HIGHmin}$	Start condition hold time
$t_{LOWSmin}$	$4.7\mu s$	Clock LOW period
$t_{HIGHmin}$	$4\mu s$	Clock HIGH period
$t_{SU;}$ t_{STA}	$t \geqslant t_{LOWmin}$	Start condition set-up time; only valid for repeated start code
$t_{HD;}$ t_{DAT}	$t = 0\mu s$	Data hold time
$t_{SU;}$ t_{DAT}	$t \geqslant 250ns$	Data setup time
t_R	$t \leqslant 1\mu s$	Rise time of both the SDA and SCL line
t_F	$t \leqslant 300ns$	Fall time of both the SDA and SCL line
$t_{SU;}$ t_{STO}	$t \geqslant t_{LOWmin}$	Stop condition setup time

NOTE:
All the timing values refer to V_{IH} and V_{IL} levels with a voltage swing of V_{SS} to V_{DD}.

Figure 5. Timing



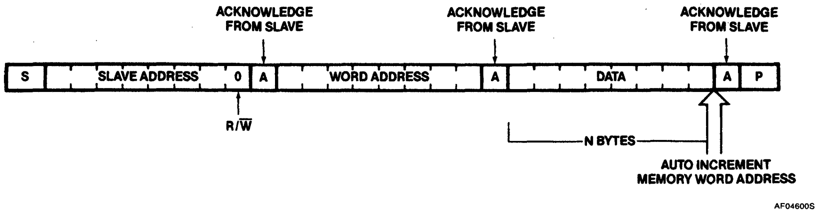
Where:

Clock t_{LOWmin}	4.7μs
$t_{HIGHmin}$	4μs
The dashed line is the acknowledgement of the receiver	
Mark-to-space ratio	1:1 (LOW-to-HIGH)
Maximum number of bytes	Unrestricted
Premature termination of transfer	Allowed by generation of STOP condition
Acknowledge clock bit	Must be provided by the master

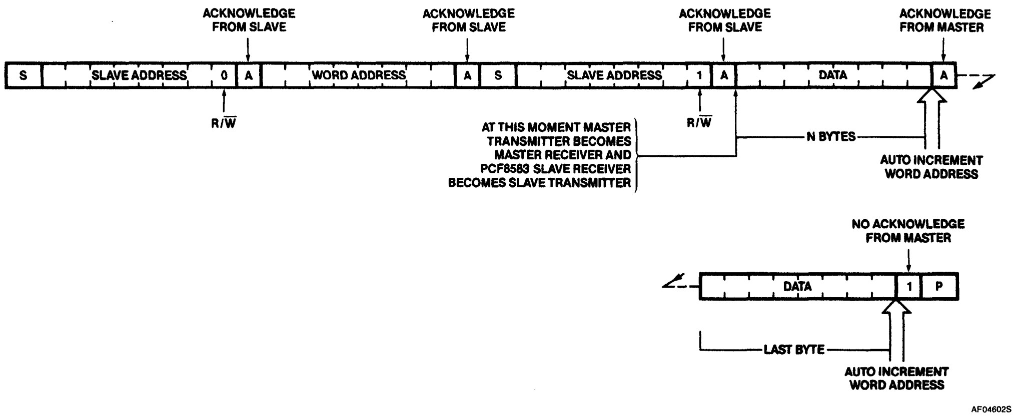
Figure 6. Complete Data Transfer in the High-Speed Mode

Bus Protocol

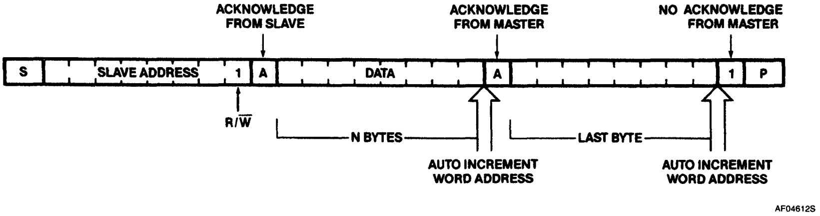
Before any data is transmitted on the I^2C bus, the device which should respond is addressed first. The addressing is always done with the first byte transmitted after the start procedure. The I^2C bus configuration for different PCF8570 READ and WRITE cycles is shown in Figure 7.



a. Master Transmits to Slave Receiver (WRITE Mode)



b. Master Reads After Setting Word Address (WRITE Word Address; READ Data)

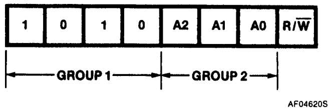


c. Master Reads Slave Immediately After First Byte (READ Mode)

Figure 7

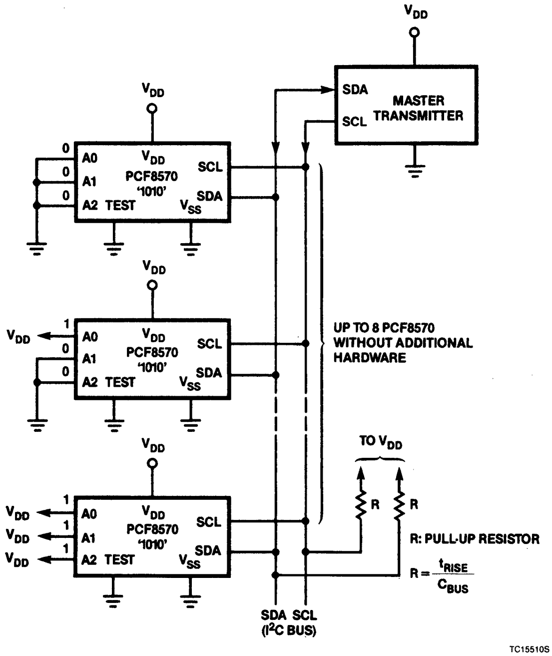
APPLICATION INFORMATION

The PCF8570 slave address has a fixed combination 1010 as group 1, while group 2 is fully programmable (see Figure 8.)



NOTE:
PCF8570A version: the slave address A0 state is X (don't care); however, the hardware address A0 input must still be connected to V_{SS} or V_{DD}.

Figure 8. PCF8570 Address



NOTE:
A0, A1, and A2 inputs must be connected to V_{DD} or V_{SS} but not left open.

Figure 9. PCF8570 Application Diagram

POWER SAVING MODE

With the condition TEST = V_{DDR}, the PCF8570 goes into the power saving mode.

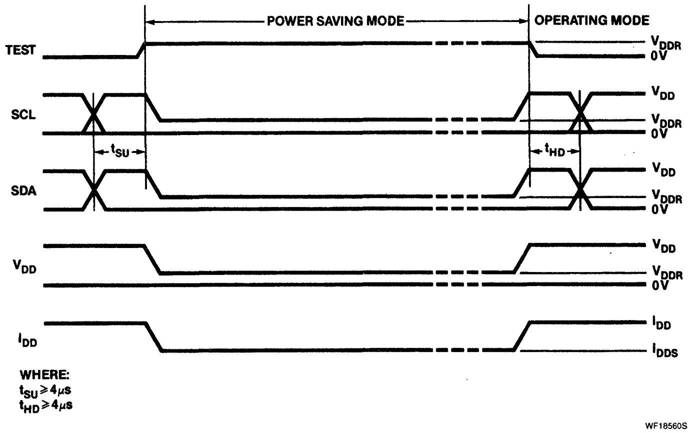
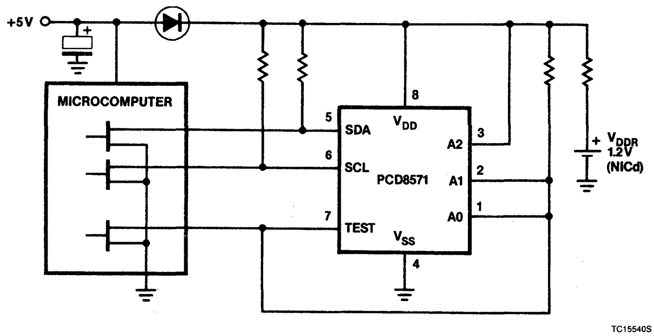


Figure 10. Timing for Power Saving Mode



NOTE:
1. In the operating mode, TEST = 0.
2. In the power saving mode, TEST = V_{DDR}.

Figure 11. Application Example for Power Saving Mode

Signetics

PCF8571
1K Serial RAM

Product Specification

Linear Products

DESCRIPTION

The PCF8571 is a low power 1024-bit static CMOS RAM organized as 128 words by 8 bits. Addresses and data are transferred serially via a two-line bidirectional bus (I^2C). The built-in word address register is incremented automatically after each written or read data byte. Three address pins — A0, A1, and A2 — are used for programming the hardware address, allowing the use of up to eight devices connected to the bus without additional hardware.

FEATURES

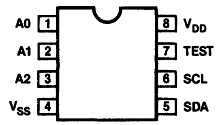
- **Operating supply voltage: 2.5V to 6V**
- **Low data retention voltage: min. 1.0V**
- **Low standby current: max. 5μA**
- **Power saving mode: typ. 50nA**
- **Serial input/output bus (I^2C)**
- **Address by 3 hardware address pins**
- **Automatic word address incrementing**
- **8-lead DIP package**

APPLICATIONS

- **Telephony**
  RAM expansion for stored numbers in repertory dialing (e.g., PCD3340 applications)
- **Radio and television channel presets**
- **Video cassette recorder**
- **General purpose**
  RAM expansion for the micro-computer families MAB8400 and PCF84C00

PIN CONFIGURATION

N, D Packages



TOP VIEW

CD11900S

PIN NO.	SYMBOL	DESCRIPTION
1	A0	Address inputs
2	A1	Address inputs
3	A2	Address inputs
4	V_{SS}	Negative supply
5	SDA	Serial data line (I^2C bus)
6	SCL	Serial clock line (I^2C bus)
7	TEST	Test input for test speed-up; must be connected to V_{SS} when not in use. (Power saving mode, see Figures 12 and 13)
8	V_{DD}	Positive supply

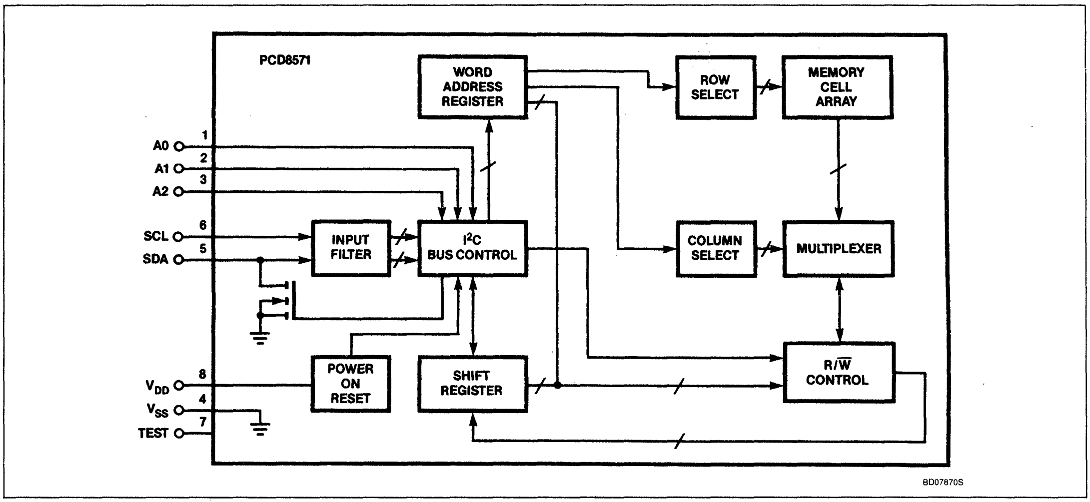
ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
8-Pin Plastic DIP (SOT-97A)	−25°C to +70°C	PCF8571PN
8-Pin Plastic SO (SOL-8; SOT-176)	−25°C to +70°C	PCF8571TD

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{DD}	Supply voltage range (Pin 8)	−0.8 to +8.0	V
V_I	Voltage range on any input	−0.8 to V_{DD} +0.8	V
$\pm I_I$	DC input current (any input)	10	mA
$\pm I_O$	DC output current (any output)	10	mA
$\pm I_{DD}$; I_{SS}	Supply current (Pin 4 or Pin 8)	50	mA
P_{TOT}	Power dissipation per package	300	mW
P_O	Power dissipation per output	50	mW
T_{STG}	Storage temperature range	−65 to +150	°C
T_A	Operating ambient temperature range	−25 to +70	°C

BLOCK DIAGRAM



DC ELECTRICAL CHARACTERISTICS V_{DD} = 2.5 to 6V; V_{SS} = 0V; T_A = −25°C to +70°C, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supply					
V_{DD}	Supply voltage	2.5		6	V
 I_{DD} I_{DDO}	Supply current at f_{SCL} = 100kHz; $V_I = V_{SS}$ or V_{DD} operating standby			 200 5	 μA μA
V_{POR}	Power-on reset voltage level at $V_{SCL} = V_{SDA} = V_{DD}$ [1]	1.5	1.9	2.3	V
Input SCL; input/output SDA					
V_{IL}	Input voltage LOW[2]	−0.8		$0.3 \times V_{DD}$	V
V_{IH}	Input voltage HIGH[2]	$0.7 \times V_{DD}$		$V_{DD} + 0.8$	V
I_{OL}	Output current LOW at V_{OL} = 0.4V	3			mA
I_{OH}	Output leakage current HIGH at $V_{OH} = V_{DD}$			100	nA
$\pm I_I$	Input leakage current (A0, A1, A2) at $V_I = V_{DD}$ or V_{SS}			100	nA
f_{SCL}	Clock frequency (Figure 5)	0		100	kHz
C_I	Input capacitance (SCL, SDA) at $V_I = V_{SS}$			7	pF
t_{SW}	Tolerable spike width on bus			100	ns
LOW V_{DD} data retention					
V_{DDR}	Supply voltage for data retention	1			V
I_{DDR}	Supply current at V_{DDR} = 1V			2	μA
Power saving mode (Figure 12)					
I_{DDS}	Supply current at T_A = 25°C; TEST = A0 = A1 = A2 = V_{DDR}		50	200	nA

NOTES:
1. The power-on reset circuit resets the I²C bus logic when $V_{DD} < V_{POR}$.
2. If the input voltages are a diode voltage above or below the supply voltage V_{DD} or V_{SS} an input current will flow: this current must not exceed ±0.5mA.

CHARACTERISTICS OF THE I²C BUS

The I²C bus is for 2-way, 2-line communication between different ICs or modules. The two lines are a serial data line (SDA) and a serial clock line (SCL). Both lines must be connected to a positive supply via a pull-up resistor when connected to the output stages of a device. Data transfer may be initiated only when the bus is not busy.

Bit Transfer

One data bit is transferred during each clock pulse. The data on the SDA line must remain stable during the HIGH period of the clock pulse, as changes in the data line at this time will be interpreted as control signals.

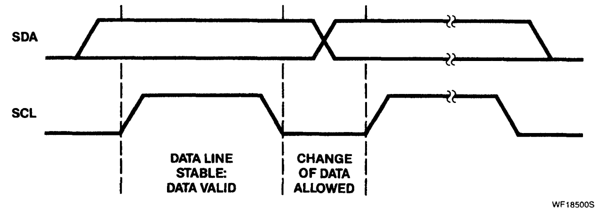


Figure 1. Bit Transfer

Start and Stop Conditions

Both data and clock lines remain HIGH when the bus is not busy. A HIGH-to-LOW transition of the data line while the clock is HIGH is defined as the start condition (S). A LOW-to-HIGH transition of the data line while the clock is HIGH is defined as the stop condition (P).

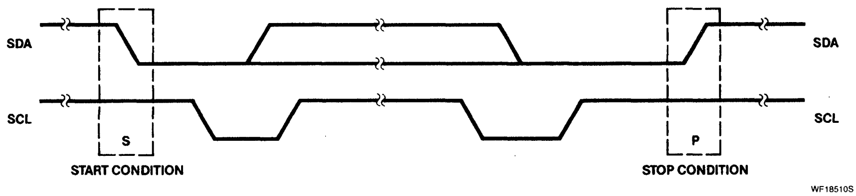


Figure 2. Definition of Start and Stop Conditions

System Configuration

A device generating a message is a "transmitter"; a device receiving a message is the "receiver". The device that controls the message is the "master" and the devices which are controlled by the master are the "slaves".

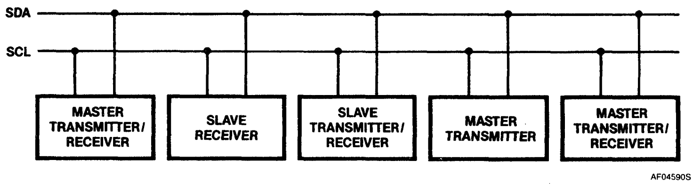


Figure 3. System Configuration

Acknowledge

The number of data bytes transferred between the start and stop conditions from transmitter to receiver is not limited. Each byte of eight bits is followed by one acknowledge bit. The acknowledge bit is a HIGH level put on the bus by the transmitter, whereas the master generates an extra acknowledge related clock pulse. A slave receiver which is addressed must generate an acknowledge after the reception of each byte. Also, a master must generate an acknowledge after the reception of each byte that has been clocked out of the slave transmitter. The device that acknowledges has to pull down the SDA line during the acknowledge clock pulse, so that the SDA line is stable LOW. During the HIGH period of the acknowledge related clock pulse, set-up and hold times must be taken into account. A master receiver must signal an end-of-data to the transmitter by *not* generating an acknowledge on the last byte that has been clocked out of the slave. In this event the transmitter must leave the data line HIGH to enable the master to generate a stop condition.

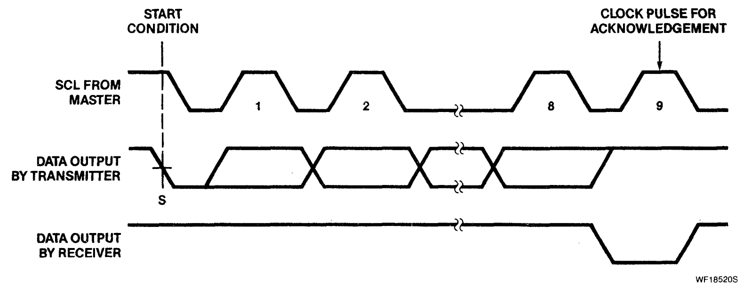
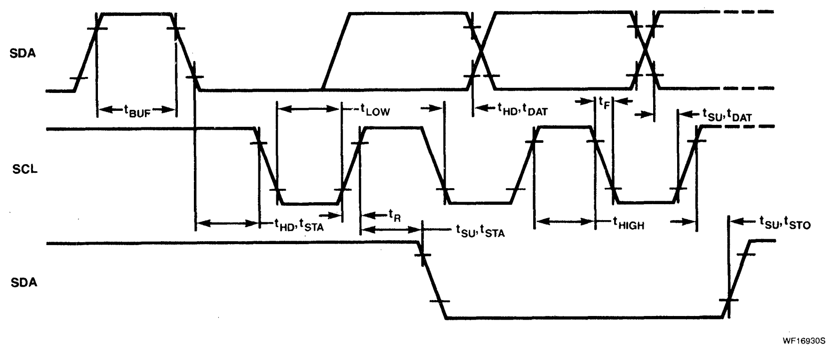


Figure 4. Acknowledgement on the I²C Bus

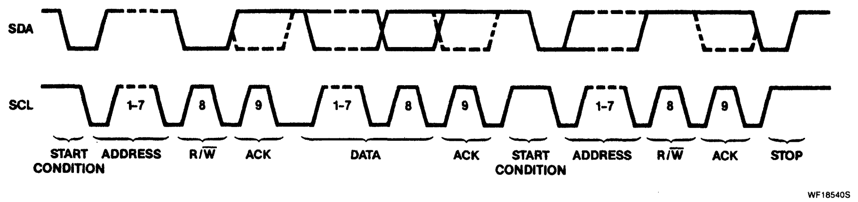


Where:

t_{BUF}	$t \geqslant t_{LOWmin}$	The minimum time the bus must be free before a new transmission can start
t_{HD}, t_{STA}	$t \geqslant t_{HIGHmin}$	Start condition hold time
t_{LOWmin}	4.7μs	Clock LOW period
$t_{HIGHmin}$	4μs	Clock HIGH period
t_{SU}, t_{STA}	$t \geqslant t_{LOWmin}$	Start condition setup time; only valid for repeated start code
t_{HD}, t_{DAT}	$t \geqslant 0\mu s$	Data hold time
t_{SU}, t_{DAT}	$t \geqslant 250ns$	Data setup time
t_R	$t \leqslant 1\mu s$	Rise time of both the SDA and SCL line
t_F	$t \leqslant 300ns$	Fall time of both the SDA and SCL line
t_{SU}, t_{STO}	$t \geqslant t_{LOWmin}$	Stop condition setup time

NOTE:
All the timing values refer to V_{IH} and V_{IL} levels with a voltage swing of V_{SS} to V_{DD}.

Figure 5. Timing



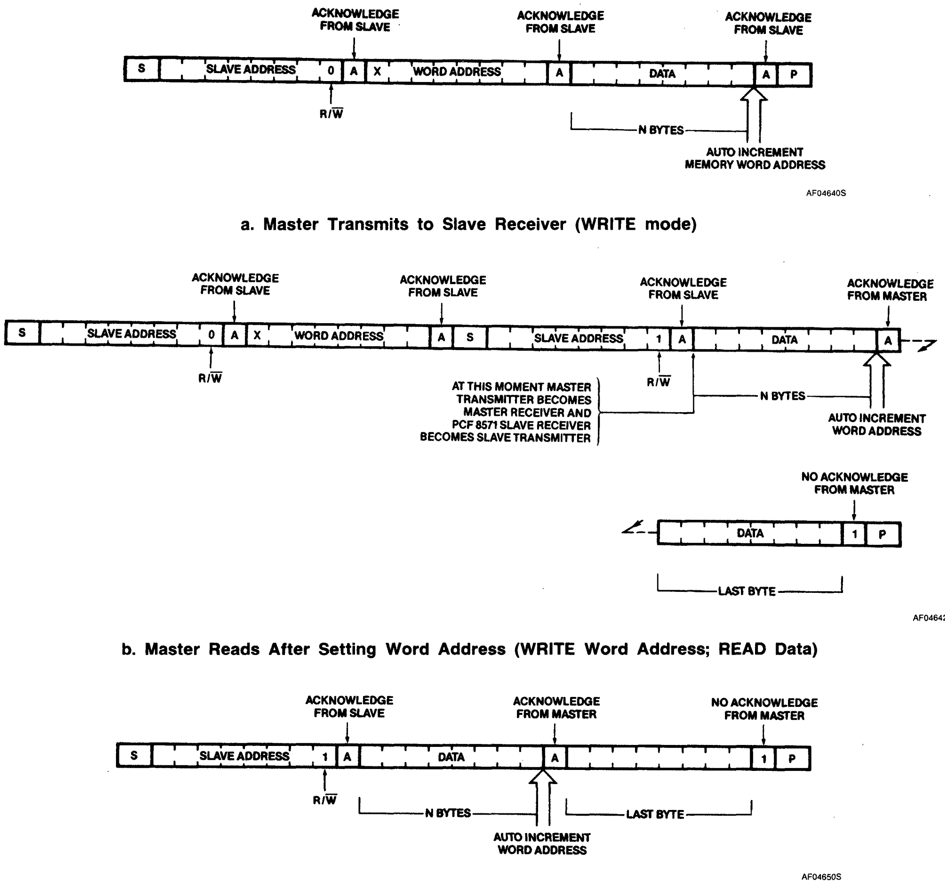
Where:

Clock t_{LOWmin}	4.7μs
$t_{HIGHmin}$	4μs
The dashed line is the acknowledgement of the receiver	
Mark-to-space ratio	1:1 (LOW-to-HIGH)
Maximum number of bytes	Unrestricted
Premature termination of transfer	Allowed by generation of STOP condition
Acknowledge clock bit	Must be provided by the master

Figure 6. Complete Data Transfer

Bus Protocol

Before any data is transmitted on the I²C bus, the device which should respond is addressed first. The addressing is always done with the first byte transmitted after the start procedure. The I²C bus configuration for different PCF8571 READ and WRITE cycles is shown in Figure 7.



a. Master Transmits to Slave Receiver (WRITE mode)

b. Master Reads After Setting Word Address (WRITE Word Address; READ Data)

c. Master Reads Slave Immediately After First Byte (READ Mode)

NOTE:
X = don't care bit

Figure 7

APPLICATION INFORMATION

The PCF8571 slave address has a fixed combination 1010 as group 1, while group 2 is fully programmable (see Figure 8).

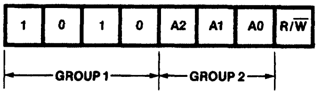
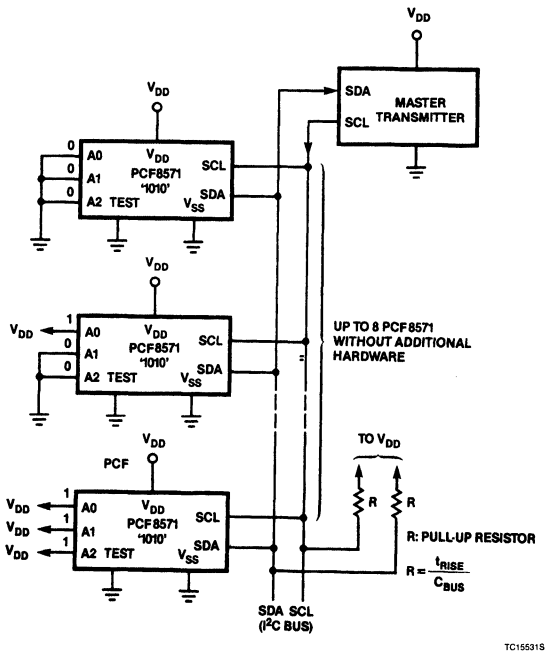


Figure 8. PCF8571 Address

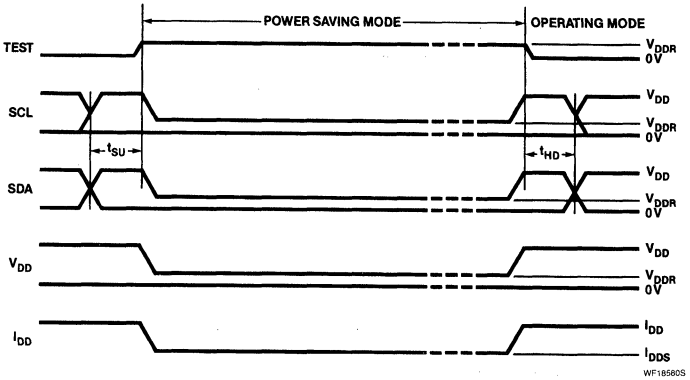


NOTES:
A0, A1, and A2 inputs must be connected to V_{DD} or V_{SS} but not left open.

Figure 9. PCF8571 Application Diagram

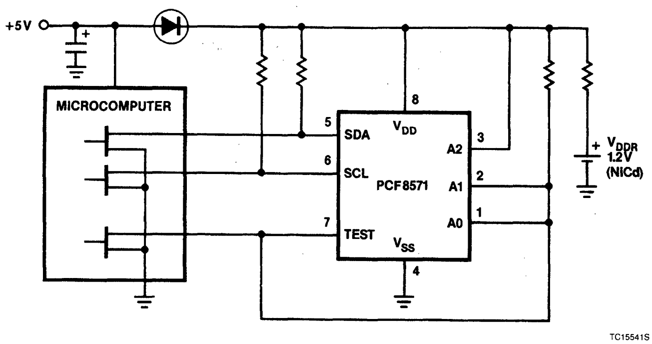
POWER SAVING MODE

With the condition TEST = A2 = A1 = A0 = V_{DDR}, the PCF8571 goes into the power saving mode.



Where:
$t_{SU} \geq 4\mu s$
$t_{HD} \geq 4\mu s$

Figure 10. Timing for Power Saving Mode



NOTES:
1. In the operating mode, TEST = 0 (A0, A1, = 0; A2 = 1).
2. In the power saving mode, TEST = A0 = A1 = A2 = V_{DDR}.

Figure 11. Application Example for Power Saving Mode

Signetics

PCF8573
Clock/Calendar with Serial I/O

Product Specification

Linear Products

DESCRIPTION

The PCF8573 is a low threshold, monolithic CMOS circuit that functions as a real-time clock/calendar in the Inter IC (I^2C) bus-oriented microcomputer systems. The device includes an addressable time counter and alarm register, both for minutes, hours, days and months. Three special control/status flags, COMP, POWF and NODA, are also available. Information is transferred serially via a two-lin bidirectional bus (I^2C). Back-up for the clock during supply interruptions is provided by a 1.2V nickel cadmium battery. The time base is generated from a 32.768kHz crystal-controlled oscillator.

FEATURES

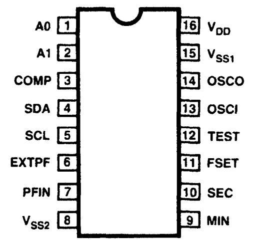
- **Serial input/output bus (I^2C) interface for minutes, hours, days and months**
- **Additional pulse outputs for seconds and minutes**
- **Alarm register for presetting a time for alarm or remote switching functions**
- **Battery back-up for clock function during supply interruption**
- **Crystal oscillator control (32.768kHz)**

APPLICATIONS

- **Automotive**
- **Telephony**

PIN CONFIGURATION

N, D Packages



TOP VIEW

CD10281S

PIN NO.	SYMBOL	DESCRIPTION
1	A0	Address input
2	A1	Address input
3	COMP	Comparator output
4	SDA	Serial data line } I^2C bus
5	SCL	Serial clock line } I^2C bus
6	EXTPF	Enable power fail flag input
7	PFIN	Power fail flag input
8	V_{SS2}	Negative supply 2 (I^2C interface)
9	MIN	One pulse per minute output
10	SEC	One pulse per second output
11	FSET	Oscillator tuning output
12	TEST	Test input; must be connected to V_{SS2} when not in use
13	OSCI	Oscillator input
14	OSCO	Oscillator input/output
15	V_{SS1}	Negative supply 1 (clock)
16	V_{DD}	Common positive supply

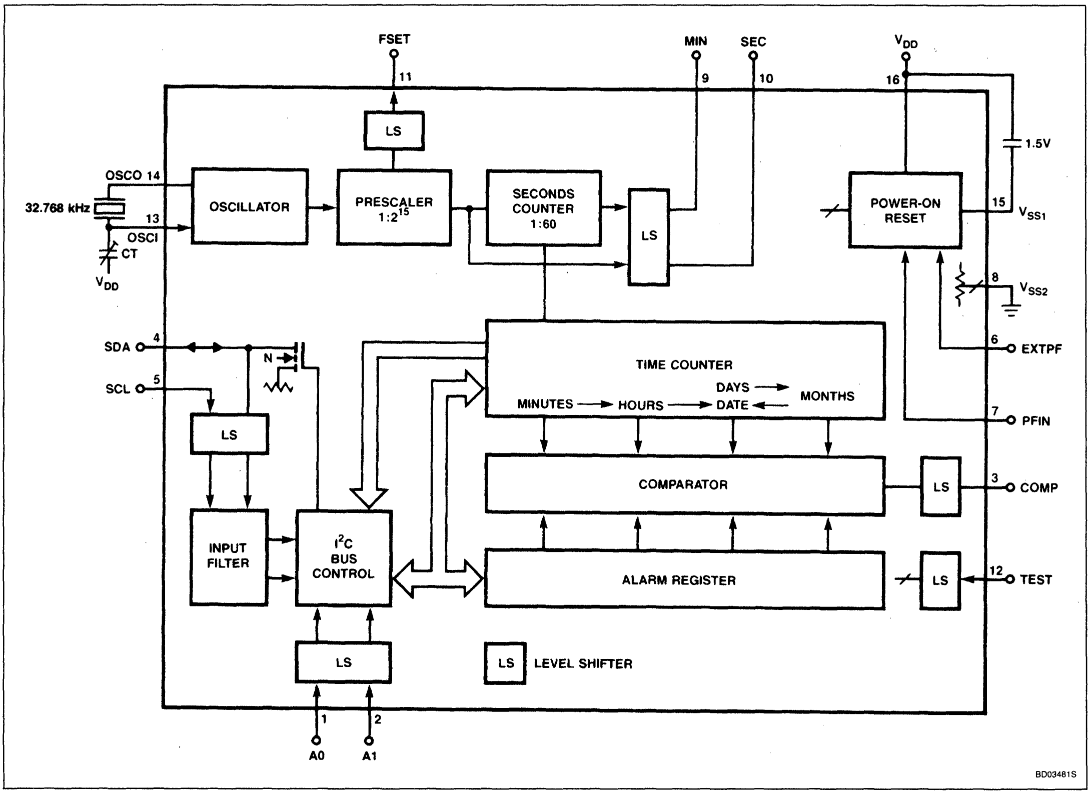
ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
16-Pin Plastic DIP (SOT-38)	-40°C to +85°C	PCF8573PN
16-Pin Plastic SOL (SOT-162A)	-40°C to +85°C	PCF8573T

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
$V_{DD} - V_{SS1}$	Supply voltage range (clock)	-0.3 to +8	V
$V_{DD} - V_{SS2}$	Supply voltage range (I^2C interface)	-0.3 to +8	V
I_{IN}	Input current	10	mA
I_{OUT}	Output current	10	mA
P_D	Maximum power dissipation per package	200	mW
T_A	Operating ambient temperature range	-40 to +85	°C
T_{STG}	Storage temperature range	-65 to +150	°C

BLOCK DIAGRAM



Clock/Calendar with Serial I/O — PCF8573

DC ELECTRICAL CHARACTERISTICS

V_{SS2} = 0V; T_A = −40 to +85°C, unless otherwise specified. Typical values at T_A = +25°C.

SYMBOL	PARAMETER	LIMITS Min	LIMITS Typ	LIMITS Max	UNIT
Supply					
$V_{DD} - V_{SS2}$	Supply voltage (I^2C interface)	2.5	5	6.0	V
$V_{DD} - V_{SS1}$	Supply voltage (clock)	1.1	1.5	$(V_{DD} - V_{SS2})$	V
	Supply current V_{SS1}				
$-I_{SS1}$	at $V_{DD} - V_{SS1}$ = 1.5V		3	10	µA
$-I_{SS1}$	at $V_{DD} - V_{SS1}$ = 5V		12	50	µA
$-I_{SS2}$	Supply current V_{SS2} at $V_{DD} - V_{SS2}$ = 5V (I_O = 0mA on all outputs)			50	µA
Inputs SCL, SDA, A0, A1, TEST					
V_{IH}	Input voltage HIGH	$0.7 \times V_{DD}$			V
V_{IL}	Input voltage LOW			$0.2 \times V_{DD}$	V
$\pm I_I$	Input leakage current at $V_I = V_{SS2}$ to V_{DD}			1	µA
Inputs EXTPF, PFIN					
$V_{IH} - V_{SS1}$	Input voltage HIGH	$0.7 \times (V_{DD} - V_{SS1})$			V
$V_{IL} - V_{SS1}$	Input voltage LOW	0		$0.2 \times (V_{DD} - V_{SS1})$	V
$\pm I_I$	Input leakage current at $V_I = V_{SS1}$ to V_{DD}			1	µA
$\pm I_I$	at T_A = 25°C; $V_I = V_{SS1}$ to V_{DD}			0.1	µA
Outputs SEC, MIN, COMP, FSET (normal buffer outputs)					
	Output voltage HIGH				
V_{OH}	at $V_{DD} - V_{SS2}$ = 2.5V; $-I_O$ = 0.1mA	$V_{DD} - 0.4$			V
V_{OH}	at $V_{DD} - V_{SS2}$ = 4 to 6V; $-I_O$ = 0.5mA	$V_{DD} - 0.4$			V
	Output voltage LOW				
V_{OL}	at $V_{DD} - V_{SS2}$ = 2.5V; I_O = 0.3mA			0.4	V
V_{OL}	at $V_{DD} - V_{SS2}$ = 4 to 6V; I_O = 1.6mA			0.4	V
Output SDA (N-Channel open drain)					
V_{OL}	Output 'ON': I_O = 3mA at $V_{DD} - V_{SS2}$ = 2.5 to 6V			0.4	V
I_O	Output 'OFF' (leakage current) at $V_{DD} - V_{SS2}$ = 6V; V_O = 6V			1	µA
Internal Threshold Voltage					
V_{TH1}	Power failure detection	1	1.2	1.4	V
V_{TH2}	Power 'ON' reset at $V_{SCL} = V_{SDA} = V_{DD}$	1.5	2.0	2.5	V

AC ELECTRICAL CHARACTERISTICS V_{SS2} = 0V; T_A = −40 to + 85°C, unless otherwise specified. Typical values at T_A = + 25°C.

SYMBOL	PARAMETER	LIMITS Min	Typ	Max	UNIT
Rise and Fall Times of Input Signals					
t_R, t_F	Input EXTPF			1	μs
t_R, t_F	Input PFIN			∞	μs
  t_R t_F	Input signals except EXTPF and PFIN between V_{IL} and V_{IH} levels rise time fall time			 1 0.3	 μs μs
Frequency at SCL					
 t_{LOW}	at $V_{DD} - V_{SS2}$ = 4 to 6V Pulse width LOW (see Figure 8)	 4.7			 μs
t_{HIGH}	Pulse width HIGH (see Figure 8)	4			μs
t_I	Noise suppression time constant at SCL and SDA input	0.25	1	2.5	μs
c_{IN}	Input capacitance (SCL, SDA)			7	pF
Oscillator					
C_{OUT}	Integrated oscillator capacitance		40		pF
R_F	Oscillator feedback resistance		3		MΩ
   f/f_{OSC}	Oscillator stability for: $\Delta(V_{DD} - V_{SS1})$ = 100mV at $V_{DD} - V_{SS1}$ = 1.55V; T_A = 25°C		 2×10^{-6}		
	Quartz crystal parameters				
	Frequency = 32.768 kHz				
R_S	Series resistance			40	kΩ
C_L	Parallel capacitance		9		pF
C_T	Trimmer capacitance	5		25	pF

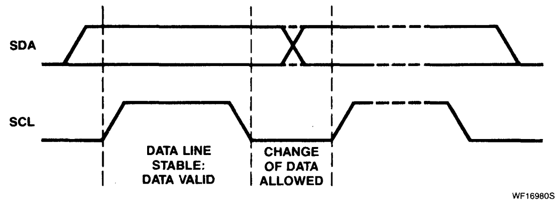


Figure 1. Bit Transfer

Table 1. Cycle Length of the Time Counter

UNIT	NUMBER OF BITS	COUNTING CYCLE	CARRY FOR FOLLOWING UNIT	CONTENT OF MONTH COUNTER
Minutes	7	00 to 59	59 → 00	
Hours	6	00 to 23	23 → 00	
Days	6	01 to 28	28 → 01 or 29 → 01	2 (see note)
		01 to 30	30 → 01	4, 6, 9, 11
		01 to 31	31 → 01	1, 3, 5, 7, 8, 10, 12
Months	5	01 to 12	12 → 01	

NOTE: Day counter may be set to 29 by a write transmission with EXECUTE ADDRESS.

FUNCTIONAL DESCRIPTION

Oscillator

The PCF8573 has an integrated crystal-controlled oscillator which provides the time base for the prescaler. The frequency is determined by a single 32.768kHz crystal connected between OSCI and OSCO. A trimmer is connected between OSCI and V_{DD}.

Prescaler and Time Counter

The prescaler provides a 128Hz signal at the FSET output for fine adjustment of the crystal oscillator without loading it. The prescaler also generates a pulse once a second to advance the seconds counter. The carry of the prescaler and the seconds counter are available at the outputs SEC and MIN, respectively, and are also readable via the I^2C bus. The mark-to-space ratio of both signals is 1:1. The time counter is advanced one count by the falling edge of output signal MIN. A transition from HIGH to LOW of output signal SEC triggers MIN to change state. The time counter counts minutes, hours, days and months, and provides a full calendar function which needs to be corrected once every four years. Cycle lengths are shown in Table 1.

Alarm Register

The alarm register is a 24-bit memory. It stores the time-point for the next setting of the status flag COMP. Details of writing and reading of the alarm register are included in the description of the characteristics of the I^2C bus.

Comparator

The comparator compares the contents of the alarm register and the time counter, each with a length of 24 bits. When these contents are equal, the flag COMP will be set 4ms after the falling edge of MIN. This set condition occurs once at the beginning of each minute. This information is latched, but can be cleared by an instruction via the I^2C bus. A clear instruction may be transmitted immediately after the flag is set, and then it will be executed. Flag COMP information is also available at the output COMP. The comparison may be based upon hours and minutes only if the internal flag NODA (no date) is set. Flag NODA can be set and cleared by separate instructions via the I^2C bus, but it is undefined until the first set or clear instruction has been received. Both COMP and NODA flags are readable via the I^2C bus.

Table 2. Power Fail Selection

EXTPF	PFIN	FUNCTION
0	0	Power fail is sensed internally
0	1	Test mode
1	0	Power fail is sensed externally
1	1	No power fail sensed

NOTE:
0: connected to V_{SS1} (LOW)
1: connected to V_{DD} (HIGH)

Power On/Power Fail Detection

If the voltage $V_{DD} - V_{SS1}$ falls below a certain value, the operation of the clock becomes undefined. Thus, a warning signal is required to indicate that faultless operation of the clock is not guaranteed. This information is latched in a flag called POWF (Power Fail) and remains latched after restoration of the correct supply voltage until a write procedure with EXECUTIVE ADDRESS has been received. The flag POWF can be set by an internally-generated power fail level-discriminator signal for application with ($V_{DD} - V_{SS1}$) greater than V_{TH1}, or by an externally-generated power fail signal for application with ($V_{DD} - V_{SS1}$) less than V_{TH1}. The external signal must be applied to the input PFIN. The input stage operates with signals of any slow rise and fall times. Internally-or externally-controlled POWF can be selected by input EXTPF as shown in Table 2.

The external power fail control operates by absence of the $V_{DD} - V_{SS2}$ supply. Therefore, the input levels applied to PFIN and EXTPF must be within the range of $V_{DD} - V_{SS1}$. A LOW level at PFIN indicates a power fail. POWF is readable via the I^2C bus. A power-on reset for the I^2C bus control is generated on-chip when the supply voltage $V_{DD} - V_{SS2}$ is less than V_{TH2}.

Interface Level Shifters

The level shifters adjust the 5V operating voltage ($V_{DD} - V_{SS2}$) of the microcontroller to the internal supply voltage ($V_{DD} - V_{SS1}$) of the clock/calendar. The oscillator and counter are not influenced by the $V_{DD} - V_{SS2}$ supply voltage. If the voltage $V_{DD} - V_{SS2}$ is absent ($V_{SS2} = V_{DD}$) the output signal of the level shifter is HIGH because V_{DD} is the common node of the $V_{DD} - V_{SS2}$ and the $V_{DD} - V_{SS1}$ supplies. Because the level shifters invert the input signal, the internal circuit behaves as if a LOW signal is present on the inputs. FSET, SEC, MIN and COMP are CMOS push-pull output stages. The driving capability of these outputs is lost when the supply voltage $V_{DD} - V_{SS2} = 0$.

CHARACTERISTICS OF THE I²C BUS

The I²C bus is for 2-way, 2-line communication between different ICs or modules. The two lines are a serial data line (SDA) and a serial clock line (SCL). Both lines must be connected to a positive supply via a pull-up resistor when connected to the output stages of a device. Data transfer may be initiated only when the bus is not busy.

Bit Transfer (see Figure 1)

One data bit is transferred during each clock pulse. The data on the SDA line must remain stable during the HIGH period of the clock pulse as changes in the data line at this time will be interpreted as control signals.

Start and Stop Conditions (see Figure 2)

Both data and clock lines remain HIGH when the bus is not busy. A HIGH-to-LOW transition of the data line while the clock is HIGH is defined as the start condition (S). A LOW-to-HIGH transition of the data line while the clock is HIGH is defined as the stop condition (P).

System Configuration (see Figure 3)

A device generating a message is a "transmitter", a device receiving a message is the "receiver". The device that controls the message is the "master" and the devices which are controlled by the master are the "slaves".

Acknowledge (see Figure 4)

The number of data bytes transferred between the start and stop conditions from transmitter to receiver is not limited. Each byte of eight bits is followed by one acknowledge bit. The acknowledge bit is a HIGH level put on the bus by the transmitter whereas the master generates an extra acknowledge-related clock pulse. A slave receiver which is addressed must generate an acknowledge after the reception of each byte. Also a master must generate an acknowledge after the reception of each byte that has been clocked out of the slave transmitter. The device that acknowledges has to pull down the SDA line during the acknowledge clock pulse. So that the SDA line is stable LOW during the HIGH period of the acknowledge related clock pulse, setup and hold times must be taken into account. A master receiver must signal an end of data to the transmitter by not generating an acknowledge on the last byte that has been clocked out of the slave. In this event, the transmitter must leave the data line HIGH to enable the master to generate a stop condition.

Timing Specifications

Masters generate a bus clock with a maximum frequency of 100kHz. Detailed timing is shown in Figure 5.

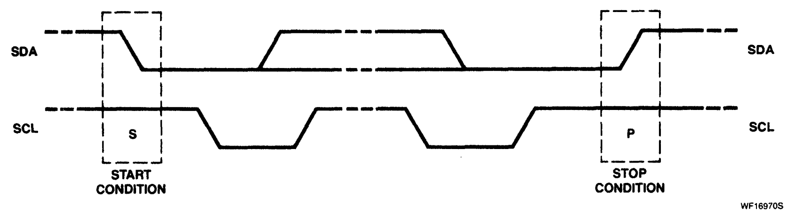


Figure 2. Definition of Start and Stop Conditions

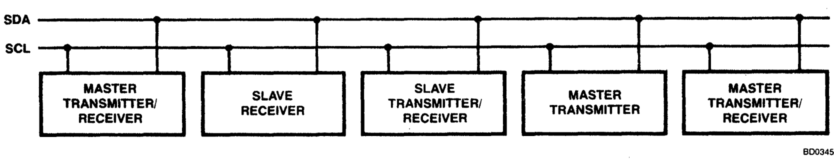


Figure 3. System Configuration

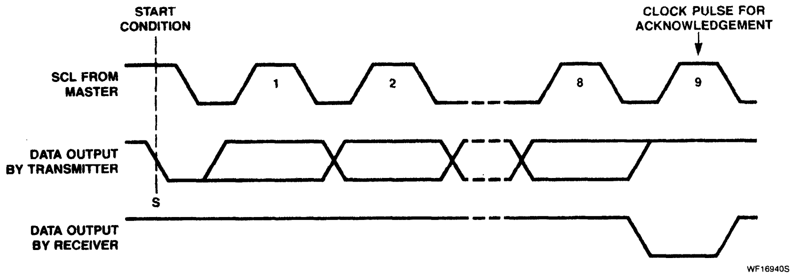
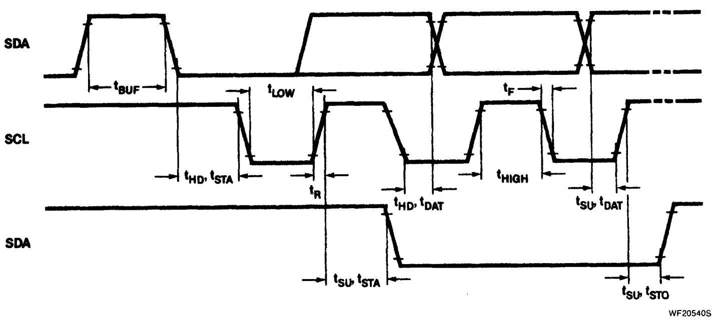


Figure 4. Acknowledgement on the I^2C Bus

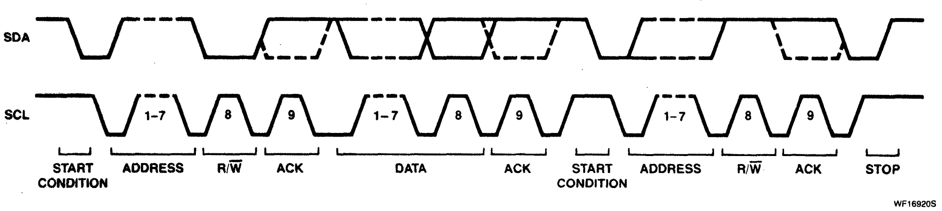


Where:

t_{BUF}	$t \geqslant t_{LOWmin}$	The minimum time the bus must be free before a new transmission can start
t_{HD}, t_{STA}	$t \geqslant t_{HIGHmin}$	Start condition hold time
t_{LOWmin}	4.7μs	Clock LOW period
$t_{HIGHmin}$	4μs	Clock HIGH period
t_{SU}, t_{STA}	$t \geqslant t_{LOWmin}$	Start condition setup time, only valid for repeated start code
t_{HD}, t_{DAT}	$t \geqslant$ 300ns	Data hold time
t_{SU}, t_{DAT}	$t \geqslant$ 250ns	Data setup time
t_R	$t \leqslant$ 1μs	Rise time of both the SDA and SCL line
t_F	$t \leqslant$ 300ns	Fall time of both the SDA and SCL line
t_{SU}, t_{STO}	$t \geqslant t_{LOWmin}$	Stop condition setup time

NOTE:
1. All the values refer to V_{IH} and V_{IL} levels with a voltage swing of V_{DD} to V_{SS2}.

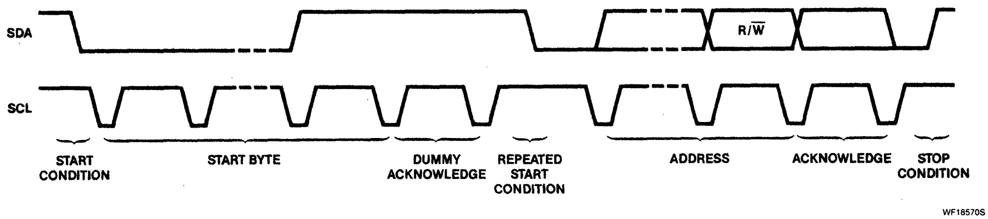
Figure 5. Timing



Where:

Clock t_{LOWmin}	4.7μs
$t_{HIGHmin}$	4μs
The dashed line is the acknowledgement of the receiver	
Mark-to-space ratio	1:1 (LOW-to-HIGH)
Max. number of bytes	unrestricted
Premature termination of transfer	allowed by generation of STOP condition
Acknowledge clock bit	must be provided by the master

Figure 6. Complete Data Transfer



Where:

Clock t_{LOWmin}	130μs ± 25μs
$t_{HIGHmin}$	390μs ± 25μs
Mark-to-space ratio	1:3 (LOW-to-HIGH)
Start byte	0000 0001
Maximum number of bytes	6
Premature termination of transfer	not allowed
Acknowledge clock bit	must be provided by master

NOTE:

1. The general characteristics and detailed specification of the I²C bus are described in a separate data sheet (serial data buses) in handbook: *ICs for digital systems in radio, audio and video equipment.*

Figure 7. Complete Data Transfer in the Low-Speed Mode

ADDRESSING

Before any data is transmitted on the I²C bus, the device which should respond is addressed first. The addressing is always done with the first byte transmitted after the start procedure.

Slave Address

The clock/calendar acts as a slave receiver or slave transmitter. Therefore, the clock signal SCL is only an input signal, but the data signal SDA is a bidirectional line. The clock calendar slave address is shown in Figure 8.

The subaddress bits A0 and A1 correspond to the two hardware address pins A0 and A1 which allows the device to have 1 of 4 different addresses.

MSB							LSB
1	1	0	1	0	A1	A0	R/$\overline{W}$

DF06250S

Figure 8. Slave Address

Clock/Calendar READ/WRITE Cycles

The I²C bus configuration for different clock/calendar READ and WRITE cycles is shown in Figures 9 and 10.

The write cycle is used to set the time counter, the alarm register and the flags. The transmission of the clock/calendar address is followed by the MODE-POINTER-WORD which contains a CONTROL-nibble (Table 3) and an ADDRESS-nibble (Table 4). The ADDRESS-nibble is valid only if the preceding CONTROL-nibble is set to EXECUTE ADDRESS. The third transmitted word contains the data to be written into the time counter or alarm register.

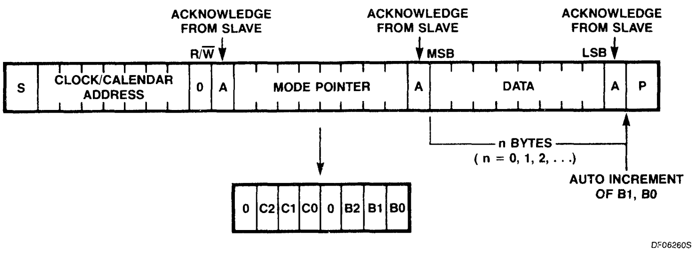


Figure 9. Master Transmitter Transmits to Clock/Calendar Slave Receiver

Table 3. CONTROL-nibble

	C2	C1	C0	FUNCTION
0	0	0	0	Execute address
0	0	0	1	Read control/status flags
0	0	1	0	Reset prescaler, including seconds counter; without carry for minute counter
0	0	1	1	Time adjust, with carry for minute counter[1]
0	1	0	0	Reset NODA flag
0	1	0	1	Set NODA flag
0	1	1	0	Reset COMP flag

NOTE:
1. If the seconds counter is below 30 there is no carry. This causes a time adjustment of max. −30 sec. From the count 30 there is a carry which adjusts the time by max. +30 sec.

At the end of each data word the address bits B1, B0 will be incremented automatically provided the preceding CONTROL-nibble is set to EXECUTE ADDRESS. There is no carry to B2.

Table 5 shows the placement of the BCD upper and lower digits in the DATA byte for writing into the addressed part of the time counter and alarm register, respectively.

Acknowledgement response of the clock calendar as slave receiver is shown in Table 6.

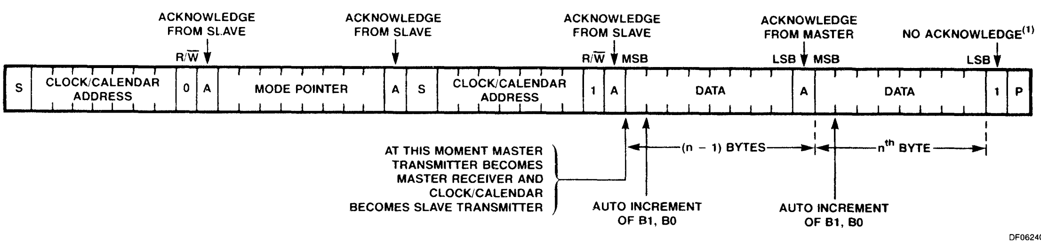
Table 4. ADDRESS-nibble

	B2	B1	B0	ADDRESSED TO:
0	0	0	0	Time counter hours
0	0	0	1	Time counter minutes
0	0	1	0	Time counter days
0	0	1	1	Time counter months
0	1	0	0	Alarm register hours
0	1	0	1	Alarm register minutes
0	1	1	0	Alarm register days
0	1	1	1	Alarm register months

Table 5. Placement of BCD Digits in the DATA Byte

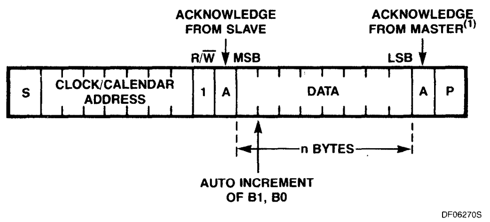
MSB			DATA				LSB	
UPPER DIGIT				LOWER DIGIT				
UD	UC	UB	UA	LD	LC	LB	LA	ADDRESSED TO:
X	X	D	D	D	D	D	D	Hours
X	D	D	D	D	D	D	D	Minutes
X	X	D	D	D	D	D	D	Days
X	X	X	D	D	D	D	D	Months

NOTE:
1. Where "X" is the don't care bit and "D" is the data bit.



NOTE:
The master receiver must signal an end-of-data to the slave transmitter by not generating an acknowledge on the last byte that has been clocked out of the slave.

Figure 10. Master Transmitter Reads Clock/Calendar After Setting Mode Pointer



NOTE:
The master receiver must signal an end of data to the slave transmitter by not generating an acknowledge on the last byte that has been clocked out of the slave.

Figure 11. Master Reads Clock/Calendar Immediately After First Byte

To read the addressed part of the time counter and alarm register, plus information from specified control/status flags, the BCD digits in the DATA byte are organized as shown in Table 7.

The status of the MODE-POINTER-WORD concerning the CONTROL-nibble remains unchanged until a write to MODE POINTER conditon occurs.

Table 6. Slave Receiver Acknowledgement

MODE POINTER								ACKNOWLEDGE ON BYTE		
								Address	Mode pointer	Data
	C2	C1	C0		B2	B1	B0			
0	0	0	0	0	X	X	X	yes	yes	yes
0	0	0	0	1	X	X	X	yes	no	no
0	0	0	1	X	X	X	X	yes	yes	no
0	0	1	0	X	X	X	X	yes	yes	no
0	0	1	1	X	X	X	X	yes	yes	no
0	1	0	0	X	X	X	X	yes	yes	no
0	1	0	1	X	X	X	X	yes	yes	no
0	1	1	0	X	X	X	X	yes	yes	no
0	1	1	1	X	X	X	X	yes	no	no
1	X	X	X	X	X	X	X	yes	no	no

NOTE:
1. Where "X" is the don't care bit.

Table 7. Organization of the BCD Digits in the DATA Byte

MSB		DATA				LSB		ADDRESSED TO:
UPPER DIGIT				LOWER DIGIT				
UD	UC	UB	UA	LD	LC	LB	LA	
0	0	D	D	D	D	D	D	Hours
0	D	D	D	D	D	D	D	Minutes
0	0	D	D	D	D	D	D	Days
0	0	0	D	D	D	D	D	Months
0	0	0	*	**	NODA	COMP	POWF	Control/status flags

NOTES:
1. Where: "D" is the data bit, * = minutes, ** = seconds.

Clock/Calendar with Serial I/O

PCF8573

APPLICATION INFORMATION

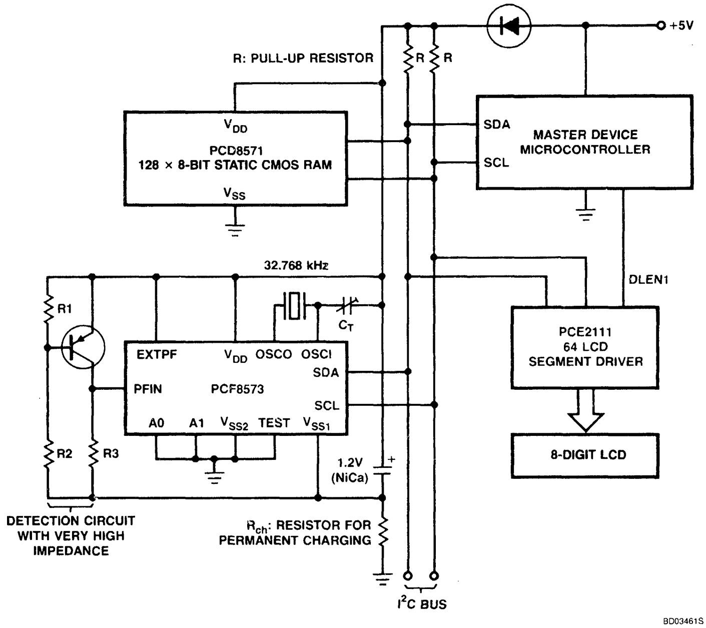


Figure 12. Application Example of the PCF8573 Clock/Calendar

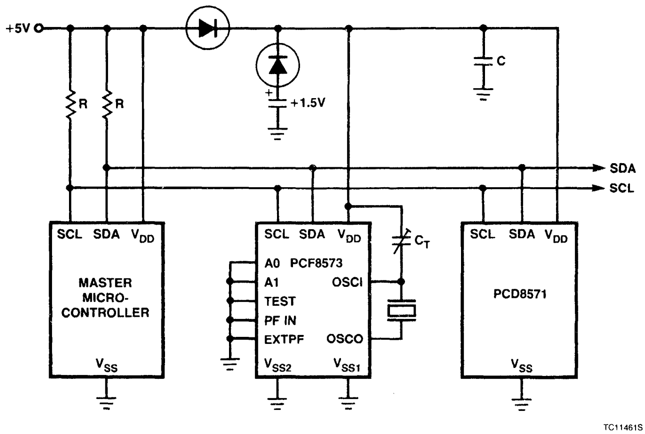


Figure 13. Application Example of the PCF8573 With Common V_{SS1} and V_{SS2} Supply

Signetics

PCF8574/A
8-Bit Remote I/O Expander

Product Specification

Linear Products

DESCRIPTION

The PCF8574 is a single-chip silicon gate CMOS circuit. It provides remote I/O expansion for the MAB8400 and PCF84CXX microcomputer families via the two-line serial bidirectional bus (I^2C). It can also interface microcomputers without a serial interface to the I^2C bus (as a slave function only). The device consists of an 8-bit quasi-bidirectional port and an I^2C interface.

The PCF8574 has low-current consumption and includes latched outputs with high-current drive capability for directly driving LEDs. It also possesses an interrupt line (INT) which is connected to the interrupt logic of the microcomputer on the I^2C bus. By sending an interrupt signal on this line, the remote I/O can inform the microcomputer if there is incoming data on its ports without having to communicate via the I^2C bus. This means that the PCF8574 can remain a simple slave device.

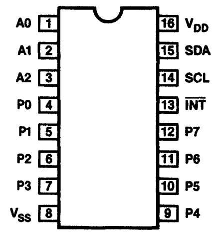
The PCF8574 and the PCF8574A versions differ only in their slave address, as shown in Figure 9.

FEATURES

- **Operating supply voltage: 2.5V to 6V**
- **Low-standby current consumption: max. 10μA**
- **Bidirectional expander**
- **Open-drain interrupt output**
- **8-bit remote I/O port for the I^2C bus**
- **Peripheral for the MAB8400 and PCF8500 microcomputer families**
- **Latched outputs with high-current drive capability for directly driving LEDs**
- **Address by 3 hardware address pins for use of up to 8 devices (up to 16 possible with mask option)**

PIN CONFIGURATION

N, D Packages



TOP VIEW

CD07880S

PIN NO.	SYMBOL	DESCRIPTION
1	A0	Address inputs
2	A1	
3	A2	
4	P0	8-bit quasi-bidirectional I/O ports
5	P1	
6	P2	
7	P3	
8	V_{SS}	
9	P4	8-bit quasi-bidirectional I/O ports
10	P5	
11	P6	
12	P7	
13	$\overline{INT}$	Interrupt output
14	SCL	Serial clock line
15	SDA	Serial data line
16	V_{DD}	Positive supply

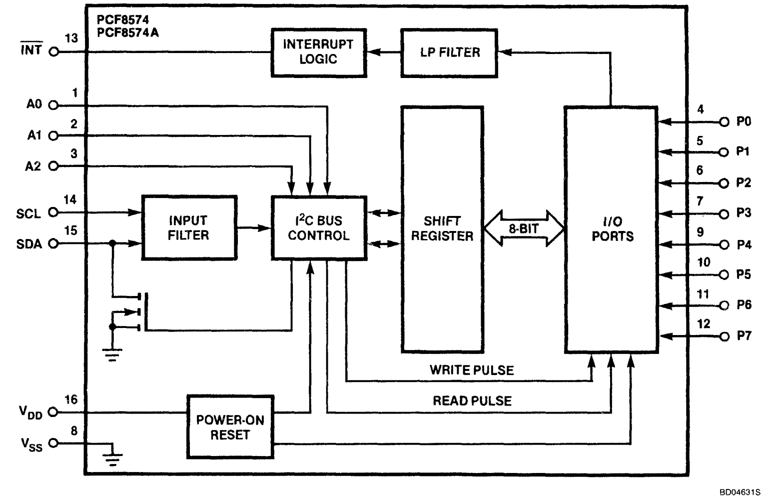
ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
16-Pin Plastic DIP (SOT-38)	-40°C to +85°C	PCF8574PN
16-Pin Plastic DIP (SOT-38)	-40°C to +85°C	PCF8574APN
16-Pin Plastic SO package (SO16L; SOT-162A)	-40°C to +85°C	PCF8574TD
16-Pin Plastic SO package (SO16L; SOT-162A)	-40°C to +85°C	PCF8574ATD

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{DD}	Supply voltage range	-0.5 to +7	V
V_I	Input voltage range (any pin)	V_{SS} - 0.5 to V_{DD} + 0.5	V
$\pm I_I$	DC current into any input	20	mA
$\pm I_O$	DC current into any output	25	mA
$\pm I_{DD}$; I_{SS}	V_{DD} or V_{SS} current	100	mA
P_D	Total power dissipation	400	mW
P_O	Power dissipation per output	100	mW
T_{STG}	Storage temperature range	-65 to +150	°C
T_A	Operating ambient temperature range	-40 to +85	°C

BLOCK DIAGRAM



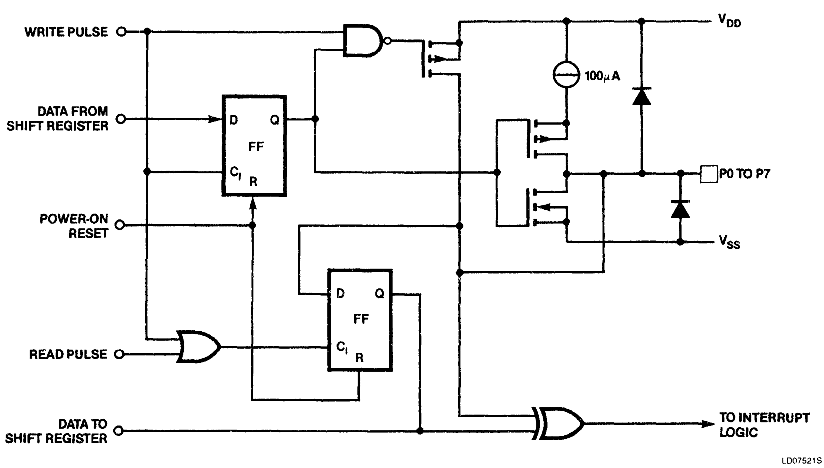


Figure 1. Simplified Schematic Diagram of Each Port

DC ELECTRICAL CHARACTERISTICS V_{DD} = 2.5 to 6V; V_{SS} = 0V; T_A = −40°C to +85°C, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supply (Pin 16)					
V_{DD}	Supply voltage	2.5		6	V
 I_{DD} I_{DDO}	Supply current at V_{DD} = 6V; no load, inputs at V_{DD}, V_{SS} operating standby		 40 1.5	 100 10	 μA μA
V_{REF}	Power-on reset voltage level[1]		1.3	2.4	V
Input SCL; input/output SDA (Pins 14; 15)					
V_{IL}	Input voltage Low	−0.5V		$0.3V_{DD}$	V
V_{IH}	Input voltage High	$0.7V_{DD}$		V_{DD} + 0.5	V
I_{OL}	Output current Low at V_{OL} = 0.4V	3			mA
$\|I_{LI}\|$	Input/output leakage current			100	nA
f_{SCL}	Clock frequency (See Figure 6)			100	kHz
t_S	Tolerable spike width at SCL and SDA input			100	ns
C_I	Input capacitance (SCL, SDA) at $V_I = V_{SS}$			7	pF
I/O ports (Pins 4 to 7; 9 to 12)					
V_{IL}	Input voltage Low	−0.5V		$0.3V_{DD}$	V
V_{IH}	Input voltage High	$0.7V_{DD}$		V_{DD} + 0.5V	V
$\pm I_{IHL}$	Maximum allowed input current through protection diode at $V_I \geqslant V_{DD}$ or $\leqslant V_{SS}$			400	μA
I_{OL}	Output current Low at V_{OL} = 1V; V_{DD} = 2.5V	10	30		mA
$-I_{OH}$	Output current High at $V_{OH} = V_{SS}$ (current source only)	30	100	300	μA
$-I_{OHt}$	Transient pull-up current High during acknowledge (see Figure 14) at $V_{OH} = V_{SS}$		0.5		mA
$C_{I/O}$	Input/output capacitance			10	pF
Port timing; $C_L \leqslant$ 100pF (see Figures 10 and 11)					
t_{PV}	Output data valid			4	μs
t_{PS}	Input data setup	0			μs
t_{PH}	Input data hold	4			μs
Interrupt $\overline{INT}$ (Pin 13)					
I_{OL}	Output current Low at V_{OL} = 0.4V	1.6			mA
$\|I_{OH}\|$	Output current High at $V_{OH} = V_{DD}$			100	nA
$\overline{INT}$ timing; $C_L \leqslant$ 100pF (see Figure 11)					
t_{IV} t_{IR}	Input data valid Reset delay			4 4	μs μs
Select inputs A0, A1, A2 (Pins 1 to 3)					
V_{IH}	Input voltage Low	−0.5V		$0.3V_{DD}$	V
V_{IH}	Input voltage High	$0.7V_{DD}$		V_{DD} + 0.5V	V
$\|I_{LI}\|$	Input leakage current at $V_I = V_{DD}$ or V_{SS}			100	nA

NOTE:
1. The power-on reset circuit resets the I^2C bus logic with $V_{DD} < V_{REF}$ and sets all ports to logic 1 (input mode with current source to V_{DD}).

CHARACTERISTICS OF THE I²C BUS

The I²C bus is for 2-way, 2-line communication between different ICs or modules. The two lines are a serial data line (SDA) and a serial clock line (SCL). Both lines must be connected to a positive supply via a pull-up resistor when connected to the output stages of a device. Data transfer may be initiated only when the bus is not busy.

Bit Transfer

One data bit is transferred during each clock pulse. The data on the SDA line must remain stable during the High period of the clock pulse, as changes in the data line at this time will be interpreted as control signals.

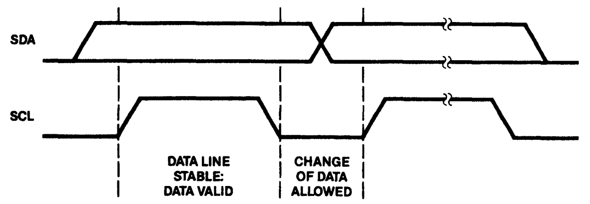


Figure 2. Bit Transfer

Start and Stop Conditions

Both data and clock lines remain High when the bus is not busy. A High-to-Low transition of the data line while the clock is High is defined as the start condition (S). A Low-to-High transition of the data line while the clock is High is defined as the stop condition (P).

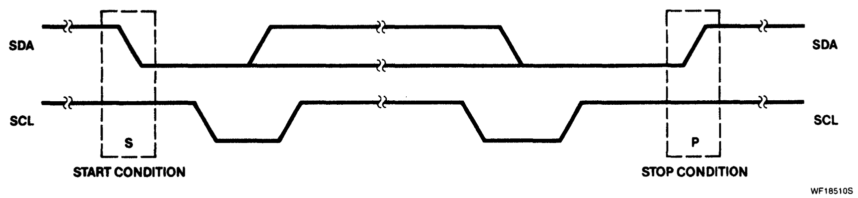


Figure 3. Definition of Start and Stop Conditions

System Configuration

A device generating a message is a "transmitter"; a device receiving a message is the "receiver". The device that controls the message is the "master" and the devices which are controlled by the master are the "slaves".

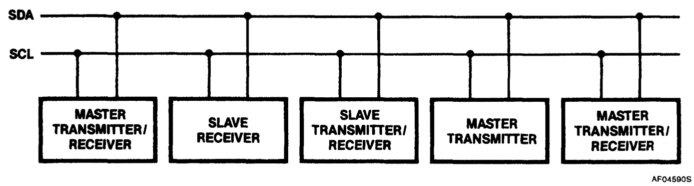


Figure 4. System Configuration

Acknowledge

The number of data bytes transferred between the start and stop conditions from transmitter to receiver is not limited. Each byte of eight bits is followed by one acknowledge bit. The acknowledge bit is a High level put on the bus by the transmitter whereas the master generates an extra acknowledge related clock pulse. A slave receiver which is addressed must generate an acknowledge after the reception of each byte. Also, a master must generate an acknowledge after the reception of each byte that has been clocked out of the slave transmitter. The device that acknowledges has to pull down the SDA line during the acknowledge clock pulse, so that the SDA line is stable Low during the High period of the acknowledgement. Related clock pulse, setup and hold times must be taken into account. A master receiver must signal an end of data to the transmitter by *not* generating an acknowledge on the last byte that has been clocked out of the slave. In this event the transmitter must leave the data line High to enable the master to generate a stop condition.

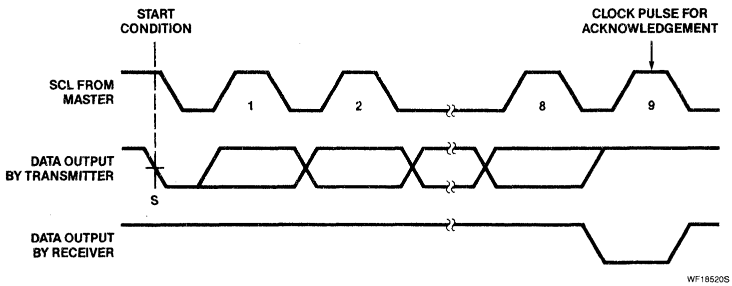
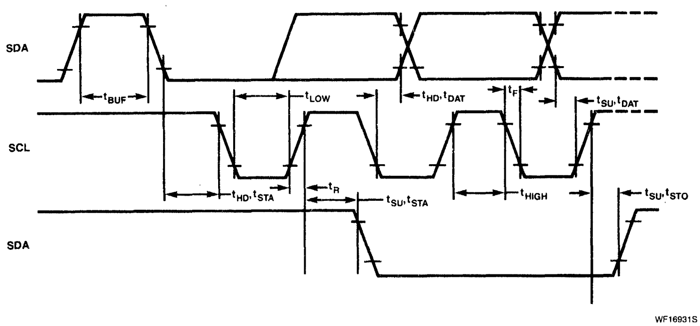


Figure 5. Acknowledgement on the I²C Bus

Timing Specifications

Masters generate a bus clock with a maximum frequency of 100kHz. Detailed timing is shown in Figure 6.

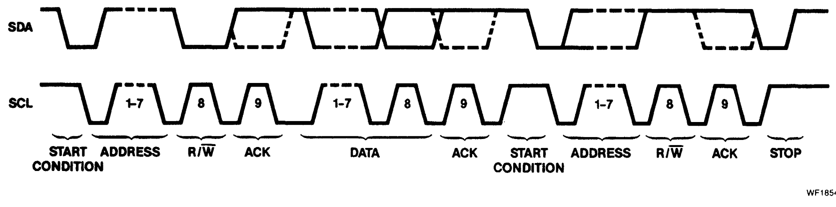


Where:

t_{BUF}	$t \geq t_{LOWmin}$	The minimum time the bus must be free before a new transmission can start
$t_{HD}; t_{STA}$	$t \geq t_{HIGHmin}$	Start condition hold time
t_{LOWmin}	$4.7\mu s$	Clock Low period
$t_{HIGHmin}$	$4\mu s$	Clock High period
$t_{SU}; t_{STA}$	$t \geq t_{LOWmin}$	Start condition set-up time; only valid for repeated start code
$t_{HD}; t_{DAT}$	$t \geq 0\mu s$	Data hold time
$t_{SU}; t_{DAT}$	$t \geq 250ns$	Data setup time
t_R	$t \leq 1\mu s$	Rise time of both the SDA and SCL line
t_F	$t \leq 300ns$	Fall time of both the SDA and SCL line
$t_{SU}; t_{STO}$	$t \geq t_{LOWmin}$	Stop condition setup time

NOTE:
All the values refer to V_{IH} and V_{IL} levels with a voltage swing of V_{SS} to V_{DD}.

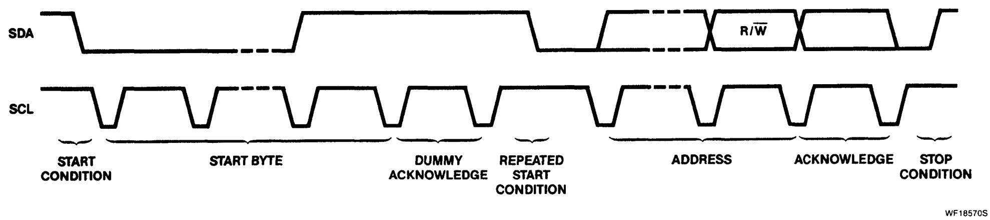
Figure 6. Timing of the I²C Bus



Where:

Clock t_{LOWmin}	4.7µs
$t_{HIGHmin}$	4µs
The dashed line is the acknowledgement of the receiver	
Mark-to-space ratio	1:1 (Low-to-High)
Maximum number of bytes	Unrestricted
Premature termination of transfer	Allowed by generation of STOP condition
Acknowledge clock bit	Must be provided by the master

Figure 7. Complete Data Transfer



Where:

Clock t_{LOWmin}	130µs ± 25µs
$t_{HIGHmin}$	390µs ± 25µs
Mark-to-space ratio	1 : 3 (Low-to-High)
Start byte	0000 0001
Maximum number of bytes	6
Premature termination of transfer	not allowed
Acknowledge clock bit	must be provided by master

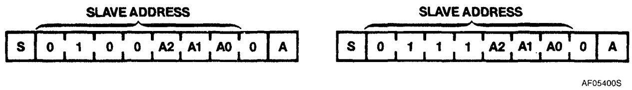
Figure 8. Complete Data Transfer

FUNCTIONAL DESCRIPTION

Addressing (See Figures 9, 10 and 11)

Each bit of the PCF8574 I/O port can be independently used as an input or an output.

Input data is transferred from the port to the microcomputer by the READ mode. Output data is transmitted to the port by the WRITE mode.



a. PCF8574 b. PCF8574A

Figure 9. PCF8574 and PCF8574A Slave Address

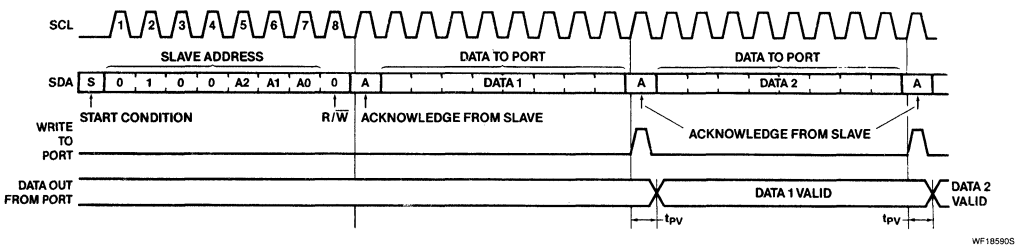
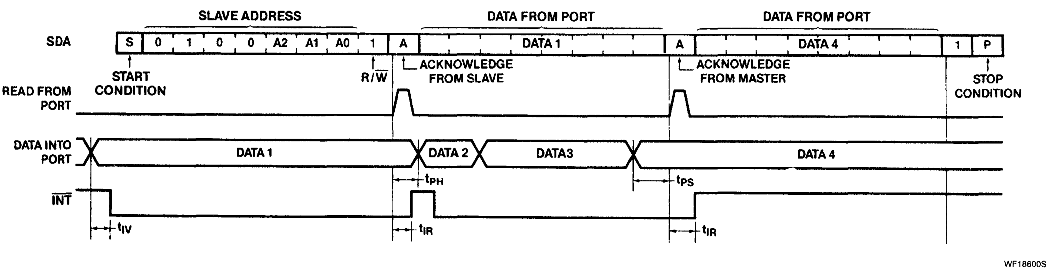


Figure 10. WRITE Mode (Output Port)



NOTE:
A Low-to-High transition of SDA while SCL is High is defined as the stop condition (P). Transfer of data can be stopped at any moment by a stop condition. When this occurs, data present at the last acknowledge phase is valid (output mode). Input data is lost.

Figure 11. READ Mode (Input Port)

Interrupt (See Figures 12 and 13)

The PCF8574/A provides an open-drain output ($\overline{INT}$) which can be fed to a corresponding input of the microcomputer. This gives these chips a type of master function which can initiate an action elsewhere in the system.

An interrupt is generated by any rising or falling edge of the port inputs in the input mode. After time t_{IV} the signal $\overline{INT}$ is valid.

Resetting and reactivating the interrupt circuit is achieved when data on the port is changed to the original setting or data is read from or written to the port which has generated the interrupt. Resetting occurs as follows:

- In the READ mode at the acknowledge bit after the rising edge of the SCL signal.
- In the WRITE mode at the acknowledge bit after the High-to-Low transition of the SCL signal.

Each change of the ports after the resettings will be detected and after the next rising clock edge, will be transmitted as $\overline{INT}$.

Reading from or writing to another device does not affect the interrupt circuit.

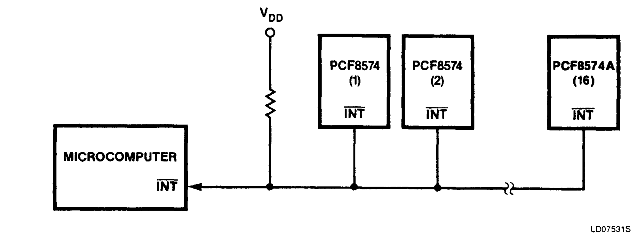


Figure 12. Application of Multiple PCF8574s With Interrupt

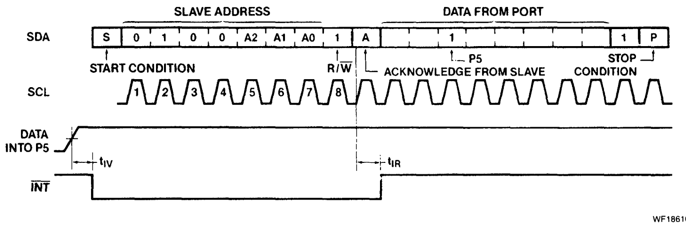


Figure 13. Interrupt Generated by a Change of Input to Port P5

Quasi-Bidirectional I/O Ports (See Figure 14)

A quasi-bidirectional port can be used as an input or output without the use of a control signal for data direction. The bit designated as an input must first be loaded with a logic 1. In this mode only a current source to V_{DD} is active. An additional strong pull-up to V_{DD} allows fast rising edges into heavily loaded outputs. These devices turn on when an output changes from Low-to-High, and are switched off by the negative edge of SCL. SCL should not remain High when a short-circuit to V_{SS} is allowed (input mode).

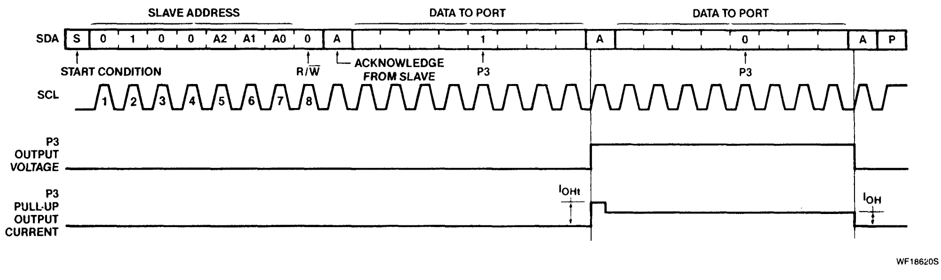


Figure 14. Transient Pull-Up Current I_{OHt} While P3 Changes From Low-to-High and Back to Low

Signetics

PCF8582A
Static CMOS EEPROM (256 × 8-bit)

Preliminary Specification

Linear Products

DESCRIPTION

The PCF8582A is 2K-bit 5V electrically erasable programmable read only memory (EEPROM) organized as 256 by 8 bits. It is designed in a floating-gate CMOS technology.

As data bytes are received and transmitted via the serial I^2C bus, an 8-pin DIP package is sufficient. Up to eight PCF8582A devices may be connected to the I^2C bus.

Chip select is accomplished by three address inputs.

FEATURES

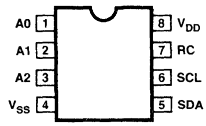
- **Non-volatile storage of 2K-bit organized as 256 × 8**
- **Only one power supply required (5V)**
- **On-chip voltage multiplier for erase/write**
- **Serial input/output bus (I^2C)**
- **Automatic word address incrementing**
- **Low power consumption**
- **One point erase/write timer**
- **Power-on reset**
- **10,000 erase/write cycles per byte**
- **10 years non-volatile data retention**
- **Infinite number of read cycles**
- **Pin-and address-compatible to PCF8570 and PCF8571**

APPLICATIONS

- **Telephony**
- **Radio and television**
- **General purpose**

PIN CONFIGURATION

N Package

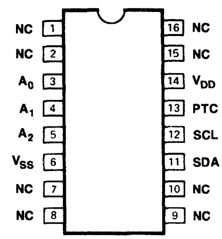


TOP VIEW

CD12090S

PIN NO.	SYMBOL	DESCRIPTION
1	A0	
2	A1	Address inputs/test
3	A2	Mode select
4	V_{SS}	Ground
5	SDA	I^2C bus lines
6	SCL	
7	PTC	Programming time control
8	V_{DD}	Positive supply

D Package



CD07001S

PIN NO.	SYMBOL	DESCRIPTION
1	NC	Not connected
2	NC	Not Connected
3	A_0	
4	A_1	Address inputs/test mode select
5	A_2	
6	V_{SS}	Logic ground
7	NC	Not connected
8	NC	Not connected
9	NC	Not connected
10	NC	Not connected
11	SDA	I^2C bus data input/output
12	SCL	I^2C bus clock input/output
13	PTC	Programming time control
14	V_{DD}	Positive supply voltage
15	NC	Not connected
16	NC	Not connected

ORDERING INFORMATION

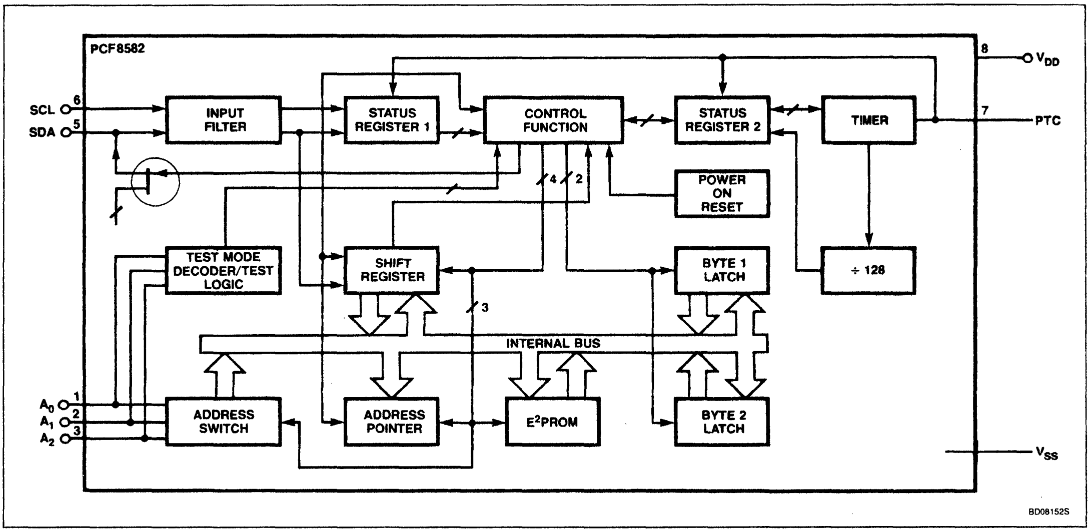
DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
8-Pin Plastic DIP (SOT-97A)	−40°C to +85°C	PCF8582APN
16-Pin Plastic SO (SO16L; SOT-162A)	−40°C to +85°C	PCF8582ATD

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{DD}	Supply voltage	-0.3 to 7	V
V_{IN}	Input voltage, at Pin 4, (input impedance 500Ω)	$V_{SS}-0.8$ to $V_{DD}+0.8$	V
T_A	Operating temperature range	-40 to +85	°C
T_{STG}	Storage temperature range	-65 to +150	°C
I_I	Current into any input pin	1	mA
I_O	Output current	10	mA

Static CMOS EEPROM | PCF8582A

BLOCK DIAGRAM



a. b.

Figure 1. RC Circuit Connections to PCF8582APN and PCF8582ATD When Using the Internal Oscillator

DC AND AC ELECTRICAL CHARACTERISTICS V_{DD} = 5V; V_{SS} = 0V; T_A = −40°C to +85°C, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
V_{DD}	Operating supply voltage	4.5	5	5.5	V
I_{DDR}	Operating supply current, READ ($V_{DD\ MAX}$, f_{SLC} = 100kHz)			0.4	mA
I_{DDW}	Operating supply current, WRITE/ERASE			2.0	mA
I_{DDO}	Standby supply current ($V_{DD\ MAX}$)			10	μA
Input PTC					
V_{IHP}	Input voltage High	V_{DD} − 0.3			V
V_{ILP}	Input voltage Low			V_{SS} + 0.3	V

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) V_{DD} = 5V; V_{SS} = 0V; T_A = −40°C to +85°C, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Input SCL					
 V_{IL} V_{IH} V	Input/output SDA: Input voltage LOW Input voltage HIGH Output voltage LOW	 −0.3 3		 1.5 V_{DD} + 0.8	 V V
V_{OL}	(I_{OL} = 3mA, V_{DD} = 4.5V)			0.4	V
I_{OH}	Output leakage current HIGH (V_{OH} = V_{DD})			1	μA
$\pm I_{IN}$	Input leakage current (A0, A1, A2, SCL)[1]			1	μA
f_{SCL}	Clock frequency	0		100	kHz
C_I	Input capacitance (SCL, SDA)			7	pF
t_I	Noise suppression time constant at SCL and SDA input	0.25	0.5	1	μs
t_{BUF}	Time the bus must be free before a new transmission can start	4.7			μs
t_{HD}, t_{STA}	Hold time start condition. After this period the first clock pulse is generated	4			μs
t_{LOW}	The LOW period of the clock	4.7			μs
t_{HIGH}	The HIGH period of the clock	4			μs
t_{SU}, t_{STA}	Setup time for start condition (only relevent for a repeated start condition)	4.7			μs
 t_{HD}, t_{DAT} t_{HD}, t_{DAT}	Hold time DATA for: CBUS compatible masters I^2C devices[2]	 5 0			 μs μs
t_{SU}, t_{DAT}	Setup time DATA	250			ns
t_R	Rise time for both SDA and SCL lines			1	μs
t_F	Fall time for both SDA and SCL lines			300	ns
t_{SU}, t_{STO}	Setup time for stop condition	4.7			μs
Erase/write timer constant					
$C_{E/W}$	Erase/write timing capacitor for erase/write cycle of 30ns[3]		3.3		nF
$R_{E/W}$	Erase/write cycle timing resistor[4]		56.0		kΩ
Programming frequency using external clock					
f_P	Frequency	2.57		12.85	kHz
t_{Low}	Period Low	10.0			μs
t_{High}	Period High	10.0			μs
t_R	Rise time			300	ns
t_F	Fall time			300	ns
t_D	Delay time	0			ns
t_S	Data retention time (T_A = 55°C)	10			years

NOTES:
1. Selection of the chip address is done by connecting the A0, A1, and A2 inputs either to V_{SS} or V_{DD}.
2. A transmitter must internally provide a hold time to bridge the undefined region (maximum 300ns) of the falling edge of SCL.
3. Maximum tolerance ±10% using internal oscillator.
4. Maximum tolerance ±5% using internal oscillator.

FUNCTIONAL DESCRIPTION

Characteristics of the I²C Bus

The I²C bus is intended for communication between different ICs. The serial bus consists of two bidirectional lines, one for data signals (SDA), and one for clock signals (SCL). Both the SDA and the SCL lines must be connected to a positive supply voltage via a pull-up resistor.

The following protocol has been defined:

Data transfer may be initiated only when the bus is not busy.

During data transfer, the data line must remain stable whenever the clock line is HIGH. Changes in the data line while the clock line is HIGH will be interpreted as control signals.

Accordingly, the following bus conditions have been defined:

Bus Not Busy — both data and clock lines remain HIGH.

Start Data Transfer — a change in the state of the data line, from HIGH to LOW, while the clock is HIGH defines the start condition.

Stop Data Transfer — a change in the state of the data line, from LOW to HIGH, defines the stop condition.

Data Valid — the state of the data line represents valid data when, after a start condition, the data line is stable for the duration of the HIGH period of the clock signal. The data on the line may be changed during the LOW period of the clock signal. There is one clock pulse per bit of data.

Each data transfer is initiated with a start condition and terminated with a stop condition; the number of the data bytes transferred between the start and stop conditions is limited to two bytes in the ERASE/WRITE mode and unlimited in the READ mode. The information is transmitted in bytes and each receiver acknowledges with a ninth bit.

Within the I²C bus specifications a low-speed mode (2kHz clock rate) and a high-speed mode (100kHz clock rate) are defined. The PCF8582A works in both modes. By definition a device that gives out a signal is called a "transmitter," and the device which receives the signal is called a "receiver". The device which controls the signal is called the "master". The devices that are controlled by the master are called "slaves".

Each word of eight bits is followed by one acknowledge bit. This acknowledge bit is a HIGH level put on the bus by the transmitter whereas the master generates an extra acknowledge-related clock puise. A slave receiver which it addresses is obliged to generate an acknowledge after the reception of each byte.

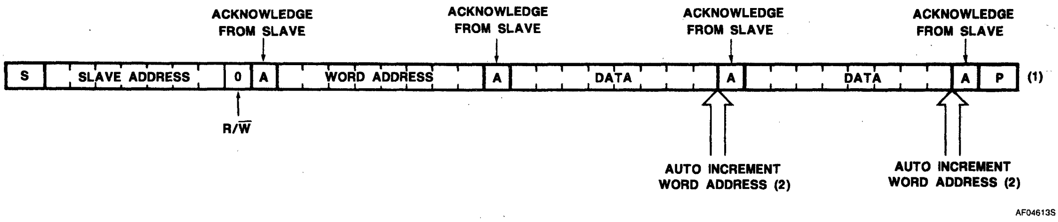
Also, a master receiver must generate an acknowledge after the reception of each byte that has been clocked out of the slave transmitter.

The device that acknowledges has to pull down the SDA line during the acknowledge clock pulse in such a way that the SDA line is stable LOW during the high period of the acknowledge related clock pulse.

Setup and hold times must be taken into account. A master receiver must signal an end-of-data to the slave transmitter by not generating an acknowledge on the last byte that has been clocked out of the slave. In this case the transmitter must leave the data line HIGH to enable the master generation of the stop condition.

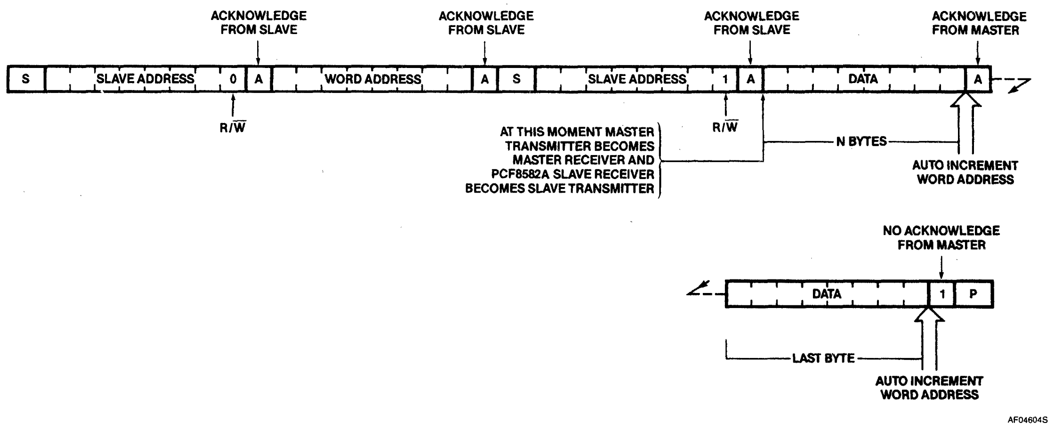
I²C Bus Protocol

The I²C bus configuration for different READ and WRITE cycles of the PCF8582A are shown in Figures 1a and 1b.

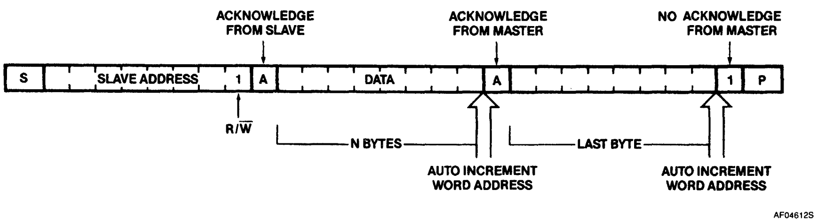


NOTES:
1. After this stop condition the erase/write cycle starts and the bus is free for another transmission; the duration of the erase/write cycle is approximately 30ms if only one byte is written, and 60ms if two bytes are written. During the erase/written cycle the slave receiver does not send an acknowledge bit if addressed via I^2C bus.
2. The second data byte is voluntary. Trying to erase/write more than two bytes is not allowed.

a. Master Transmitter Transmits to PCF8582A Slave Receiver (ERASE/WRITE Mode)



b. Master Reads PCF8582A Slave After Setting Word Address (Write Word Address; READ Data)



The slave address is defined in accordance with the I^2C bus specification as:

1	0	1	0	A2	A1	A0	R/$\overline{W}$

AF05430S

NOTE:
1. The device can be used as read only without the programming clock.

c. Master Reads PCF8582A Slave Immediately After First Byte (READ Mode)[1]

Figure 2

I²C BUS TIMING

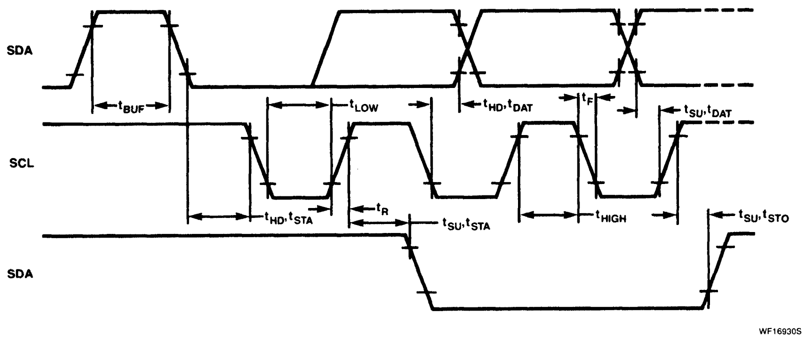


Figure 3. Timing Requirements for the I²C Bus

Signetics

SAB3035
FLL Tuning and Control Circuit

Product Specification

Linear Products

DESCRIPTION

The SAB3035 provides closed-loop digital tuning of TV receivers, with or without AFC, as required. It also controls up to 8 analog functions, 4 general purpose I/O ports, and 4 high-current outputs for tuner band selection.

The IC is used in conjunction with a microcomputer from the MAB8400 family and is controlled via a two-wire, bidirectional I^2C bus.

FEATURES

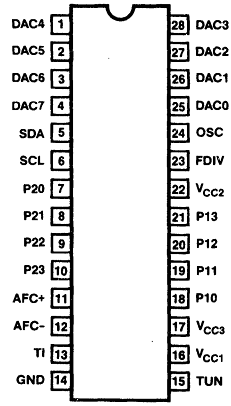
- **Combined analog and digital circuitry minimizes the number of additional interfacing components required**
- **Frequency measurement with resolution of 50kHz**
- **Selectable prescaler divisor of 64 or 256**
- **32V tuning voltage amplifier**
- **4 high-current outputs for direct band selection**
- **8 static digital-to-analog converters (DACs) for control of analog functions**
- **Four general purpose input/output (I/O) ports**
- **Tuning with control of speed and direction**
- **Tuning with or without AFC**
- **Single-pin, 4MHz on-chip oscillator**
- **I^2C bus slave transceiver**

APPLICATIONS

- **Satellite receivers**
- **Television receivers**
- **CATV converters**

PIN CONFIGURATION

N Package



TOP VIEW

CD11950S

PIN NO.	SYMBOL	DESCRIPTION
1	DAC4	Outputs of static DACs
2	DAC5	
3	DAC6	
4	DAC7	
5	SDA	Serial data line } I^2C bus
6	SCL	Serial clock line
7	P20	General purpose input/output ports
8	P21	
9	P22	
10	P23	
11	AFC+	AFC inputs
12	AFC-	
13	TI	Tuning voltage amplifier inverting input
14	GND	Ground
15	TUN	Tuning voltage amplifier output
16	V_{CC1}	+12V supply voltage
17	V_{CC3}	+32V supply for tuning voltage amplifier
18	P10	High-current band-selection output ports
19	P11	
20	P12	
21	P13	
22	V_{CC2}	Positive supply for high-current band-selection output circuits
23	FDIV	Input from prescaler
24	OSC	Crystal oscillator input
25	DAC0	Outputs of static DACs
26	DAC1	
27	DAC2	
28	DAC3	

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
28-Pin Plastic DIP (SOT-117)	-20°C to +70°C	SAB3035N

ABSOLUTE MAXIMUM RATINGS

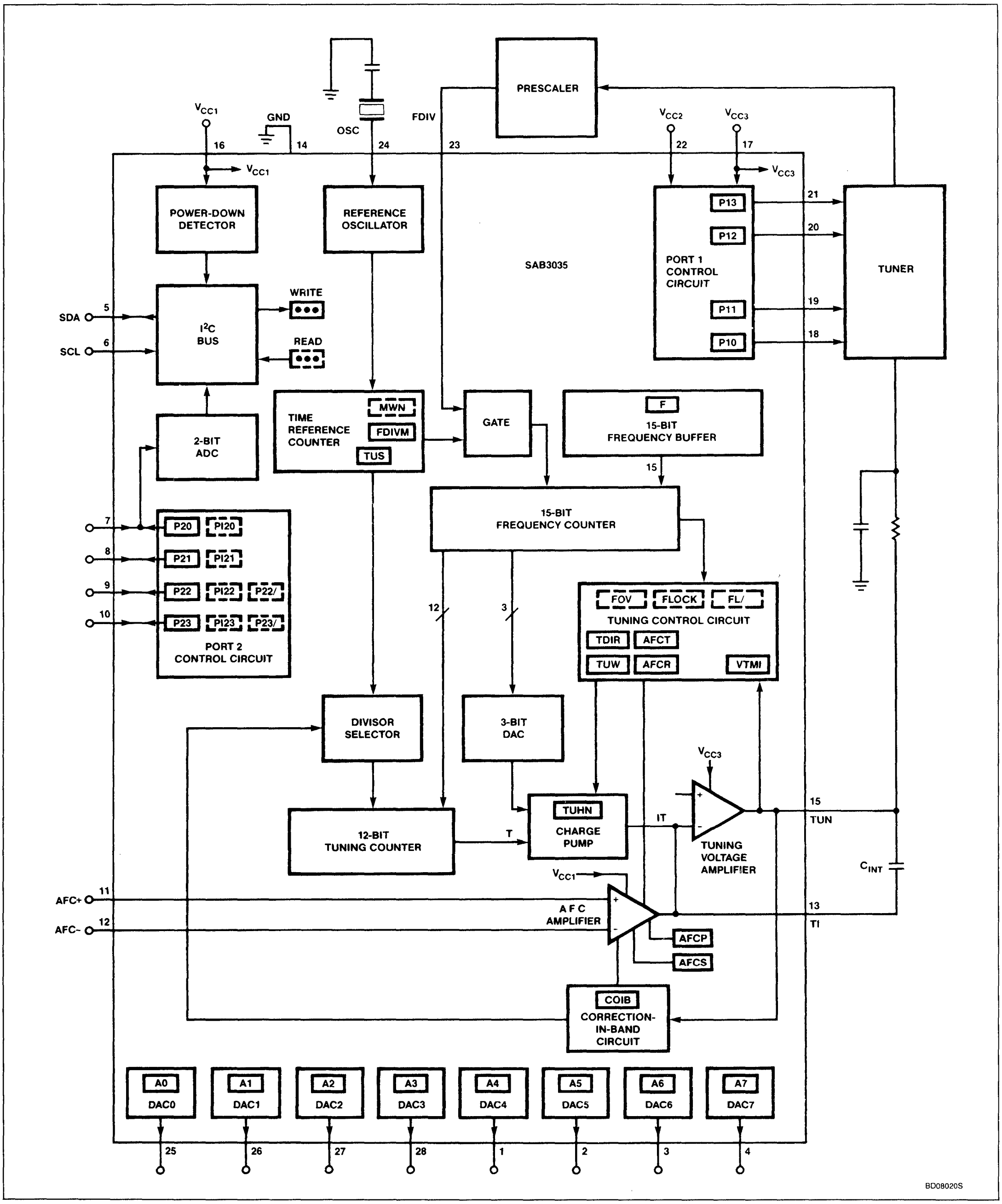
SYMBOL	PARAMETER	RATING	UNIT
	Supply voltage ranges:		
V_{CC1}	(Pin 16)	-0.3 to +18	V
V_{CC2}	(Pin 22)	-0.3 to +18	V
V_{CC3}	(Pin 17)	-0.3 to +36	V
	Input/output voltage ranges:		
V_{SDA}	(Pin 5)	-0.3 to +18	V
V_{SCL}	(Pin 6)	-0.3 to +18	V
V_{CC2X}	(Pins 7 to 10)	-0.3 to +18	V
$V_{AFC+,\ AFC-}$	(Pins 11 and 12)	-0.3 to V_{CC1} [1]	V
V_{TI}	(Pin 13)	-0.3 to V_{CC1} [2]	V
V_{TUN}	(Pin 15)	-0.3 to V_{CC3} [1]	V
V_{CC1X}	(Pins 18 to 21)	-0.3 to V_{CC2} [2]	V
V_{FDIV}	(Pin 23)	-0.3 to V_{CC1} [1]	V
V_{OSC}	(Pin 24)	-0.3 to +5	V
V_{DACX}	(Pins 1 to 4 and 25 to 28)	-0.3 to V_{CC1} [1]	V
P_{TOT}	Total power dissipation	1000	mW
T_{STG}	Storage temperature range	-65 to +150	°C
T_A	Operating ambient temperature range	-20 to +70	°C

NOTES:
1. Pin voltage may exceed supply voltage if current is limited to 10mA.
2. Pin voltage must not exceed 18V but may exceed V_{CC2} if current is limited to 200mA.

FLL Tuning and Control Circuit

SAB3035

BLOCK DIAGRAM



FLL Tuning and Control Circuit SAB3035

DC AND AC ELECTRICAL CHARACTERISTICS $T_A = 25°C$; V_{CC1}, V_{CC2}, V_{CC3} at typical voltages, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
V_{CC1} V_{CC2} V_{CC3}	Supply voltages	10.5 4.7 30	12 13 32	13.5 16 35	V V V
I_{CC1} I_{CC2} I_{CC3}	Supply currents (no outputs loaded)	20 0 0.2	32 0.6	50 0.1 2	mA mA mA
I_{CC2A} I_{CC3A}	Additional supply currents (A) See Note 1	−2 0.2		I_{OHP1X} 2	mA mA
P_{TOT}	Total power dissipation		400		mW
T_A	Operating ambient temperature	−20		+70	°C
I²C bus inputs/outputs SDA input (Pin 5) SCL input (Pin 6)					
V_{IH}	Input voltage HIGH[2]	3		$V_{CC1} - 1$	V
V_{IL}	Input voltage LOW	−0.3		1.5	V
I_{IH}	Input current HIGH[2]			10	μA
I_{IL}	Input current LOW[2]			10	μA
	SDA output (Pin 5, open-collector)				
V_{OL}	Output voltage LOW at $I_{OL} = 3mA$			0.4	V
I_{OL}	Maximum output sink current		5		mA
Open-collector I/O ports P20, P21, P22, P23 (Pins 7 to 10, open-collector)					
V_{IH}	Input voltage HIGH	2		16	V
V_{IL}	Input voltage LOW	−0.3		0.8	V
I_{IH}	Input current HIGH			25	μA
$-I_{IL}$	Input current LOW			25	μA
V_{OL}	Output voltage LOW at $I_{OL} = 2mA$			0.4	V
I_{OL}	Maximum output sink current		4		mA
AFC amplifier Inputs AFC+, AFC− (Pins 11, 12)					
  g00 g01 g10 g11	Transconductance for input voltages up to 1V differential: AFCS1 AFCS2 0 0 0 1 1 0 1 1	 100 15 30 60	 250 25 50 100	 800 35 70 140	 nA/V μA/V μA/V μA/V
ΔM_g	Tolerance of transconductance multiplying factor (2, 4, or 8) when correction-in-band is used	−20		+20	%
V_{IOFF}	Input offset voltage	−75		+75	mV
V_{COM}	Common-mode input voltage	3		$V_{CC1} - 2.5$	V
CMRR	Common-mode rejection ratio		50		dB
PSRR	Power supply (V_{CC1}) rejection ratio		50		dB
I_I	Input current			500	nA

FLL Tuning and Control Circuit SAB3035

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $T_A = 25°C$; V_{CC1}, V_{CC2}, V_{CC3} at typical voltage, unless otherwise specified.

SYMBOL	PARAMETER		LIMITS Min	Typ	Max	UNIT
Tuning voltage amplifier Input TI, output TUN (Pins 13, 15)						
V_{TUN}	Maximum output voltage at $I_{LOAD} = \pm 2.5mA$		$V_{CC3} - 1.6$		$V_{CC3} - 0.4$	V
	Minimum output voltage at $I_{LOAD} = \pm 2.5mA$:					
	VTMI1	VTMI0				
V_{TM00}	0	0	300		500	mV
V_{TM10}	1	0	450		650	mV
V_{TM11}	1	1	650		900	mV
$-I_{TUNH}$	Maximum output source current		2.5		8	mA
I_{TUNL}	Maximum output sink current			40		mA
I_{TI}	Input bias current		−5		+5	nA
PSRR	Power supply V_{CC3} rejection ratio			60		dB
	Minimum charge IT to tuning voltage amplifier					
	TUHN1	TUHN0				
CH_{00}	0	0	0.4	1	1.7	μA/μs
CH_{01}	0	1	4	8	14	μA/μs
CH_{10}	1	0	15	30	48	μA/μs
CH_{11}	1	1	130	250	370	μA/μs
ΔCH	Tolerance of charge (or ΔV_{TUN}) multiplying factor when COIB and/or TUS are used		−20		+20	%
	Maximum current I into tuning amplifier					
	TUHN1	TUHN0				
I_{T00}	0	0	1.7	3.5	5.1	μA
I_{T01}	0	1	15	29	41	μA
I_{T10}	1	0	65	110	160	μA
I_{T11}	1	1	530	875	1220	μA
Correction-in-band						
ΔV_{CIB}	Tolerance of correction-in-band levels 12V, 18V, and 24V		−15		+15	%
Band-select output ports P10, P11, P12, P13 (Pins 18 to 21)						
V_{OH}	Output voltage HIGH at $-I_{OH} = 50mA$[3]		$V_{CC2} - 0.6$			V
V_{OL}	Output voltage LOW at $I_{OL} = 2mA$				0.4	V
$-I_{OH}$	Maximum output source current[3]			130	200	mA
I_{OL}	Maximum output sink current			5		mA
FDIV input (Pin 23)						
$V_{FDIV\ (P\text{-}P)}$	Input voltage (peak-to-peak value) t_{RISE} and $t_{FALL} \leq 40ns$		0.1		2	V
	Duty cycle		40		60	%
f_{MAX}	Maximum input frequency		14.5			MHz
Z_I	Input impedance			8		kΩ
C_I	Input capacitance			5		pF
OSC input (Pin 24)						
R_X	Crystal resistance at resonance (4MHz)				150	Ω

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $T_A = 25°C$; V_{CC1}, V_{CC2}, V_{CC3} at typical voltage, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
DAC outputs 0 to 7 (Pins 25 to 28 and 1 to 4)					
V_{DH}	Maximum output voltage (no load) at $V_{CC1} = 12V$[4]	10		11.5	V
V_{DL}	Minimum output voltage (no load) at $V_{CC1} = 12V$[4]	0.1		1	V
ΔV_D	Positive value of smallest step (1 least significant bit)	0		350	mV
	Deviation from linearity			0.5	V
Z_O	Output impedance at $I_{LOAD} = \pm 2mA$			70	Ω
$-I_{DH}$	Maximum output source current			6	mA
I_{DL}	Maximum output sink current		8		mA
Power-down reset					
V_{PD}	Maximum supply voltage V_{CC1} at which power-down reset is active	7.5		9.5	V
t_R	V_{CC1} rise time during power-up (up to V_{PD})	5			μs
Voltage level for valid module address					
	Voltage level at P20 (Pin 7) for valid module address as a function of MA1, MA0 MA1 MA0				
V_{VA00}	0 0	−0.3		16	V
V_{VA01}	0 1	−0.3		0.8	V
V_{VA10}	1 0	2.5		$V_{CC1} - 2$	V
V_{VA11}	1 1	$V_{CC1} - 0.3$		V_{CC1}	V

NOTES:
1. For each band-select output which is programmed at logic 1, sourcing a current I_{OHP1X}, the additional supply currents (A) shown must be added to I_{CC2} and I_{CC3}, respectively.
2. If $V_{CC1} < 1V$, the input current is limited to 10μA at input voltages up to 16V.
3. At continuous operation the output current should not exceed 50mA. When the output is short-circuited to ground for several seconds, the device may be damaged.
4. Values are proportional to V_{CC1}.

FUNCTIONAL DESCRIPTION

The SAB3035 is a monolithic computer interface which provides tuning and control functions and operates in conjunction with a microcomputer via an I^2C bus.

Tuning

This is performed using frequency-locked loop digital control. Data corresponding to the required tuner frequency is stored in a 15-bit frequency buffer. The actual tuner frequency, divided by a factor of 256 (or by 64) by a prescaler, is applied via a gate to a 15-bit frequency counter. This input (FDIV) is measured over a period controlled by a time reference counter and is compared with the contents of the frequency buffer. The result of the comparison is used to control the tuning voltage so that the tuner frequency equals the contents of the frequency buffer multiplied by 50kHz within a programmable tuning window (TUW).

The system cycles over a period of 6.4ms (or 2.56ms), controlled by the time reference counter which is clocked by an on-chip 4MHz reference oscillator. Regulation of the tuning voltage is performed by a charge pump frequency-locked loop system. The charge IT flowing into the tuning voltage amplifier is controlled by the tuning counter, 3-bit DAC, and the charge pump circuit. The charge IT is linear with the frequency deviation Δf in steps of 50kHz. For loop gain control, the relationship $\Delta IT/\Delta f$ is programmable. In the normal mode (when control bits TUHNO and TUHN1 are both at logic 1, see OPERATION), the minimum charge IT at $\Delta f = 50kHz$ equals $250\mu A/\mu s$ (typical).

By programming the tuning sensitivity bits (TUS), the charge IT can be doubled up to 6 times. If correction-in-band (COIB) is programmed, the charge can be further doubled up to three times in relation to the tuning voltage level. From this, the maximum charge IT at Δf=50kHz equals $2^6 \times 2^3 \times 250\mu A/\mu s$ (typical).

The maximum tuning current I is $875\mu A$ (typical). In the tuning-hold (TUHN) mode (TUHN is Active-LOW), the tuning current I is reduced and, as a consequence, the charge into the tuning amplifier is also reduced.

An in-lock situation can be detected by reading FLOCK. When the tuner oscillator frequency is within the programmable tuning window (TUW), FLOCK is set to logic 1. If the frequency is also within the programmable AFC hold range (AFCR), which always occurs if AFCR is wider than TUW, control bit AFCT can be set to logic 1. When set, digital tuning will be switched off, AFC will be switched on and FLOCK will stay at logic 1 as long as the oscillator frequency is within AFCR. If the frequency of the tuning oscillator does not remain within AFCR, AFCT is cleared automatically and the system reverts to digital tuning. To be able to detect this situation, the occurrence of positive and negative transitions in the FLOCK signal can be read (FL/1N and FL/0N). AFCT can also be cleared by programming the AFCT bit to logic 0.

The AFC has programmable polarity and transconductance; the latter can be doubled up to 3 times, depending on the tuning voltage level if correction-in-band is used.

The direction of tuning is programmable by using control bits TDIRD (tuning direction down) and TDIRU (tuning direction up). If a tuner enters a region in which oscillation stops, then, providing the prescaler remains stable, no FDIV signal is supplied to CITAC. In this situation the system will tune up, moving away from frequency lock-in. This situation is avoided by setting TDIRD which causes the system to tune down. In normal operation TDIRD must be cleared.

If a tuner stops oscillating and the prescaler becomes unstable by going into self-oscillation at a very high frequency, the system will react by tuning down, moving away from frequency lock-in. To overcome this, the system can be forced to tune up at the lowest sensitivity (TUS) value, by setting TDIRU.

Setting both TDIRD and TDIRU causes the digital tuning to be interrupted and AFC to be switched on.

The minimum tuning voltage which can be generated during digital tuning is programmable by VTMI to prevent the tuner from being driven into an unspecified low tuning voltage region.

Control

For tuner band selection there are four outputs — P10 to P13 — which are capable of sourcing up to 50mA at a voltage drop of less than 600mV with respect to the separate power supply input V_{CC2}

For additional digital control, four open-collector I/O ports — P20 to P23 — are provided. Ports P22 and P23 are capable of detecting positive and negative transitions in their input signals. With the aid of port P20, up to three independent module addresses can be programmed.

Eight 6-bit digital-to-analog converters — DACO to DAC7 — are provided for analog control.

Reset

CITAC goes into the power-down reset mode when V_{CC1} is below 8.5V (typical). In this mode all registers are set to a defined state. Reset can also be programmed.

OPERATION

Write

CITAC is controlled via a bidirectional two-wire I^2C bus. For programming, a module address, R/$\overline{W}$ bit (logic 0), an instruction byte, and a data/control byte, are written into CITAC in the format shown in Figure 1.

	MODULE ADDRESS										INSTRUCTION BYTE									DATA/CONTROL BYTE								
S	1	1	0	0	0	MA 1	MA 0	0	A	I_7	I_6	I_5	I_4	I_3	I_2	I_1	I_0	A	D_7	D_6	D_5	D_4	D_3	D_2	D_1	D_0	A	P
	msb							R/$\overline{W}$		msb									msb									

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Figure 1. I^2C Bus Write Format

The module address bits MA1, MA0 are used to give a 2-bit module address as a function of the voltage at port P20 as shown in Table 1.

Acknowledge (A) is generated by CITAC only when a valid address is received and the device is not in the power-down reset mode (V_{CC1} > 8.5V (typical)).

Tuning

Tuning is controlled by the instruction and data/control bytes as shown in Figure 2.

Frequency

Frequency is set when Bit I_7 of the instruction byte is set to logic 1; the remainder of this byte together with the data/control byte are loaded into the frequency buffer. The frequency to which the tuner oscillator is regulated equals the decimal representation of the 15-bit word multiplied by 50kHz. All frequency bits are set to logic 1 at reset.

Tuning Hold

The TUHN bits are used to decrease the maximum tuning current and, as a consequence, the minimum charge IT (at $\Delta f = 50kHz$) into the tuning amplifier.

Table 1. Valid Module Addresses

MA1	MA0	P20
0	0	Don't care
0	1	GND
1	0	½ V_{CC1}
1	1	V_{CC1}

Table 2. Tuning Current Control

TUHN1	TUHN0	TYP. I_{MAX} (μA)	TYP. IT_{MIN} ($\mu A/\mu s$)	TYP. ΔV_{TUNmin} at $C_{INT} = 1\mu F$ (μV)
0	0	3.5[1]	1[1]	1[1]
0	1	29	8	8
1	0	110	30	30
1	1	875	250	250

NOTE:
1. Values after reset.

During tuning but before lock-in, the highest current value should be selected. After lock-in the current may be reduced to decrease the tuning voltage ripple.

The lowest current value should not be used for tuning due to the input bias current of the tuning voltage amplifier (maximum 5nA). However, it is good practice to program the lowest current value during tuner band switching.

	INSTRUCTION BYTE								DATA/CONTROL BYTE							
	I_7	I_6	I_5	I_4	I_3	I_2	I_1	I_0	D_7	D_6	D_5	D_4	D_3	D_2	D_1	D_0
FREQ	1	F14	F13	F12	F11	F10	F9	F8	F7	F6	F5	F4	F3	F2	F1	F0
TCD0	0	0	1	0	1	0	0	1	AFCT	VTMI0	AFCR1	AFCR0	TUHN1	TUHN0	TUW1	TUW0
TCD1	0	0	1	0	1	0	1	0	VTMI1	COIB1	COIB0	AFCS1	AFCS0	TUS2	TUS1	TUS0
TCD2	0	0	1	0	1	0	1	1	0	0	0	0	AFCP	FDIVM	TDIRD	TDIRU

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Figure 2. Tuning Control Format

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Table 3. Minimum Charge IT as a Function of TUS $\Delta f = 50kHz$; TUHN0 = Logic 1; TUHN1 = Logic 1

TUS2	TUS1	TUS0	TYP. IT_{MIN} (mA/μs)	TYP. ΔV_{TUNmin} at $C_{INT} = 1\mu F$ (mV)
0	0	0	0.25[1]	0.25[1]
0	0	1	0.5	0.5
0	1	0	1	1
0	1	1	2	2
1	0	0	4	4
1	0	1	8	8
1	1	0	16	16

NOTE:
1. Values after reset.

Table 4. Programming Correction-In-Band

COIB1	COIB0	CHARGE MULTIPLYING FACTORS AT TYPICAL VALUES OF V_{TUN} AT:			
		< 12V	12 to 18V	18 to 24V	> 24V
0	0	1[1]	1[1]	1[1]	1[1]
0	1	1	1	1	2
1	0	1	1	2	4
1	1	1	2	4	8

NOTE:
1. Values after reset.

Table 5. Tuning Window Programming

TUW1	TUW0	$\lvert\Delta f\rvert$ (kHz)	TUNING WINDOW (kHz)
0	0	0[1]	0[1]
0	1	50	100
1	0	150	300

NOTE:
1. Values after reset.

Table 6. AFC Hold Range Programming

AFCR1	AFCR0	$\lvert\Delta f\rvert$ (kHz)	AFC HOLD RANGE (kHz)
0	0	0[1]	0[1]
0	1	350	700
1	0	750	1500

NOTE:
1. Values after reset.

Table 7. Transconductance Programming

AFCS1	AFCS0	TYP. TRANSCONDUCTANCE (μA/V)
0	0	0.25[1]
0	1	25
1	0	50
1	1	100

NOTE:
1. Value after reset.

Tuning Sensitivity

To be able to program an optimum loop gain, the charge IT can be programmed by changing T using tuning sensitivity (TUS). Table 3 shows the minimum charge IT obtained by programming the TUS bits at Δf=50kHz; TUHN0 and TUHN1 = logic 1.

Correction-In-Band

This control is used to correct the loop gain of the tuning system to reduce in-band variations due to a non-linear voltage/frequency characteristic of the tuner. Correction-in-band (COIB) controls the time T of the charge equation IT and takes into account the tuning voltage V_{TUN} to give charge multiplying factors as shown in Table 4.

The transconductance multiplying factor of the AFC amplifier is similar when COIB is used, except for the lowest transconductance which is not affected.

Tuning Window

Digital tuning is interrupted and FLOCK is set to logic 1 (in-lock) when the absolute deviation $\lvert\Delta f\rvert$ between the tuner oscillator frequency and the programmed frequency is smaller than the programmed TUW value (see Table 5). If $\lvert\Delta f\rvert$ is up to 50kHz above the values listed in Table 5, it is possible for the system to be locked depending on the phase relationship between FDIV and the reference counter.

AFC

When AFCT is set to logic 1 it will not be cleared and the AFC will remain on as long as $\lvert\Delta f\rvert$ is less than the value programmed for the AFC hold range AFCR (see Table 6). It is possible for the AFC to remain on for values of up to 50kHz more than the programmed value depending on the phase relationship between FDIV and the reference counter.

Transconductance

The transconductance (g) of the AFC amplifier is programmed via the AFC sensitivity bits AFCS as shown in Table 7.

	INSTRUCTION BYTE								DATA/CONTROL BYTE							
	I_7	I_6	I_5	I_4	I_3	I_2	I_1	I_0	D_7	D_6	D_5	D_4	D_3	D_2	D_1	D_0
POD	0	0	1	0	1	0	0	0	P23	P22	P21	P20	P13	P12	P11	P10
DACX	0	0	1	0	0	X2	X1	X0	0	0	AX5	AX4	AX3	AX2	AX1	AX0

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Figure 3. Control Programming

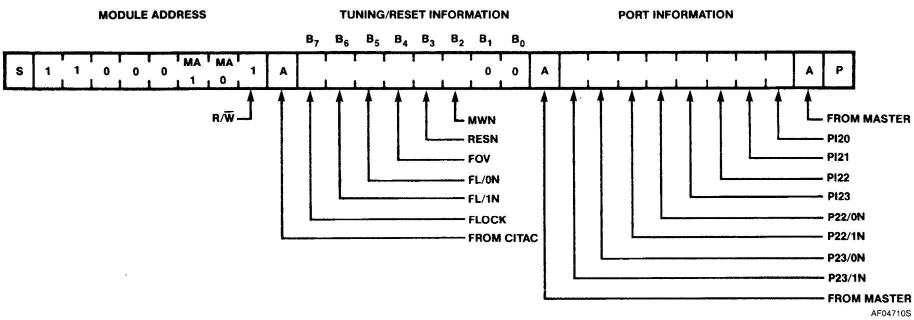

Figure 4. Information Byte Format

AFC Polarity

If a positive differential input voltage is applied to the (switched on) AFC amplifier, the tuning voltage V_{TUN} falls when the AFC polarity bit AFCP is at logic 0 (value after reset). At AFCP = logic 1, V_{TUN} rises.

Minimum Tuning Voltage

Both minimum tuning voltage control bits, VTMI1 and VTMI0, are at logic 0 after reset. Further details are given in the DC Electrical Characteristics table.

Frequency Measuring Window

The frequency measuring window which is programmed must correspond with the division factor of the prescaler in use (see Table 8).

Tuning Direction

Both tuning direction bits, TDIRU (up) and TDIRD (down), are at logic 0 after reset.

Control

The instruction bytes POD (port output data) and DACX (digital-to-analog converter control) are shown in Figure 3, together with the corresponding data/control bytes. Control is implemented as follows:

P13, P12, P11, P10 — Band select outputs. If a logic 1 is programmed on any of the POD bits D_3 to D_0, the relevant output goes HIGH. All outputs are LOW after reset.

P23, P22, P21, P20 — Open-collector I/O ports. If a logic 0 is programmed on any of the POD bits D_7 to D_4, the relevant output is forced LOW. All outputs are at logic 1 after reset (high impedance state).

DACX — Digital-to-analog converters. The digital-to-analog converter selected corresponds to the decimal equivalent of the DACX bits X2, X1, X0. The output voltage of the selected DAC is set by programming the bits AX5 to AX0; the lowest output voltage is programmed with all data AX5 to AX0 at logic 0, or after reset has been activated.

Read

Information is read from CITAC when the R/$\overline{W}$ bit is set to logic 1. An acknowledge must be generated by the master after each data byte to allow transmission to continue. If no acknowledge is generated by the master, the slave (CITAC) stops transmitting. The format of the information bytes is shown in Figure 4.

Table 8. Frequency Measuring Window Programming

FDIVM	PRESCALER DIVISION FACTOR	CYCLE PERIOD (ms)	MEASURING WINDOW (ms)
0	256	6.4[1]	5.12[1]
1	64	2.56	1.28

NOTE:
1. Values after reset.

Tuning/Reset Information Bits

FLOCK — Set to logic 1 when the tuning oscillator frequency is within the programmed tuning window.

FL/1N — Set to logic 0 (Active-LOW) when FLOCK changes from 0 to 1 and is reset to logic 1 automatically after tuning information has been read.

FL/ON — As for FL/1N, but is set to logic 0 when FLOCK changes from 1 to 0.

FOV — Indicates frequency overflow. When the tuner oscillator frequency is too high with respect to the programmed frequency, FOV is at logic 1, and when too low, FOV is at logic 0. FOV is not valid when TDIRU and/or TDIRD are set to logic 1.

RESN — Set to logic 0 (Active-LOW) by a programmed reset or a power-down reset. It is reset to logic 1 automatically after tuning/reset information has been read.

MWN — MWN (frequency measuring window, Active-LOW) is at logic 1 for a period of 1.28ms, during which time the results of frequency measurement are processed. This time is independent of the cycle period. During the remaining time, MWN is at logic 0 and the received frequency is measured.

When slightly different frequencies are programmed repeatedly and AFC is switched on, the received frequency can be measured using FOV and FLOCK. To prevent the frequency counter and frequency buffer being loaded at the same time, frequency should be programmed only during the period of MWN = logic 0.

Port Information Bits

P23/1N, P22/1N — Set to logic 0 (Active-LOW) at a LOW-to-HIGH transition in the input voltage on P23 and P22, respectively. Both are reset to logic 1 after the port information has been read.

P23/0N, P22/0N — As for P23/1N and P22/1N, but are set to logic 0 at a HIGH-to-LOW transition.

PI23, PI21, PI20, PI — Indicate input voltage levels at P23, P22, P21, and P20, respectively. A logic 1 indicates a HIGH input level.

Reset

The programming to reset all registers is shown in Figure 5. Reset is activated only at data byte HEX06. Acknowledge is generated at every byte, provided that CITAC is not in the power-down reset mode. After the general call address byte, transmission of more than one data byte is not allowed.

GENERAL CALL ADDRESS | HEX06

S	0	0	0	0	0	0	0	0	A	0	0	0	0	0	1	1	0	A	P

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Figure 5. Reset Programming

I^2C BUS TIMING (Figure 6)

I^2C bus load conditions are as follows:

4kΩ pull-up resistor to +5V; 200pF capacitor to GND.

All values are referred to $V_{IH} = 3V$ and $V_{IL} = 1.5V$.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
t_{BUF}	Bus free before start	4			μs
t_{SU}, t_{STA}	Start condition setup time	4			μs
t_{HD}, t_{STA}	Start condition hold time	4			μs
t_{LOW}	SCL, SDA LOW period	4			μs
t_{HIGH}	SCL HIGH period	4			μs
t_R	SCL, SDA rise time			1	μs
t_F	SCL, SDA fall time			0.3	μs
t_{SU}, t_{DAT}	Data setup time (write)	1			μs
t_{HD}, t_{DAT}	Data hold time (write)	1			μs
t_{SU}, t_{CAC}	Acknowledge (from CITAC) setup time			2	μs
t_{HD}, t_{CAC}	Acknowledge (from CITAC) hold time	0			μs
t_{SU}, t_{STO}	Stop condition setup time	4			μs
t_{SU}, t_{RDA}	Data setup time (read)			2	μs
t_{HD}, t_{RDA}	Data hold time (read)	0			μs
t_{SU}, t_{MAC}	Acknowledge (from master) setup time	1			μs
t_{HD}, t_{MAC}	Acknowledge (from master) hold time	2			μs

NOTE:

Timings t_{SU}, t_{DAT} and t_{HD}, t_{DAT} deviate from the I^2C bus specification.

After reset has been activated, transmission may only be started after a 50μs delay.

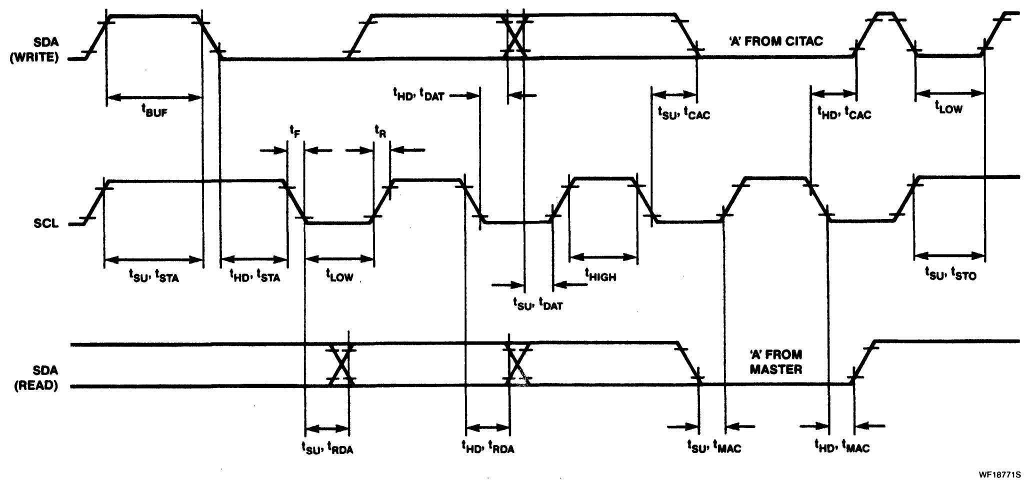

Figure 6. I²C Bus Timing SAB3035

Signetics

AN157
Microcomputer Peripheral IC Tunes and Controls a TV Set

Application Note

Linear Products

Author: K.H. Seidler

The necessity for television set manufacturers to reduce costs, provide more features, simplify tuning and incorporate remote control has led to a need for all-electronic digital tuning and control circuits. Naturally enough, component manufacturers would prefer to meet the need with a dedicated integrated system which they can make in large quantities. This, however, is impractical because it would not allow the set manufacturers to satisfy the widely varying requirements of the TV market. The most suitable system is therefore one controlled by a standard microcomputer (e.g., one from the MAB/SCN8400 family), so that the variants can be accommodated by software. The only additional components that then need to be separately integrated are those required for interfacing and for performing functions that cannot be handled by the microcomputer because of speed, voltage or power consumption considerations. To minimize costs and maximize performance, however, the partitioning of the remaining functions and their allocation to various integrated circuits peripheral to the microcomputer must be carefully considered.

Figure 1 illustrates the control and tuning functions in a basic TV set, and shows how the circuitry is positioned within the cabinet. Some of the functions are concentrated around the microcomputer and mounted close to the front panel to reduce the cost of the wiring to the local keyboard and displays. The tuning and analog controls are on the main chassis. The only link between the microcomputer and the main chassis is a 2-wire bidirectional I^2C bus which allows the microcomputer to read tuning status and other information from the main chassis, and to write data regarding required frequency and analog control settings to the main chassis.

The foregoing considerations have led to the design of the SAB3035 integrated Computer Interface for Tuning and Analog Control (CITAC). The SAB3035 is an I^2C bus-compatible microcomputer peripheral IC for digital frequency-locked loop (FLL) tuning and control of analog functions associated with the TV picture and sound. This is shown in block form in Figure 2. The IC incorporates a frequency synthesizer using the charge pump FLL principle and contains the following circuits:

- 15-bit frequency counter with a resolution of 50kHz
- Charge pump and 30V tuning-voltage amplifier
- AFC amplifier
- Logic circuitry for programming the currents for the charge pump and AFC amplifier
- Four high-current band switches
- Four general-purpose I/O ports for additional control functions
- A single-pin crystal-controlled 4MHz reference oscillator
- Receiving/transmitting logic for the 2-wire I^2C bus
- Eight static DACs for control of analog functions associated with the picture and sound.

FUNCTIONAL DESCRIPTION

I^2C Bus

The SAB3035 is microcomputer-controlled via an asynchronous, Inter-IC (I^2C) bus. The bus is a two-wire, bidirectional serial interconnect which allows integrated circuits to communicate with each other and pass control and data from one IC to another. The communication commences after a start code incorporating an IC address and ceases on receipt of a stop code. Every byte of transmitted data must be acknowledged by the IC that receives it. Data to be read must be clocked out of the IC by the microcomputer. The address byte includes a control bit which defines the read/write mode.

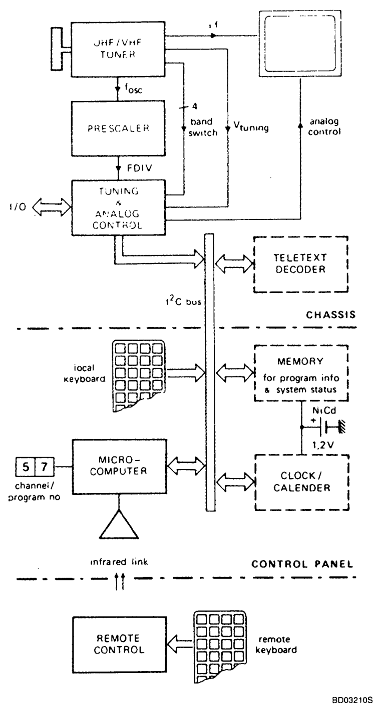

Figure 1. Basic TV Control System

Microcomputer Peripheral IC Tunes and Controls a TV Set

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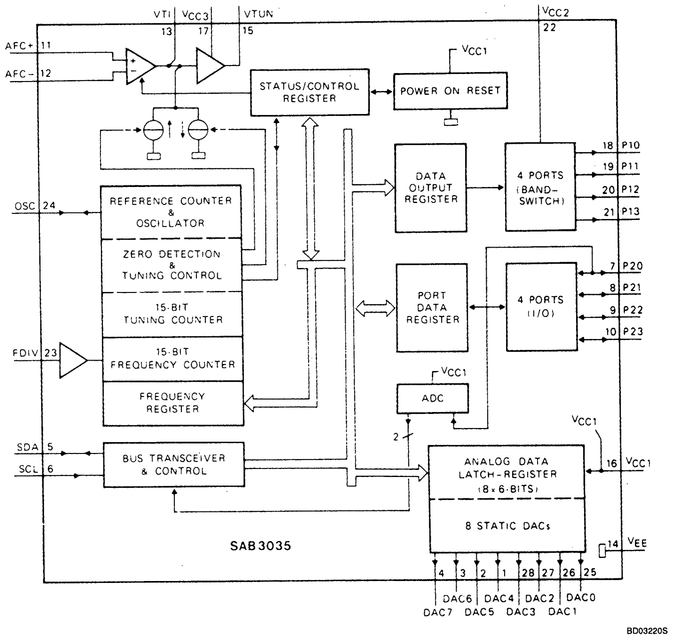


Figure 2. Block Diagram of the SAB3035

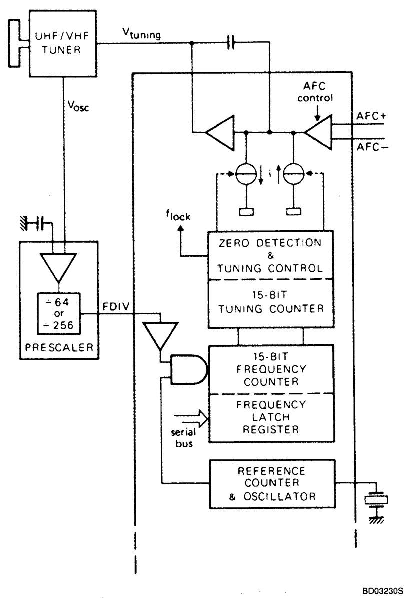


Figure 3. Block Diagram of the SAB3035

Frequency Synthesis Tuning System

Figure 3 is the block diagram of the frequency synthesizing system comprising a frequency-locked loop (FLL) and an external prescaler which divides the frequency of the voltage-controlled local oscillator in the TV tuner by 64 or 256. The tuning section comprises a 15-bit programmable frequency counter, a 15-bit tuning counter, tuning control and zero detection logic, a reference counter and a charge pump followed by a low-pass filter amplifier.

FDIV Input accepts frequency-divided local oscillator signals with a level of more than 100mV and a frequency of up to 16MHz. The frequency measurement period is defined by passing the internally-amplified signal from FDIV through a gate which is controlled by the reference counter. The reference counter is driven by a crystal-controlled oscillator, the low level output of which is almost free from high-order harmonics. This oscillator also generates the internal clock for the IC. Before starting the frequency measurement cycle, the 15 bits of data in the latch register, which represent the required local oscillator frequency, are loaded into the frequency counter. Pulses from the prescaler then decrement the frequency counter for the duration of the measurement period.

Microcomputer Peripheral IC Tunes and Controls a TV Set

AN157

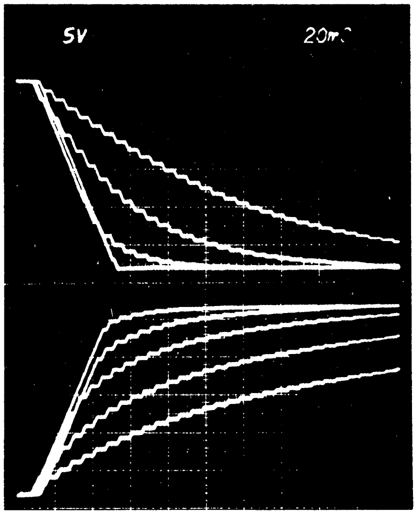
The contents of the frequency counter at the end of the measurement period indicate whether or not the frequency of the local oscillator in the tuner is the same as the desired frequency, which was preloaded into the frequency counter. If the frequency counter contents is zero after the measurement period, a flag (FLOCK), which can be read by the microcomputer serial bus, is set to indicate that the local-oscillator is correctly tuned.

A frequency counter contents of other than zero at the end of the measurement period indicates that the tuner local oscillator frequency is either too high (contents below zero) or too low (contents above zero). If it is too high, an overflow flag which initiates the "tuning down" function is set. To generate the tuning voltage correction, the tuning counter is loaded with the remaining contents of the frequency counter at the end of the measurement period, and then decremented to zero by an internal clock. The duration of the pulse applied to the charge pump is proportional to the time taken to decrement the tuning counter to zero, and therefore also proportional to the tuning error. The frequency correction has a resolution of 50kHz.

The frequency measurement method of tuning used in the SAB3035 can also be easily combined with analog AFC to allow tracking of a drifting transmitter frequency within a limited range. The required tuning mode (with or without AFC) is selected and controlled by software. By not testing some of the LSBs of the contents of the frequency counter, tune-in "windows" of ± 100kHz or ± 200kHz can be defined. The corresponding AFC "windows" are ± 400kHz or ± 800kHz. The SAB3035 also contains the AFC control logic and amplifier. To allow matching to a wide variety of tuners, the tuning loop gain and tuning speed can be adjusted over a wide range. To minimize sound on picture, a "tuning hold" mode is selectable in which the charge pump and AFC currents can be reduced when correct tuning has been achieved.

Bandswitching

The IC also incorporates four 50 mA current sources with outputs at ports P10 to P13 for executing band switching instructions from the microcomputer. Bandswitching data is stored in the data output register. The supply voltage for the current sources is derived from a separate input (V_{CC2}) and is therefore independent of the logic supply voltage (V_{CC1}).



WF16220S

NOTES:
Decreasing frequency (top)
Increasing frequency (bottom)

Figure 4. Using Some of the Selectable Charge Pump Currents for Making 50kHz Tuning Steps in the UHF Band

I/O Ports

There are four bidirectional ports P20 to P23 for additional control signals to or from the TV receiver. Typical examples of these additional controls are stereo/dual sound, search tuning and switching for external video sources. The output data for ports P20 to P23 is stored in the port data register.

Input data must be present during the read cycle. Two of the inputs are edge-triggered. Each input signal transition is stored and can be read by the microcomputer via the serial data bus. The stored data is cleared after each read cycle.

Analog Controls

The SAB3035 includes eight static DACs for controlling analog functions associated with the TV picture and sound (volume, tone, brightness, contrast, color saturation, etc.). External RC networks are not necessary to complete the D/A conversion. The control data for the DACs is derived from the serial data bus and stored in eight 6-bit latch registers. The output voltage range at DAC0 to DAC7 is 0.5V to 10.5V and can be adjusted in 64 increments.

ACKNOWLEDGEMENTS

Special thanks are due to F.A.v.d.Kerkhof and B.Strassenburg for their contributions, and to M.F.Geurts for the electrical design of the SAB3035.

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Microcomputer Peripheral IC Tunes and Controls a TV Set

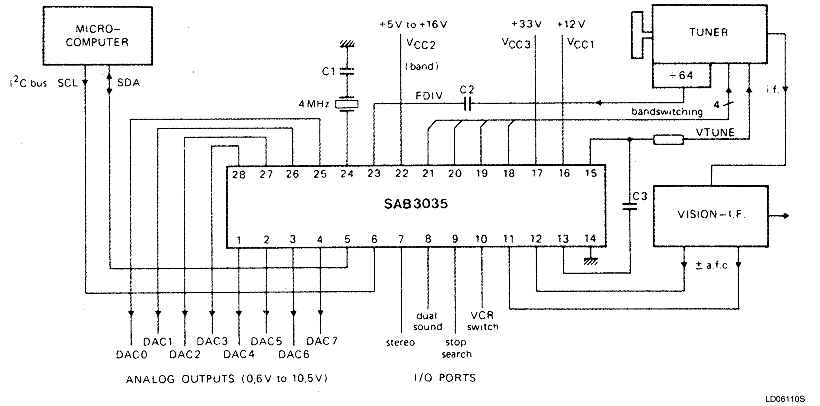


Figure 5. This Typical Example of the SAB3035 in a TV Tuning and Control System Shows how the Peripheral Components Have Been Reduced to Three Capacitors, a Resistor and a 4MHz Crystal

NOTE:
Originally published as Technical Publication 097, Electronic Components and Applications, Vol. 5 No. 2, February, 1983, the Netherlands.

Signetics

SAB3036
FLL Tuning and Control Circuit

Product Specification

Linear Products

DESCRIPTION

The SAB3036 provides closed-loop digital tuning of TV receivers, with or without AFC, as required. It also controls 4 general purpose I/O ports and 4 high-current outputs for tuner band selection.

The IC is used in conjunction with a microcomputer from the MAB8400 family and is controlled via a two-wire, bidirectional I^2C bus.

FEATURES

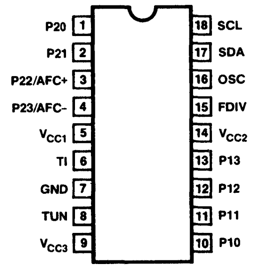
- **Combined analog and digital circuitry minimizes the number of additional interfacing components required**
- **Frequency measurement with resolution of 50kHz**
- **Selectable prescaler divisor of 64 or 256**
- **32V tuning voltage amplifier**
- **4 high-current outputs for direct band selection**
- **Four general purpose input/output (I/O) ports**
- **Tuning with control of speed and direction**
- **Tuning with or without AFC**
- **Single-pin, 4MHz on-chip oscillator**
- **I^2C bus slave transceiver**

APPLICATIONS

- **TV receivers**
- **Satellite receivers**
- **CATV converters**

PIN CONFIGURATION

N Package



TOP VIEW

CD11960S

PIN NO.	SYMBOL	DESCRIPTION
1	P20	General purpose
2	P21	Input/output ports
3	P22/AFC+	General purpose input/output
4	P23/AFC−	ports and AFC inputs
5	V_{CC1}	+12V supply voltage
6	TI	Tuning voltage amplifier inverting input
7	GND	Ground
8	TUN	Tuning voltage amplifier output
9	V_{CC3}	+32V supply for tuning voltage amplifier
10	P10	
11	P11	High-current band-selection
12	P12	output ports
13	P13	
14	V_{CC2}	Positive supply for high-current band-selection output circuits
15	FDIV	Input from prescaler
16	OSC	Crystal oscillator input
17	SDA	Serial data line I^2C bus
18	SCL	Serial clock line

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
18-Pin Plastic DIP (SOT-102HE)	−20°C to +70°C	SAB3036N

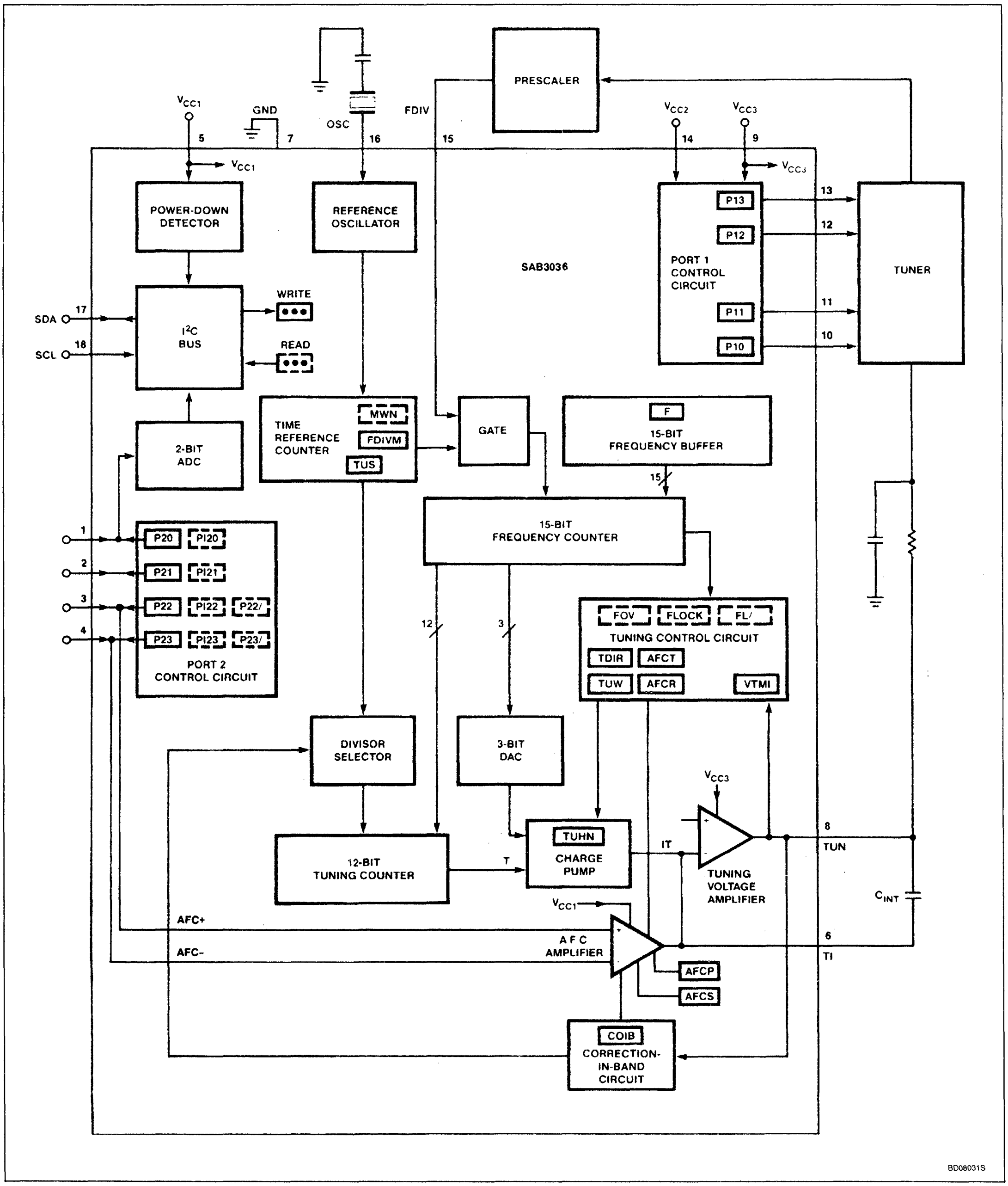
ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
	Supply voltage ranges:		
V_{CC1}	(Pin 5)	−0.3 to +18	V
V_{CC2}	(Pin 14)	−0.3 to +18	V
V_{CC3}	(Pin 9)	−0.3 to +36	V
	Input/output voltage ranges:		
V_{SDA}	(Pin 17)	−0.3 to +18	V
V_{SCL}	(Pin 18)	−0.3 to +18	V
$V_{P20,\ P21}$	(Pins 1 and 2)	−0.3 to +18	V
$V_{P22,\ P23,\ AFC}$	(Pins 3 and 4)	−0.3 to V_{CC1} [1]	V
V_{TI}	(Pin 6)	−0.3 to V_{CC1} [1]	V
V_{TUN}	(Pin 8)	−0.3 to V_{CC3}	V
V_{P1X}	(Pins 10 to 13)	−0.3 to V_{CC2} [2]	V
V_{FDIV}	(Pin 15)	−0.3 to V_{CC1} [1]	V
V_{OSC}	(Pin 16)	−0.3 to +5	V
P_{TOT}	Total power dissipation	1000	mW
T_{STG}	Storage temperature range	−65 to +150	°C
T_A	Operating ambient temperature range	−20 to +70	°C

NOTES:

1. Pin voltage may exceed supply voltage if current is limited to 10mA.
2. Pin voltage must not exceed 18V but may exceed V_{CC2} if current is limited to 200mA.

BLOCK DIAGRAM



DC AND AC ELECTRICAL CHARACTERISTICS $T_A = 25°C$; V_{CC1}, V_{CC2}, V_{CC3} at typical voltages, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
V_{CC1} V_{CC2} V_{CC3}	Supply voltages	10.5 4.7 30	12 13 32	13.5 16 35	V V V
I_{CC1} I_{CC2} I_{CC3}	Supply currents (no outputs loaded)	14 0 0.2	23 0.6	40 0.1 2	mA mA mA
I_{CC2A} I_{CC3A}	Additional supply currents (A)[1]	−2 0.2		I_{OHP1X} 2	mA mA
P_{TOT}	Total power dissipation		300		mW
T_A	Operating ambient temperature	−20		+70	°C
I^2C bus inputs/outputs SDA input (Pin 17); SCL input (Pin 18)					
V_{IH}	Input voltage HIGH[2]	3		$V_{CC1} - 1$	V
V_{IL}	Input voltage LOW	−0.3		1.5	V
I_{IH}	Input current HIGH[2]			10	μA
I_{IL}	Input current LOW[2]			10	μA
	SDA output (Pin 17, open-collector)				
V_{OL}	Output voltage LOW at $I_{OL} = 3mA$			0.4	V
I_{OL}	Maximum output sink current		5		mA
Open-collector I/O ports P20, P21, P22, P23 (Pins 1 to 4, open-collector)					
V_{IH}	Input voltage HIGH (P20, P21)	2		16	V
V_{IH}	Input voltage HIGH (P22, P23) AFC switched off	2		$V_{CC1} - 2$	V
V_{IL}	Input voltage LOW	−0.3		0.8	V
I_{IH}	Input current HIGH			25	μA
$-I_{IL}$	Input current LOW			25	μA
V_{OL}	Output voltage LOW at $I_{OL} = 2mA$			0.4	V
I_{OL}	Maximum output sink current		4		mA
AFC amplifier Inputs AFC+, AFC– (Pins 3, 4)					
  g_{00} g_{01} g_{10} g_{11}	Transconductance for input voltage up to 1V differential: AFCS1 AFCS2 0 0 0 1 1 0 1 1	 100 15 30 60	 250 25 50 100	 800 35 70 140	 nA/V μA/V μA/V μA/V
ΔM_g	Tolerance of transconductance multiplying factor (2, 4 or 8) when correction-in-band is used	−20		+20	%
V_{IOFF}	Input offset voltage	−75		+75	mV
V_{COM}	Common-mode input voltage	3		$V_{CC1} - 2.5$	V
CMRR	Common-mode rejection ratio		50		dB
PSRR	Power supply (V_{CC1}) rejection ratio		50		dB
I_I	Input current (P22 and P23 programmed HIGH)			500	nA

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $T_A = 25°C$; V_{CC1}, V_{CC2}, V_{CC3} at typical voltages, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Tuning voltage amplifier Input TI, output TUN (Pins 6, 8)					
V_{TUN}	Maximum output voltage at $I_{LOAD} = \pm 2.5mA$	$V_{CC3} - 1.6$		$V_{CC3} - 0.4$	V
	Minimum output voltage at $I_{LOAD} = \pm 2.5mA$: VTMI1 VTMI0				
V_{TM00}	0 0	300		500	mV
V_{TM10}	1 0	450		650	mV
V_{TM11}	1 1	650		900	mV
$-I_{TUNH}$	Maximum output source current	2.5		8	mA
I_{TUNL}	Maximum output sink current		40		mA
I_{TI}	Input bias current	-5		+5	nA
PSRR	Power supply (V_{CC3}) rejection ratio		60		dB
	Minimum charge IT to tuning voltage amplifier TUHN1 TUHN0				
CH_{00}	0 0	0.4	1	1.7	μA/μs
CH_{01}	0 1	4	8	14	μA/μs
CH_{10}	1 0	15	30	48	μA/μs
CH_{11}	1 1	130	250	370	μA/μs
ΔCH	Tolerance of charge (or ΔV_{TUN}) multiplying factor when COIB and/or TUS are used	-20		+20	%
	Maximum current I into tuning amplifier TUHN1 TUHN0				
I_{T00}	0 0	1.7	3.5	5.1	μA
I_{T01}	0 1	15	29	41	μA
I_{T10}	1 0	65	110	160	μA
I_{T11}	1 1	530	875	1220	μA
Correction-in-band					
ΔV_{CIB}	Tolerance of correction-in-band levels 12V, 18V and 24V	-15		+15	%
Band-select output ports P10, P11, P12, P13 (Pins 10 to 13)					
V_{OH}	Output voltage HIGH at $-I_{OH} = 50mA$[3]	$V_{CC2} - 0.6$			V
V_{OL}	Output voltage LOW at $I_{OL} = 2mA$			0.4	V
$-I_{OH}$	Maximum output source current[3]		130	200	mA
I_{OL}	Maximum output sink current		5		mA
FDIV input (Pin 15)					
$V_{FDIV\ (P-P)}$	Input voltage (peak-to-peak value) (t_{RISE} and $t_{FALL} \leq 40ns$)	0.1		2	V
	Duty cycle	40		60	%
f_{MAX}	Maximum input frequency	16			MHz
Z_I	Input impedance		8		kΩ
C_I	Input capacitance		5		pF

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $T_A = 25°C$; V_{CC1}, V_{CC2}, V_{CC3} at typical voltages, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
OSC input (Pin 24)					
R_X	Crystal resistance at resonance (4MHz)			150	Ω
Power-down reset					
V_{PD}	Maximum supply voltage V_{CC1} at which power-down reset is active	7.5		9.5	V
t_R	V_{CC1} rise time during power-up (up to V_{PD})	5			μs
Voltage level for valid module address					
	Voltage level at P20 (Pin 1) for valid module address as a function of MA1, MA0 MA1 MA0				
V_{VA00}	0 0	−0.3		16	V
V_{VA01}	0 1	−0.3		0.8	V
V_{VA10}	1 0	2.5		$V_{CC1} - 2$	V
V_{VA11}	1 1	$V_{CC1} - 0.3$		V_{CC1}	V

NOTES:
1. For each band-select output which is programmed at logic 1, sourcing a current I_{OHP1X}, the additional supply currents (A) shown must be added to I_{CC2} and I_{CC3}, respectively.
2. If $V_{CC1} < 1V$, the input current is limited to 10μA at input voltages up to 16V.
3. At continuous operation the output current should not exceed 50mA. When the output is short-circuited to ground for several seconds the device may be damaged.
4. Values are proportional to V_{CC1}.

FUNCTIONAL DESCRIPTION

The SAB3036 is a monolithic computer interface which provides tuning and control functions and operates in conjunction with a microcomputer via an I^2C bus.

Tuning

This is performed using frequency-locked loop digital control. Data corresponding to the required tuner frequency is stored in a 15-bit frequency buffer. The actual tuner frequency, divided by a factor of 256 (or by 64) by a prescaler, is applied via a gate to a 15-bit frequency counter. This input (FDIV) is measured over a period controlled by a time reference counter and is compared with the contents of the frequency buffer. The result of the comparison is used to control the tuning voltage so that the tuner frequency equals the contents of the frequency buffer multiplied by 50kHz within a programmable tuning window (TUW).

The system cycles over a period of 6.4ms (or 2.56ms), controlled by the time reference counter which is clocked by an on-chip 4MHz reference oscillator. Regulation of the tuning voltage is performed by a charge pump frequency-locked loop system. The charge IT flowing into the tuning voltage amplifier is controlled by the tuning counter, 3-bit DAC and the charge pump circuit. The charge IT is linear with the frequency deviation Δf in steps of 50kHz. For loop gain control, the relationship $\Delta IT/\Delta f$ is programmable. In the normal mode (when control bits TUHN0 and TUHN1 are both at logic 1, see OPERATION), the minimum charge IT at $\Delta f = 50kHz$ equals $250\mu A$ μs (typical).

By programming the tuning sensitivity bits (TUS), the charge IT can be doubled up to 6 times. If correction-in-band (COIB) is programmed, the charge can be further doubled up to three times in relation to the tuning voltage level. From this, the maximum charge IT at $\Delta f = 50kHz$ equals $2^6 \times 2^3 \times 250\mu A\mu s$ (typical).

The maximum tuning current I is $875\mu A$ (typical). In the tuning-hold (TUHN) mode (TUHN is Active-LOW), the tuning current I is reduced and as a consequence the charge into the tuning amplifier is also reduced.

An in-lock situation can be detected by reading FLOCK. When the tuner oscillator frequency is within the programmable tuning window (TUW), FLOCK is set to logic 1. If the frequency is also within the programmable AFC hold range (AFCR), which always occurs if AFCR is wider than TUW, control bit AFCT can be set to logic 1. When set, digital tuning will be switched off, AFC will be switched on and FLOCK will stay at logic 1 as long as the oscillator frequency is within AFCR. If the frequency of the tuning oscillator does not remain within AFCR, AFCT is cleared automatically and the system reverts to digital tuning. To be able to detect this situation, the occurrence of positive and negative transitions in the FLOCK signal can be read (FL/1N and FL/0N). AFCT can also be cleared by programming the AFCT bit to logic 0.

The AFC has programmable polarity and transconductance; the latter can be doubled up to 3 times, depending on the tuning voltage level if correction-in-band is used.

The direction of tuning is programmable by using control bits TDIRD (tuning direction down) and TDIRU (tuning direction up). If a tuner enters a region in which oscillation stops, then, providing the prescaler remains stable, no FDIV signal is supplied to CITAC. In this situation the system will tune up, moving away from frequency lock-in. This situation is avoided by setting TDIRD which causes the system to tune down. In normal operation TDIRD must be cleared.

If a tuner stops oscillating and the prescaler becomes unstable by going into self-oscillation at a very high frequency, the system will react by tuning down, moving away from frequency lock-in. To overcome this, the system can be forced to tune up at the lowest sensitivity (TUS) value, by setting TDIRU.

Setting both TDIRD and TDIRU causes the digital tuning to be interrupted and AFC to be switched on.

The minimum tuning voltage which can be generated during digital tuning is programmable by VTMI to prevent the tuner being driven into an unspecified low tuning voltage region.

Control

For tuner band selection there are four outputs — P10 to P13 — which are capable of sourcing up to 50mA at a voltage drop of less than 600mV with respect to the separate power supply input V_{CC2}.

For additional digital control, four open-collector I/O ports — P20 to P23 — are provided. Ports P22 and P23 are capable of detecting positive and negative transitions in their input signals and are connected with the AFC+ and AFC– inputs, respectively. The AFC amplifier must be switched off when P22 and/or P23 are used. When AFC is used, P22 and P23 must be programmed HIGH (high impedance state). With the aid of port P20, up to three independent module addresses can be programmed.

Reset

CITAC goes into the power-down reset mode when V_{CC1} is below 8.5V (typical). In this mode all registers are set to a defined state. Reset can also be programmed.

OPERATION

Write

CITAC is controlled via a bidirectional two-wire I^2C bus. For programming, a module address, R/$\overline{W}$ bit (logic 0), an instruction byte and a data/control byte are written into CITAC in the format shown in Figure 1.

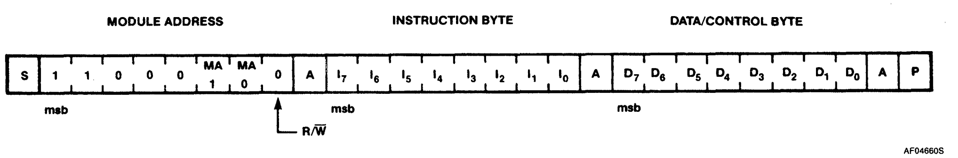


Figure 1. I^2C Bus Write Format

The module address bits MA1, MA0 are used to give a 2-bit module address as a function of the voltage at port P20 as shown in Table 1.

Acknowledge (A) is generated by CITAC only when a valid address is received and the device is not in the power-down reset mode ($V_{CC1} > 8.5V$ (typical)).

Tuning

Tuning is controlled by the instruction and data/control bytes as shown in Figure 2.

Frequency

Frequency is set when Bit I_7 of the instruction byte is set to logic 1; the remainder of this byte together with the data/control byte are loaded into the frequency buffer. The frequency to which the tuner oscillator is regulated equals the decimal representation of the 15-bit word multiplied by 50kHz. All frequency bits are set to logic 1 at reset.

Tuning Hold

The TUHN bits are used to decrease the maximum tuning current and, as a consequence, the minimum charge IT (at $\Delta f = 50kHz$) into the tuning amplifier.

During tuning but before lock-in, the highest current value should be selected. After lock-in the current may be reduced to decrease the tuning voltage ripple.

The lowest current value should not be used for tuning due to the input bias current of the tuning voltage amplifier (maximum 5nA). However, it is good practice to program the lowest current value during tuner band switching.

Tuning Sensitivity

To be able to program an optimum loop gain, the charge IT can be programmed by changing T using tuning sensitivity (TUS). Table 3 shows the minimum charge IT obtained by programming the TUS bits at $\Delta f = 50kHz$; TUHN0 and TUHN1 = logic 1.

Table 1. Valid Module Addresses

MA1	MA0	P20
0	0	Don't care
0	1	GND
1	0	½ V_{CC1}
1	1	V_{CC1}

Table 2. Tuning Current Control

TUHN1	TUHN0	TYP. I_{MAX} (μA)	TYP. IT_{MIN} ($\mu A/\mu s$)	TYP. ΔV_{TUNmin} at $C_{INT} = 1\mu F$ (μV)
0	0	3.5[1]	1[1]	1[1]
0	1	29	8	8
1	0	110	30	30
1	1	875	250	250

NOTE:
1. Values after reset.

Table 3. Minimum Charge IT as a Function of TUS $\Delta f = 50kHz$; TUHN0 = Logic 1; TUHN1 = Logic 1

TUS2	TUS1	TUS0	TYP. IT_{MIN} ($mA/\mu s$)	TYP. ΔV_{TUNmin} at $C_{INT} = 1\mu F$ (mV)
0	0	0	0.25[1]	0.25[1]
0	0	1	0.5	0.5
0	1	0	1	1
0	1	1	2	2
1	0	0	4	4
1	0	1	8	8
1	1	0	16	16

NOTE:
1. Values after reset.

INSTRUCTION BYTE

	I_7	I_6	I_5	I_4	I_3	I_2	I_1	I_0
FREQ	1	F14	F13	F12	F11	F10	F9	F8
TCD0	0	0	1	0	1	0	0	1
TCD1	0	0	1	0	1	0	1	0
TCD2	0	0	1	0	1	0	1	1

DATA/CONTROL BYTE

	D_7	D_6	D_5	D_4	D_3	D_2	D_1	D_0
FREQ	F7	F6	F5	F4	F3	F2	F1	F0
TCD0	AFCT	VTMI0	AFCR1	AFCR0	TUHN1	TUHN0	TUW1	TUW0
TCD1	VTMI1	COIB1	COIB0	AFCS1	AFCS0	TUS2	TUS1	TUS0
TCD2	0	0	0	0	AFCP	FDIVM	TDIRD	TDIRU

AF04670S

Figure 2. Tuning Control Format

Correction-In-Band
This control is used to correct the loop gain of the tuning system to reduce in-band variations due to a non-linear voltage/frequency characteristic of the tuner. Correction-in-band (COIB) controls the time T of the charge equation IT and takes into account the tuning voltage V_{TUN} to give charge multiplying factors as shown in Table 4.

The transconductance multiplying factor of the AFC amplifier is similar when COIB is used, except for the lowest transconductance which is not affected.

Tuning Window
Digital tuning is interrupted and FLOCK is set to logic 1 (in-lock) when the absolute deviation $|\Delta f|$ between the tuner oscillator frequency and the programmed frequency is smaller than the programmed TUW value (see Table 5). If $|\Delta f|$ is up to 50kHz above the values listed in Table 5, it is possible for the system to be locked depending on the phase relationship between FDIV and the reference counter.

AFC
When AFCT is set to logic 1 it will not be cleared and the AFC will remain on as long as $|\Delta f|$ is less than the value programmed for the AFC hold range AFCR (see Table 6). It is possible for the AFC to remain on for values of up to 50kHz more than the programmed value depending on the phase relationship between FDIV and the reference counter.

Transconductance
The transconductance (g) of the AFC amplifier is programmed via the AFC sensitivity bits AFCS as shown in Table 7.

AFC Polarity
If a positive differential input voltage is applied to the (switched on) AFC amplifier, the tuning voltage V_{TUN} falls when the AFC polarity bit AFCP is at logic 0 (value after reset). At AFCP = logic 1, V_{TUN} rises.

Minimum Tuning Voltage
Both minimum tuning voltage control bits, VTMI1 and VTMI0, are at logic 0 after reset. Further details are given in CHARACTERISTICS.

Frequency Measuring Window
The frequency measuring window which is programmed must correspond with the division factor of the prescaler in use (see Table 8).

Tuning Direction
Both tuning direction bits, TDIRU (up) and TDIRD (down), are at logic 0 after reset.

Table 4. Programming Correction-In-Band

COIB1	COIB0	CHARGE MULTIPLYING FACTORS AT TYPICAL VALUES OF V_{TUN} AT:			
		< 12V	12 to 18V	18 to 24V	> 24V
0	0	1[1]	1[1]	1[1]	1[1]
0	1	1	1	1	2
1	0	1	1	2	4
1	1	1	2	4	8

NOTE:
1. Values after reset.

Table 5. Tuning Window Programming

TUW1	TUW0	$|\Delta f|$ (kHz)	TUNING WINDOW (kHz)
0	0	0[1]	0[1]
0	1	50	100
1	0	150	300

NOTE:
1. Values after reset.

Table 6. AFC Hold Range Programming

AFCR1	AFCR0	$|\Delta f|$ (kHz)	AFC HOLD RANGE (kHz)
0	0	0[1]	0[1]
0	1	350	700
1	0	750	1500

NOTE:
1. Values after reset.

Table 7. Transconductance Programming

AFCS1	AFCS0	TYP. TRANSCONDUCTANCE (μA/V)
0	0	0.25[1]
0	1	25
1	0	50
1	1	100

NOTE:
1. Value after reset.

Table 8. Frequency Measuring Window Programming

FDIVM	PRESCALER DIVISION FACTOR	CYCLE PERIOD (ms)	MEASURING WINDOW (ms)
0	256	6.4[1]	5.12[1]
1	64	2.56	1.28

NOTE:
1. Values after reset.

Control

The instruction byte POD (port output data) is shown in Figure 3, together with the corresponding data/control byte. Control is implemented as follows:

P13, P12, P11, P10 — Band select outputs. If a logic 1 is programmed on any of the POD bits D_3 to D_0, the relevant output goes HIGH. All outputs are LOW after reset.

P23, P22, P21, P20 — Open-collector I/O ports. If a logic 0 is programmed on any of the POD bits D_7 to D_4, the relevant output is forced LOW. All outputs are at logic 1 after reset (high impedance state).

Read

Information is read from CITAC when the R/$\overline{W}$ bit is set to logic 1. An acknowledge must be generated by the master after each data byte to allow transmission to continue. If no acknowledge is generated by the master the slave (CITAC) stops transmitting. The format of the information bytes is shown in Figure 4.

Tuning/Reset Information Bits

FLOCK — Set to logic 1 when the tuning oscillator frequency is within the programmed tuning window.

FL/1N — Set to logic 0 (Active-LOW) when FLOCK changes from 0 to 1 and is reset to logic 1 automatically after tuning information has been read.

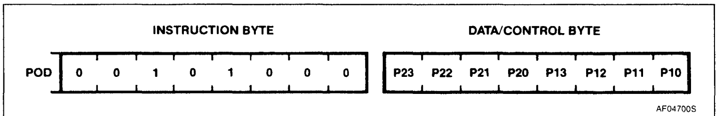


Figure 3. Control Programming

FL/ON — As for FL/1N but is set to logic 0 when FLOCK changes from 1 to 0.

FOV — Indicates frequency overflow. When the tuner oscillator frequency is too high with respect to the programmed frequency, FOV is at logic 1, and when too low, FOV is at logic 0. FOV is not valid when TDIRU and/or TDIRD are set to logic 1.

RESN — Set to logic 0 (Active-LOW) by a programmed reset or a power-down reset. It is reset to logic 1 automatically after tuning/reset information has been read.

MWN — MWN (frequency measuring window, Active-LOW) is at logic 1 for a period of 1.28ms, during which time the results of frequency measurement are processed. This time is independent of the cycle period. During the remaining time, MWN is at logic 0 and the received frequency is measured.

When slightly different frequencies are programmed repeatedly and AFC is switched on, the received frequency can be measured using FOV and FLOCK. To prevent the frequency counter and frequency buffer being loaded at the same time, frequency should be programmed only during the period of MWN = logic 0.

Port Information Bits

P23/1N, P22/1N — Set to logic 0 (Active-LOW) at a LOW-to-HIGH transition in the input voltage on P23 and P22, respectively. Both are reset to logic 1 after the port information has been read.

P23/0N, P22/0N — As for P23/1N and P22/1N but are set to logic 0 at a HIGH-to-LOW transition.

P123, P122, P121, P120 — Indicate input voltage levels at P23, P22, P21 and P20, respectively. A logic 1 indicates a HIGH input level.

Reset

The programming to reset all registers is shown in Figure 5. Reset is activated only at data byte HEX06. Acknowledge is generated at every byte, provided that CITAC is not in the power-down-reset mode. After the general call address byte, transmission of more than one data byte is not allowed.

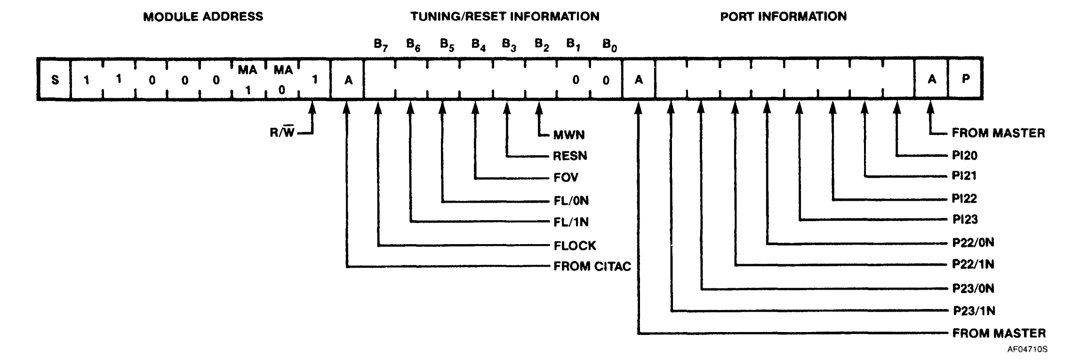


Figure 4. Information Byte Format

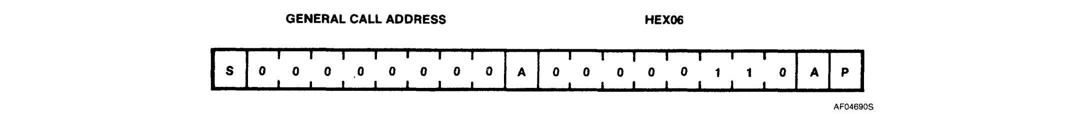


Figure 5. Reset Programming

FLL Tuning and Control Circuit SAB3036

I^2C Bus Timing

I^2C bus load conditions are as follows:

4kΩ pull-up resistor to +5V; 200pF capacitor to GND.

All values are referred to V_{IH} = 3V and V_{IL} = 1.5V.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
t_{BUF}	Bus free before start	4			μs
t_{SU}, t_{STA}	Start condition setup time	4			μs
t_{HD}, t_{STA}	Start condition hold time	4			μs
t_{LOW}	SCL, SDA LOW period	4			μs
t_{HIGH}	SCL HIGH period	4			μs
t_R	SCL, SDA rise time			1	μs
t_F	SCL, SDA fall time			0.3	μs
t_{SU}, t_{DAT}	Data setup time (write)	1			μs
t_{HD}, t_{DAT}	Data hold time (write)	1			μs
t_{SU}, t_{CAC}	Acknowledge (from CITAC) setup time			2	μs
t_{HD}, t_{CAC}	Acknowledge (from CITAC) hold time	0			μs
t_{SU}, t_{STO}	Stop condition setup time	4			μs
t_{SU}, t_{RDA}	Data setup time (read)			2	μs
t_{HD}, t_{RDA}	Data hold time (read)	0			μs
t_{SU}, t_{MAC}	Acknowledge (from master) setup time	1			μs
t_{HD}, t_{MAC}	Acknowledge (from master) hold time	2			μs

NOTE:
1. Timings t_{SU}, t_{DAT} and t_{HD}, t_{DAT} deviate from the I^2C bus specification.
   After reset has been activated, transmission may only be started after a 50μs delay.

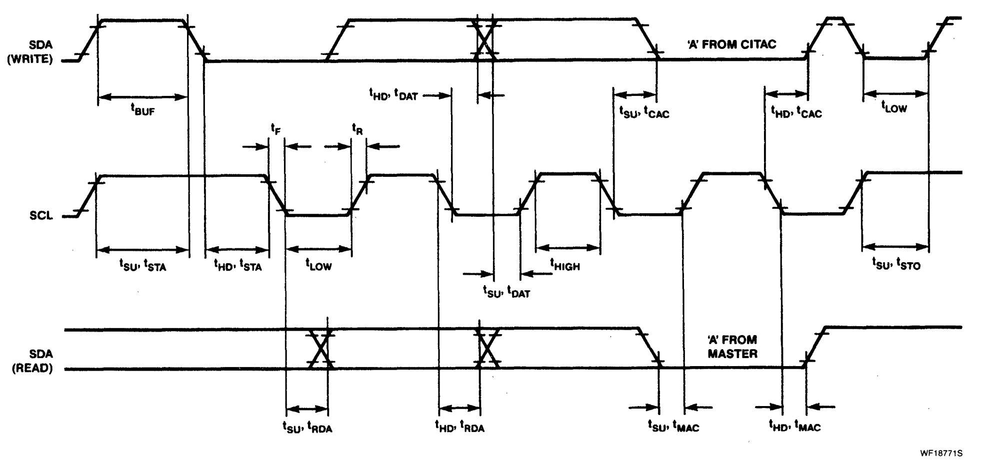


Figure 6. I^2C Bus Timing SAB3036

Signetics

SAB3037
FLL Tuning and Control Circuit

Product Specification

Linear Products

DESCRIPTION

The SAB3037 provides closed-loop digital tuning of TV receivers, with or without AFC, as required. It also controls up to 4 analog functions, 4 general purpose I/O ports and 4 high-current outputs for tuner band selection.

The IC is used in conjunction with a microcomputer from the MAB8400 family and is controlled via a two-wire, bi-directional I^2C bus.

FEATURES

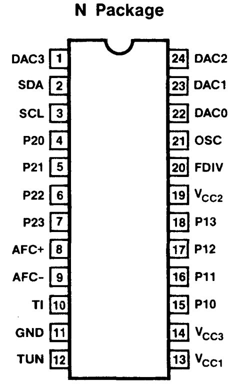
- **Combined analog and digital circuitry minimizes the number of additional interfacing components required**
- **Frequency measurement with resolution of 50kHz**
- **Selectable prescaler divisor of 64 or 256**
- **32V tuning voltage amplifier**
- **4 high-current outputs for direct band selection**
- **4 static digital to analog convertors (DACs) for control of analog functions**
- **Four general purpose input/output (I/O) ports**
- **Tuning with control of speed and direction**
- **Tuning with or without AFC**
- **Single-pin, 4MHz on-chip oscillator**
- **I^2C bus slave transceiver**

APPLICATIONS

- **TV receivers**
- **Satellite receivers**
- **CATV converters**

PIN CONFIGURATION

N Package



TOP VIEW

CD11970S

PIN NO.	SYMBOL	DESCRIPTION
1	DAC3	Output of static DAC
2	SDA	Serial data line } I^2C bus
3	SCL	Serial clock line } I^2C bus
4	P20	General purpose input/output ports
5	P21	
6	P22	
7	P23	
8	AFC+	AFC inputs
9	AFC−	
10	TI	Tuning voltage amplifier inverting input
11	GND	Ground
12	TUN	Tuning voltage amplifier output
13	V_{CC1}	+12V supply voltage
14	V_{CC3}	+32V supply for tuning voltage amplifier
15	P10	High-current band-selection output ports
16	P11	
17	P12	
18	P13	
19	V_{CC2}	Positive supply for high-current band-selection output circuits
20	FDIV	Input from prescaler
21	OSC	Crystal oscillator input
22	DAC0	Outputs of static DACs
23	DAC1	
24	DAC2	

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
24-Pin Plastic DIP (SOT-101A)	−20°C to +70°C	SAB3037N

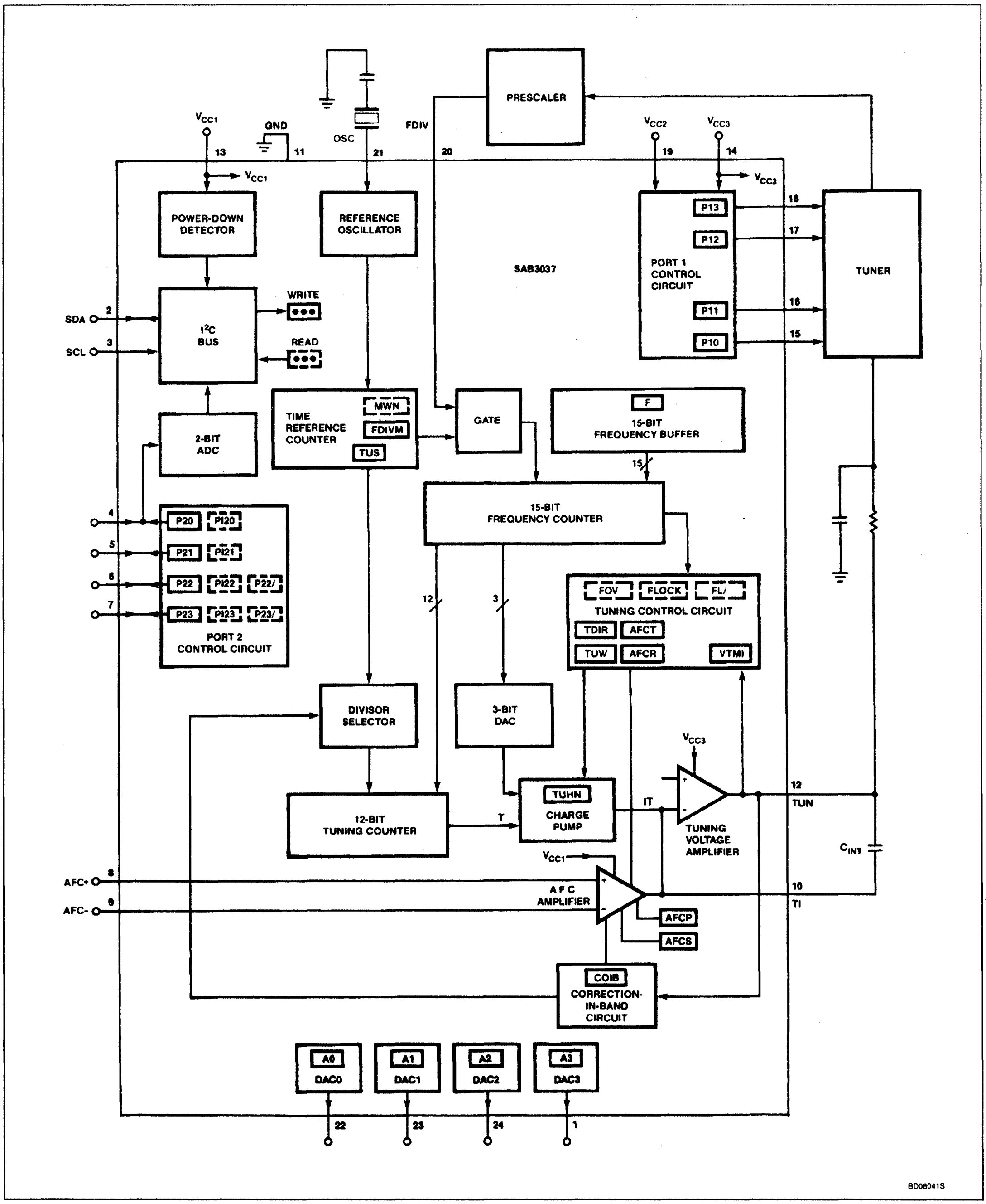
ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
	Supply voltage ranges:		
V_{CC1}	(Pin 13)	−0.3 to +18	V
V_{CC2}	(Pin 19)	−0.3 to +18	V
V_{CC3}	(Pin 14)	−0.3 to +36	V
	Input/output voltage ranges:		
V_{SDA}	(Pin 2)	−0.3 to +18	V
V_{SCL}	(Pin 3)	−0.3 to +18	V
V_{P2X}	(Pins 4 to 7)	−0.3 to +18	V
$V_{AFC+,\ AFC-}$	(Pins 8 and 9)	−0.3 to V_{CC1} [1]	V
V_{TI}	(Pin 10)	−0.3 to V_{CC1} [1]	V
V_{TUN}	(Pin 12)	−0.3 to V_{CC3} [3]	V
V_{P1X}	(Pins 15 to 18)	−0.3 to V_{CC2} [3]	V
V_{FDIV}	(Pin 20)	−0.3 to V_{CC1} [1]	V
V_{OSC}	(Pin 21)	−0.3 to +5	V
V_{DACX}	(Pins 1 and 22 to 24)	−0.3 to V_{CC} [1]	V
P_{TOT}	Total power dissipation	1000	mW
T_{STG}	Storage temperature range	−65 to +150	°C
T_A	Operating ambient temperature range	−20 to +70	°C

NOTES:

1. Pin voltage may exceed supply voltage if current is limited to 10mA.
2. Pin voltage must not exceed 18V but may exceed V_{CC2} if current is limited to 200mA.

BLOCK DIAGRAM



FLL Tuning and Control Circuit SAB3037

DC AND AC ELECTRICAL CHARACTERISTICS $T_A = 25°C$; V_{CC1}, V_{CC2}, V_{CC3} at typical voltages, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
V_{CC1} V_{CC2} V_{CC3}	Supply voltages	10.5 4.7 30	12 13 32	13.5 16 35	V V V
I_{CC1} I_{CC2} I_{CC3}	Supply currents (no outputs loaded)	18 0 0.2	30 0.6	45 0.1 2	mA mA mA
I_{CC2A} I_{CC3A}	Additional supply currents (A)[1]	−2 0.2		I_{OHP1X} 2	mA mA
P_{TOT}	Total power dissipation		380		mW
T_A	Operating ambient temperature	−20		+70	°C
I^2C bus inputs/outputs SDA input (Pin 2); SCL input (Pin 3)					
V_{IH}	Input voltage HIGH[2]	3		$V_{CC} - 1$	V
V_{IL}	Input voltage LOW	−0.3		1.5	V
I_{IH}	Input current HIGH[2]			10	µA
I_{IL}	Input current LOW[2]			10	µA
	SDA output (Pin 2, open-collector)				
V_{OL}	Output voltage LOW at $I_{OL} = 3mA$			0.4	V
I_{OL}	Maximum output sink current		5		mA
Open-collector I/O ports P20, P21, P22, P23 (Pins 4 to 7, open-collector)					
V_{IH}	Input voltage HIGH	2		16	V
V_{IL}	Input voltage LOW	−0.3		0.8	V
I_{IH}	Input current HIGH			25	µA
$-I_{IL}$	Input current LOW			25	µA
V_{OL}	Output voltage LOW at $I_{OL} = 2mA$			0.4	V
I_{OL}	Maximum output sink current		4		mA
AFC amplifier Inputs AFC+, AFC− (Pins 8, 9)					
  g00 g01 g10 g11	Transconductance for input voltages up to 1V differential: AFCS1 AFCS2 0 0 0 1 1 0 1 1	 100 15 30 60	 250 25 50 100	 800 35 70 140	 nA/V µA/V µA/V µA/V
ΔM_g	Tolerance of transconductance multiplying factor (2, 4 or 8) when correction-in-band is used	−20		+20	%
V_{IOFF}	Input offset voltage	−75		+75	mV
V_{COM}	Common-mode input voltage	3		$V_{CC1} - 2.5$	V
CMRR	Common-mode rejection ratio		50		dB
PSRR	Power supply (V_{CC1}) rejection ratio		50		dB
I_I	Input current			500	nA

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $T_A = 25°C$; V_{CC1}, V_{CC2}, V_{CC3} at typical voltages, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Tuning voltage amplifier Input TI, output TUN (Pins 10, 12)					
V_{TUN}	Maximum output voltage at $I_{LOAD} = \pm 2.5mA$	$V_{CC3} - 1.6$		$V_{CC3} - 0.4$	V
	Minimum output voltage at $I_{LOAD} = \pm 2.5mA$: VTMI1 VTMI0				
V_{TM00}	0 0	300		500	mV
V_{TM10}	1 0	450		650	mV
V_{TM11}	1 1	650		900	mV
$-I_{TUNH}$	Maximum output source current	2.5		8	mA
I_{TUNL}	Maximum output sink current		40		mA
I_{TI}	Input bias current	−5		+5	nA
PSRR	Power supply V_{CC3} rejection ratio		60		dB
	Minimum charge IT to tuning voltage amplifier TUHN1 TUHN0				
CH_{00}	0 0	0.4	1	1.7	μA/μs
CH_{01}	0 1	4	8	14	μA/μs
CH_{10}	1 0	15	30	48	μA/μs
CH_{11}	1 1	130	250	370	μA/μs
ΔCH	Tolerance of charge (or ΔV_{TUN}) multiplying factor when COIB and/or TUS are used	−20		+20	%
	Maximum current I into tuning amplifier TUHN1 TUHN0				
I_{T00}	0 0	1.7	3.5	5.1	μA
I_{T01}	0 1	15	29	41	μA
I_{T10}	1 0	65	110	160	μA
I_{T11}	1 1	530	875	1220	μA
Correction-in-band					
ΔV_{CIB}	Tolerance of correction-in-band levels 12V, 18V, and 24V	−15		+15	%
Band-select output ports P10, P11, P12, P13 (Pins 15 to 18)					
V_{OH}	Output voltage HIGH at $-I_{OH} = 50mA$[3]	$V_{CC2} - 0.6$			V
V_{OL}	Output voltage LOW at $I_{OL} = 2mA$			0.4	V
$-I_{OH}$	Maximum output source current[3]		130	200	mA
I_{OL}	Maximum output sink current		5		mA
FDIV input (Pin 20)					
$V_{FDIV\ (P-P)}$	Input voltage (peak-to-peak value) (t_{RISE} and $t_{FALL} \leq 40ns$)	0.1		2	V
	Duty cycle	40		60	%
f_{MAX}	Maximum input frequency	14.5			MHz
Z_I	Input impedance		8		kΩ
C_I	Input capacitance		5		pF
OSC input (Pin 21)					
R_X	Crystal resistance at resonance (4MHz)			150	Ω

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $T_A = 25°C$; V_{CC1}, V_{CC2}, V_{CC3} at typical voltages, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
DAC outputs 0 to 3 (Pins 22 to 24 and Pin 1)					
V_{DH}	Maximum output voltage (no load) at $V_{CC1} = 12V^4$	10		11.5	V
V_{DL}	Minimum output voltage (no load) at $V_{CC1} = 12V^4$	0.1		1	V
ΔV_D	Positive value of smallest step (1 least significant bit)	0		350	mV
	Deviation from linearity			0.5	V
Z_O	Output impedance at $I_{LOAD} = \pm 2mA$			70	Ω
$-I_{DH}$	Maximum output source current			6	mA
I_{DL}	Maximum output sink current		8		mA
Power-down reset					
V_{PD}	Maximum supply voltage V_{CC1} at which power-down reset is active	7.5		9.5	V
t_R	V_{CC1} rise time during power-up (up to V_{PD})	5			μs
Voltage level for valid module address					
	Voltage level at P20 (Pin 4) for valid module address as a function of MA1, MA0 MA1 MA0				
V_{VA00}	0 0	−0.3		16	V
V_{VA01}	0 1	−0.3		0.8	V
V_{VA10}	1 0	2.5		$V_{CC1} - 2$	V
V_{VA11}	1 1	$V_{CC1} - 0.3$		V_{CC1}	V

NOTES:
1. For each band-select output which is programmed at logic 1, sourcing a current I_{OHP1X}, the additional supply currents (A) shown must be added to I_{CC2} and I_{CC3}, respectively.
2. If $V_{CC1} < 1V$, the input current is limited to 10μA at input voltages up to 16V.
3. At continuous operation the output current should not exceed 50mA. When the output is short-circuited to ground for several seconds the device may be damaged.
4. Values are proportional to V_{CC1}.

FUNCTIONAL DESCRIPTION

The SAB3037 is a monolithic computer interface which provides tuning and control functions and operates in conjunction with a microcomputer via an I^2C bus.

Tuning

This is performed using frequency-locked loop digital control. Data corresponding to the required tuner frequency is stored in a 15-bit frequency buffer. The actual tuner frequency, divided by a factor of 256 (or by 64) by a prescaler, is applied via a gate to a 15-bit frequency counter. This input (FDIV) is measured over a period controlled by a time reference counter and is compared with the contents of the frequency buffer. The result of the comparison is used to control the tuning voltage so that the tuner frequency equals the contents of the frequency buffer multiplied by 50kHz within a programmable tuning window (TUW).

The system cycles over a period of 6.4ms (or 2.56ms), controlled by the time reference counter which is clocked by an on-chip 4MHz reference oscillator. Regulation of the tuning voltage is performed by a charge pump frequency-locked loop system. The charge IT flowing into the tuning voltage amplifier is controlled by the tuning counter, 3-bit DAC and the charge pump circuit. The charge IT is linear with the frequency deviation Δf in steps of 50kHz. For loop gain control, the relationship $\Delta IT/\Delta f$ is programmable. In the normal mode (when control bits TUHN0 and TUHN1 are both at logic 1 (see OPERATION) the minimum charge IT at $\Delta f = 50kHz$ equals $250\mu A/\mu s$ (typical).

By programming the tuning sensitivity bits (TUS), the charge IT can be doubled up to 6 times. If correction-in-band (COIB) is programmed, the charge can be further doubled up to three times in relation to the tuning voltage level. From this, the maximum charge IT at $\Delta f = 50kHz$ equals $2^6 \times 2^3 \times 250\mu A/\mu s$ (typical).

The maximum tuning current I is $875\mu A$ (typical). In the tuning-hold (TUHN) mode (TUHN is Active-LOW), the tuning current I is reduced and as a consequence the charge into the tuning amplifier is also reduced.

An in-lock situation can be detected by reading FLOCK. When the tuner oscillator frequency is within the programmable tuning window (TUW), FLOCK is set to logic 1. If the frequency is also within the programmable AFC hold range (AFCR), which always occurs if AFCR is wider than TUW, control bit AFCT can be set to logic 1. When set, digital tuning will be switched off, AFC will be switched on and FLOCK will stay at logic 1 as long as the oscillator frequency is within AFCR. If the frequency of the tuning oscillator does not remain within AFCR, AFCT is cleared automatically and the system reverts to digital tuning. To be able to detect this situation, the occurrence of positive and negative transitions in the FLOCK signal can be read (FL/1N and FL/0N). AFCT can also be cleared by programming the AFCT bit to logic 0.

The AFC has programmable polarity and transconductance; the latter can be doubled up to 3 times, depending on the tuning voltage level if correction-in-band is used.

The direction of tuning is programmable by using control bits TDIRD (tuning direction down) and TDIRU (tuning direction up). If a tuner enters a region in which oscillation stops, then, providing the prescaler remains stable, no FDIV signal is supplied to CITAC. In this situation the system will tune up, moving away from frequency lock-in. This situation is avoided by setting TDIRD which causes the system to tune down. In normal operation TDIRD must be cleared.

If a tuner stops oscillating and the prescaler becomes unstable by going into self-oscillation at a very high frequency, the system will react by tuning down, moving away from frequency lock-in. To overcome this, the system can be forced to tune up at the lowest sensitivity (TUS) value, by setting TDIRU.

Setting both TDIRD and TDIRU causes the digital tuning to be interrupted and AFC to be switched on.

The minimum tuning voltage which can be generated during digital tuning is programmable by VTMI to prevent the tuner from being driven into an unspecified low tuning voltage region.

Control

For tuner band selection there are four outputs — P10 to P13 — which are capable of sourcing up to 50mA at a voltage drop of less than 600mV with respect to the separate power supply input V_{CC2}.

For additional digital control, four open-collector I/O ports — P20 to P23 — are provided. Ports P22 and P23 are capable of detecting positive and negative transitions in their input signals. With the aid of port P20, up to three independent module addresses can be programmed.

Four 6-bit digital-to-analog converters — DAC0 to DAC3 — are provided for analog control.

Reset

CITAC goes into the power-down reset mode when V_{CC1} is below 8.5V (typical). In this mode all registers are set to a defined state. Reset can also be programmed.

OPERATION

Write

CITAC is controlled via a bidirectional two-wire I^2C bus. For programming, a module address, R/$\overline{W}$ bit (logic 0), an instruction byte and a data/control byte are written into CITAC in the format shown in Figure 1.

	MODULE ADDRESS									INSTRUCTION BYTE									DATA/CONTROL BYTE									
S	1	1	0	0	0	MA 1	MA 0	0	A	I_7	I_6	I_5	I_4	I_3	I_2	I_1	I_0	A	D_7	D_6	D_5	D_4	D_3	D_2	D_1	D_0	A	P
	msb							R/$\overline{W}$		msb									msb									

AF04660S

Figure 1. I^2C Bus Write Format

The module address bits MA1, MA0 are used to give a 2-bit module address as a function of the voltage at port P20 as shown in Table 1.

Acknowledge (A) is generated by CITAC only when a valid address is received and the device is not in the power-down reset mode (V_{CC1} > 8.5V (typical)).

Tuning

Tuning is controlled by the instruction and data/control bytes as shown in Figure 2.

Frequency

Frequency is set when Bit I_7 of the instruction byte is set to logic 1; the remainder of this byte together with the data/control byte are loaded into the frequency buffer. The frequency to which the tuner oscillator is regulated equals the decimal representation of the 15-bit word multiplied by 50kHz. All frequency bits are set to logic 1 at reset.

Tuning Hold

The TUHN bits are used to decrease the maximum tuning current and, as a consequence, the minimum charge IT (at Δf = 50kHz) into the tuning amplifier.

Table 1. Valid Module Addresses

MA1	MA0	P20
0	0	Don't care
0	1	GND
1	0	½ V_{CC1}
1	1	V_{CC1}

Table 2. Tuning Current Control

TUHN1	TUHN0	TYP. I_{MAX} (μA)	TYP. IT_{MIN} (μA/μs)	TYP. ΔV_{TUNmin} at C_{INT} = 1μF (μV)
0	0	3.5[1]	1[1]	1[1]
0	1	29	8	8
1	0	110	30	30
1	1	875	250	250

NOTE:
1. Values after reset.

During tuning but before lock-in, the highest current value should be selected. After lock-in the current may be reduced to decrease the tuning voltage ripple.

The lowest current value should not be used for tuning due to the input bias current of the tuning voltage amplifier (maximum 5nA). However, it is good practice to program the lowest current value during tuner band switching.

	INSTRUCTION BYTE								DATA/CONTROL BYTE							
	I_7	I_6	I_5	I_4	I_3	I_2	I_1	I_0	D_7	D_6	D_5	D_4	D_3	D_2	D_1	D_0
FREQ	1	F14	F13	F12	F11	F10	F9	F8	F7	F6	F5	F4	F3	F2	F1	F0
TCD0	0	0	1	0	1	0	0	1	AFCT	VTMI0	AFCR1	AFCR0	TUHN1	TUHN0	TUW1	TUW0
TCD1	0	0	1	0	1	0	1	0	VTMI1	COIB1	COIB0	AFCS1	AFCS0	TUS2	TUS1	TUS0
TCD2	0	0	1	0	1	0	1	1	0	0	0	0	AFCP	FDIVM	TDIRD	TDIRU

AF04670S

Figure 2. Tuning Control Format

Table 3. Minimum Charge IT as a Function of TUS $\Delta f = 50kHz$; TUHN0 = Logic 1; TUHN1 = Logic 1

TUS2	TUS1	TUS0	TYP. IT_{MIN} (mA/μs)	TYP. ΔV_{TUNmin} at $C_{INT} = 1\mu F$ (mV)
0	0	0	0.25[1]	0.25[1]
0	0	1	0.5	0.5
0	1	0	1	1
0	1	1	2	2
1	0	0	4	4
1	0	1	8	8
1	1	0	16	16

NOTE:
1. Values after reset.

Table 4. Programming Correction-In-Band

COIB1	COIB0	CHARGE MULTIPLYING FACTORS AT TYPICAL VALUES OF V_{TUN} AT:			
		< 12V	12 to 18V	18 to 24V	> 24V
0	0	1[1]	1[1]	1[1]	1[1]
0	1	1	1	1	2
1	0	1	1	2	4
1	1	1	2	4	8

NOTE:
1. Values after reset.

Table 5. Tuning Window Programming

TUW1	TUW0	$\lvert\Delta f\rvert$ (kHz)	TUNING WINDOW (kHz)
0	0	0[1]	0[1]
0	1	50	100
1	0	150	300

NOTE:
1. Values after reset.

Table 6. AFC Hold Range Programming

AFCR1	AFCR0	$\lvert\Delta f\rvert$ (kHz)	AFC HOLD RANGE (kHz)
0	0	0[1]	0[1]
0	1	350	700
1	0	750	1500

NOTE:
1. Values after reset.

Table 7. Transconductance Programming

AFCS1	AFCS0	TYP. TRANSCONDUCTANCE (μA/V)
0	0	0.25[1]
0	1	25
1	0	50
1	1	100

NOTE:
1. Values after reset.

Tuning Sensitivity

To be able to program an optimum loop gain, the charge IT can be programmed by changing T using tuning sensitivity (TUS). Table 3 shows the minimum charge IT obtained by programming the TUS bits at $\Delta f = 50kHz$; TUHN0 and TUHN1 = logic 1.

Correction-In-Band

This control is used to correct the loop gain of the tuning system to reduce in-band variations due to a non-linear voltage/frequency characteristic of the tuner. Correction-in-band (COIB) controls the time T of the charge equation IT and takes into account the tuning voltage V_{TUN} to give charge multiplying factors as shown in Table 4.

The transconductance multiplying factor of the AFC amplifier is similar when COIB is used, except for the lowest transconductance which is not affected.

Tuning Window

Digital tuning is interrupted and FLOCK is set to logic 1 (in-lock) when the absolute deviation $|\Delta f|$ between the tuner oscillator frequency and the programmed frequency is smaller than the programmed TUW value (see Table 5). If $|\Delta f|$ is up to 50kHz above the values listed in Table 5, it is possible for the system to be locked depending on the phase relationship between FDIV and the reference counter.

AFC

When AFCT is set to logic 1 it will not be cleared and the AFC will remain on as long as $|\Delta f|$ is less than the value programmed for the AFC hold range AFCR (see Table 6). It is possible for the AFC to remain on for values of up to 50kHz more than the programmed value depending on the phase relationship between FDIV and the reference counter.

Transconductance

The transconductance (g) of the AFC amplifier is programmed via the AFC sensitivity bits AFCS as shown in Table 7.

FLL Tuning and Control Circuit SAB3037

	INSTRUCTION BYTE								DATA/CONTROL BYTE							
	I_7	I_6	I_5	I_4	I_3	I_2	I_1	I_0	D_7	D_6	D_5	D_4	D_3	D_2	D_1	D_0
POD	0	0	1	0	1	0	0	0	P23	P22	P21	P20	P13	P12	P11	P10
DACX	0	0	1	0	0	0	X1	X0	0	0	AX5	AX4	AX3	AX2	AX1	AX0

AF04681S

Figure 3. Control Programming

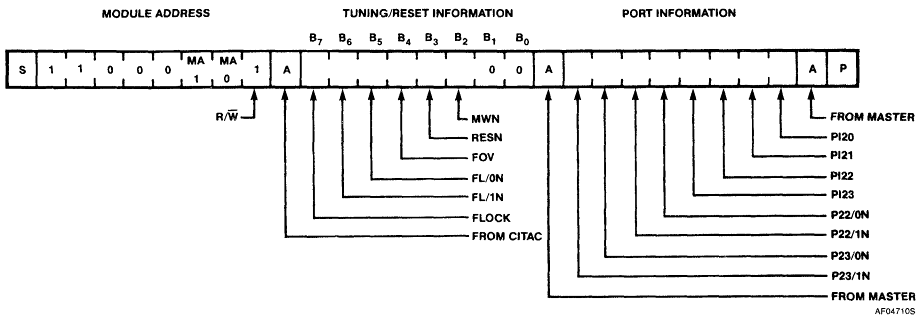


Figure 4. Information Byte Format

AFC Polarity

If a positive differential input voltage is applied to the (switched-on) AFC amplifier, the tuning voltage V_{TUN} falls when the AFC polarity bit AFCP is at logic 0 (value after reset). At AFCP = logic 1, V_{TUN} rises.

Minimum Tuning Voltage

Both minimum tuning voltage control bits, VTMI1 and VTMI0, are at logic 0 after reset. Further details are given in the DC Electrical Characteristics table.

Frequency Measuring Window

The frequency measuring window which is programmed must correspond with the division factor of the prescaler in use (see Table 8).

Tuning Direction

Both tuning direction bits, TDIRU (up) and TDIRD (down), are at logic 0 after reset.

Control

The instruction bytes POD (port output data) and DACX (digital-to-analog converter control) are shown in Figure 5, together with the corresponding data/control bytes. Control is implemented as follows:

P13, P12, P11, P10 — Band select outputs. If a logic 1 is programmed on any of the POD bits D_3 to D_0, the relevant output goes High. All outputs are Low after reset.

P23, P22, P21, P20 — Open-collector I/O ports. If a logic 0 is programmed on any of the POD bits D_7 to D_4, the relevant output is forced LOW. All outputs are at logic 1 after reset (high impedance state).

DACX — Digital-to-analog converters. The digital-to-analog converter selected corresponds to the decimal equivalent of the DACX bits X1, X0. The output voltage of the selected DAC is set by programming the bits AX5 to AX0; the lowest output voltage is programmed with all data AX5 to AX0 at logic 0, or after reset has been activated.

Table 8. Frequency Measuring Window Programming

FDIVM	PRESCALER DIVISION FACTOR	CYCLE PERIOD (ms)	MEASURING WINDOW (ms)
0	256	6.4[1]	5.12[1]
1	64	2.56	1.28

NOTE:
1. Values after reset.

Read

Information is read from CITAC when the R/$\overline{W}$ bit is set to logic 1. An acknowledge must be generated by the master after each data byte to allow transmission to continue. If no acknowledge is generated by the master, the slave (CITAC) stops transmitting. The format of the information bytes is shown in Figure 4.

Tuning/Reset Information Bits

FLOCK — Set to logic 1 when the tuning oscillator frequency is within the programmed tuning window.

FL/1N — Set to logic 0 (Active-LOW) when FLOCK changes from 0 to 1 and is reset to logic 1 automatically after tuning information has been read.

FL/0N — As for FL/1N but is set to logic 0 when FLOCK changes from 1 to 0.

FOV — Indicates frequency overflow. When the tuner oscillator frequency is too high with respect to the programmed frequency, FOV is at logic 1, and when too low, FOV is at logic 0. FOV is not valid when TDIRU and/or TDIRD are set to logic 1.

RESN — Set to logic 0 (Active-LOW) by a programmed reset or a power-down-reset. It is reset to logic 1 automatically after tuning/reset information has been read.

MWN — MWN (frequency measuring window, Active-LOW) is at logic 1 for a period of 1.28ms, during which time the results of frequency measurement are processed. This time is independent of the cycle period. During the remaining time, MWN is at logic 0 and the received frequency is measured.

When slightly different frequencies are programmed repeatedly and AFC is switched on, the received frequency can be measured using FOV and FLOCK. To prevent the frequency counter and frequency buffer being loaded at the same time, frequency should be programmed only during the period of MWN = logic 0.

Port Information Bits

P23/1N, P22/1N — Set to logic 0 (Active-LOW) at a LOW-to-HIGH transition in the input voltage on P23 and P22, respectively. Both are reset to logic 1 after the port information has been read.

P23/0N, P22/0N — As for P23/1N and P22/1N but are set to logic 0 at a HIGH-to-LOW transition.

PI23, PI22, PI21, PI20 — Indicate input voltage levels at P23, P22, P21 and P20, respectively. A logic 1 indicates a HIGH input level.

Reset

The programming to reset all registers is shown in Figure 5. Reset is activated only at data byte HEX 06. Acknowledge is generated at every byte, provided that CITAC is not in the power-down reset mode. After the general call address byte, transmission of more than one data byte is not allowed.

GENERAL CALL ADDRESS — HEX06

S	0	0	0	0	0	0	0	0	A	0	0	0	0	0	1	1	0	A	P

AF04690S

Figure 5. Reset Programming

I^2C BUS TIMING (Figure 6)

I^2C bus load conditions are as follows:

4kΩ pull-up resistor to +5V; 200pF capacitor to GND.

All values are referred to V_{IH} = 3V and V_{IL} = 1.5V.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
t_{BUF}	Bus free before start	4			µs
t_{SU}, t_{STA}	Start condition setup time	4			µs
t_{HD}, t_{STA}	Start condition hold time	4			µs
t_{LOW}	SCL, SDA LOW period	4			µs
t_{HIGH}	SCL HIGH period	4			µs
t_R	SCL, SDA rise time			1	µs
t_F	SCL, SDA fall time			0.3	µs
t_{SU}, t_{DAT}	Data setup time (write)	1			µs
t_{HD}, t_{DAT}	Data hold time (write)	1			µs
t_{SU}, t_{CAC}	Acknowledge (from CITAC) setup time			2	µs
t_{HD}, t_{CAC}	Acknowledge (from CITAC) hold time	0			µs
t_{SU}, t_{STO}	Stop condition setup time	4			µs
t_{SU}, t_{RDA}	Data setup time (read)			2	µs
t_{HD}, t_{RDA}	Data hold time (read)	0			µs
t_{SU}, t_{MAC}	Acknowledge (from master) setup time	1			µs
t_{HD}, t_{MAC}	Acknowledge (from master) hold time	2			µs

NOTE:

1. Timings t_{SU}, t_{DAT} and t_{HD}, t_{DAT} deviate from the I^2C bus specification.
   After reset has been activated, transmission may only be started after a 50µs delay.

FLL Tuning and Control Circuit SAB3037

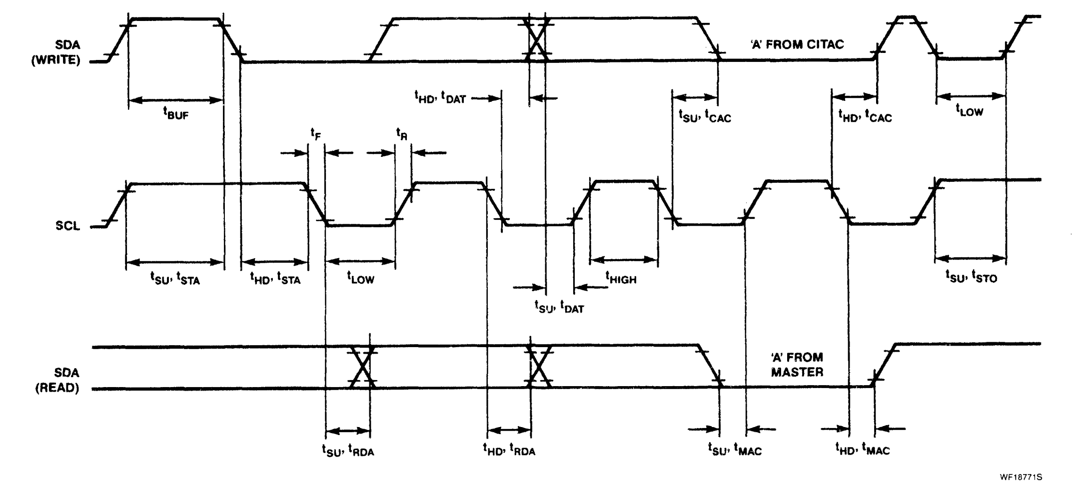


Figure 6. I²C Bus Timing SAB3037

Signetics

TDA5030A
VHF Mixer/Oscillator Circuit

Product Specification

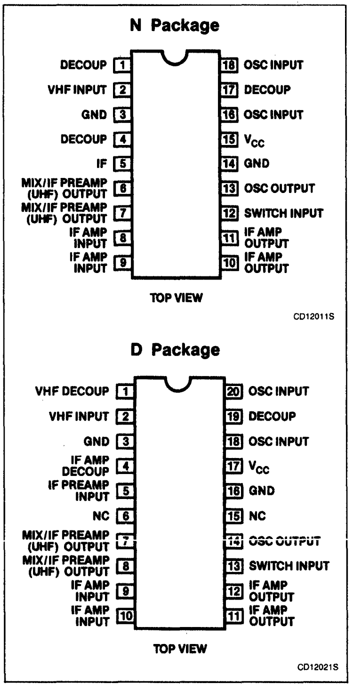
Linear Products

DESCRIPTION
The TDA5030A performs the VHF mixer, VHF oscillator, SAW filter IF amplifier, and UHF IF amplifier functions in television tuners.

FEATURES
- **A balanced VHF mixer**
- **An amplitude-controlled VHF local oscillator**
- **A surface acoustic wave filter IF amplifier**
- **A UHF IF preamplifier**
- **A buffer stage for driving an external prescaler with the local oscillator signal**
- **A voltage stabilizer**
- **A UHF/VHF switching circuit**

APPLICATIONS
- **Mixer/oscillator**
- **TV tuners**
- **CATV**
- **LAN**
- **Demodulator**

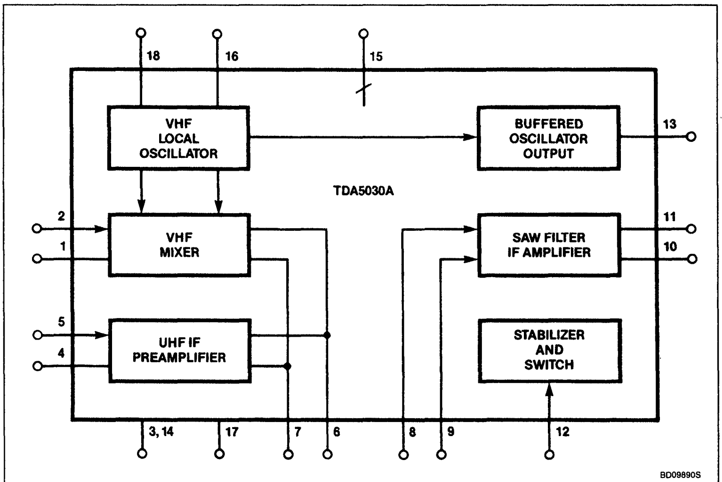
PIN CONFIGURATIONS



ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
18-Pin Plastic DIP (SOT-102A)	−25°C to +85°C	TDA5030AN
20-Pin Plastic SO DIP (SOT-163A)	−25°C to +85°C	TDA5030ATD

BLOCK DIAGRAM



NOTE:
Pinout is for 18-pin N package.

VHF Mixer/Oscillator Circuit

TDA5030A

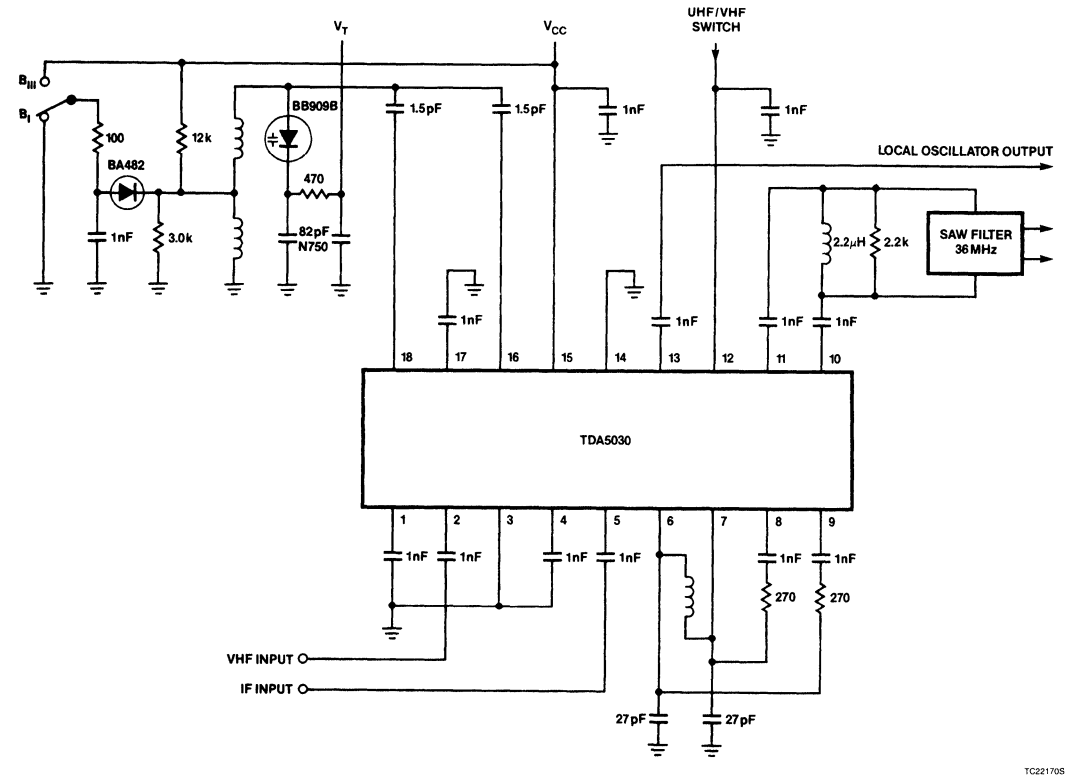


Figure 1. Test Circuit

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{CC}	Supply voltage (Pin 15)	14	V
V_I	Input voltage (Pin 1, 2, 4, and 5)	0 to 5	V
V_{12}	Switching voltage (Pin 12)	0 to V_{CC}+0.3	V
$-I_{10, 11, 13}$	Output currents	10	mA
t_{SS}	Storage-circuit time on outputs (Pin 10 and 11)	10	s
T_{STG}	Storage temperature range	-65 to +150	°C
T_A	Operating ambient temperature range	-25 to +85	°C
T_J	Junction temperature	+125	°C
θ_{JA}	Thermal resistance from junction to ambient	+55	°C/W

DC AND AC ELECTRICAL CHARACTERISTICS

Measured in circuit of Figure 1; V_{CC} = 12V; T_A = 25°C, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supply					
V_{CC}	Supply voltage	10		13.2	V
I_{CC}	Supply current		42	55	mA
V_{12}	Switching voltage VHF	0		2.5	V
V_{12}	Switching voltage UHF	9.5		V_{CC}+0.3	V
I_{12}	Switching current UHF			0.7	mA
VHF mixer (including IF amplifier)					
f_R	Frequency range	50		470	MHz
NF	Noise figure (Pin 2) 50MHz 225MHz 300MHz		 7.5 9 10	 9 10 12	 dB dB dB
G	Optimum source admittance (Pin 2) 50MHz 225MHz 300MHz		 0.5 1.1 1.2		 ms ms ms
G_I	Input conductance (Pin 2) 50MHz 225MHz 300MHz		 0.23 0.5 0.67		 ms ms ms
C_I	Input capacitance (Pin 2) 50MHz		 2.5		 pF
V_{2-3}	Input voltage for 1% cross-modulation (in channel); $R_P > 1k\Omega$; tuned circuit with C_P = 22pF; f_{RES} = 36MHz	97	99		dBμV
V_{2-14}	Input voltage for 10kHz pulling (in channel) at < 300MHz	100			dBμV
A_V	Voltage gain	22.5	24.5	26.5	dB
UHF preamplifier (including IF amplifier)					
G_I	Input conductance (Pin 5)		0.3		ms
C_I	Input capacitance (Pin 5)		3.0		pF
NF	Noise figure		5	6	dB
V_{5-14}	Input voltage for 1% cross-modulation (in channel)	88	90		dBμV
A_V	Voltage gain	31.5	33.5	35.5	dB
G_5	Optimum source admittance		3.3		ms

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) Measured in circuit of Figure 1; $V_{CC} = 12V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
VHF mixer					
$Yc_{2-6,\ 7}$	Conversion transadmittance		5.7		ms
Z_O	Output impedance		1.6		kΩ
VHF oscillator					
f_R	Frequency range	70		520	MHz
Δf	Frequency shift $\Delta V_{CC} = 10\%$; 70 to 330MHz			200	kHz
Δf	Frequency drift ΔT = 15k; 70 to 330MHz			250	kHz
Δf	Frequency drift from 5sec to 15min after switching on			200	kHz
SAW filter IF amplifier					
$Z_{8,\ 9}$	Input impedance $Z_{10,\ 11} = 2k\Omega$, f = 36MHz		340+j100		Ω
$Z_{8,\ 9-10,\ 11}$	Transimpedance		2.2		kΩ
$Z_{10,\ 11}$	Output impedance $Z_{8,\ 9} = 1.6k\Omega$; f = 36MHz		50+j40		Ω
VHF local oscillator buffer stage					
	Output voltage				
V_{13}	$R_L = 75\Omega$; f < 100MHz	14	20		mV
V_{13}	$R_L = 75\Omega$; f > 100MHz	10	20		mV
Z_{13}	Output impedance f = 100MHz		90		Ω
$\frac{RF}{(RF+LO)}$	RF signal on LO output; $R_L = 50\Omega$; $V_I = 1V$; $f \leqslant 225MHz$			10	dB

Signetics

Section 5
Remote Control Systems

Linear Products

5

INDEX

Signetics

SAA3004
Infrared Transmitter

Product Specification

Linear Products

DESCRIPTION

The SAA3004 transmitter IC is designed for infrared remote control systems. It has a total of 448 commands which are divided into 7 subsystem groups with 64 commands each. The subsystem code may be selected by a press button, a slider switch or hard wired.

The SAA3004 generates the pattern for driving the output stage. These patterns are pulse distance coded. The pulses are infrared flashes or modulated. The transmission mode is defined in conjunction with the subsystem address. Modulated pulses allow receivers with narrow-band preamplifiers for improved noise rejection to be used. Flashed pulses require a wide-band preamplifier within the receiver.

FEATURES

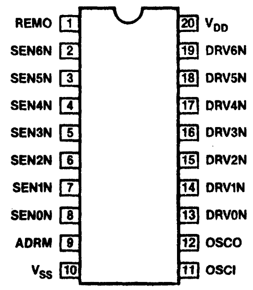
- **Flashed or modulated transmission**
- **7 subsystem addresses**
- **Up to 64 commands per subsystem address**
- **High-current remote output at V_{DD} = 6V ($-I_{OH}$ = 40mA)**
- **Low number of additional components**
- **Key release detection by toggle bits**
- **Very low standby current (< 2μA)**
- **Operational current < 2mA at 6V supply**
- **Wide supply voltage range (4 to 11V)**
- **Ceramic resonator controlled frequency (typ. 450kHz)**
- **Encapsulation: 20-lead plastic DIP or 20-lead plastic mini-pack (SO-20)**

APPLICATIONS

- **TV**
- **Audio**

PIN CONFIGURATION

N, D Packages



TOP VIEW

CD12000S

PIN NO.	SYMBOL	DESCRIPTION
1	REMO	Remote data output
2	SEN6N	Key matrix sense inputs
3	SEN5N	
4	SEN4N	
5	SEN3N	
6	SEN2N	
7	SEN1N	
8	SEN0N	
9	ADRM	Address mode control input
10	V_{SS}	Ground
11	OSCI	Oscillator input
12	OSCO	Oscillator output
13	DRV0N	Key matrix drive outputs
14	DRV1N	
15	DRV2N	
16	DRV3N	
17	DRV4N	
18	DRV5N	
19	DRV6N	
20	V_{DD}	Positive supply

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
20-Pin Plastic DIP (SOT-146C1)	−20°C to +70°C	SAA3004PN
20-Pin Plastic SOL (SOT-163AC3)	−20°C to +70°C	SAA3004TD

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{DD}	Supply voltage range	-0.5 to +15	V
V_I	Input voltage range	-0.5 to V_{DD} + 0.5	V
V_O	Output voltage range	-0.5 to V_{DD} + 0.5	V
±I	DC current into any input or output	10	mA
$-I_{(REMO)M}$	Peak REMO output current during 10μs; duty factor = 1%	300	mA
P_{TOT}	Power dissipation per package for T_A = −20 to +70°C	200	mW
T_{STG}	Storage temperature range	-65 to +150	°C
T_A	Operating ambient temperature range	-20 to +70	°C

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DC ELECTRICAL CHARACTERISTICS V_{SS} = 0V; T_A=25°C, unless otherwise specified.

SYMBOL	PARAMETER	V_{DD} (V)	LIMITS			UNIT
			Min	Typ	Max	
V_{DD}	Supply voltage T_A = 0 to +70°C		4		11	V
 I_{DD} I_{DD}	Supply current; active f_{OSC} = 455kHz; REMO output unloaded	 6 9		 1 3		 mA mA
 I_{DD} I_{DD}	Supply current; inactive (stand-by mode) T_A = 25°C	 6 9			 2 2	 μA μA
f_{OSC}	Oscillator frequency (ceramic resonator)	4 to 11	400		500	kHz
Keyboard matrix						
	Inputs SEN0N to SEN6N					
V_{IL}	Input voltage LOW	4 to 11			0.2 × V_{DD}	V
V_{IH}	Input voltage HIGH	4 to 11	0.8 × V_{DD}			V
$-I_I$ $-I_I$	Input current V_I = 0V	4 11	10 30		100 300	μA μA
 I_I	Input leakage current V_I = V_{DD}	 11			 1	 μA
	Outputs DRV0N to DRV6N					
 V_{OL} V_{OL}	Output voltage "ON" I_O = 0.1mA I_O = 1.0mA	 4 11			 0.3 0.5	 V V
 I_O	Output current "OFF" V_O = 11V	 11			 10	 μA
Control input ADRM						
V_{IL}	Input voltage LOW				0.8 × V_{DD}	V
V_{IH}	Input voltage HIGH		0.2 × V_{DD}			V
	Input current (switched P-and N-channel pull-up/pull-down)					
I_{IL} I_{IL}	Pull-up active standby voltage: 0V	4 11	10 30		100 300	μA μA
I_{IH} I_{IH}	Pull-down active standby voltage: V_{DD}	4 11	10 30		100 300	μA μA
Data output REMO						
V_{OH} V_{OH}	Output voltage HIGH $-I_{OH}$ = 40mA	6 9	3 6			V V
V_{OL} V_{OL}	Output voltage LOW I_{OL} = 0.3mA	6 9			0.2 0.1	V V
Oscillator						
 I_I	Input current OSCI at V_{DD}	 6	 0.8		 2.7	 μA
 V_{OH}	Output voltage HIGH $-I_{OL}$ = 0.1mA	 6			 V_{DD} − 0.6	 V
 V_{OL}	Output voltage LOW I_{OH} = 0.1mA	 6			 0.6	 V

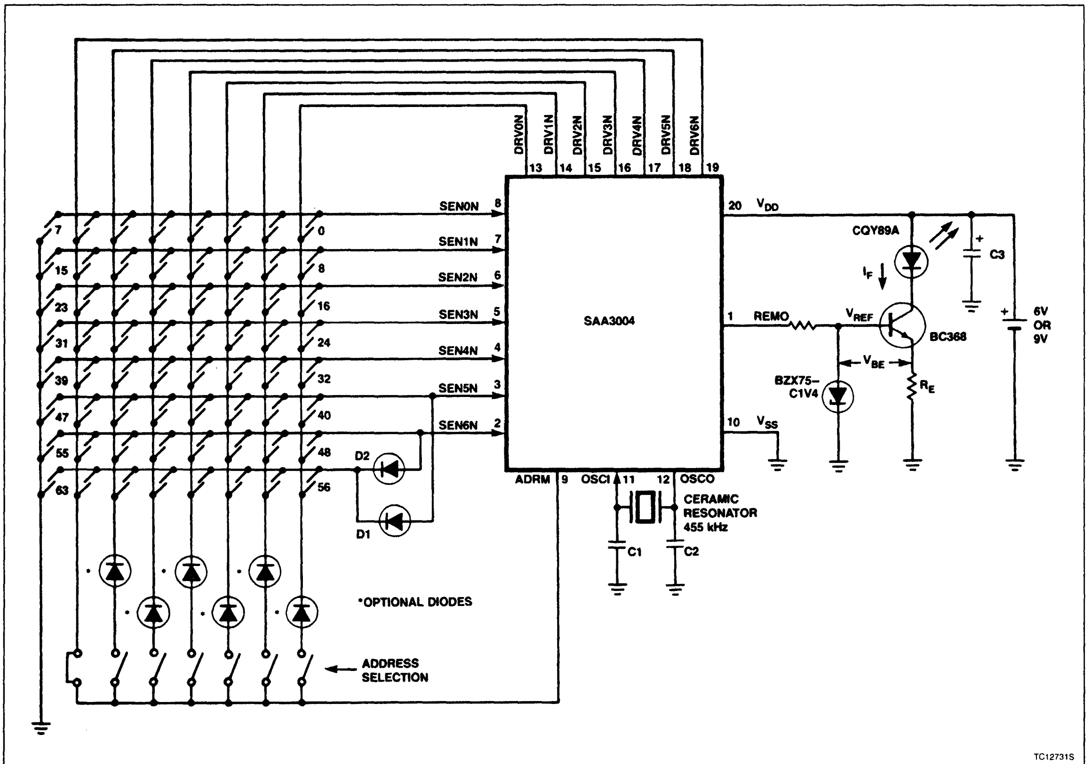


Figure 1. Transmitter With SAA3004

INPUTS AND OUTPUTS

Key Matrix Inputs and Outputs (DRV0N to DRV6N and SEN0N to SEN6N)

The transmitter keyboard is arranged as a scanned matrix. The matrix consists of 7 driver outputs and 7 sense inputs as shown in Figure 1. The driver outputs DRV0N to DRV6N are open-drain N-channel transistors and they are conductive in the stand-by mode. The 7 sense inputs (SEN0N to SEN6N) enable the generation of 56 command codes. With 2 external diodes all 64 commands are addressable. The sense inputs have P-channel pull-up transistors, so that they are HIGH until they are pulled LOW by connecting them to an output via a key depression to initiate a code transmission.

Address Mode Input (ADRM)

The subsystem address and the transmission mode are defined by connecting the ADRM input to one or more driver outputs (DRV0N to DRV6N) of the key matrix. If more than one driver is connected to ADRM, they must be decoupled by a diode. This allows the definition of seven subsystem addresses as shown in Table 3. If driver DRV6N is connected to ADRM the data output format of REMO is modulated or if not connected, flashed.

The ADRM input has switched pull-up and pull-down loads. In the stand-by mode only the pull-down device is active. Whether ADRM is open (subsystem address 0, flashed mode) or connected to the driver outputs, this input is LOW and will not cause unwanted dissipation. When the transmitter becomes active by pressing a key, the pull-down device is switched off and the pull-up device is switched on, so that the applied driver signals are sensed for the decoding of the subsystem address and the mode of transmission.

The arrangement of the subsystem address coding is such that only the driver DRVnN with the highest number (n) defines the subsystem address, e.g., if driver DRV2N and DRV4N are connected to ADRM, only DRVN4N will define the subsystem address. This option can be used in transmitters for more than one subsystem address. The transmitter may be hard-wired for subsystem address 2 by connecting DRV1N to ADRM. If now DRV3N is added to ADRM by a key or a switch, the transmitted subsystem address changes to 4.

A change of the subsystem address will not start a transmission.

Remote Control Signal Output (REMO)

The REMO signal output stage is a push-pull type. In the HIGH state a bipolar emitter-follower allows a high output current. The timing of the data output format is listed in Tables 1 and 2.

The information is defined by the distance t_b between the leading edges of the flashed pulses or the first edge of the modulated pulses (see Figure 3).

The format of the output data is given in Figures 2 and 3. In the flashed transmission mode, the data word starts with two toggle bits, T1 and T0, followed by three bits for defining the subsystem address S2, S1 and S0, and six bits F, E, D, C, B and A, which are defined by the selected key.

In the modulated transmission mode the first toggle bit, T1, is replaced by a constant reference time bit (REF). This can be used as a reference time for the decoding sequence.

The toggle bits function as an indication for the decoder that the next instruction has to be considered as a new command.

The codes for the subsystem address and the selected key are given in Tables 3 and 4.

Oscillator Input/Output (OSCI and OSCO)

The external components must be connected to these pins when using an oscillator with a ceramic resonator. The oscillator frequency may vary between 400kHz and 500kHz as defined by the resonator.

FUNCTIONAL DESCRIPTION

Keyboard Operation

In the standby mode all drivers (DRV0N to DRV6N) are on. Whenever a key is pressed, one or more of the sense inputs (SENnN) are tied to ground. This will start the power-up sequence. First the oscillator is activated and after the debounce time t_{DB} (see Figure 4) the output drivers (DRV0N to DRV6N) become active successively.

Within the first scan cycle the transmission mode, the applied subsystem address and the selected command code are sensed and loaded into an internal data latch. In contradiction to the command code the subsystem address is sensed only within the first scan cycle. If the applied subsystem address is changed while the command key is pressed, the transmitted subsystem address is not altered.

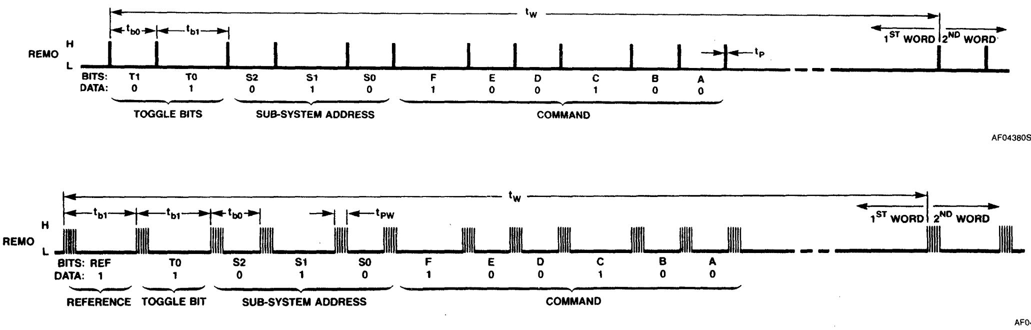
In a multiple keystroke sequence (see Figure 5), the command code is always altered in accordance with the sensed key.

Multiple Keystroke Protection

The keyboard is protected against multiple keystrokes. If more than one key is pressed at the same time, the circuit will not generate a new output at REMO (see Figure 5). In case of a multiple keystroke the scan repetition rate is increased to detect the release of a key as soon as possible.

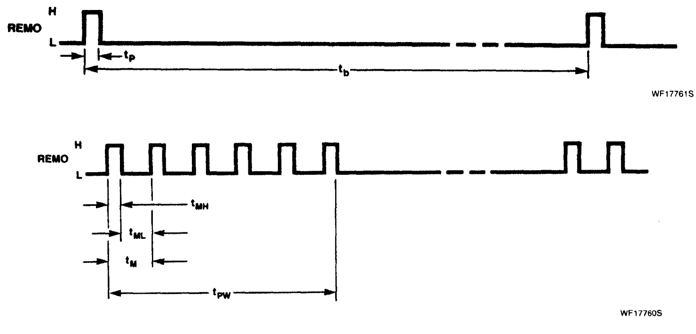
There are two restrictions caused by the special structure of the keyboard matrix:

- The keys switching to ground (code numbers 7, 15, 23, 31, 39, 47, 55 and 63) and the keys connected to SEN5N and SEN6N are not covered completely by the multiple key protection. If one sense input is switched to ground, further keys on the same sense line are ignored.
- SEN5N and SEN6N are not protected against multiple keystroke on the same driver line, because this condition has been used for the definition of additional codes (code numbers 56 to 63).



NOTES:
a. Flashed mode: transmission with 2 toggle bits and 3 address bits, followed by 6 command bits (pulses are flashed).
b. Modulated mode: transmission with reference time, toggle bit and 3 address bits, followed by 6 command bits (pulses are modulated).

Figure 2. Data Format of REMO Output; REF = Reference Time; T0 and T1 = Toggle Bits; S0, S1 and S2 = System Address; A, B, C, D, E, and F = Command Bits



NOTES:
1. Flashed pulse.
2. Modulated pulse (t_{PW} = (5 × t_M) + t_{MH}.

Figure 3. REMO Output Waveform

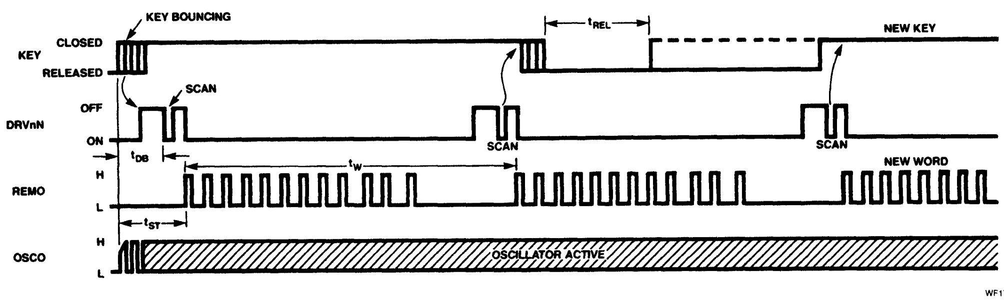
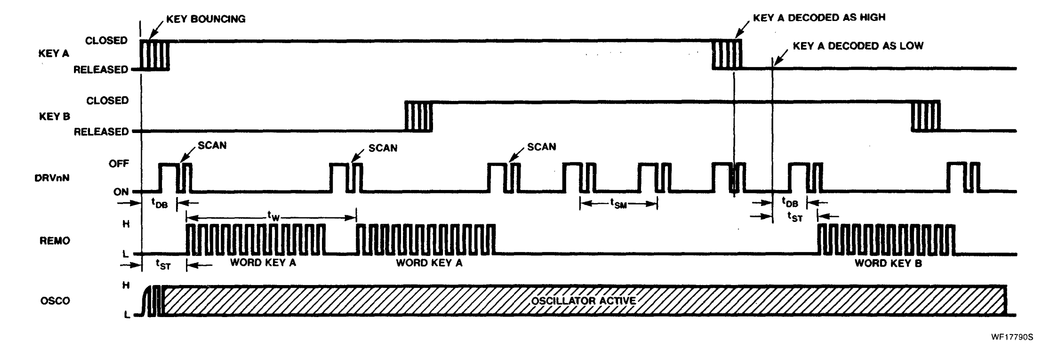


Figure 4. Single Key-Stroke Sequence

Output Sequence (Data Format)

The output operation will start when the selected code is found. A burst of pulses, including the latched address and command codes, is generated at the output REMO as long as a key is pressed. The format of the output pulse train is given in Figures 2 and 3. The operation is terminated by releasing the key or if more than one key is pressed at the same time. Once a sequence is started, the transmitted words will always be completed after the key is released.

The toggle bits T0 and T1 are incremented if the key is released for a minimum time t_{REL} (see Figure 4). The toggle bits remain unchanged within a multiple keystroke sequence.



NOTES:
1. Scan rate multiple key-stroke: $t_{SM} = 6$ to $10 \times t_O$.
2. For t_{DB}, t_{ST} and t_W see Figure 4.

Figure 5. Multiple Key-Stroke Sequence

Table 1. Pulse Train Timing

MODE	t_O (ms)	t_P (μs)	t_M (μs)	t_{ML} (μs)	t_{MH} (μs)	t_W (ms)
Flashed	2.53	8.8				121
Modulated	2.53		26.4	17.6	8.8	121

NOTES:

f_{OSC}	455kHz	$t_{OSC} = 2.2\mu s$
t_P	$4 \times t_{OSC}$	Flashed pulse width
t_M	$12 \times t_{OSC}$	Modulation period
t_{ML}	$8 \times t_{OSC}$	Modulation period LOW
t_{MH}	$4 \times t_{OSC}$	Modulation period HIGH
t_O	$1152 \times t_{OSC}$	Basic unit of pulse distance
t_W	$55\ 296 \times t_{OSC}$	Word distance

Table 2. Pulse Train Separation (t_B)

CODE	t_B
Logic "0"	$2 \times t_O$
Logic "1"	$3 \times t_O$
Reference time	$3 \times t_O$
Toggle bit time	$2 \times t_O$ or $3 \times t_O$

Table 3. Transmission Mode and Subsystem Address Election

MODE	SUBSYSTEM ADDRESS				DRIVER DRVnN FOR n =						
	#	S2	S1	S0	0	1	2	3	4	5	6
F	0	1	1	1							
L	1	0	0	0	o						
A	2	0	0	1	X	o					
S	3	0	1	0	X	X	o				
H	4	0	1	1	X	X	X	o			
E	5	1	0	0	X	X	X	X	o		
D	6	1	0	1	X	X	X	X	X	o	
M											
O	0	1	1	1							o
D	1	0	0	0	o						o
U	2	0	0	1	X	o					o
L	3	0	1	0	X	X	o				o
A	4	0	1	1	X	X	X	o			o
T	5	1	0	0	X	X	X	X	o		o
E	6	1	0	1	X	X	X	X	X	o	o
D											

NOTES:

o	= Connected to ADRM
Blank	= Not connected to ADRM
X	= Don't care

Table 4. Key Codes

MATRIX DRIVE	MATRIX SENSE	CODE F	E	D	C	B	A	MATRIX POSITION
DRV0N	SEN0N	0	0	0	0	0	0	0
DRV1N	SEN0N	0	0	0	0	0	1	1
DRV2N	SEN0N	0	0	0	0	1	0	2
DRV3N	SEN0N	0	0	0	0	1	1	3
DRV4N	SEN0N	0	0	0	1	0	0	4
DRV5N	SEN0N	0	0	0	1	0	1	5
DRV6N	SEN0N	0	0	0	1	1	0	6
V_{SS}	SEN0N	0	0	0	1	1	1	7
1	SEN1N	0	0	1		2		8 to 15
1	SEN2N	0	1	0		2		16 to 23
1	SEN3N	0	1	1		2		24 to 31
1	SEN4N	1	0	0		2		32 to 39
1	SEN5N	1	0	1		2		40 to 47
1	SEN6N	1	1	0		2		48 to 55
1	SEN5N and SEN6N	1	1	1		2		56 to 63

NOTES:
1. The complete matrix drive as shown above for SEN0N is also applicable for the matrix sense inputs SEN1N to SEN6N and the combined SEN5N/SEN6N.
2. The C, B and A codes are identical to SEN0N as given above.

The subsystem address and the transmission modes are defined by connecting the ADRM input to one or more driver outputs (DRV0N to DRV6N) of the key matrix. If more than one driver is connected to ADRM, they must be decoupled by a diode.

Signetics

Linear Products

AN1731
Low-Power Remote Control IR Transmitter and Receiver Preamplifiers

Application Note

LOW-POWER IR TRANSMITTER SAA3004

The SAA3004 is a new MOS transmitter IC for infrared remote control systems in which the received commands are decoded by a microcomputer. It can transmit up to 448 commands, divided into 7 subsystem groups of 64 commands each and is therefore suitable for single or multi-system use. To allow remote control systems with a variety of ranges, noise immunities, and costs to be built, two operating modes are available: unmodulated (single pulse per bit) or modulated (burst of 6 pulses per bit). The subsystem address and mode of operation may be selected by keyboard contacts for multi-system use, or may be hard-wired for single system use. The output from the SAA3004 is Pulse Distance Modulated (PDM) for maximum power economy and the high level of output current available (40mA with a 6V supply) allows the IC to drive an IR LED via a very simple amplifier using a single external transistor.

Compared with earlier IR transmitter ICs, the SAA3004 operates over a much wider supply voltage range (4V to 11V), consumes less current during operation (1mA typical with a 6V supply), has a lower standby current ($< 2\mu A$), and requires a minimum number of external components. The low current consumption is largely due to the fairly low oscillator frequency (455kHz).

Transmission Formats

The formats of the two transmission modes are shown in Figure 1.

At least one complete 11-bit word is generated for each legal detected keystroke. The logic state of a bit is defined by the interval between consecutive output pulses or bursts, measured from leading edge to leading edge. The word is repeated as long as a key remains pressed. When a key is released, the transmission ceases as soon as the current word has been transmitted.

In the unmodulated mode, only one pulse per bit is generated and passed to output pin REMO. For this mode, the IR preamplifier in the receiver can be a broadband type and therefore inexpensive. However, the interference immunity and range of the remote control will not be as high as that for a transmitter in the modulated mode in conjunction with a narrow-band IR receiver.

In the modulated mode, each bit is transmitted as a burst of 6 pulses at a repetition rate of about 38kHz. Since this frequency lies between the first and second harmonics of the TV line frequency, a narrow-band IR receiver tuned to 38kHz should be used in the equipment being controlled. Although such a receiver is more expensive than a broadband one, the remote control will be less sensitive to interference and will have a longer range. However, if these requirements are not stringent, a broadband receiver could also be used to receive transmissions in the modulated mode.

Remote control systems normally detect a command continuously from the moment it is received. To distinguish between multiple keystrokes and new commands, it is then necessary to detect the length of the transmitted data words. The disadvantage of this method is that a repeated command can be seen as a new one if the data stream is interrupted by an external influence. In the SAA3004, this problem is eliminated by incorporating toggle bits in the data stream. The toggle bits change state after each key release according to the truth table given in Table 1. The toggle bits therefore inform the remote control receiver that new data is arriving so that the microcomputer can easily distinguish between new data words and repeated ones. It can also count the number of identical commands if they are issued more than once in sequence. This is an important facility for selection of Teletext pages with repeated digits, resetting clock/calendars and programming VCRs.

Figure 1a is a pulse diagram of the output signal from the SAA3004 in the unmodulated mode. The data word consists of 2 toggle bits (T1 and T0), 3 address bits (S2, S1, and S0) and 6 command bits (F, E, D, C, B, and A). Toggle Bit T1 provides additional protection against interference. If the second keystroke in a sequence of three is disturbed, the decoding part of the receiver will recognize the same data twice; the fact that T1 has changed state will indicate that a new command is being transmitted.

Figure 2 shows the timing of a single bit for each transmission mode.

A complete message always consists of 12 pulses, the timing of which is directly related to the oscillator period t_{OSC}. The pulse timing data for $f_{OSC} = 455kHz$ is as follows.

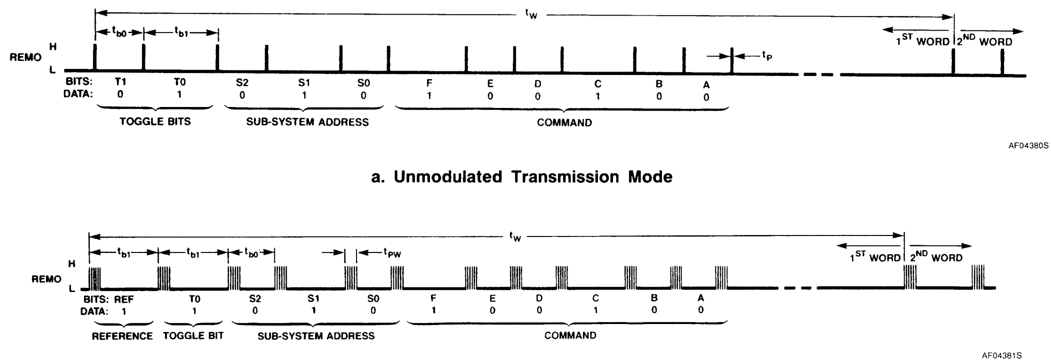
Oscillator period	$t_{OSC} = 2.2\mu s$
Pulse width	$V_{CC} = t_{MH} = 4t_{OSC} = 8.8\mu s$
Low period of modulation pulses	$t_{ML} = 8t_{OSC} = 17.6\mu s$
Modulated pulse burst period	$t_M = 12t_{OSC} = 26.4\mu s$
Duration of modulated pulse burst	$t_{PW} = 64t_{OSC} = 141\mu s$
Interval between pulses	$t_0 = 1152t_{OSC} = 2.53ms$
Data word repetition period	$t_W = 48T_0 = 121ms$
Logic '0' pulse or burst spacing	$t_{B0} = 2T_0 = 5.06ms$
Logic '1' pulse or burst spacing	$t_{B1} = 3T_0 = 7.6ms$

The data word format and timing shown in Figure 1b for the modulated mode of transmission is the same as that previously described for the unmodulated mode. In this case, however, each bit consists of a $141\mu s$ burst of 6 pulses, and toggle bit T1 is replaced by a reference pulse with a permanent logic 1, the timing of which is ($t_{REF} = t_{B1} = 7.6ms$). This allows a lower stability oscillator to be used in the transmitter because t_{REF} can be used as a reference for decoding in the equipment being controlled.

Functional Description of the SAA3004

A detailed functional block diagram of the SAA3004 is given in Figure 3 and the key sequencing diagram is given in Figure 4, which shows that, during standby, all the drive outputs are LOW. When a keystroke is detected (one or more sense inputs LOW) by the sense detector, the sequence control block enables the oscillator which starts to generate clock pulses. The oscillator increments the scan counter which, after debouncing time ($t_{DB} > 4T_0$) has elapsed, sequentially activates the drive outputs at intervals of $t_{OSC}/72$ ($158\mu s$ for $f_{OSC} = 455kHz$). See Figure 5.

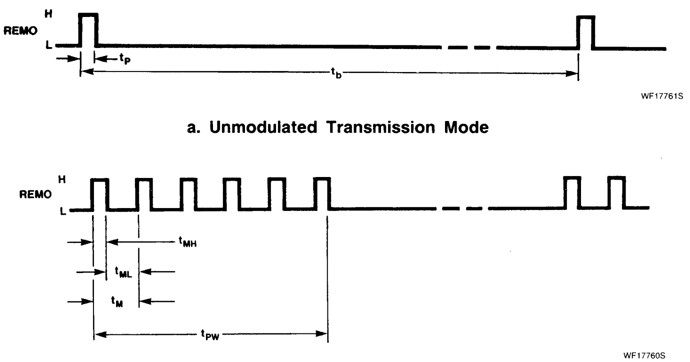
The activated key position is stored in the data memory together with the subsystem address (determined by which of the drive outputs 1 – 5 is connected to ADRM) and the output mode (whether or not drive output 6 is connected to ADRM). However, unlike the command code, the subsystem address is only sensed during the first scan cycle and does not cause any output when it is changed. The stored data, together with the toggle bits, are applied to the data multiplexer, the serial output from which is converted into the correct pulse distances by the modulation counter. The pulses are then fed to



a. Unmodulated Transmission Mode

b. Modulated Transmission Mode

Figure 1. Transmission Format at Output REMO



a. Unmodulated Transmission Mode

b. Modulated Transmission Mode

Figure 2. Timing of a Single Bit at Output REMO

output REMO via the output modulator. After a key is released, the oscillator stops and the circuits return to the standby state to conserve battery power as soon as the output sequence is completed.

The SAA3004 has built-in protection against multiple keystrokes (two or more keys pressed at a time). In this event, the IC reacts as shown in Figure 6. At the end of any current output sequence, output REMO becomes inactive, and the keyboard scanning interval t_W = 121ms is reduced to t_{SM} (about 20ms). This ensures that a key release is detected as soon as possible. Also, the toggle bits remain unchanged during multiple keystrokes.

Table 1. Sequence of Toggle Bits

KEY SEQUENCE	T0	T1
n	0	1
n+1	1	1
n+2	0	0
n+3	1	0
n+4	0	1
n+5	1	1
.	.	.
.	.	.
.	.	.

A Practical IR Transmitter

An example of a complete IR remote control transmitter is given in Figure 7.

Forty-nine of the keys (7 × 7 matrix) are connected directly between driver lines DRV0N to DRV6N and sense lines SEN0N to SEN6N. Expanding the keyboard for 64 commands is done in three steps. First, seven keys are added to switch each of the sense lines to ground. Next, seven keys are added to switch each of the drive lines to SEN5N and SEN6N via diodes D_1 and D_2. The final key is added to switch sense lines SEN5N and SEN6N to ground via diodes D_1 and D_2.

In standby, the drive lines are LOW and the sense lines are HIGH. A scan cycle starts as soon as one of the sense inputs is forced LOW by a keystroke. If the keystroke is detected as being legal (only one key pressed), the appropriate command is decoded according to the scheme in Table 2, and the correct data word is fed to output REMO. Bits ABC in Table 2 indicate which of the seven driver outputs is activated and bits DEF indicate which of the seven sense inputs has detected a LOW level.

Address mode input ADRM selects the subsystem address and determines the transmission mode (modulated or unmodulated). The subsystem address and mode of operation depend on which of the seven drive lines is connected to ADRM as shown in Table 3. The address is selected either by closing an address switch to connect a drive output to input ADRM before pressing a command key, or by installing a permanent link between one of the drive outputs and input ADRM. With no address selected, the basic address (address bits S2, S1, and S0 all equal 1) is automatically generated.

Mode selection is made via a link between drive line DRV6N and input ADRM. The

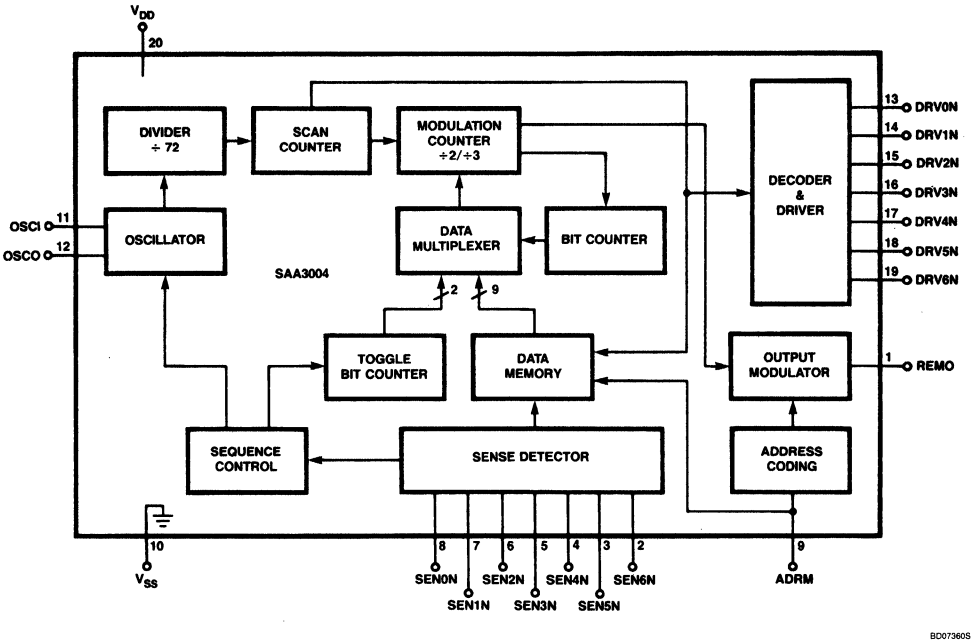
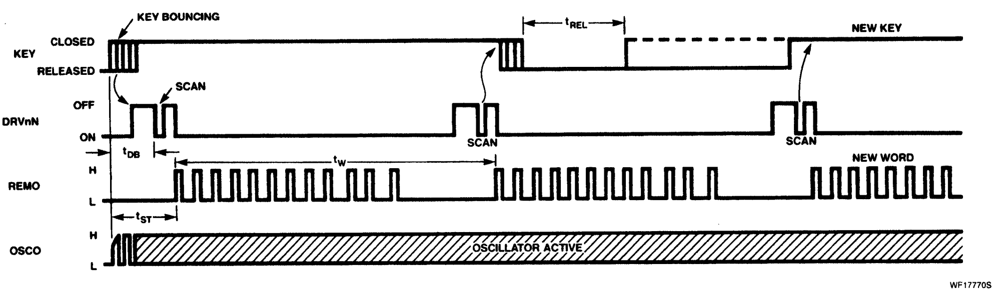


Figure 3. Block Diagram of Remote Control Transmitter SAA3004



NOTE:
$T_0 = 1152t_{OSC}$, debounce time $t_{DB} = 4$ to $9 \times t_0$, start time $t_{ST} = 5$ to $10 \times t_0$, minimum release time $t_{REL} = t_0$.

Figure 4. Single Keystroke Sequence

transmission is modulated with the link fitted or unmodulated without it.

Capacitors C_1 and C_2 associated with the oscillator must be chosen with regard to low current consumption and quick starting over the whole supply voltage range.

The output stage of the SAA3004 shown in Figure 8 provides a current output of up to 40mA with a 6V supply, sufficient to drive a very simple single transistor amplifier to provide current for an infrared LED. When the output stage is driven by a HIGH level, the NPN transistor conducts and pulls output pin REMO HIGH (3V min. with a 6V supply). When the output stage is driven by a LOW level, the NPN transistor is turned off and the n-channel output FET conducts and pulls output pin REMO LOW (200mV maximum with a 6V supply). In this state, the output stage can sink a typical current of 300μA.

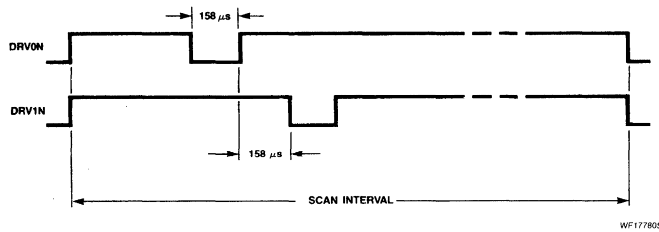
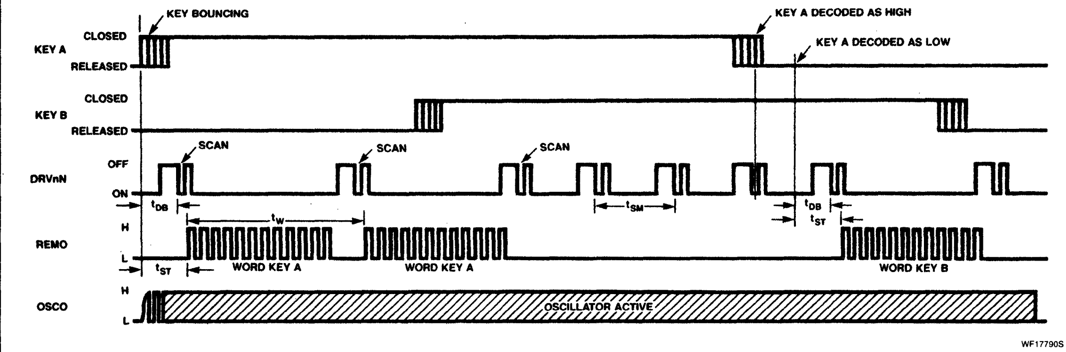


Figure 5. Timing at Outputs DRV0N to DRV6N

Table 2. Key Codes

MATRIX POS.	CODE F	E	D	C	B	A	MATRIX POS.	CODE F	E	D	C	B	A
0	0	0	0	0	0	0	32	1	0	0	0	0	0
1	0	0	0	0	0	1	33	1	0	0	0	0	1
2	0	0	0	0	1	0	34	1	0	0	0	1	0
3	0	0	0	0	1	1	35	1	0	0	0	1	1
4	0	0	0	1	0	0	36	1	0	0	1	0	0
5	0	0	0	1	0	1	37	1	0	0	1	0	1
6	0	0	0	1	1	0	38	1	0	0	1	1	0
7	0	0	0	1	1	1	39	1	0	0	1	1	1
8	0	0	1	0	0	0	40	1	0	1	0	0	0
9	0	0	1	0	0	1	41	1	0	1	0	0	1
10	0	0	1	0	1	0	42	1	0	1	0	1	0
11	0	0	1	0	1	1	43	1	0	1	0	1	1
12	0	0	1	1	0	0	44	1	0	1	1	0	0
13	0	0	1	1	0	1	45	1	0	1	1	0	1
14	0	0	1	1	1	0	46	1	0	1	1	1	0
15	0	0	1	1	1	1	47	1	0	1	1	1	1
16	0	1	0	0	0	0	48	1	1	0	0	0	0
17	0	1	0	0	0	1	49	1	1	0	0	0	1
18	0	1	0	0	1	0	50	1	1	0	0	1	0
19	0	1	0	0	1	1	51	1	1	0	0	1	1
20	0	1	0	1	0	0	52	1	1	0	1	0	0
21	0	1	0	1	0	1	53	1	1	0	1	0	1
22	0	1	0	1	1	0	54	1	1	0	1	1	0
23	0	1	0	1	1	1	55	1	1	0	1	1	1
24	0	1	1	0	0	0	56	1	1	1	0	0	0
25	0	1	1	0	0	1	57	1	1	1	0	0	1
26	0	1	1	0	1	0	58	1	1	1	0	1	0
27	0	1	1	0	1	1	59	1	1	1	0	1	1
28	0	1	1	1	0	0	60	1	1	1	1	0	0
29	0	1	1	1	0	1	61	1	1	1	1	0	1
30	0	1	1	1	1	0	62	1	1	1	1	1	0
31	0	1	1	1	1	1	63	1	1	1	1	1	1



NOTE:
$t_0 = 1152 t_{OSC}$, debounce time $t_{DB} = 4$ to $9 \times t_0$, scan rate $t_{SM} = 6$ to $10 \times t_0$.

Figure 6. Multiple Keystroke Sequence

Table 3. Transmission Mode and Subsystem Address Selection

OUTPUT FORMAT	SUBSYSTEM ADDRESS				DRIVE OUTPUT DRVnN n =						
	No.	S2	S1	S0	0	1	2	3	4	5	6
unmodulated	1	1	1	1							
	2	0	0	0	X						
	3	0	0	1	–	X					
	4	0	1	0	–	–	X				
	5	0	1	1	–	–	–	X			
	6	1	0	0	–	–	–	–	X		
	7	1	0	1	–	–	–	–	–	X	
modulated	1	1	1	1							X
	2	0	0	0	X						X
	3	0	0	1	–	X					X
	4	0	1	0	–	–	X				X
	5	0	1	1	–	–	–	X			X
	6	1	0	0	–	–	–	–	X		X
	7	1	0	1	–	–	–	–	–	X	X

NOTES:
X Connected to ADRM.
– Allowed connection to ADRM without any influence on the subsystem address.

Power Consumption Considerations

The intensity of IR radiation I_E, and therefore the transmitter range, is proportional to the LED forward current I_F. The peak value of I_F in the circuit of Figure 7 is determined by the value of emitter resistor R_E and is given by:

$$I_F = (V_{REF} - V_{BE})/R_E.$$

However, since the output is pulsed, the battery life is mainly determined by the average value of the forward current. This average LED current is the peak current multiplied by the duty factor of the output signal. The duty factor is the ratio of the total HIGH time of a data word (12 pulses each of width $T_P = 8.8\mu s$) to the data word repetition period ($t_W = 121$ms).

In the unmodulated mode, the average LED current is:

$$I_{Fav} = I_F(12t_P/t_W) = 8.7 I_F \times 10^{-4}.$$

In the modulated mode, each pulse is a burst of six $8.8\mu s$ pulses. The total HIGH time of a data word is therefore six times that for the unmodulated mode so that the duty factor is multiplied by six.

In the modulated mode, the average LED current is therefore:

$$I_{Fav} = 52 I_F \times 10^{-4}.$$

At first glance, the higher required average current for the modulated mode makes it appear unattractive because of increased battery drain. However, if a narrow-band receiver is used with a modulated transmitter,

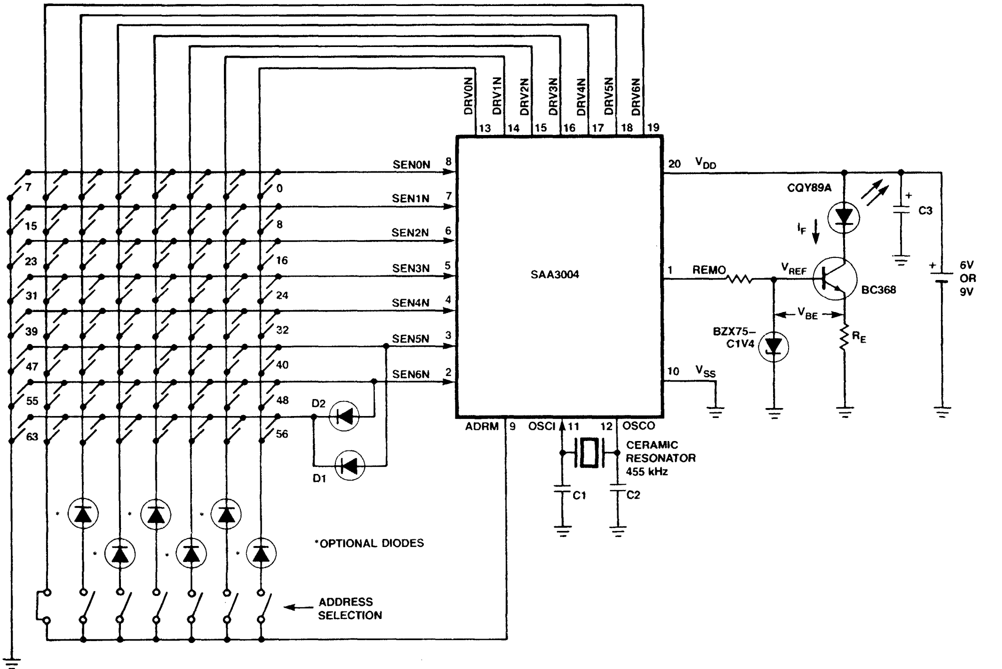


Figure 7. A Complete Remote Control Infrared Transmitter Using SAA3004

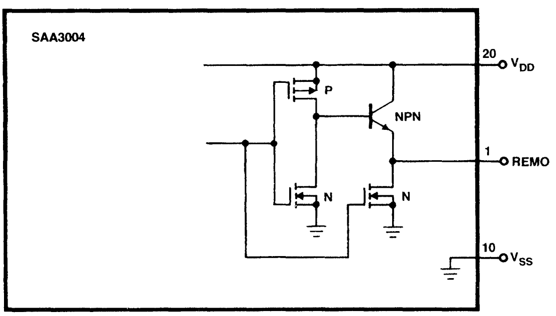


Figure 8. Output Stage of the SAA3004

this will not be the case because the resonance peak of the tuned circuit at the input makes a narrow-band receiver more sensitive to infrared radiation and less sensitive to interference than a broadband receiver. For a given remote control range, then, the required forward current for the transmitter LED is less than that required for an LED in an unmodulated transmitter used with a broadband receiver. This is confirmed by the range measurement results given at the end of this publication.

The total current drain from the battery when the transmitter is in use is the sum of I_{Fav}, the very small leakage current of the battery buffer electrolytic capacitor C_3, and the current drain of the SAA3004 (typically 1mA with a 6V supply or 3mA with a 9V supply). During standby, the maximum current drain of the SAA3004 is 2μA, regardless of the supply voltage.

INFRARED RECEIVER PREAMPLIFIERS TDA3047 AND TDA3048

The TDA3047 and TDA3048 are bipolar preamplifier ICs for infrared remote control receivers. The ICs differ only in the polarity of the output signal; the TDA3047 is active HIGH and the TDA3048 is active LOW. This choice of polarity allows the preamplifier IC to be selected to suit the microprocessor in the system being controlled. For example, if an 8048 microprocessor is used on interrupt level (active-LOW input INT), the TDA3048 is the correct choice. Power consumption of the ICs is only 10mW from a 5V supply, which is

Low-Power Remote Control IR Transmitter and Receiver Preamplifiers

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considerably less than that of earlier preamplifier ICs. Operation from a 5V supply means that the preamplifiers can use the same supply as the microprocessor in the equipment being controlled.

Both ICs are excellent for use in narrow-band IR receivers which are necessary to achieve high noise immunity and long range for the reception of a modulated data stream. The ICs can also be used in inexpensive broadband IR receivers for the reception of unmodulated data or modulated data if noise immunity and long range are not of major importance.

The 66dB AGC range of the ICs ensures stable amplification of a wide range of signal levels, thus allowing remote-control systems to operate over a wide range of transmitter-to-receiver distances.

The ICs in a Narrow-Band IR Receiver

The functional block diagram of the TDA3047/48 in a narrow-band IR receiver is shown in Figure 9. Figure 10 shows some of the internal circuitry connected to the IC pins.

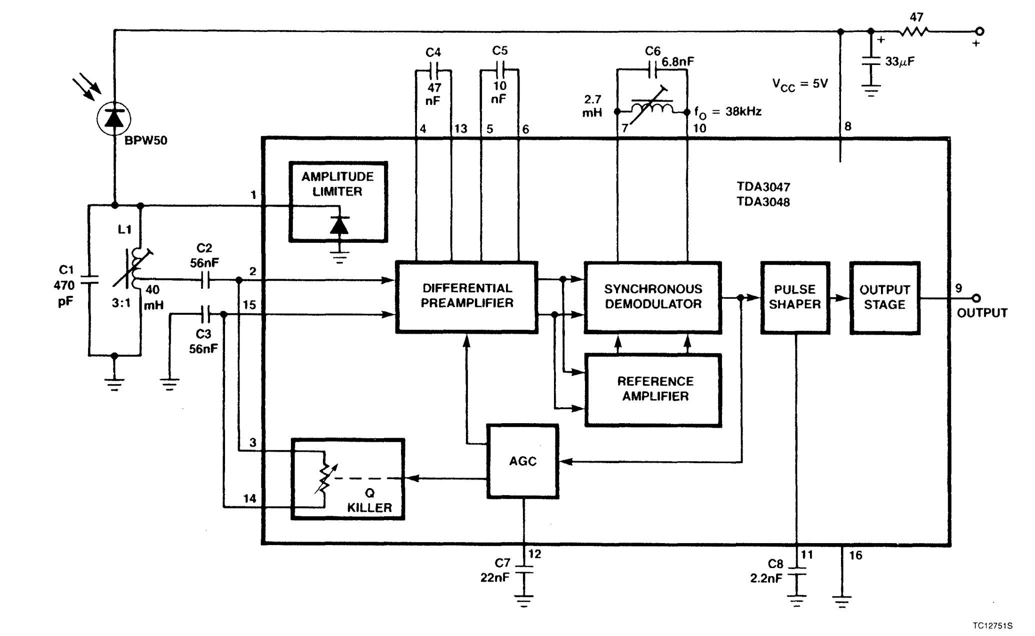
The input signal from the photodiode is coupled to input Pins 2 and 15 via a 38kHz parallel tuned circuit with a Q of about 10 giving a bandwidth of about 3kHz. This considerably improves selectivity and attenuates continuous IR interference caused, for example, by sunlight. The low resistance of L_1 (125Ω) ensures that the photodiode never saturates. The tapping point for the coil (3:1) is chosen to match the input resistance of the IC (16kΩ) and is optimum for low-level signals (Q-killer inactive) so that the operating range of the remote-control system remains almost independent of component value spreads or frequency tolerance in either the transmitter or the receiver.

Alternatively, L_1 could be capacitively tapped as shown in Figure 11. The total capacitance of C_{1a} and C_{1b} must be that required to tune the circuit to 38kHz (470pF with a 40mH coil). The ratio C_{1a}/C_{1b} must be 3:1. Values of 2.2nF for C_{1a} and 560pF for C_{1b} meet these requirements and give about the same Q as the input tuned circuit given in Figure 9.

The signal from the tuned circuit is capacitively-coupled to Pins 2 and 15 of the IC and is then amplified by an internal two-stage gain-controlled differential amplifier. The first stage of the differential amplifier has a maximum gain of 56dB, and the second stage has a maximum gain of 26dB, giving overall gain of more than 80dB. Feedback capacitors C_4 and C_5 stabilize the first and second stage, respectively. Together, they set the lower frequency limit of the circuit, C_4 having the most effect because the first stage has the higher gain. The values of both capacitors should be chosen such that IR interference is suppressed, bearing in mind that incandescent lamps radiate IR at multiples of 100Hz. The upper frequency limit of the amplifier is set by internal capacitance and is above 1MHz.

The amplified signal is fed to a synchronous demodulator and a reference amplifier that limits high amplitude input signals. The 2.7mH coil in the 38kHz demodulator tuned circuit has a Q of about 7 in conjunction with the resistance between Pins 7 and 10 (6kΩ).

After multiplication of the input and reference signals, the demodulated signal is fed to a pulse shaper and an AGC circuit. A Q-killer in the AGC loop damps the Q of the input tuned circuit for high level inputs so that the circuit can handle large variations of signal amplitude. An absolute maximum input level of about 600mV is set by the limiter at Pin 1. The AGC acquisition time and the time constant of the pulse shaper are determined by C_7 at Pin



NOTE:
Capacitor C_5 should be kept well clear of the input pins.

Figure 9. TDA3047/3048 in a Narrow-Band IR Receiver

12 and C_8 at Pin 11, respectively. The time constant at Pin 12 is equal to the duration of one data bit. The time constant at Pin 11 sets the delay between the pulse shaper and the output stage. The value of C_8 must be low enough to ensure that, with a charging time of one pulse width (8.8µs from the SAA3004 transmitter), the threshold of the pulse shaper (about 4V) can be exceeded. If the value of C_8 is too low, however, short duration interference pulses can easily trigger the pulse shaper. The value of C_8 is therefore a compromise between the receiver sensitivity and immunity to interference.

The ICs in a Broadband IR Receiver

The TDA3047 and TDA3048 are shown in a broadband IR receiver circuit in Figure 12. This circuit is similar to the previously described narrow-band receiver except that the Q-Killer and amplitude limiter are not necessary. (Pins 1, 3, 14 are not used.) Also, the IR photodiode is simply connected between two 12kΩ load resistors and connected to the IC inputs via 10nF capacitors instead of via a tuned circuit.

CONTROL SYSTEM RANGE MEASUREMENTS

Measurements have been made with both IR receivers in conjunction with an IR transmitter based on the SAA3004 to determine the operating range.

As previously explained, when the SAA3004 transmitter in the unmodulated mode drives a single infrared LED with a constant peak forward current I_F of 2A, the average current, which is proportional to the infrared radiation, is:

$$I_{Fav} = 8.7 I_F \times 10^{-4} = 1.7\text{mA}.$$

Under these conditions, the range of the remote-control was 11m with a narrow-band receiver and 12m with a broadband receiver.

Under the same conditions in the modulated mode, the average current is:

$$I_{Fav} = 52 I_F \times 10^{-4} = 10.4\text{mA}.$$

Under these conditions, the range of the remote-control was 25m with a narrow-band receiver and 16m with a broadband receiver.

To allow direct comparison between the two transmission modes, the average LED current for the modulated mode was reduced to 1.7mA. Under these conditions, the range of the remote-control was 11m with a narrow-band receiver and 8m with a broadband receiver.

Originally published as Technical Publication 167, March 22, 1985, The Netherlands.

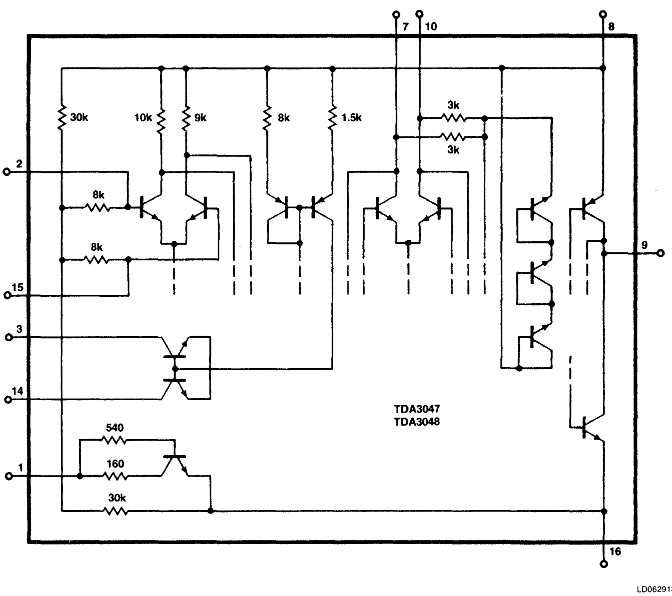


Figure 10. Internal Connectors to Pins of the TDA3047/3048

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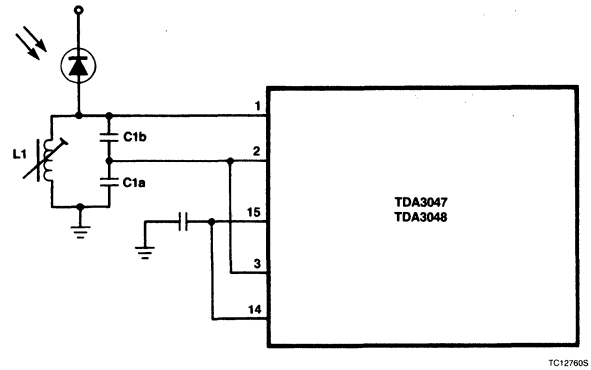


Figure 11. Alternative Input Coupling for a Narrow-Band IR Receiver

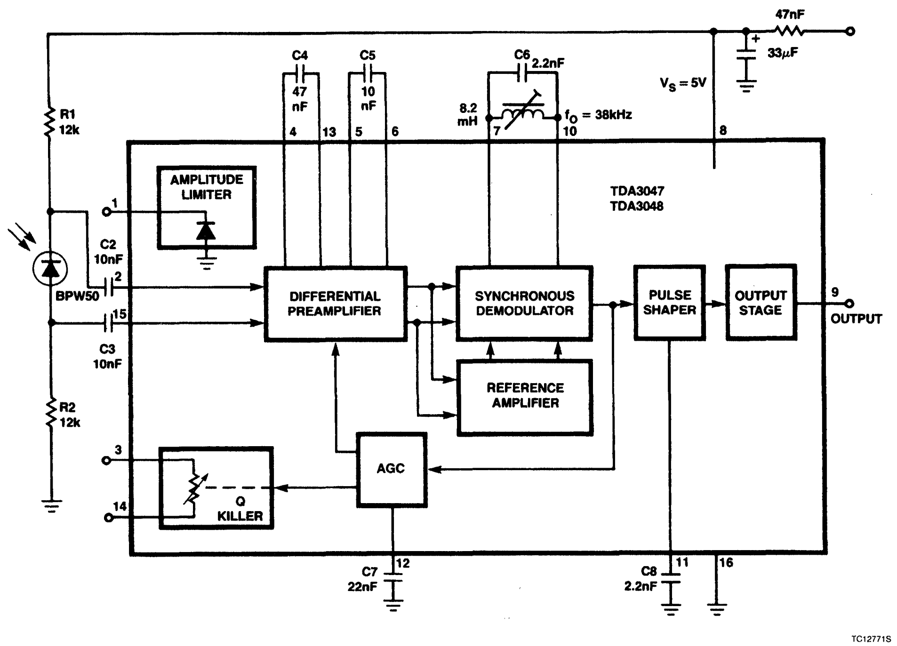


Figure 12. TDA3047/3048 in a Broadband IR Receiver

SAA3006
Infrared Transmitter

Product Specification

Linear Products

DESCRIPTION

The SAA3006 is intended as a general purpose (RC-5) infrared remote control system for use where only low supply voltages are available. The device can generate 2048 different commands and utilizes a keyboard with a single-pole switch per key. The commands are arranged so that 32 systems can be addressed, each system containing 64 different commands.

The circuit response to legal (one key pressed at a time) and illegal (more than one key pressed at a time) keyboard operation is specified later in this publication (see KEY ACTIVITIES).

FEATURES

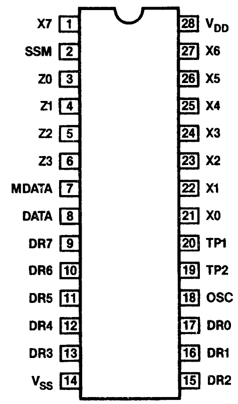
- **Low supply voltage requirements**
- **Very low current consumption**
- **For infrared transmission link**
- **Transmitter for 32 × 64 commands**
- **One transmitter controls 32 systems**
- **Transmission biphase technique**
- **Short transmission times; speed-up of system reaction time**
- **Single-pin oscillator input**
- **Input protection**
- **Test mode facility**

APPLICATIONS

- **Audio**
- **TV**

PIN CONFIGURATION

N Package



TOP VIEW

CD12050S

PIN NO.	SYMBOL	DESCRIPTION
1	X7	Keyboard command inputs with P-channel pull-up transistors
21	X0	
22	X1	
23	X2	
24	X3	
25	X4	
26	X5	
27	X6	
2	SSM	System mode selection input
3	Z0	Keyboard system inputs with P-channel pull-up transistors
4	Z1	
5	Z2	
6	Z3	
7	MDATA	Remote signal outputs (3-state outputs)
8	DATA	
9	DR7	Scan driver outputs with open-drain N-channel transistors
10	DR6	
11	DR5	
12	DR4	
13	DR3	
15	DR2	
16	DR1	
17	DR0	
14	V_{SS}	Negative supply (ground)
18	OSC	Oscillator input
19	TP2	Test input/output
20	TP1	Test input
28	V_{DD}	Positive supply

ORDERING INFORMATION

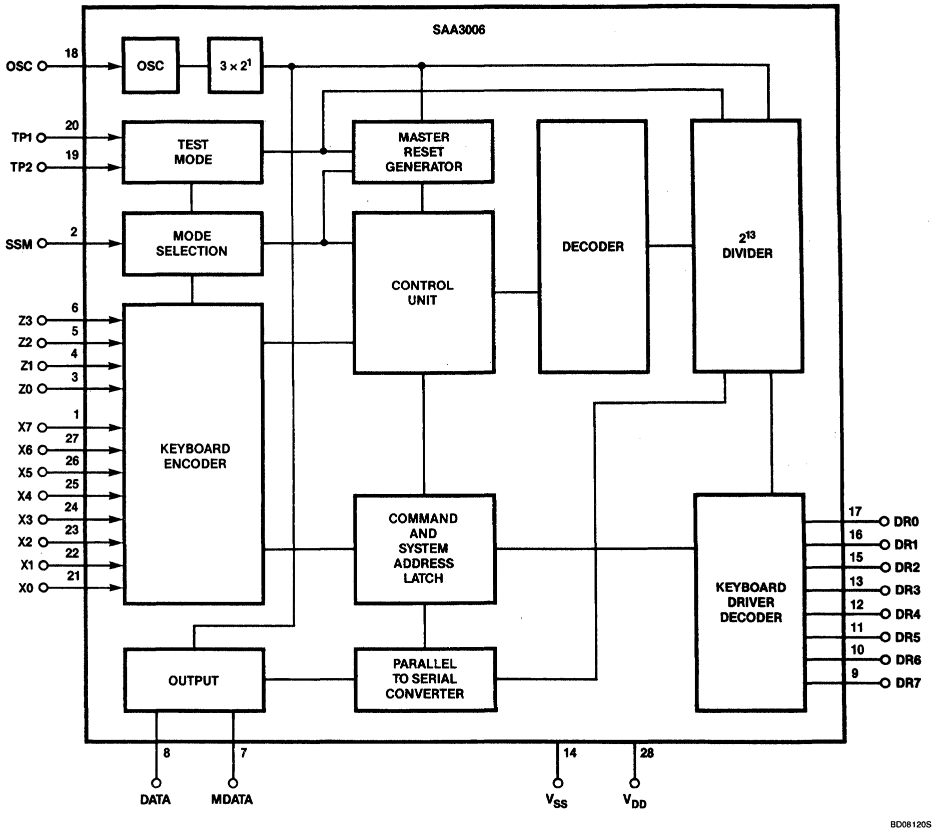
DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
28-Pin Plastic DIP (SOT-117)	−25°C to +85°C	SAA3006PN

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{DD}	Supply voltage range with respect to V_{SS}	−0.5 to +8.5	V
V_I	Input voltage range	−0.5 to (V_{DD} + 0.5)	V[1]
$+I_I$	Input current	10	mA
V_O	Output voltage range	−0.5 to (V_{DD} + 0.5)	V[1]
$+I_O$	Output current	10	mA
P_O	Power dissipation output OSC	50	mW
P_O	Power dissipation per output (all other outputs)	100	mW
P_{TOT}	Total power dissipation per package	200	mW
T_A	Operating ambient temperature range	-25 to +85	°C
T_{STG}	Storage temperature range	-65 to +150	°C

NOTE:
1. V_{DD}+0.5V not to exceed 9V.

BLOCK DIAGRAM



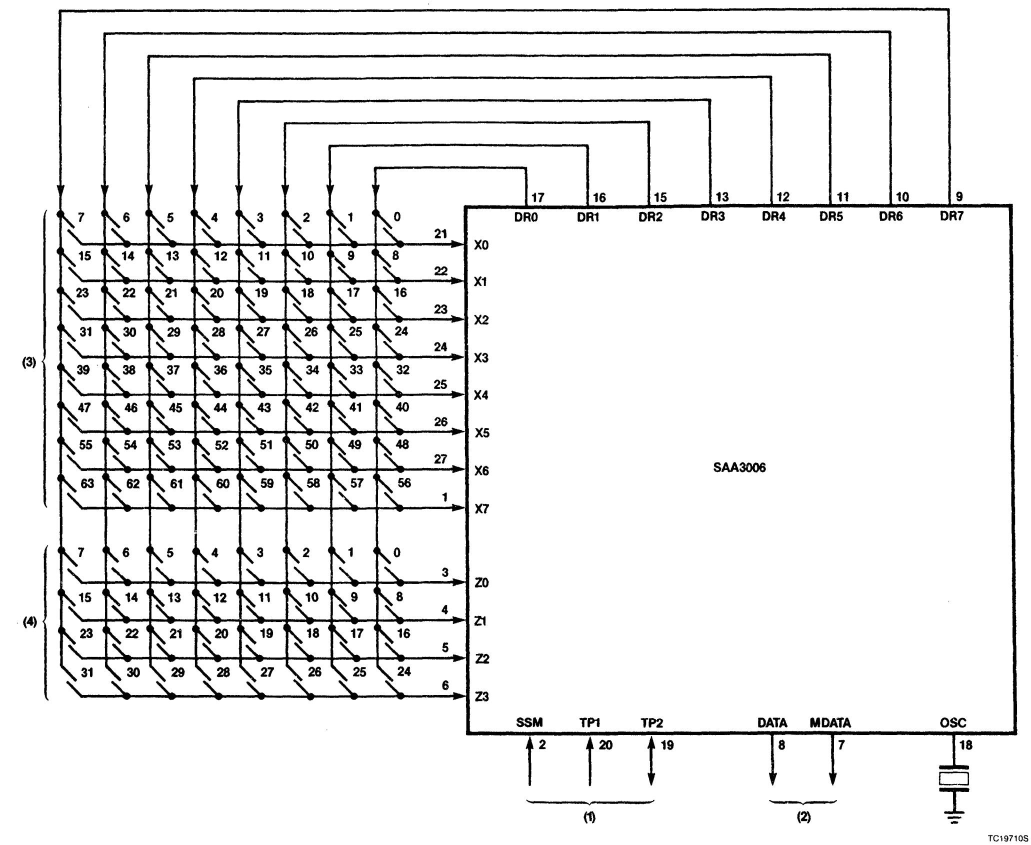
DC ELECTRICAL CHARACTERISTICS $V_{SS} = 0V$; $T = -25$ to $85°C$, unless otherwise specified.

SYMBOL	PARAMETER	V_{DD} (V)	LIMITS			UNIT
			Min	Typ	Max	
V_{DD}	Supply voltage		2		7	V
	Supply current at $I_O = 0mA$ for all outputs; X0 to X7 and Z3 at V_{DD}; all other inputs at V_{DD} or V_{SS}; excluding leakage current from open-drain N-channel outputs					
I_{DD}	$T_A = 25°C$	7			10	μA
Inputs Keyboard inputs X and Z with P-channel pull-up transistors						
$-I_I$	Input current (each input) at $V_I = 0V$; TP = SSM = LOW	2 to 7	10		600	μA
V_{IH}	Input voltage HIGH	2 to 7	$0.7 \times V_{DD}$		V_{DD}	V
V_{IL}	Input voltage LOW	2 to 7	0		$0.3 \times V_{DD}$	V
I_{IR} $-I_{IR}$	Input leakage current at $T_A = 25°C$; TP = HIGH; $V_I = 7V$ $V_I = 0V$				1 1	μA μA
SSM, TP1 and TP2						
V_{IH}	Input voltage HIGH	2 to 7	$0.7 \times V_{DD}$		V_{DD}	V
V_{IL}	Input voltage LOW	2 to 7	0		$0.3 \times V_{DD}$	V
I_{IR} $-I_{IR}$	Input leakage current at $T_A = 25°C$; $V_I = 7V$ $V_I = 0V$				1 1	μA μA
OSC						
$-I_I$	Input leakage current at $T_A = 25°C$; $V_I = 0V$; TP1 = HIGH; Z2 = Z3 = LOW	2 to 7			2	μA
Outputs DATA and MDATA						
V_{OH}	Output voltage HIGH at $-I_{OH} = 0.4mA$	2 to 7	$V_{DD} - 0.3$			V
V_{OL}	Output voltage LOW at $I_{OL} = 0.6mA$	2 to 7			0.3	V
I_{OR} $-I_{OR}$	Output leakage current at: $V_O = 7V$ $V_O = 0V$				10 20	μA μA
I_{OR} $-I_{OR}$	$T_A = 25°C$; $V_O = 7V$ $V_O = 0V$				1 2	μA μA
DR0 to DR7, TP2						
V_{OL}	Output voltage LOW at $I_{OL} = 0.3mA$	2 to 7			0.3	V
I_{OR} I_{OR}	Output leakage current at $V_O = 7V$ at $V_O = 7V$; $T_A = 25°C$	7			10 1	μA μA

5

DC ELECTRICAL CHARACTERISTICS (Continued) $V_{SS} = 0V$; $T = -25$ to $85°C$, unless otherwise specified.

SYMBOL	PARAMETER	V_{DD} (V)	LIMITS Min	Typ	Max	UNIT
OSC						
I_{OSC}	Oscillator current at OSC = V_{DD}	7	4.5		30	μA
Oscillator						
f_{OSC}	Maximum oscillator frequency at $C_L = 40pF$ (Figures 4 and 5)	2			450	kHz
f_{OSC}	Free-running oscillator frequency at $T_A = 25°C$	2	10		120	kHz



NOTES:
1. Control inputs for operating modes, test modes and reset.
2. Remote signal outputs.
3. Keyboard command code matrix 8 × 8.
4. Keyboard system code matrix 4 × 8.

Figure 1. Keyboard Interconnection

FUNCTIONAL DESCRIPTION

Combined System Mode (SSM = LOW)

The X and Z lines are active-HIGH in the quiescent state. Legal key operation either in the X-DR or Z-DR matrix starts the debounce cycle. When the contact is made for two bit times without interruption, the oscillator enable signal is latched and the key may be released. Interruption within the two bit times resets the internal action. At the end of the debounce time, the DR outputs are switched off and two scan cycles are started, switching on the DR-outputs one by one. When a Z or X input senses a LOW level, a latch enable signal is fed to the system address or command latches, depending on whether sensing was found in the Z or X input matrix. After latching a system address number, the device will generate the last command (i.e., all command bits '1') in the chosen system as long as the key is pressed. Latching of a command number causes the device to generate this command together with the system address number stored in the system address latch. Releasing the key will reset the internal action if no data is transmitted at that time. Once the transmission is started, the signal will be finished completely.

Single System Mode (SSM = HIGH)

The X lines are active-HIGH in the quiescent state; the pull-up transistors of the Z lines are switched off and the inputs are disabled. Only legal key operation in the X-DR matrix starts the debounce cycle. When the contact is made for two bit times without interruption, the oscillator enable signal is latched and the key may be released. Interruption within the two bit times resets the internal action. At the end of the debounce time, the pull-up transistors in the X lines are switched off. Those in the Z lines are switched on during the first scan cycle. The wired connection in the Z matrix is then translated into a system address number and stored in the system address latch. At the end of the first scan cycle the pull-up transistors in the Z lines are switched off and the inputs are disabled again, while the transistors in the X lines are switched on. The second scan cycle produces the command number which, after latching, is transmitted together with the system address number.

Inputs

The command inputs X0 to X7 carry a logical '1' in the quiescent state by means of an internal pull-up transistor. When SSM is LOW, the system inputs Z0 to Z3 also carry a logical '1' in the quiescent state by means of an internal pull-up transistor.

When SSM is HIGH, the transistors are switched off and no current flows via the wired connection in the Z-DR matrix.

Oscillator

The oscillator is formed by a ceramic resonator (cataloq number 2422 540 98021 or equivalent) feeding the single-pin input OSC. Direct connection is made for supply voltages in the range 2 to 5.25V but it is necessary to fit a 10kΩ resistor in series with the resonator when using supply voltages in the range 2.6 to 7V.

Key Release Detection

An extra control bit is added which will be complemented after key release. In this way the decoder gets an indication that shows if the next code is to be considered as a new command. This is very important for multi-digit entry (e.g., by channel numbers or Teletext/Viewdata pages). The control bit will only be complemented after finishing at least one code transmission. The scan cycles are repeated before every code transmission, so that, even by 'takeover' of key operation during the code transmission, the correct system and command numbers are generated.

Outputs

The output DATA carries the generated information according to the format given in Figure 2 and Tables 2 and 3. The code is transmitted in biphase; definitions of logical '1' and '0' are given in Figure 3.

The code consists of four parts:

- Start part formed by 2 bits (two times a logical '1')
- Control part formed by 1 bit
- System part formed by 5 bits
- Command part formed by 6 bits.

The output MDATA carries the same information as output DATA but is modulated on a carrier frequency of 1/12 the oscillator frequency, so that each bit is presented as a burst of 32 pulses. To reduce power consumption, the carrier frequency has a 25% duty cycle.

In the quiescent state, both outputs are non-conducting (3-state outputs). The scan drivers DR0 to DR7 are of the open-drain N-channel type and are conducting in the quiescent state of the circuit. After a legal key operation all the driver outputs go into the high ohmic state; a scanning procedure is then started so that the outputs are switched into the conducting state one after the other.

Reset Action

The circuit will be reset immediately when a key release occurs during:

- Debounce time
- Between two codes.

When a key release occurs during scanning of the matrix, a reset action will be accomplished if:

- The key is released while one of the driver outputs is in the low-ohmic '0' state
- The key is released before detection of that key
- There is no wired connection in the Z-DR matrix while SSM is HIGH.

Test Pin

The test pins TP1 and TP2 are used for testing in conjunction with inputs Z2 and Z3 as shown in Table 1.

Table 1. Test Functions

TP1	TP2	Z2	Z3	FUNCTION
LOW	LOW	Matrix input	Matrix input	Normal
LOW	HIGH	Matrix input	Matrix input	Scan + output frequency 6 times faster than normal
HIGH	Output f_{OSC}^6	LOW	LOW	Reset
HIGH	Output f_{OSC}^6	HIGH	HIGH	Output frequency 3×2^7 faster than normal

5

KEY ACTIVITIES

Every connection of one X input and one DR output is recognized as a legal keyboard operation and causes the device to generate the corresponding code.

Activating more than one X input at a time is an illegal keyboard operation and no circuit action is taken (oscillator does not start).

When SSM is LOW, every connection of one Z input and one DR output is recognized as a legal keyboard operation and causes the device to generate the corresponding code.

Activating two or more Z inputs, or Z inputs and X inputs, at one time is an illegal keyboard operation and no circuit action is taken.

When SSM is HIGH, a wired connection must be made between a Z input and a DR output. If no connection is made, the code is not generated.

When one X or Z input is connected to more than one DR output, the last scan signal is considered legal.

The maximum allowable value of the contact series resistance of the keyboard switches is 7kΩ.

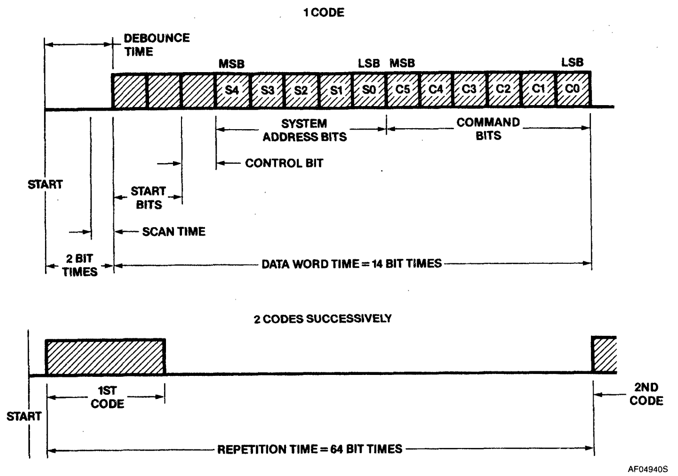
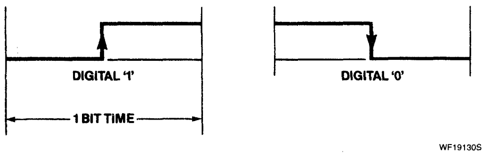


Figure 2. DATA Output Format (RC-5)



NOTE:
1. Bit time = $3 \times 2^8 \times t_{OSC}$ (typically 1.778ms) where t_{OSC} is the oscillator period time.

Figure 3. Biphase Transmission Code

Infrared Transmitter SAA3006

Table 2. Command Matrix X-DR

CODE NO	X-LINES X								DR-LINES DR								COMMAND BITS C					
	0	1	2	3	4	5	6	7	0	1	2	3	4	5	6	7	5	4	3	2	1	0
0	•								•								0	0	0	0	0	0
1	•									•							0	0	0	0	0	1
2	•										•						0	0	0	0	1	0
3	•											•					0	0	0	0	1	1
4	•												•				0	0	0	1	0	0
5	•													•			0	0	0	1	0	1
6	•														•		0	0	0	1	1	0
7	•															•	0	0	0	1	1	1
8		•							•								0	0	1	0	0	0
9		•								•							0	0	1	0	0	1
10		•									•						0	0	1	0	1	0
11		•										•					0	0	1	0	1	1
12		•											•				0	0	1	1	0	0
13		•												•			0	0	1	1	0	1
14		•													•		0	0	1	1	1	0
15		•														•	0	0	1	1	1	1
16			•						•								0	1	0	0	0	0
17			•							•							0	1	0	0	0	1
18			•								•						0	1	0	0	1	0
19			•									•					0	1	0	0	1	1
20			•										•				0	1	0	1	0	0
21			•											•			0	1	0	1	0	1
22			•												•		0	1	0	1	1	0
23			•													•	0	1	0	1	1	1
24				•					•								0	1	1	0	0	0
25				•						•							0	1	1	0	0	1
26				•							•						0	1	1	0	1	0
27				•								•					0	1	1	0	1	1
28				•									•				0	1	1	1	0	0
29				•										•			0	1	1	1	0	1
30				•											•		0	1	1	1	1	0
31				•												•	0	1	1	1	1	1

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Table 2. Command Matrix X-DR (Continued)

CODE NO	X-LINES X								DR-LINES DR								COMMAND BITS C					
	0	1	2	3	4	5	6	7	0	1	2	3	4	5	6	7	5	4	3	2	1	0
32					•				•								1	0	0	0	0	0
33					•					•							1	0	0	0	0	1
34					•						•						1	0	0	0	1	0
35					•							•					1	0	0	0	1	1
36					•								•				1	0	0	1	0	0
37					•									•			1	0	0	1	0	1
38					•										•		1	0	0	1	1	0
39					•											•	1	0	0	1	1	1
40						•			•								1	0	1	0	0	0
41						•				•							1	0	1	0	0	1
42						•					•						1	0	1	0	1	0
43						•						•					1	0	1	0	1	1
44						•							•				1	0	1	1	0	0
45						•								•			1	0	1	1	0	1
46						•									•		1	0	1	1	1	0
47						•										•	1	0	1	1	1	1
48							•		•								1	1	0	0	0	0
49							•			•							1	1	0	0	0	1
50							•				•						1	1	0	0	1	0
51							•					•					1	1	0	0	1	1
52							•						•				1	1	0	1	0	0
53							•							•			1	1	0	1	0	1
54							•								•		1	1	0	1	1	0
55							•									•	1	1	0	1	1	1
56								•	•								1	1	1	0	0	0
57								•		•							1	1	1	0	0	1
58								•			•						1	1	1	0	1	0
59								•				•					1	1	1	0	1	1
60								•					•				1	1	1	1	0	0
61								•						•			1	1	1	1	0	1
62								•							•		1	1	1	1	1	0
63								•								•	1	1	1	1	1	1

Table 3. System Matrix Z-DR

SYSTEM NO	Z-LINES Z				DR-LINES DR								SYSTEM BITS S				
	0	1	2	3	0	1	2	3	4	5	6	7	4	3	2	1	0
0	●				●								0	0	0	0	0
1	●					●							0	0	0	0	1
2	●						●						0	0	0	1	0
3	●							●					0	0	0	1	1
4	●								●				0	0	1	0	0
5	●									●			0	0	1	0	1
6	●										●		0	0	1	1	0
7	●											●	0	0	1	1	1
8		●			●								0	1	0	0	0
9		●				●							0	1	0	0	1
10		●					●						0	1	0	1	0
11		●						●					0	1	0	1	1
12		●							●				0	1	1	0	0
13		●								●			0	1	1	0	1
14		●									●		0	1	1	1	0
15		●										●	0	1	1	1	1
16			●		●								1	0	0	0	0
17			●			●							1	0	0	0	1
18			●				●						1	0	0	1	0
19			●					●					1	0	0	1	1
20			●						●				1	0	1	0	0
21			●							●			1	0	1	0	1
22			●								●		1	0	1	1	0
23			●									●	1	0	1	1	1
24				●	●								1	1	0	0	0
25				●		●							1	1	0	0	1
26				●			●						1	1	0	1	0
27				●				●					1	1	0	1	1
28				●					●				1	1	1	0	0
29				●						●			1	1	1	0	1
30				●							●		1	1	1	1	0
31				●								●	1	1	1	1	1

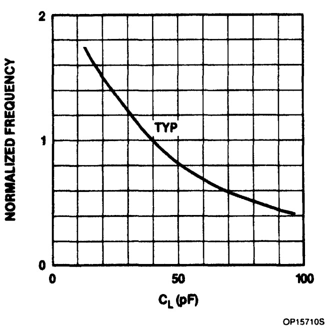


Figure 4. Typical Normalized Input Frequency as a Function of the Load (Keyboard) Capacitance

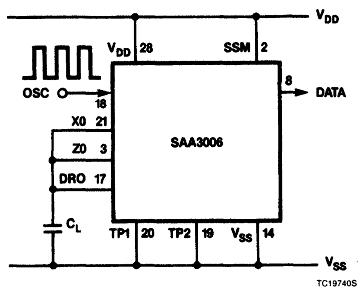


Figure 5. Test Circuit for Measurement of Maximum Oscillator Frequency

HANDLING

Inputs and outputs are protected against electrostatic charge in normal handling. However, to be totally safe, it is desirable to take normal precautions appropriate to handling MOS devices.

SAA3027
Infrared Remote Control Transmitter (RC-5)

Product Specification

Linear Products

DESCRIPTION

The SAA3027 is intended for a general purpose (RC-5) infrared remote control system. The device can generate 2048 different commands and utilizes a keyboard with a single-pole switch per key. The commands are arranged so that 32 systems can be addressed, each system containing 64 different commands.

The circuit response to legal (one key pressed at a time) and illegal (more than one key pressed at a time) keyboard operation is specified later in this publication (see KEY ACTIVITIES).

FEATURES

- **Transmitter for 32 × 64 commands**
- **One transmitter controls 32 systems**
- **Very low current consumption**
- **For infrared transmission link**
- **Transmission by biphase technique**
- **Short transmission times; speed-up of system reaction time**
- **LC oscillator; no crystal required**
- **Input protection**
- **Test mode facility**

APPLICATION

- **Remote control systems**

ORDERING INFORMATION

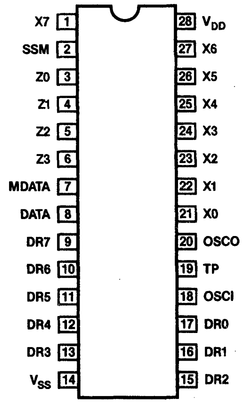
DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
28-Pin Plastic DIP (SOT-117)	−25°C to +85°C	SAA3027PN

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{DD}	Supply voltage range with respect to V_{SS}	−0.5 to +15	V
V_I	Input voltage range	−0.5 to (V_{DD} + 0.5)	V
$\pm I_I$	Input current	10	mA
V_O	Output voltage range	−0.5 to (V_{DD} + 0.5)	V
$\pm I_O$	Output current	10	mA
P_O	Power dissipation output OSCO	50	mW
P_O	Power dissipation per output (all other outputs)	100	mW
P_{TOT}	Total power dissipation per package	200	mW
T_A	Operating ambient temperature range	−25 to +85	°C
T_{STG}	Storage temperature range	−65 to +150	°C

PIN CONFIGURATION

N Package

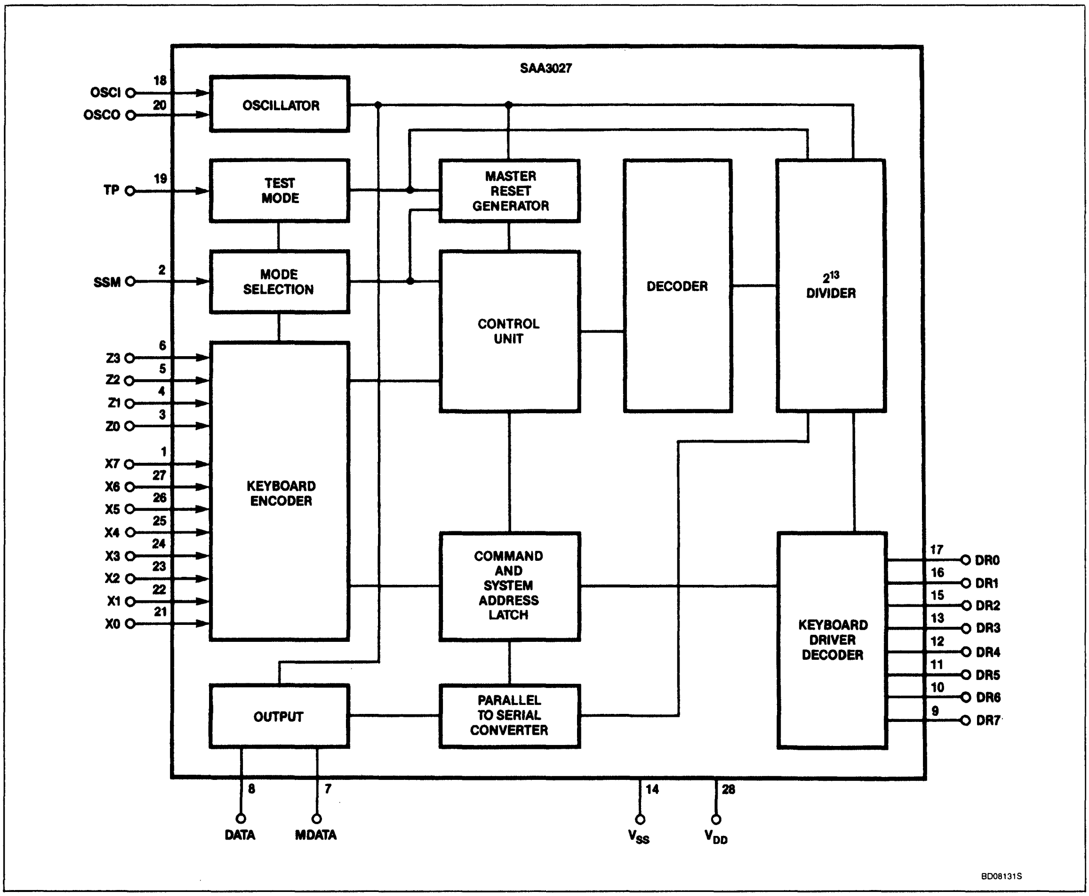


TOP VIEW

CD12060S

PIN NO.	SYMBOL	DESCRIPTION
1	X7	Keyboard command inputs with P-channel pull-up transistors
21	X0	
22	X1	
23	X2	
24	X3	
25	X4	
26	X5	
27	X6	
2	SSM	System mode selection input
3	Z0	Keyboard system inputs with P-channel pull-up transistors
4	Z1	
5	Z2	
6	Z3	
7	MDATA	Remote signal outputs (3-state outputs)
8	DATA	
9	DR7	Scan driver outputs with open-drain N-channel transistors
10	DR6	
11	DR5	
12	DR4	
13	DR3	
15	DR2	
16	DR1	
17	DR0	
14	V_{SS}	Negative supply (ground)
18	OSCI	Oscillator input
19	TP	Test pin
20	OSCO	Oscillator output
28	V_{DD}	Positive supply

BLOCK DIAGRAM



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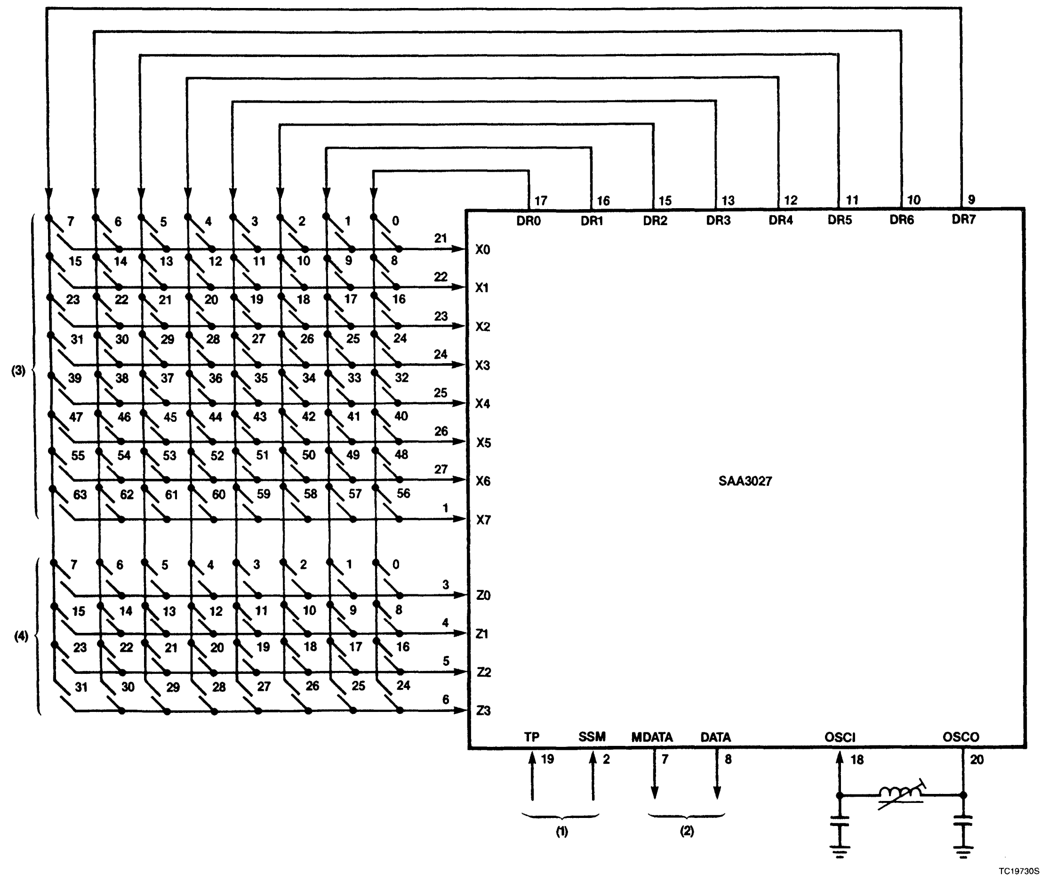
DC AND AC ELECTRICAL CHARACTERISTICS V_{SS} = 0V; T_A = −25°C to 85°C, unless otherwise specified.

SYMBOL	PARAMETER	V_{DD} (V)	LIMITS			UNIT
			Min	Typ	Max	
V_{DD}	Supply voltage		4.75		12.6	V
	Supply current at I_O = 0mA for all outputs; X0 to X7 and Z3 at V_{DD}; all other inputs at V_{DD} or V_{SS}; excluding leakage current from open drain N-channel outputs;					
I_{DD}	T_A = 25°C	12.6			10	μA
Inputs Keyboard inputs X and Z with P-channel pull-up transistors						
$-I_I$	Input current (each input) at V_I = 0V; TP = SSM = LOW	4.75 to 12.6	10		300	μA
V_{IH}	Input voltage HIGH	4.75 to 12.6	0.7 × V_{DD}		V_{DD}	V
V_{IL}	Input voltage LOW	4.75 to 12.6	0		0.3 × V_{DD}	V
	Input leakage current at T_A = 25°C; TP = HIGH;					
I_{IR}	V_I = 12.6V	12.6			1	μA
$-I_{IR}$	V_I = 0V	12.6			1	μA
SSM, TP and OSCI inputs						
V_{IH}	Input voltage HIGH	4.75 to 12.6	0.7 × V_{DD}		V_{DD}	V
V_{IL}	Input voltage LOW	4.75 to 12.6	0		0.3 × V_{DD}	V
	Input leakage current at T_A = 25°C;					
I_{IR}	V_I = 12.6V	12.6			1	μA
$-I_{IR}$	V_I = 0V	12.6			1	μA
Outputs DATA, MDATA						
V_{OH}	Output voltage HIGH at $-I_{OH}$ = 0.8mA	4.75 to 12.6	V_{DD} −0.6			V
V_{OL}	Output voltage LOW at I_{OL} = 0.8mA	4.75 to 12.6			0.4	V
	Output leakage current at:					
I_{OR}	V_O = 12.6V	12.6			10	μA
$-I_{OR}$	V_O = 0V	12.6			20	μA
	T_A = 25°C;					
I_{OR}	V_O = 12.6V	12.6			1	μA
$-I_{OR}$	V_O = 0V	12.6			2	μA
DR0 to DR7 outputs						
V_{OL}	Output voltage LOW at I_{OL} = 0.35mA	4.75 to 12.6			0.4	V
	Output leakage current					
I_{OR}	at V_O = 12.6V	12.6			10	μA
I_{OR}	at V_O = 12.6V; T_A = 25°C	12.6			1	μA
OSCO output						
V_{OH}	Output voltage HIGH at $-I_{OH}$ = 0.2mA; OSCI = V_{SS}	4.75 to 12.6	V_{DD} − 0.6			V
V_{OL}	Output voltage LOW at $-I_{OL}$ = 0.45mA; OSCI = V_{DD}	4.75 to 12.6			0.5	V
Oscillator						
f_{OSCI}	Maximum oscillator frequency at C_L = 40pF (Figures 4 and 5)	4.75	75	72		kHz
f_{OSCI}		6	120	72		kHz
f_{OSCI}		12.6	300	72		kHz

Handling

Inputs and outputs are protected against electrostatic charge in normal handling. However, to be totally safe, it is desirable to take normal precautions appropriate to handling MOS devices.



NOTES:
(1) Programming inputs for operating modes, test mode and reset.
(2) Remote signal outputs.
(3) Keyboard command code matrix 8 × 8.
(4) Keyboard system code matrix 4 × 8.

Figure 1. Keyboard Interconnection

FUNCTIONAL DESCRIPTION

Combined System Mode (SSM = LOW)

The X and Z-lines are active HIGH in the quiescent state. Legal key operation either in the X-DR or Z-DR matrix starts the debounce cycle. When the contact is made for two bit times without interruption, the oscillator-enable signal is latched and the key may be released. Interruption within the two bit times resets the internal action. At the end of the debounce time, the DR-outputs are switched off and two scan cycles are started, switching on the DR-outputs one by one. When a Z or X-input senses a LOW level, a latch-enable signal is fed to the system address or command latches; depending on whether sensing was found in the Z or X-input matrix. After latching a system address number, the device will generate the last command (i.e., all command bits '1') in the chosen system as long as the key is pressed. Latching of a command number causes the device to generate this command together with the system address number stored in the system address latch. Releasing the key will reset the internal action if no data is transmitted at that time. Once the transmission is started, the signal will be finished completely.

Single System Mode (SSM = HIGH)

The X-lines are active HIGH in the quiescent state; the pull-up transistors of the Z-lines are switched off and the inputs are disabled. Only legal key operation in the X-DR matrix starts the debounce cycle. When the contact is made for two bit times without interruption, the oscillator-enable signal is latched and the key may be released. Interruption within the two bit times resets the internal action. At the end of the debounce time, the pull-up transistors in the X-lines are switched off; those in the Z-lines are switched on during the first scan cycle. The wired connection in the Z-matrix is then translated into a system address number and stored in the system address latch. At the end of the first scan cycle the pull-up transistors in the Z-lines are switched off and the inputs are disabled again, while the transistors in the X-lines are switched on. The second scan cycle produces the command number which, after latching, is transmitted together with the system address number.

Inputs

The command inputs X0 to X7 carry a logical '1' in the quiescent state by means of an internal pull-up transistor. When SSM is LOW, the system inputs Z0 to Z3 also carry a logical '1' in the quiescent state by means of an internal pull-up transistor.

When SSM is HIGH, the transistors are switched off and no current flows via the wired connection in the Z-DR matrix.

Oscillator

OSCI and OSCO are the input/output, respectively, of a two-pin oscillator. The oscillator is formed externally by one inductor and two capacitors and operates at 72kHz (typical).

Key-Release Detection

An extra control bit is added which will be complemented after key-release. In this way the decoder gets an indication that shows if the next code is to be considered as a new command. This is very important for multi-digit entry (e.g. by channel numbers or Teletext/Viewdata pages). The control bit will only be complemented after finishing at least one code transmission. The scan cycles are repeated before every code transmission, so that, even by 'take-over' of key operation during code transmission, the correct system and command numbers are generated.

Outputs

The output DATA carries the generated information according to the format given in Figure 2 and Tables 1 and 2. The code is transmitted in biphase; definitions of logical '1' and '0' are given in Figure 3.

The code consists of four parts:

- Start part formed by 2 bits (two times a logical '1')
- Control part formed by 1 bit
- System part formed by 5 bits
- Command part formed by 6 bits

The output MDATA carries the same information as output DATA but is modulated on a carrier frequency of half the oscillator frequency, so that each bit is presented as a burst of 32 oscillator periods. To reduce power consumption, the carrier frequency has a 25% duty cycle.

In the quiescent state, both outputs are non-conducting (3-state outputs). The scan drivers DR0 to DR7 are of the open drain N-channel type and are conducting in the quiescent state of the circuit. After a legal key operation, a scanning procedure is started so that they are switched into the conducting state one after the other.

Reset Action

The circuit will be reset immediately when a key release occurs during:

- Debounce time
- Between two codes

When a key release occurs during scanning of the matrix, a reset action will be accomplished if:

- The key is released while one of the driver outputs is in the low-ohmic '0' state;
- The key is released before detection of that key;
- There is no wired connection in the Z-DR matrix while SSM is HIGH.

Test Pin

The test pin TP is an input which can be used for testing purposes.

When LOW, the circuit operates normally.

When HIGH, all pull-up transistors are switched off, the control bit is set to zero and the output data is 2^6 times faster than normal.

When Z2 = Z3 = LOW, the counter will be reset to zero.

Key Activities

Every connection of one X-input and one DR-output is recognized as a legal keyboard operation and causes the device to generate the corresponding code.

Activating more than one X-input at a time is an illegal keyboard operation and no circuit action is taken (oscillator does not start).

When SSM is LOW, every connection of one Z-input and one DR-output is recognized as a legal keyboard operation and causes the device to generate the corresponding code.

Activating two or more Z-inputs, or Z-inputs and X-inputs, at one time is an illegal keyboard operation and no circuit action is taken.

When SSM is HIGH, a wired connection must be made between a Z-input and a DR-output. If no connection is made, the code is not generated.

When one X or Z-input is connected to more than one DR-output, the last scan signal is considered legal.

The maximum allowable value of the contact series resistance of the keyboard switches is 10kΩ

Z2 or Z3 must be connected to V_{DD} to avoid unwanted supply current.

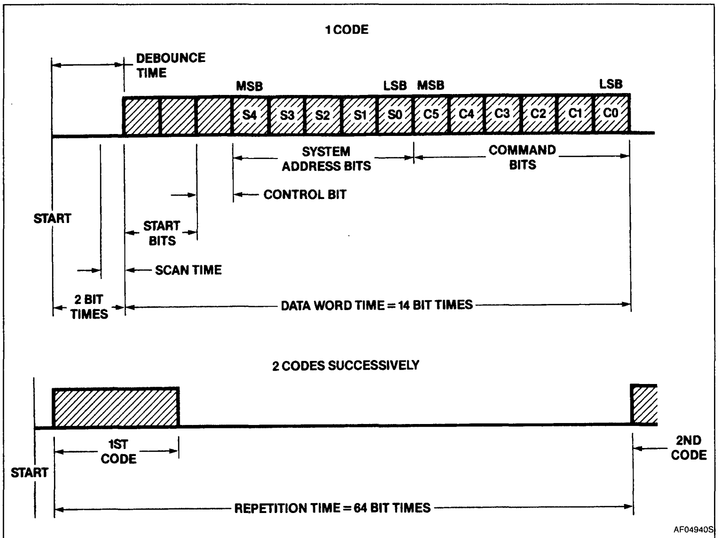
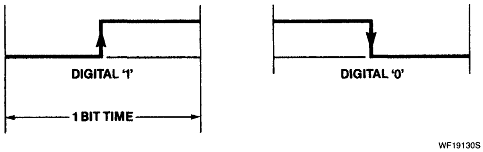


Figure 2. DATA Output Format (RC-5)



NOTE:
1. Bit Time = $2^7 \times T_{OSC}$ = 1.778ms (Typical), where T_{OSC} is the oscillator period time.

Figure 3. Biphase Transmission Code

Table 1. Command Matrix X-DR

CODE NO	X-LINES X								DR-LINES DR								COMMAND BITS C					
	0	1	2	3	4	5	6	7	0	1	2	3	4	5	6	7	5	4	3	2	1	0
0	•								•								0	0	0	0	0	0
1	•									•							0	0	0	0	0	1
2	•										•						0	0	0	0	1	0
3	•											•					0	0	0	0	1	1
4	•												•				0	0	0	1	0	0
5	•													•			0	0	0	1	0	1
6	•														•		0	0	0	1	1	0
7	•															•	0	0	0	1	1	1
8		•							•								0	0	1	0	0	0
9		•								•							0	0	1	0	0	1
10		•									•						0	0	1	0	1	0
11		•										•					0	0	1	0	1	1
12		•											•				0	0	1	1	0	0
13		•												•			0	0	1	1	0	1
14		•													•		0	0	1	1	1	0
15		•														•	0	0	1	1	1	1
16			•						•								0	1	0	0	0	0
17			•							•							0	1	0	0	0	1
18			•								•						0	1	0	0	1	0
19			•									•					0	1	0	0	1	1
20			•										•				0	1	0	1	0	0
21			•											•			0	1	0	1	0	1
22			•												•		0	1	0	1	1	0
23			•													•	0	1	0	1	1	1
24				•					•								0	1	1	0	0	0
25				•						•							0	1	1	0	0	1
26				•							•						0	1	1	0	1	0
27				•								•					0	1	1	0	1	1
28				•									•				0	1	1	1	0	0
29				•										•			0	1	1	1	0	1
30				•											•		0	1	1	1	1	0
31				•												•	0	1	1	1	1	1

Table 1. Command Matrix X-DR (Continued)

CODE NO	X-LINES X								DR-LINES DR								COMMAND BITS C					
	0	1	2	3	4	5	6	7	0	1	2	3	4	5	6	7	5	4	3	2	1	0
32					●				●								1	0	0	0	0	0
33					●					●							1	0	0	0	0	1
34					●						●						1	0	0	0	1	0
35					●							●					1	0	0	0	1	1
36					●								●				1	0	0	1	0	0
37					●									●			1	0	0	1	0	1
38					●										●		1	0	0	1	1	0
39					●											●	1	0	0	1	1	1
40						●			●								1	0	1	0	0	0
41						●				●							1	0	1	0	0	1
42						●					●						1	0	1	0	1	0
43						●						●					1	0	1	0	1	1
44						●							●				1	0	1	1	0	0
45						●								●			1	0	1	1	0	1
46						●									●		1	0	1	1	1	0
47						●										●	1	0	1	1	1	1
48							●		●								1	1	0	0	0	0
49							●			●							1	1	0	0	0	1
50							●				●						1	1	0	0	1	0
51							●					●					1	1	0	0	1	1
52							●						●				1	1	0	1	0	0
53							●							●			1	1	0	1	0	1
54							●								●		1	1	0	1	1	0
55							●									●	1	1	0	1	1	1
56								●	●								1	1	1	0	0	0
57								●		●							1	1	1	0	0	1
58								●			●						1	1	1	0	1	0
59								●				●					1	1	1	0	1	1
60								●					●				1	1	1	1	0	0
61								●						●			1	1	1	1	0	1
62								●							●		1	1	1	1	1	0
63								●								●	1	1	1	1	1	1

Table 2. System Matrix Z-DR

SYSTEM NO	Z-LINES Z				DR-LINES DR								SYSTEM BITS S				
	0	1	2	3	0	1	2	3	4	5	6	7	4	3	2	1	0
0	●				●								0	0	0	0	0
1	●					●							0	0	0	0	1
2	●						●						0	0	0	1	0
3	●							●					0	0	0	1	1
4	●								●				0	0	1	0	0
5	●									●			0	0	1	0	1
6	●										●		0	0	1	1	0
7	●											●	0	0	1	1	1
8		●			●								0	1	0	0	0
9		●				●							0	1	0	0	1
10		●					●						0	1	0	1	0
11		●						●					0	1	0	1	1
12		●							●				0	1	1	0	0
13		●								●			0	1	1	0	1
14		●									●		0	1	1	1	0
15		●										●	0	1	1	1	1
16			●		●								1	0	0	0	0
17			●			●							1	0	0	0	1
18			●				●						1	0	0	1	0
19			●					●					1	0	0	1	1
20			●						●				1	0	1	0	0
21			●							●			1	0	1	0	1
22			●								●		1	0	1	1	0
23			●									●	1	0	1	1	1
24				●	●								1	1	0	0	0
25				●		●							1	1	0	0	1
26				●			●						1	1	0	1	0
27				●				●					1	1	0	1	1
28				●					●				1	1	1	0	0
29				●						●			1	1	1	0	1
30				●							●		1	1	1	1	0
31				●								●	1	1	1	1	1

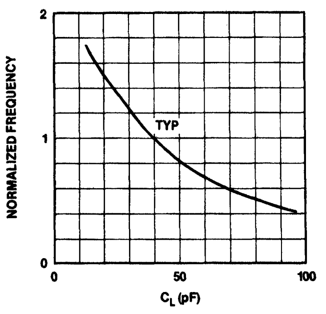


Figure 4. Typical Normalized Input Frequency as a Function of the Load (Keyboard) Capacitance

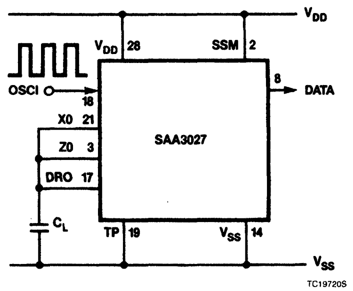


Figure 5. Test Circuit for Measurement of Maximum Oscillator Frequency

Signetics

Linear Products

SAA3028
Remote Control Receiver/Transcoder

Product Specification

DESCRIPTION

The SAA3028 is intended for use in general purpose (RC-5) remote control systems. The main function of this integrated circuit is to convert RC-5 biphase coded signals into equivalent binary values. Two input circuits are available: one for RC-5 coded signals only; the other selectable to accept RC-5 coded signals only, or RC-5 (extended) coded signals only. The input used is that at which an active code is first detected. Coded signals not in RC-5/RC-5(ext) format are rejected. Data input and output is by serial transfer, the output interface being compatible for I^2C bus operation.

FEATURES

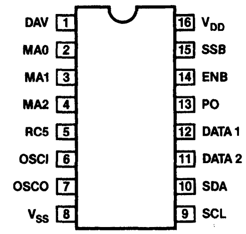
- **Converts RC-5 or RC-5(ext) biphase coded signals into binary equivalents**
- **Two data inputs: one fixed (RC-5); one selectable (RC-5/RC-5(ext))**
- **Rejects all codes not in RC-5/RC-5(ext) format**
- **I^2C output interface capability**
- **Power-off facility**
- **Master/slave addressable for multi-transmitter/receiver applications in RC-5(ext) mode**
- **Power-on reset for defined start-up**

APPLICATION

- **Remote control systems**

PIN CONFIGURATION

N Package



TOP VIEW

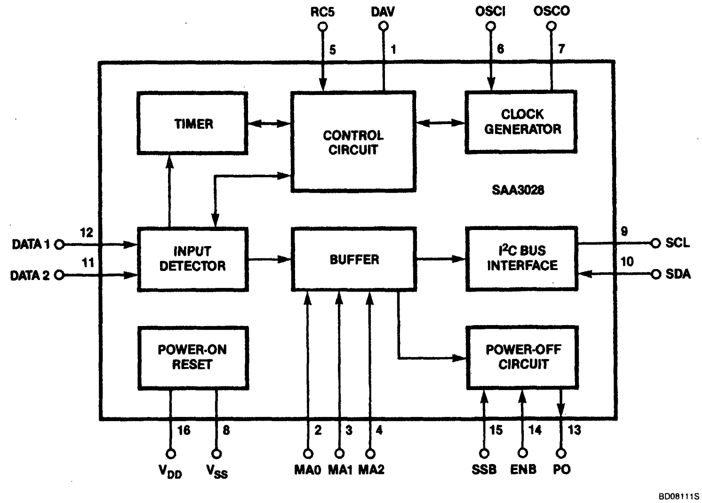
CD12040S

PIN NO.	SYMBOL	DESCRIPTION
1	DAV	Data valid output with open drain N-channel transistor
2	MA0	Master address inputs
3	MA1	Master address inputs
4	MA2	Master address inputs
5	RC5	Data 2 input select
6	OSCI	Oscillator input
7	OSCO	Oscillator output
8	V_{SS}	Negative supply (ground)
9	SCL	Serial clock line (I^2C bus)
10	SDA	Serial data line (I^2C bus)
11	DATA 2	Data 2 input
12	DATA 1	Data 1 input
13	PO	Power-off signal output with open drain N-channel transistor
14	ENB	Enable input
15	SSB	Set standby input
16	V_{DD}	Positive supply (+5V)

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
16-Pin Plastic DIP (SOT-38Z)	−25°C to 85°C	SAA3028N

BLOCK DIAGRAM



ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{DD}	Supply voltage range with respect to V_{SS}	−0.5 to +15	V
V_I	Input voltage range	−0.5 to (V_{DD} + 0.5)	V[1]
$\pm I_I$	Input current	10	mA
V_O	Output voltage range	−0.5 to (V_{DD} + 0.5)	V[1]
$\pm I_O$	Output current	10	mA
P_O	Power dissipation output OSCO	50	mW
P_O	Power dissipation per output (all other outputs)	100	mW
P_{TOT}	Total power dissipation per package	200	mW
T_A	Operating ambient temperature range	−25 to +85	°C
T_{STG}	Storage temperature range	−55 to +150	°C

NOTE:
1. V_{DD}+0.5 not to exceed 15V.

DC ELECTRICAL CHARACTERISTICS V_{SS}=0V; T_A = −25°C to 85°C, unless otherwise specified.

SYMBOL	PARAMETER	V_{DD} (V)	LIMITS			UNIT
			Min	Typ	Max	
V_{DD}	Supply voltage		4.5		5.5	V
I_{DD}	Supply current; quiescent at T_A = 25°C	5.5			200	μA
Inputs MA0, MA1, MA2, DATA 1, DATA 2, RC5, SCL, ENB, SSB, OSCI						
V_{IH}	Input voltage HIGH	4.5 to 5.5	0.7 × V_{DD}		V_{DD}	V
V_{IL}	Input voltage LOW	4.5 to 5.5	0		0.3 × V_{DD}	V
I_I	Input leakage current at V_I = 5.5V; T_A = 25°C	5.5			1	μA
$-I_I$	Input leakage current at V_I = 0V; T_A = 25°C	5.5			1	μA
Outputs DAV, PO						
V_{OL}	Output voltage LOW at I_{OL} = 1.6mA	4.5 to 5.5			0.4	V
I_{OR}	Output leakage current at V_O = 5.5V; T_A = 25°C	5.5			1	μA
OSCO						
V_{OH}	Output voltage HIGH at $-I_{OH}$ = 0.2mA	4.5 to 5.5	V_{DD} − 0.5			V
V_{OL}	Output voltage LOW at I_{OL} = 0.3mA	4.5 to 5.5			0.4	V
	Output leakage current at T_A = 25°C;					
I_{OR}	V_O = 5.5V	5.5			1	μA
I_{OR}	V_O = 0V	5.5			1	μA
SDO						
V_{OL}	Output voltage LOW at I_{OL} = 2mA	4.5 to 5.5			0.4	V
I_{OR}	Output leakage current at V_O = 5.5V; T_A = 25°C	5.5			1	μA
Oscillator						
f_{OSCI}	Maximum oscillator frequency (Figure 6)	4.75	500			kHz

HANDLING

Inputs and outputs are protected against electrostatic charge in normal handling. However, to be totally safe, it is desirable to take normal precautions appropriate to handling MOS devices.

FUNCTIONAL DESCRIPTION

Input Function

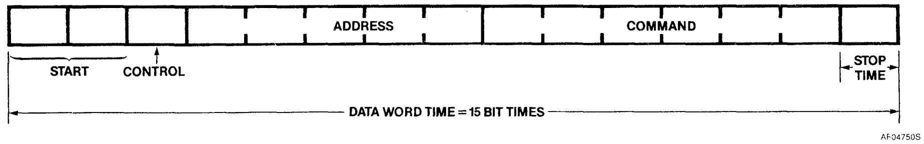
The two data inputs are accepted into the buffer as follows:

DATA 1: Only biphase coded signals which conform to the RC-5 format are accepted at this input.

DATA 2: This input performs according to the logic state of the select input RC5. When RC5 = HIGH, DATA 2 input will accept only RC-5 coded signals. When RC5 = LOW, DATA 2 input will accept only RC-5(ext) coded signals.

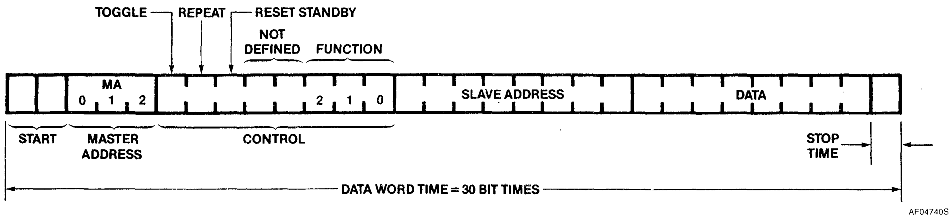
The input detector selects the input, DATA 1 or DATA 2, in which a HIGH-to-LOW transition is first detected. The selected input is then accepted by the buffer for code conversion. All signals received that are not in the RC-5 or RC-5(ext) format are rejected.

Formats of RC-5 and RC-5(ext) biphase coded signals are shown in Figures 1 and 2, respectively; the codes commence from the left of the formats shown. The bit-times of the biphase codes are defined in Figure 3.



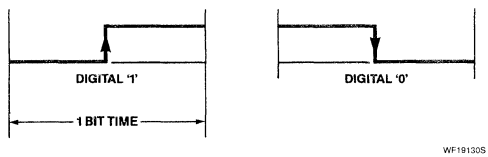
NOTE:
Stop time = 1.5 bit-times (nominal).

Figure 1. RC-5 Code Format: the First Start Bit is Used Only for Detection and Input Gain-Setting



NOTE:
Stop time = 1.5 bit-times (nominal).

Figure 2. RC-5 (extended) Code Format: the First Start Bit is Used Only for Detection and Input Gain-Setting



NOTE:
RC-5 bit-time = $2^7 \times t_{OSC}$ = 1.778ms (typical), RC-5(ext) bit-time = $2^6 \times t_{OSC}$ = 0.89ms (typical), where t_{OSC} = the oscillator period time.

Figure 3. Biphase Code Definition

More information is added to the input data held in the buffer in order to make it suitable for transmission via the I^{2}C interface. The information now held in the buffer is as shown in the table.

RC-5 BUFFER CONTENTS		RC-5(EXT) BUFFER CONTENTS	
• Data valid indicator	1 Bit	• Data valid indicator	1 Bit
• Format indicator	1 Bit	• Format indicator	1 Bit
• Input indicator	1 Bit	• Input indicator	1 Bit
• Control	1 Bit	• Master address	3 Bits
• Address data	5 Bits	• Control	8 Bits
• Command data	6 Bits	• Slave address	8 Bits
		• Data	8 Bits

The information assembled in the buffer is subjected to the following controls before being made available at the I_2C interface:

ENB = HIGH Enables the set standby input SSB.

SSB = LOW Causes power-off output PO to go HIGH.

PO = HIGH This occurs when the set standby input SSB = LOW and allows the existing values in the buffer to be overwritten by the new binary equivalent values. After ENB = LOW, SSB is don't care.

PO = LOW This occurs according to the type of code being processed, as follows: RC-5: When the binary equivalent value is transferred to the buffer. RC-5(ext): When the reset standby bit is active and the master address bits are equal in value to the MAO, MA1, MA2 inputs. At power-on, PO is reset to LOW.

DAV = HIGH This occurs when the buffer contents are valid. If the buffer is not empty, or an output transfer is taking place, then the new binary values are discarded.

Output Function

The data is assembled in the buffer in the format shown in Figure 4 for RC-5 binary equivalent values, or in the format shown in Figure 5 for RC-5(ext) binary equivalent values. The data is output serially, starting from the left of the formats shown in Figures 4 and 5.

The output signal DAV, derived in the buffer from the data valid bit, is provided to facilitate use of the transcoder on an interrupt basis. This output is reset to LOW during power-on.

The I^{2}C interface allows transmission on a bidirectional, two-wire I^{2}C bus. The interface is a slave transmitter with a built-in slave address, having a fixed 7-bit binary value of 0100110. Serial output of the slave address onto the I^{2}C bus starts from the left-hand bit.

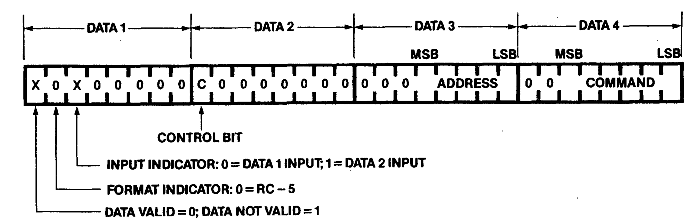


Figure 4. RC-5 Binary Equivalent Value Format

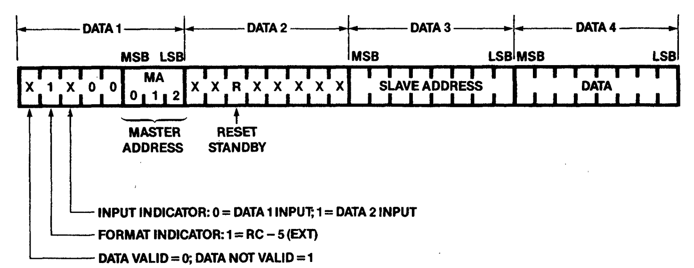
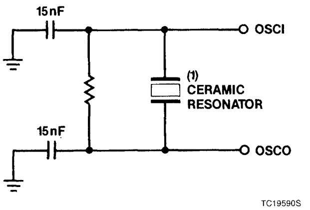


Figure 5. RC-5(ext) Binary Equivalent Value Format

Oscillator

The oscillator can comprise a ceramic resonator circuit as shown in Figure 6. The typical frequency of oscillation is 455kHz.



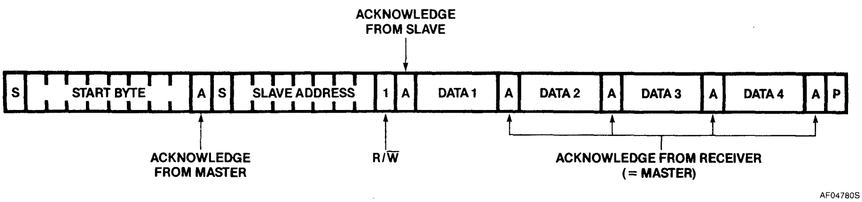
NOTE:
(1) Catalog number of ceramic resonator: 2422 540 98008.

Figure 6. Oscillator Circuit

FUNCTIONAL DESCRIPTION

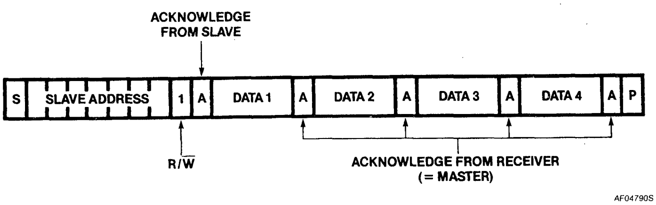
I^{2}C Bus Transmission

Formats for I^{2}C transmission in low-and high-speed modes are shown respectively in Figures 7 and 8.



NOTES:
When R/$\overline{W}$ bit = 0, the slave generates a NACK (negative acknowledge), leaves the data line HIGH and waits for a stop (P) condition.
When the receiver generates a NACK, the slave leaves the data line HIGH and waits for P (the slave acting as if all data has been transmitted).
When all data has been transmitted, the data line remains HIGH and the slave waits for P.

Figure 7. Format for Transmission in I^{2}C Low-Speed Mode



NOTES:
When R/$\overline{W}$ bit = 0, the slave generates a NACK (negative acknowledge), leaves the data line HIGH and waits for a stop (P) condition.
When the receiver generates a NACK, the slave leaves the data line HIGH and waits for P (the slave acting as if all data has been transmitted).
When all data has been transmitted, the data line remains HIGH and the slave waits for P.

Figure 8. Format for Transmission in I^{2}C High-Speed Mode

Signetics

TDA3047
IR Preamplifier

Product Specification

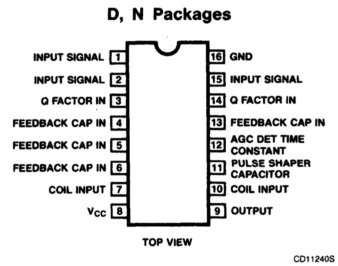
Linear Products

DESCRIPTION
The TDA3047 is for infrared reception with low power consumption.

FEATURES
- **HF amplifier with a control range of 66dB**
- **Synchronous demodulator and reference amplifier**
- **AGC detector**
- **Pulse shaper**
- **Q-factor killing of the input selectivity, which is controlled by the AGC circuit**
- **Input voltage limiter**

APPLICATION
- **IR remote control systems**

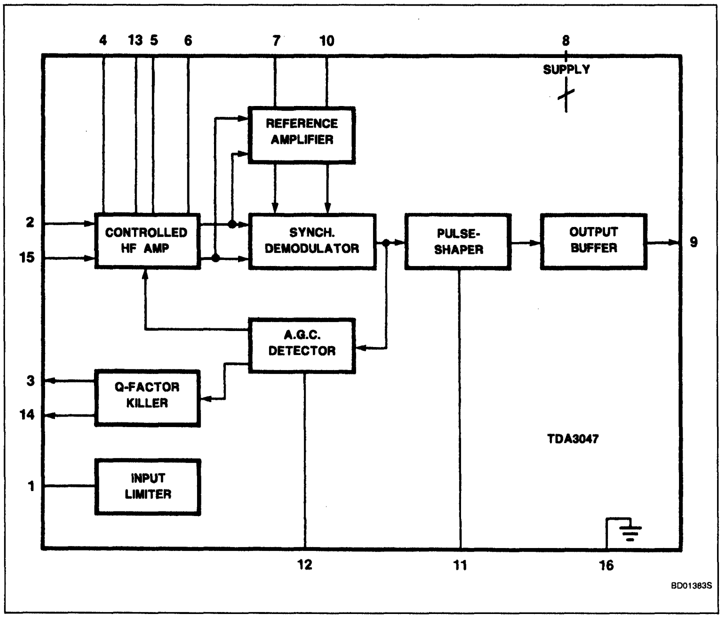
PIN CONFIGURATION



ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
16-Pin Plastic DIP (SOT-38)	−25°C to +125°C	TDA3047N
16-Pin Plastic SO (SOT-109A)	0 to +70°C	TDA3047TD

BLOCK DIAGRAM



ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{CC}	Supply voltage (Pin 8)	13.2	V
I_{11}	Output current pulse shaper (Pin 11)	10	mA
	Voltages between pins[1]		
V_{2-15}	Pins 2 and 15	4.5	V
V_{4-13}	Pins 4 and 13	4.5	V
V_{5-6}	Pins 5 and 6	4.5	V
V_{7-10}	Pins 7 and 10	4.5	V
V_{9-11}	Pins 9 and 11	4.5	V
T_{STG}	Storage temperature range	−65 to +150	°C
T_A	Operating ambient temperature range	−25 to +125	°C

NOTE:
1. All pins except Pin 11 are short-circuit protected.

5

DC ELECTRICAL CHARACTERISTICS $V_{CC} = V_8 = 5V$; $T_A = 25°C$, measured in Figure 3, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supply (Pin 8)					
V_{CC}	Supply voltage	4.65	5.0	5.35	V
$I_{CC} = I_8$	Supply current	1.2	2.1	3.0	mA
Controlled HF amplifier (Pins 2 and 15)					
	Minimum input signal (peak-to-peak value)				
$V_{2-15(P-P)}$	at f = 36kHz[1]		15	25	μV
$V_{2-15(P-P)}$	at f = 36kHz[2]			5	μV
	AGC control range (without Q-killing)	60	66		dB
$V_{2-15(P-P)}$	Input signal for correct operation (peak-to-peak value)[3]	0.02		200	mV
$V_{2-15(P-P)}$	Q-killing inactive ($I_3 = I_{14} < 0.5\mu A$) peak-to-peak value)			140	μV
$V_{2-15(P-P)}$	Q-killing active ($I_{14} = I_3 = max.$) (peak-to-peak value)	28			mV
	Q-killing range		Figure 1		
Inputs					
V_2	Input voltage (Pin 2)	2.25	2.45	2.65	V
V_{15}	Input voltage (Pin 15)	2.25	2.45	2.65	V
R_{2-15}	Input resistance (Pin 2)	10	15	20	kΩ
C_{2-15}	Input capacitance (Pin 2)		3		pF
V_{1-16}	Input limiting (Pin 1) at $I_1 = 3mA$		0.8	0.9	V
Outputs					
$-V_{9-8}$	Output voltage HIGH (Pin 9) at $-I_9 = 75\mu A$		0.1	0.5	V
V_9	Output voltage LOW (Pin 9) at $I_9 = 75\mu A$		0.1	0.5	V
	Output current; output voltage HIGH				
$-I_9$	at $V_9 = 4.5V$	75	120		μA
$-I_9$	at $V_9 = 3.0V$	75	130		μA
$-I_9$	at $V_9 = 1.0V$	75	140		μA
I_9	Output current; output voltage LOW at $V_9 = 0.5V$	75	120		μA
R_{7-10}	Output resistance between Pins 7 and 10	3.1	4.7	6.2	kΩ

DC ELECTRICAL CHARACTERISTICS (Continued) $V_{CC} = V_8 = 5V$; $T_A = 25°C$, measured in Figure 3, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Pulse shaper (Pin 11)					
V_{11}	Trigger level in positive direction (voltage Pin 9 changes from HIGH to LOW)	3.75	3.9	4.05	V
V_{11}	Trigger level in negative direction (voltage Pin 9 changes from LOW to HIGH)	3.4	3.55	3.7	V
ΔV_{11}	Hysteresis of trigger levels	0.25	0.35	0.45	V
AGC detector (Pin 12)					
$-I_{12}$	AGC capacitor charge current	3.3	4.7	6.1	μA
I_{12}	AGC capacitor discharge current	67	100	133	μA
Q-factor killer (Pins 3 and 14)					
$-I_3$	Output current (Pin 3) at $V_{12-16} = 2V$	2.5	7.5	15	μA
$-I_{14}$	Output current (Pin 14) at $V_{12-16} = 2V$	2.5	7.5	15	μA

NOTES:
1. Voltage Pin 9 is HIGH; $-I_9 = 75\mu A$.
2. Voltage Pin 9 remains LOW.
3. Undistorted output pulse with 100% AM input.

FUNCTIONAL DESCRIPTION

General

The circuit operates from a 5V supply and has a current consumption of 2mA. The output is a current source which can drive or suppress current of $> 75\mu A$ with a voltage swing of 4.5V. The Q-killer circuit eliminates distortion of the output pulses due to the decay of the tuned input circuit at high input voltages. The input circuit is protected against signals of $> 600mV$ by an input limiter. The typical input is an AM signal at a frequency of 36kHz. Figures 2 and 3 show the circuit diagrams for the application of narrow-band and wide-band receivers, respectively. Circuit description of the eight sections shown in the Block Diagram are given below.

Controlled HF Amplifier

The input signal is amplified by the gain-controlled amplifier. This circuit comprises three DC amplifier stages connected in cascade. The overall gain of the circuit is approximately 83dB and the gain control range is in the order of 66dB. Gain control is initially active in the second amplifier stage and is transferred to the first stage as limiting in the second stage occurs, thus maintaining optimum signal-to-noise ratio. Offset voltages in the DC coupled amplifier are minimized by two negative feedback loops. These also allow the circuit to have some series resistance of the decoupling capacitor. The output signal of the amplifier is applied to the reference amplifier and to the synchronous demodulator inputs.

Reference Amplifier

The reference amplifier amplifies and limits the input signal. The voltage gain is approximately 0dB. The output signal of this amplifier is applied to the synchronous demodulator.

Synchronous Demodulator

In the synchronous demodulator, the input signal and reference signal are multiplied. The demodulator output current is $25\mu A$ peak-to-peak. The output signal of the demodulator is fed to the input of the AGC detector and to the input of the pulse-shaper circuit.

AGC Detector

The AGC detector comprises two NPN transistors operating as a differential pair. The top level of the output signal from the synchronous demodulator is detected by the AGC circuit. Noise pulses are integrated by an internal capacitor. The output signal is amplified and applied to the first and second stages of the amplifier and to the Q-factor killer circuit.

Pulse-Shaper

The pulse-shaper comprises two NPN transistors operating as a differential pair connected in parallel with the AGC differential pair. The slicing level of the pulse shaper is lower than the slicing level of the AGC detector. The output of the pulse-shaper is determined by the voltage of the capacitor connected to Pin 11 which is applied directly to the output buffer.

Output Buffer

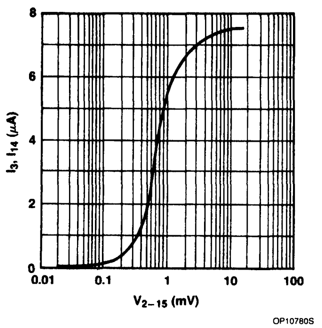
The voltage of the pulse-shaper capacitor is fed to the base of the first transistor of a differential pair. To obtain a correct RC-5 code, a hysteresis circuit protects the output against spikes. The output at Pin 9 is active HIGH.

Q-factor Killer

Figure 2 shows the Q-factor killer in the narrow-band application. In this application it is necessary to decrease the Q-factor of the input selectivity particularly when large input signals occur at Pins 2 and 15. In the narrow-band application the output of the Q-factor killer can be directly coupled to the input; Pin 3 to Pin 2, and Pin 14 to Pin 15.

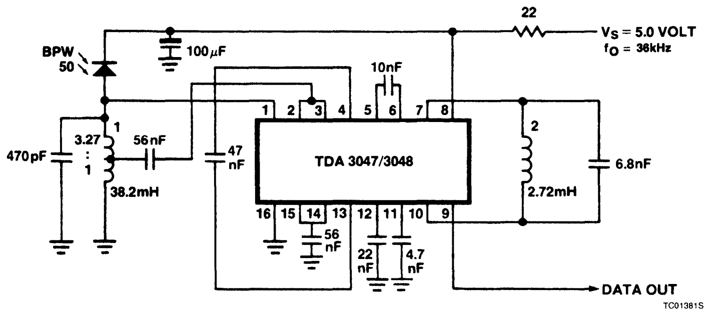
Input Limiter

In the narrow-band application, high voltage peaks can occur on the input of the selectivity circuit. The input limiter limits these voltage peaks to approximately 0.7V. Limiting is 0.9V maximum at $I_1 = 3mA$.



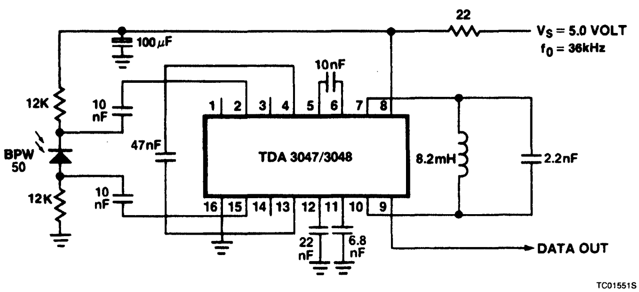
NOTE:
$I_{3,\ 14}$ is measured to ground, $V_{2\text{-}15(P\text{-}P)}$ is a symmetrical square wave. Measured in Figure 3; V_{CC} = 5V.

Figure 1. Typical Q-Factor Killer Current (Pins 3 and 14) as a Function of the Peak-to-Peak Input Voltage (V_2-15)



NOTES:
1. Q = 16
2. Q = 6

Figure 2. Narrow-Band Receiver Using TDA3047



NOTE:
For better sensitivity, both 12kΩ resistors may have a higher value.

Figure 3. Wide-Band Receiver With TDA3047

Signetics

TDA3048
IR Preamplifier

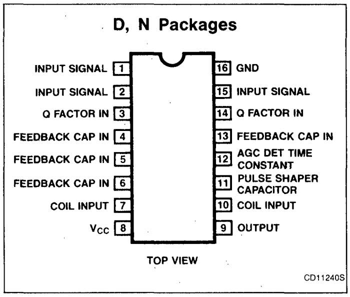
Product Specification

Linear Products

DESCRIPTION
The TDA3048 is for infrared reception with low power consumption.

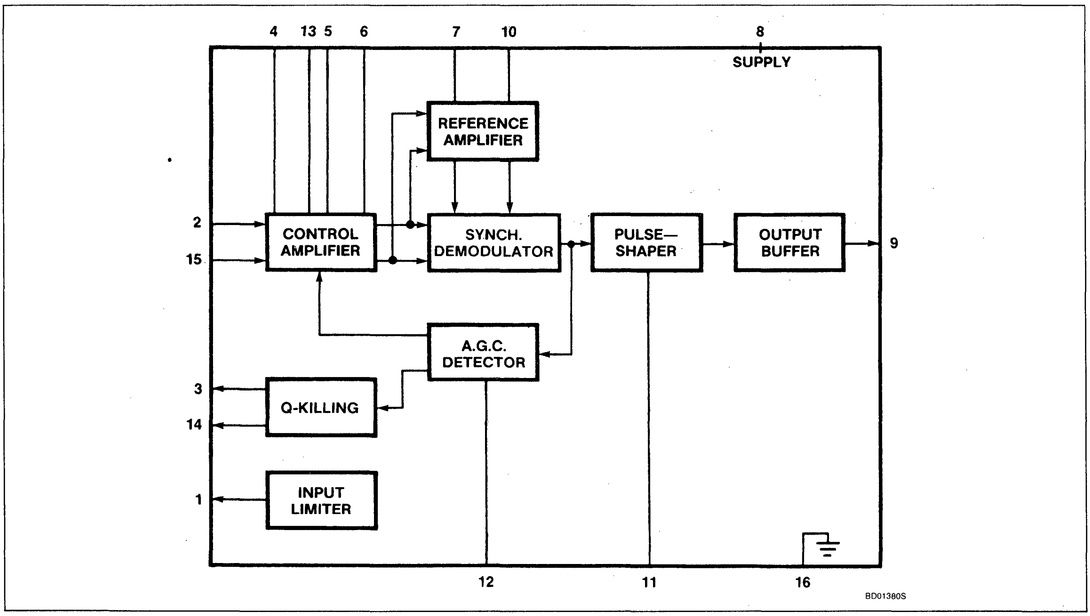
FEATURES
- **HF amplifier with a control range of 66dB**
- **Synchronous demodulator and reference amplifier**
- **AGC detector**
- **Pulse shaper**
- **Q-factor killing of the input selectivity, which is controlled by the AGC circuit**
- **Input voltage limiter**

APPLICATION
- **IR Remote control systems**

PIN CONFIGURATION


ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
16-Pin Plastic DIP (SOT-38)	−25°C to +125°C	TDA3048N
16-Pin Plastic SO (SOT-109A)	0 to +70°C	TDA3048TD

BLOCK DIAGRAM


IR Preamplifier TDA3048

FUNCTIONAL DESCRIPTION

General

The circuit operates from a 5V supply and has a current consumption of 2mA. The output is a current source which can drive or suppress a current of > 75μA with a voltage swing of 4.5V. The Q-killer circuit eliminates distortion of the output pulses due to the decay of the tuned input circuit at high input voltages. The input circuit is protected against signals of > 600mV by an input limiter. The typical input is an AM signal at a frequency of 36kHz. Figures 2 and 3 show the circuit diagrams for the application of narrow-band and wide-band receivers, respectively. Circuit description of the eight sections shown in the Block Diagram are given below.

Controlled HF Amplifier

The input signal is amplified by the gain-controlled amplifier. This circuit comprises three DC amplifier stages connected in cascade. The overall gain of the circuit is approximately 83dB and the gain control range is in the order of 66dB. Gain control is initially active in the second amplifier stage and is transferred to the first stage as limiting in the second stage occurs, thus maintaining optimum signal-to-noise ratio. Offset voltages in the DC coupled amplifier are minimized by two negative feedback loops. These also allow the circuit to have some series resistance of the decoupling capacitor. The output signal of the amplifier is applied to the reference amplifier and to the synchronous demodulator inputs.

Reference Amplifier

The reference amplifier amplifies and limits the input signal. The voltage gain is approximately 0dB. The output signal of this amplifier is applied to the synchronous demodulator.

Synchronous Demodulator

In the synchronous demodulator, the input signal and reference signal are multiplied. The demodulator output current is 25μA peak-to-peak. The output signal of the demodulator is fed to the input of the AGC detector and to the input of the pulse-shaper circuit.

AGC Detector

The AGC detector comprises two NPN transistors operating as a differential pair. The top level of the output signal from the synchronous demodulator is detected by the AGC circuit. Noise pulses are integrated by an internal capacitor. The output signal is amplified and applied to the first and second stages of the amplifier and to the Q-factor killer circuit.

Pulse-Shaper

The pulse-shaper comprises two NPN transistors operating as a differential pair connected in parallel with the AGC differential pair. The slicing level of the pulse shaper is lower than the slicing level of the AGC detector. The output of the pulse-shaper is determined by the voltage of the capacitor connected to Pin 11, which is applied directly to the output buffer.

Output Buffer

The voltage of the pulse-shaper capacitor is fed to the base of the first transistor of a differential pair. To obtain a correct RC-5 code, a hysteresis circuit protects the output against spikes. The output at Pin 9 is active LOW.

Q-Factor Killer

Figure 2 shows the Q-factor killer in the narrow-band application. In this application it is necessary to decrease the Q-factor of the input selectivity particularly when large input signals occur at Pins 2 and 15. In the narrow-band application the output of the Q-factor killer can be directly coupled to the input; Pin 3 to Pin 2 and Pin 14 to Pin 15.

Input Limiter

In the narrow-band application, high voltage peaks can occur on the input of the selectivity circuit. The input limiter limits these voltage peaks to approximately 0.7V. Limiting is 0.9V max. at I_1 = 3mA.

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{CC}	Supply voltage (Pin 8)	13.2	V
I_{11}	Output current pulse shaper (Pin 11)	10	mA
	Voltages between pins[1]		
V_{2-15}	Pins 2 and 15	4.5	V
V_{4-13}	Pins 4 and 13	4.5	V
V_{5-6}	Pins 5 and 6	4.5	V
V_{7-10}	Pins 7 and 10	4.5	V
V_{9-11}	Pins 9 and 11	4.5	V
T_{STG}	Storage temperature range	-65 to +150	°C
T_A	Operating ambient temperature range	-25 to +125	°C

NOTE:
1. All pins except Pin 11 are short-circuit protected.

5

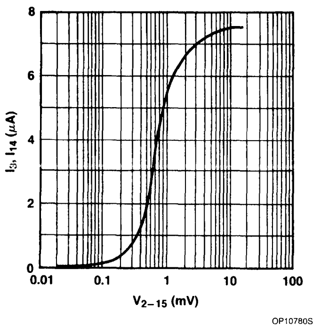
DC ELECTRICAL CHARACTERISTICS $V_{CC} = V_8 = 5V$; $T_A = 25°C$; measured in Figure 3, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supply (Pin 8)					
V_{CC}	Supply voltage	4.65	5.0	5.35	V
I_{CC}	Supply current	1.2	2.1	3.0	mA
Controlled HF amplifier (Pins 2 and 15)					
 V_{2-15} V_{2-15}	Minimum input signal (peak-to-peak value) at f = 36kHz[1] at f = 36kHz[2]		 15	 25 5	 μV μV
	AGC control range (without Q-killing)	60	66		dB
V_{2-15}	Input signal for correct operation (peak-to-peak value)[3]	0.02		200	mV
V_{2-15}	Q-killing inactive ($I_3 = I_{14} < 0.5\mu A$) (peak-to-peak value)			140	μV
V_{2-15}	Q-killing active ($I_{14} = I_3 =$ max.) (peak-to-peak) value	28			mV
	Q-killing range	See Figure 1			
Inputs					
V_2	Input voltage (Pin 2)	2.25	2.45	2.65	V
V_{15}	Input voltage (Pin 15)	2.25	2.45	2.65	V
R_{2-15}	Input resistance (Pin 2)	10	15	20	kΩ
C_{2-15}	Input capacitance (Pin 2)		3		pF
V_{1-16}	Input limiting (Pin 1) at $I_1 = 3mA$		0.8	0.9	V
Outputs					
$-V_{9-8}$	Output voltage HIGH (Pin 9) at $-I_9 = 75\mu A$		0.1	0.5	V
V_9	Output voltage LOW (Pin 9) at $I_9 = 75\mu A$		0.1	0.5	V
 I_9 I_9 I_9	Output current; output voltage LOW $-V_{9-8} = 4.5V$ $-V_{9-8} = 3.0V$ $-V_{9-8} = 1.0V$	 75 75 75	 120 130 140		 μA μA μA
$-I_9$	Output current; output voltage HIGH $-V_{9-8} = 0.5V$	 75	 120		 μA
R_{7-10}	Output resistance between Pins 7 and 10	3.1	4.7	6.2	kΩ
Pulse shaper (Pin 11)					
V_{11}	Trigger level in positive direction (voltage Pin 9 changes from HIGH to LOW)	3.75	3.9	4.05	V
V_{11}	Trigger level in negative direction (voltage Pin 9 changes from LOW to HIGH)	3.4	3.55	3.7	V
ΔV_{11}	Hysteresis of trigger levels	0.25	0.35	0.45	V
AGC detector (Pin 12)					
$-I_{12}$	AGC capacitor charge current	3.3	4.7	6.1	μA
I_{12}	AGC capacitor discharge current	67	100	133	μA
Q-factor killer (Pins 3 and 14)					
$-I_3$	Output current (Pin 3) at $V_{12} = 2V$	2.5	7.5	15	μA
$-I_{14}$	Output current (Pin 14) at $V_{12} = 2V$	2.5	7.5	15	μA

NOTES:
1. Voltage Pin 9 is LOW; $I_9 = 75\mu A$.
2. Voltage Pin 9 remains HIGH.
3. Undistorted output pulse with 100% AM input.

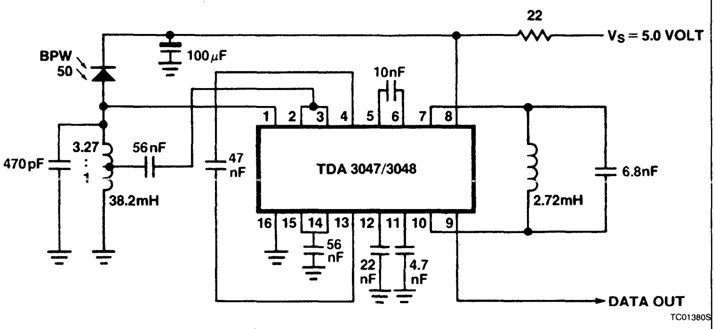
IR Preamplifier

TDA3048



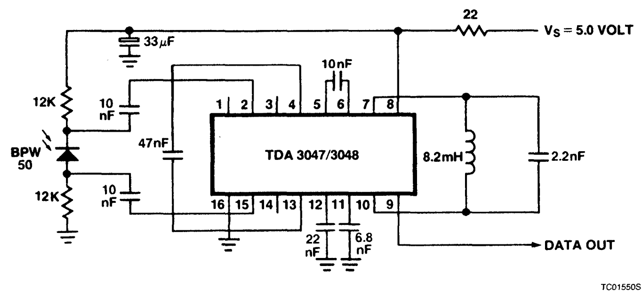
NOTE:
$I_{3,\ 14}$ is measured to ground, $V_{2-15(P-P)}$ is a symmetrical square wave measured in Figure 3; V_{CC} = 5V.

Figure 1. Typical Q-Factor Killer Current (Pins 3 and 14) as a Function of the Peak-to-Peak Input Voltage



NOTE:
N1 = 3.21
N2 = 1
Q = 16

Figure 2. Narrow-Band Receiver Using TDA3048



NOTE:
For better sensitivity both 12kΩ resistors may have a higher value.

Figure 3. Wide-Band Receiver With TDA3048

Signetics

AN172
Circuit Description of the Infrared Receiver TDA3047/TDA3048

Linear Products

Application Note

Author: A.J.E. Bretveld

INTRODUCTION

As a successor of the current integrated circuits TCA440 and NE555 for receiving infrared remote-controlled signals, a new integrated circuit has been developed.

In comparison with the TCA440-NE555 combination, this IC is aimed to have a higher replacement value and improved performance. The TDA3048 is equal to the TDA3047 except for the polarity of the output signal.

GENERAL DESIGN CONSIDERATIONS

The target of this development is to make a receiver integrated circuit for infrared remote-controlled signals which functions optimally in a narrow-band application.

This integrated circuit shall have the following advantages in comparison with the present TCA440-NE555 combination:

- A higher replacement value
- A considerable saving of the current consumption
- An improvement of the specification (less spread)
- Less periphery and no adjustment points
- Total spread on pulse widening < 10% by a standard RC-5 signal.

Besides, the IC is also suitable to be used in a RC-5 extended receiver and in a wide band receiver.

A standard bipolar process with single layer interconnect and without collector wall has been used.

Due to the low currents, a collector wall is not necessary.

FUNCTIONAL DESCRIPTION OF THE BLOCK PARTS

Figure 1 shows the block diagram of the TDA3047 and TDA3048.

Amplifier

The input signal is amplified by the gain-controlled amplifier. The output signal of the amplifier is fed to the synchronous demodulator inputs and to the reference amplifier.

Reference Amplifier

The reference amplifier amplifies and limits the input signal. The output signal of this amplifier is fed to the synchronous demodulator.

Synchronous Demodulator

In the synchronous demodulator, the input signal and reference signal are multiplied. The output signal of the demodulator is fed to the input of a pulse-shaper circuit and to the input of the AGC circuit.

AGC Circuit

The output signal of the synchronous demodulator is fed to the AGC circuit. The top level of the signal is detected by the AGC detector. Noise pulses are integrated by an internal capacitor. The output signal from the AGC detector is amplified and supplied to the first and second stage of the amplifier and to the Q-killing circuit.

Pulse-shaper Circuit

The output of the synchronous demodulator is also fed to the pulse-shaper circuit. The slicing level of the pulse-shaper is lower than the slicing level of the AGC detector.

The output of the pulse-shaper is fed to the output buffer.

Output Buffer

The output buffer gives for the TDA3047 an active-high level and for the TDA3048 an active-low level on the output pin. To obtain a correct RC-5 code a hysteresis circuit protects the output against spikes.

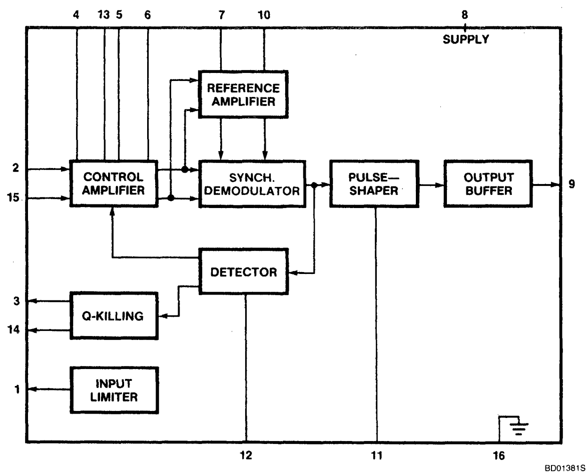


Figure 1. Block Diagram of the TDA3047/3048

Circuit Description of the Infrared Receiver TDA3047/TDA3048 AN172

Q-Killing Circuit

In the narrow-band application it is necessary to degenerate the Q of the input selectivity particularly when large signals occur at the input.

The output of the Q-killing circuit can be directly coupled to the input.

Input Voltage Limiter

In the narrow-band application high voltage peaks can occur on the input selectivity. The input limiter limits these voltage peaks to about 0.7V.

APPLICATION

The narrow-band application diagram has been given in Figure 2 and a lower performance wide-band application diagram in Figure 3.

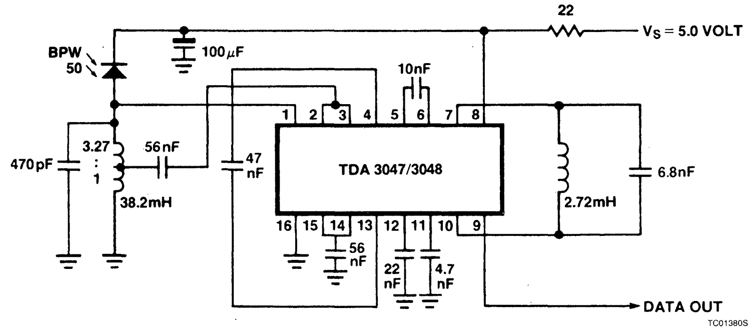


Figure 2. Narrow-Band Application Diagram of the TDA3047/3048

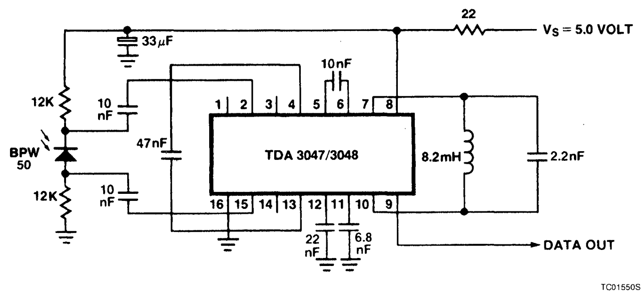


Figure 3. Wide-Band Application Diagram of the TDA 3047/3048

Signetics

Linear Products

AN173
Low Power Preamplifiers for IR Remote Control Systems

Application Note

INTRODUCTION

The monolithic integrated bipolar circuits TDA3047 and TDA3048 are amplifiers intended for use in infrared remote control systems. Both circuits are excellent and applicable as narrow-band amplifiers, especially for those types of remote control concepts which use the modulated transmission technique. Under certain conditions both ICs are also applicable as broadband amplifiers. The only difference between the ICs is polarity of the output signal. This type of IR amplifier offers the following advantages:

- Low power consumption, typically 10.5mV
- Gain-controlled amplification, control range 66dB
- High amplification factor, > 80dB, ensures a long range
- Great stability in signal handling
- Demodulation via a synchronous demodulator
- Automatic limitation of large input signals, 600mV
- Independent of large input amplitude variations with a Q-killer
- Applicable as narrow-or broadband amplifier

This circuit proves to be a reliable device with regard to interference from other IR sources such as light bulbs, etc.

The automatic gain control (AGC) ensures very good stability in amplification of large or low input signals, which correspond to short or long distances from transmitter to receiver.

FUNCTIONAL DESCRIPTION

The functional block diagram is shown in Figure 1. The input signal is applied to the gain-controlled multi-stage differential preamplifier, capacitively-coupled via C_2 and C_3. The capacitors C_4 and C_5 stabilize the differential preamplifier. Hereafter the signal is fed to a synchronous demodulator and the reference amplifier, which limits the input signal. After multiplication of the input and reference signal by the demodulator, the signal is applied to a pulse-shaper, whose time constant is controlled by C_8. The same signal is also used for the feedback loop, resulting in an automatic gain control defined by the amplitude of the input signal. The AGC acquisition time is set by C_7. The Q-killer limits the amplification of the tuned input circuit in conjunction with input amplitude. In this way the behavior of this device on large amplitude variations ensures a great stability in the signal handling. A maximum input limitation is achieved via the amplitude limiter, typically activated by a 600mV input signal.

The differential preamplifier has, in principle, two stages, as shown in Figure 2. Each stage is stabilized via an external feedback capacitor. Both define the lower boundary of the frequency, with the greatest influence from C_4 because stage 1 has the highest gain. Both capacitors should be specified so that interference from low frequencies is suppressed. For instance, bulbs radiate infrared frequencies at (n)(100Hz).

The highest boundary in frequency of this amplifier is greater than 1MHz and is given by the internal capacitance of this device.

IR AMPLIFIER

For remote control systems two different types of amplifiers are available. Both are described in the following sections.

Narrow-Band Amplifier

The diagram of Figure 3 shows the TDA3047/48 in such an application. Pin 15, one of the differential inputs, is grounded for AC, while the second input, Pin 2, is connect-

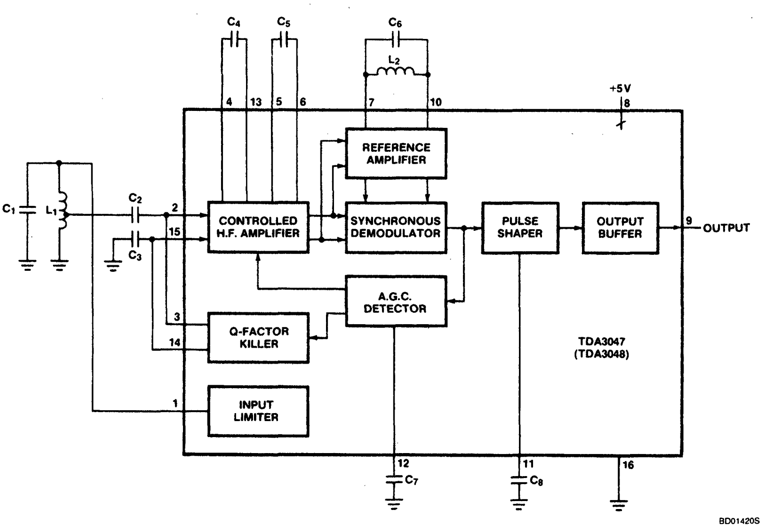


Figure 1. Functional Block Diagram

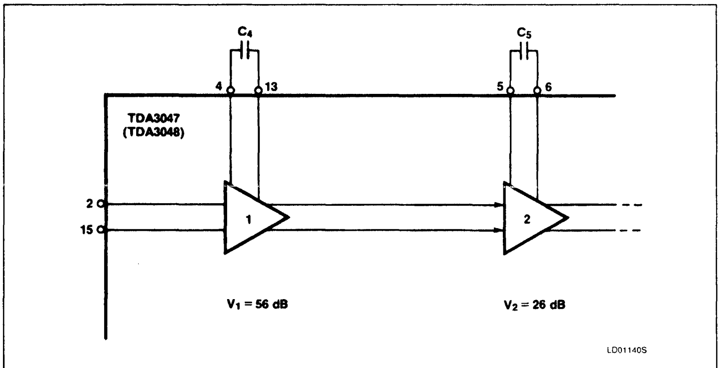


Figure 2

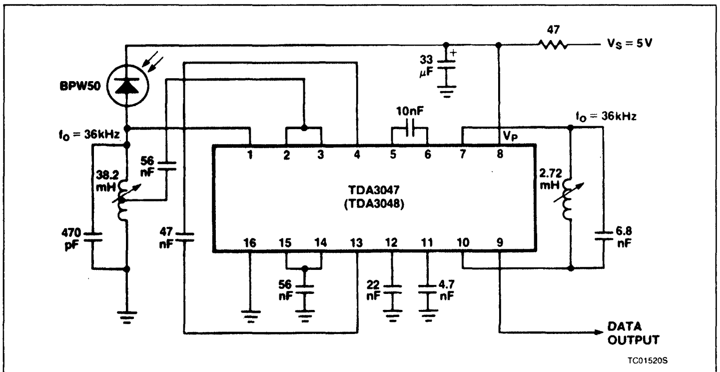


Figure 3

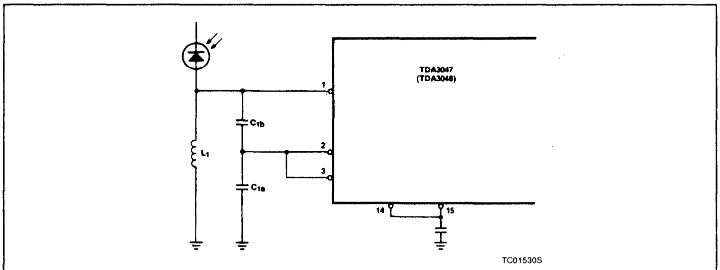


Figure 4

ed to the tuned input circuit via a capacitor of 0.056µF. The input voltage is taken with a transformer ratio N = 1:3. Direct coupling to the top will only lower the quality Q factor of the tuned input circuit, due to the relatively low input resistor, R_{IN}, of the IC.

The selectivity is obtained with the tuned input circuit and strongly reduces IR interferences. The effect of direct IR radiation is also avoided. Due to the low ohmic resistance of the coil, the IR receiving diode will never become saturated. The center frequency of the input tank must be equal to the modulation frequency of the transmitter used.

For this frequency (f_O) the input tank has a high impedance. Small variations of the current of the IR receiving diode at f_O result directly in large input signals.

This frequency (f_O) is equal to 37.5kHz for the SAA3004 transmitting chip. The RC combination of 47Ω and 0.33µF suppresses the unwanted current variations caused by the supply line.

The Q of the tuned input circuit is practically defined by the transformer ratio and the input resistor R_{IN} of the IC. The effect of R_{IN} to the quality Q_i of the coil is negligible, because R_{IN} is relatively low (typically 16kΩ).

The transformer ratio must be adjusted for small signals, so that the range is hardly influenced by component spread and/or tolerances in frequency at both sides in the system. The Q can be calculated from:

$$Q = \frac{1}{R_{L1}\sqrt{\frac{C_1}{L_1}} + \frac{1}{R_P}\sqrt{\frac{L_1}{C_1}}}$$

where R_{L1} is the ohmic resistance of the coil and the parallel resistor $R_P = n^2\ R_{IN1}$.

With the component values shown in Figure 4 and a given $R_{L1} = 125\Omega$, $R_{IN} = 16k\Omega$, the factor Q is calculated as Q = 13. The bandwidth is now known from

$$\Delta f = \frac{f_O}{Q} = 2.9kHz$$

The transformer ratio can also be realized with two capacitors in series, as shown in Figure 4, where the total capacity is equal to the required one.

The ratio is $n = \frac{C_{1a} + C_{1b}}{C_{1b}}$

With values of $C_{1a} = 2.2nF$, $C_{1b} = 560pF$ and $L_1 = 40mH$, about the same input quality will be obtained.

The AGC acquisition time and the time constant of the pulse-shaper are defined by the capacitors C_7 and C_8, respectively. The time constant at Pin 12 equals the length of a received data bit and C_8 delays the pulse-shaper output to the output stage.

The Q_S of the tuned circuit of the synchronous demodulator is practically given by the internal resistance, R_{IN2}, between Pins 7 and 10 and is calculated from

$$Q_S = \frac{1}{R_{L2}\sqrt{\frac{C_6}{L_2}} + \frac{1}{R_{In2}}\sqrt{\frac{L_2}{C_6}}}$$

with 12Ω for R_{L2} and 5kΩ for R_{IN}, $Q_S \simeq 7$. The quality Q_S is continuously limited. With a relatively high value for Q_S, the acquisition time will be increased and this will delay the pulse edges. By amplification of "biphase" modulated signals, disturbances could occur in the decoding. For correct decoding of

Low Power Preamplifiers for IR Remote Control Systems — AN173

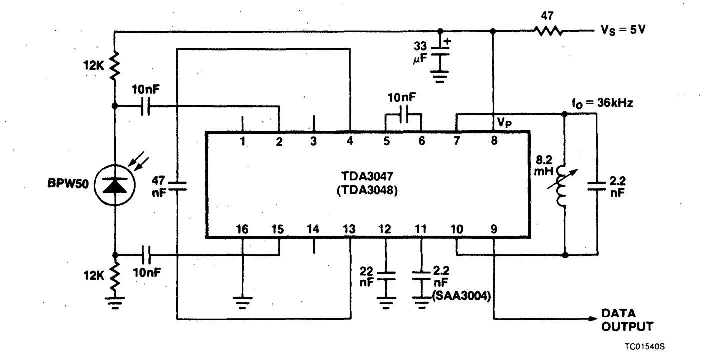


Figure 5

"biphase" coded data, a nearly exact position of the pulse edges is required.

Broadband Amplifier

The application as broadband amplifier is shown in Figure 5. The IR receiving diode is now positioned between both differential inputs, while the series resistors of 12kΩ are the work resistors. The Q killer and Amplitude Limiter do not have any function here and are not used. Also the resonance frequency, f_O, of the tuned demodulator circuit equals the modulation frequency of the remote transmitter.

The charge current to capacitor C_8 is equal to

$$I_{C8} = (C_8)\frac{\Delta V_{C8}}{\Delta t}$$

where Δt is the charge time and $\Delta\ VC_8$ is the voltage increment. IC_8 is generated by an internal current source.

The voltage increment at C_8 is proportional to Δt, with IC_8 constant and expressed as

$$\Delta V_{C8} = \frac{(I_{C8})(\Delta t)}{C_8}$$

The pulse width, Δt, of the demodulated signal must be large enough that VC_8 exceeds the threshold voltage of the pulse-shaper.

Given the format of the received data, C_8 will have different values

	Pulse Width	C_8
SAA3004	8.8µs	2.2nF

A 2.2nF capacitor in the SAA3004 remote control system is an optimum one.

The SAA3004, used in unmodulated mode, has a pulse width of 8.8µs. C_8 must have a low value so that the threshold voltage of the pulse-shaper is exceeded. On the other hand, if C_8 becomes too small, interference pulses will easily trigger the pulse-shaper. The selection of C_8 is a compromise between the sensitivity of the amplifier and the immunity against interference. Such a compromise is a 2.2nF capacitor for the unmodulated mode of the SAA3004, including the tolerances of the internal current sources. Given the technology, small tolerances are not possible.

Correct operation can not be guaranteed for the combination of a small pulse width (8.8µs) and a low source current. However, practical tests did show that correct operation of the SAA3004, in the unmodulated mode in combination with this type of preamplifier, can be realized.

CONSIDERATIONS FOR AMPLIFIER SELECTION

The narrow- or broadband application is defined by the following points:

- Modulation mode of the transmitter
- Requirements for the reach in distance
- Reliability (insensitivity to interference)
- Price-attractive total remote control system

Either modulated or unmodulated data transmission is possible with the SAA3004.

In the unmodulated mode, the logic representation of the data word is defined by the time intervals between the generated output pulses, each of 8.8µs width. In the modulated output mode, each active output stage has a burst of 6 clock periods.

The ground wave of this output, with a frequency of 38kHz, contains the IR power generated.

The greatest sensitivity is realized with a narrow-band amplifier, whose tuned input circuit is selected for this ground wave frequency.

In the unmodulated transmission mode, the single output pulse represents a continuous frequency spectrum, in which the generated IR power is divided. A broadband amplifier is then required.

The greatest range, with constant-current through the IR transmission diode(s), will be obtained with a narrow-band amplifier, because the signal-to-noise ratio is the largest value.

When IR interference is absent, the combination of modulated transmission mode and the narrow-band amplifier is the most preferable. With lower requirements for the reliability, less range, etc., the broadband amplifier is the most effective solution for both types of modulation modes.

RANGE

To give some idea what range can be expected, a number of measurements are made with the remote transmitters SAA3004.

With Various IR Output Powers

Transmitter SAA3004 drives 1 IR-transmitting diode with a peak current IC≅2A. In the modulated mode, the power product per bit equals

$(m)\ (I_F)\ (n)\ (t_P)$

where m = number of diodes, n = number of pulses per bit, and t_P = pulse width.

The power product for each bit is:

- Modulated mode $(m)\ (I_F)\ (n)\ (t_P) = (1)\ (2)\ (6)\ (8.8) = 106\mu A/sec$
- Unmodulated mode $(m)\ (I_F)\ (n)\ (t_P) = (1)\ (2)\ (1)\ (8.8) = 18\mu A/sec$

This power product is proportional to the generated IR power. Table 1 indicates the results of the measurements. Optic lenses will increase the distances about 10%.

With Equal Output Power

These measurements are done with one transmitting diode for each transmitter type

Table 1. Distance Reach With Various Power Products

	SAA3004	
	Modulated	**Unmodulated**
Power product	106µA/sec	18µA/sec
Narrow-band C_8 = 4.7nF	25mt	11mt
Broadband C_8 = 2.2nF	16mt	12mt

Table 2. Distance Reach With Constant Power Product of 18µA/sec

	SAA3004	
	Modulated	**Unmodulated**
Narrow-band C_8 = 4.7nF	11mt	11mt
Broadband C_8 = 2.2nF	8mt	12mt

Table 3. Application Possibilities

	SAA3004	
	Unmodulated	**Modulated**
Narrow-band	No sense; no selectivity	Great distance reach, high selectivity, reliable
Broadband	Function only possible with small width output pulse; less reliable	Low reach, low selectivity; interference.

and the power product/bit constant at 18µA/sec. Table 2 is comprised of the results from these measurements.

Results of the Measurements

The results of the measurements can be summarized as follows:

a. Only the combinations "modulated and narrow-band amplifier" are reasonable.

b. With the peak current I_F through one IR-transmitting diode, the range with one IR diode is limited.

c. A maximum range is obtained using the modulated mode of data transmitting, but the loss of power in the transmitter is of subordinate importance.

POWER DISSIPATION

In comparison with older types of preamplifiers, the power consumption is enormously reduced. For instance, the TDB2033 consumed 204mW at 12V supply, while the TDA3047/48 only takes 10mW at 5V supply, which is very useful for "standby" mode. A second advantage is the 5V supply which can also be used by the decoding microcomputer.

POSSIBLE APPLICATION COMBINATIONS

In Table 3, the different combinations are given for remote control systems operating in the modulated or unmodulated mode.

OUTPUT SIGNAL

As indicated in the introduction, the TDA3047 has an active-high output signal, while an active-low output is generated by the TDA3048. This choice in polarity is made available for maximum cooperation with the decoding part. If, for example, an 8048 microcomputer is used on interrupt level, with active-low at input $\overline{INT}$, the TDA3048 is then the correct amplifier. If the $\overline{INT}$ input is active-High, the TDA3047 outputs the proper high level.

PC BOARD DESIGN

Special attention must be given to the placement of C_5. The greatest distance must be realized between the position of this capacitor and the inputs 2 and/or 15. Ground connections and screening must also be done with great accuracy.

Signetics

Linear Products

Section 6
Television Subsystems

INDEX

Signetics

TDA4501
Small-Signal Subsystem IC for Color TV

Product Specification

Linear Products

DESCRIPTION

The integration into a single package of all small-signal functions (except the tuner) required for color TV reception is achieved in the TDA4501. The only additional circuits needed to complete the receiver are a tuner, the deflection output stages, and a color decoder. The TDA3563 or 67, NTSC color decoder, and TDA3653, vertical output, are ideal complements for the TDA4501.

The IC includes a vision IF amplifier with synchronous demodulator and AFC circuit, an AGC detector with tuner output, an integral three-level sandcastle pulse generator, and fully synchronized vertical and horizontal drive outputs. A triggered vertical divider automatically adapts to a 50 or 60Hz vertical signal and eliminates the need for an external vertical frequency control.

Signal strength-dependent, time constant switches in the horizontal phase detector make external VCR switching unnecessary.

Sound signals are demodulated and amplified within the IC in a circuit which includes volume control and muting.

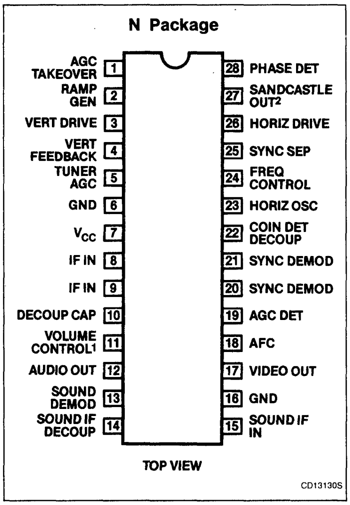
FEATURES

- **Vision IF amplifier with synchronous demodulator**
- **AGC detector for negative modulation**
- **AGC output to tuner**
- **AFC circuit**
- **Video and audio preamplifiers**
- **Sound IF amplifier and demodulator**
- **Choice of sound volume control or horizontal oscillator starting function**
- **Horizontal synchronization circuit with two control loops**
- **Triggered divider system for vertical synchronization and sawtooth generation giving automatic amplitude adjustment for 50 or 60Hz vertical signal**
- **Transmitter identification circuit with mute output**
- **Sandcastle pulse generator**

APPLICATION

- **Color TV**

PIN CONFIGURATION



ORDERING INFORMATION

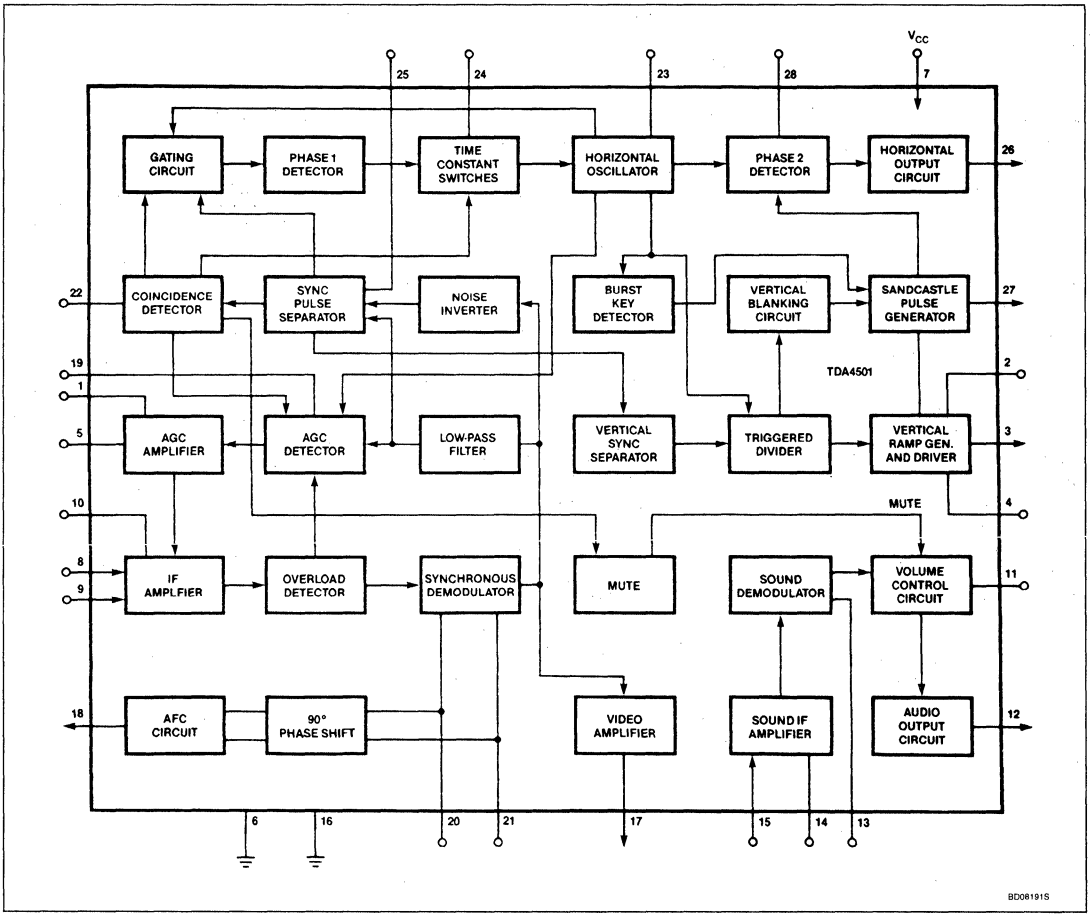
DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
28-Pin Plastic DIP (SOT-117)	−25°C to +65°C	TDA4501N

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
$V_{CC} = V_{7-6}$	Supply voltage (Pin 7)	13.2	V
P_{TOT}	Total power dissipation	1.7	W
T_A	Operating ambient temperature range	−25 to +65	°C
T_{STG}	Storage temperature range	−65 to +150	°C

Small-Signal Subsystem IC for Color TV TDA4501

BLOCK DIAGRAM



DC AND AC ELECTRICAL CHARACTERISTICS $V_{CC} = V_{7-6} = 10.5V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supplies					
V_{CC}	Supply voltage (Pin 7)	9.5	10.5	13.2	V
I_{CC}	Supply current (Pin 7)		120		mA
V_{11-6}	Supply voltage (Pin 11)		10.5		V
I_{11}	Supply current (Pin 11) for horizontal oscillator start		6		mA
Vision IF amplifier (Pins 8 and 9)					
V_{8-9}	Input sensitivity at 38.9MHz[1]	40	70	120	μV
V_{8-9}	Input sensitivity at 45.75MHz[1]		90		μV
R_{8-9}	Differential input resistance (Pin 8 to 9)		1.3		kΩ
C_{8-9}	Differential input capacitance (Pin 8 to 9)		5		pF
	AGC range		60		dB
V_{8-9}	Maximum input signal	50	70		mV
ΔV_{17-6}	Expansion of output signal for 50dB variation of input signal with V_{8-9} at 150μV (0dB)		1		dB
Video amplifier					
V_{17-6}	Output level for zero signal input (zero point of switched demodulator)		4.5		V
V_{17-6}	Output signal top sync level[2]		1.4		V
$V_{17-6(P-P)}$	Amplitude of video output signal (peak-to-peak value)		2.8		V
$I_{17(INT)}$	Internal bias current of output transistor (NPN emitter-follower)	1.4	2.0		mA
BW	Bandwidth of demodulated output signal		6		MHz
dG_{17}	Differential gain (Figure 3)		6		%
$d\rho$	Differential phase (Figure 3)		4		%
	Video non-linearity complete video signal amplitude			10	%

	Intermodulation (Figure 4) at gain control = 45dB
	f = 1.1MHz; blue;
	f = 1.1MHz; yellow;
	f = 3.3MHz; blue;
	f = 3.3MHz; yellow
		
		55
		50
		60
		55
			
			60
			54
			66
			59
					
					dB
					dB
					dB
					dB

  S/N S/N	Signal-to-noise ratio[3] $Z_S = 75\Omega$ $V_I = 10mV$ End of gain control range	 50 50	 54 56		 dB dB
	Residual carrier signal		7	30	mV
	Residual 2nd harmonic of carrier signal		3	30	mV

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DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $V_{CC} = V_{7-6} = 10.5V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS Min	Typ	Max	UNIT
Tuner AGC[4]					
V_{1-6}	Take-over voltage (Pin 1) for positive-going tuner AGC (NPN tuner)		3.5		V
$V_{1-6(RMS)}$	Starting point takeover; V = 5V		0.4	2	mV
$V_{1-6(RMS)}$	Starting point takeover; V = 1.2V	50	70		mV
V_{1-6}	Take-over voltage (Pin 1) for negative-going tuner AGC (PNP tuner)		8		V
$V_{1-6(RMS)}$	Starting point takeover; V = 9.5V		0.3	2	mV
$V_{1-6(RMS)}$	Starting point takeover; V = 5.6V	50	70		mV
$I_{5\ MAX}$	Maximum output swing	2	3		mA
$V_{5-6(SAT)}$	Output saturation voltage I = 2mA			300	mV
I_5	Leakage current			1	μA
ΔV_I	Input signal variation complete tuner control	0.5	2	4	dB
AFC circuit (Pin 18)[5]					
$V_{18-6(P-P)}$	AFC output voltage swing	9		10	V
$\pm I_{18}$	Available output current		1		mA
	Control steepness 100% picture carrier 10% picture carrier	 20 	 40 15	 80 	 mV/kHz mV/kHz
V_{18-6}	Output voltage at nominal tuning of the reference-tuned circuit		5.25		V
V_{18-6}	Output voltage without input signal	2.7	5.25	8.5	V
Sound circuit					
V_{15LIM}	Input limiting voltage $V_O = V_O$ maximum −3dB; $Q_L = 16$ $f_{AF} = 1kHz$; $f_C = 5.5MHz$		400		μV
R_{15-6}	Input resistance $V_{I(RMS)} = 1mV$		2.6		kΩ
C_{15-6}	Input capacitance $V_{I(RMS)} = 1mV$		6		pF
 AMR AMR	AM rejection (Figures 7 and 8) $V_I = 10mV$ $V_I = 50mV$		 35 43		 dB dB
$V_{12-6(RMS)}$	AF output signal $\Delta f = 7.5kHz$; minimum distortion	220	320		mV
Z_{12-6}	AF output impedance		150		Ω
THD	Total harmonic distortion $\Delta f = 27.5kHz$		1		%
 RR RR	Ripple rejection $f_K = 100Hz$, volume control 20dB when muted		 22 26		 dB dB
V_{12-6}	Output voltage Mute condition		2.6		V
S/N	Signal-to-noise ratio weighted noise (CCIR 468)		47		dB

Small-Signal Subsystem IC for Color TV | TDA4501

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $V_{CC} = V_{7-6} = 10.5V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Volume control					
V_{11-6}	Voltage (Pin 11 disconnected)		4.8		V
I_{11}	Current (Pin 11 short-circuited)		1		mA
R_{11-6}	External control resistor		10		kΩ
	Suppression output signal during Mute condition		66		dB
Horizontal synchronization					
	Slicing level sync separator		30		%
	Holding range PLL	800	1100	1500	Hz
	Catching range PLL	600	1000		Hz
	Control sensitivity video-to-oscillator; at weak signal at strong signal during scan during vertical retrace and during catching		 2 3 6		 kHz/μs kHz/μs kHz/μs
Second control loop (positive edge)					
$\Delta t_D/\Delta t_O$	Control sensitivity		300		μs
t_D	Control range		25		μs
	Phase adjustment via second control loop; control sensitivity maximum allowed phase shift		 25 ±2		 μA/μs μs
Horizontal oscillator (Pin 23)					
f_{FR}	Free-running frequency R = 35kΩ; C = 2.7nF		15,625		Hz
	Spread with fixed external components			4	%
Δf_{FR}	Frequency variation due to change of supply voltage from 8 to 12V		0	0.5	%
Δf_{FR}	Frequency variation with temperature			1×10^{-4}	K^{-1}
Δf_{FR}	Maximum frequency shift			10	%
Δf_{FR}	Maximum frequency deviation ($V_{7-6} = 8V$)			10	%
Horizontal output (Pin 26)					
V_{26-6}	Output voltage HIGH			13.2	V
V_{26-6}	Output voltage at which protection commences			15.8	V
V_{26-6}	Output voltage LOW at $I_{26} = 10mA$		0.3	0.5	V
δ_0	Duty cycle of horizontal output signal		45		%
t_R, t_F	Rise and fall times of output pulse		150		ns

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $V_{CC} = V_{7-6} = 10.5V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Flyback input and sandcastle output					
I_{27}	Input current required during flyback pulse	0.1		2	mA
V_{27-6}	Output voltage during burst key pulse	7.5			V
V_{27-6}	Output voltage during horizontal blanking	3.5	4.0	4.5	V
V_{27-6}	Output voltage during vertical blanking	1.8	2.2	2.6	V
	Width of burst key pulse	3.1	3.5	3.9	μs
	Width of horizontal blanking pulse	flyback pulse width			
	Width of vertical blanking pulse 50Hz working 60Hz working		 21 17		 lines lines
	Delay between start of sync pulse at video output and rising edge of burst key pulse		5.2		μs
Coincidence detector mute output (Pin 22)					
V_{22-6}	Voltage for in-sync condition		9.5		V
V_{22-6}	Voltage for no-sync condition no signal		1.0	1.5	V
V_{22-6}	Switching level to switch phase detector from slow to fast	4.9	5.3	5.8	V
	Fast-to-slow hysteresis		1		V
V_{22-6}	Switching level to activate mute function (transmitter identification)	2.25	2.5	2.75	V
$I_{22(P-P)}$	Output current for in-sync condition (peak-to-peak value)	0.7	1.0		mA
Vertical ramp generator (Pin 2)					
I_2	Input current during scan		12		mA
I_2	Discharge current during retrace		0.5		mA
V_{2-6}	Minimum voltage		1.5		V
Vertical output (Pin 3)					
I_3	Output current			10	mA
R_{3-6}	Output impedance		400		Ω
Feedback input (Pin 4)					
 V_{4-6} $V_{4-6(P-P)}$	Input voltage DC component AC component (peak-to-peak value)		 3 1.2		 V V
I_4	Input current			12	μA
	Internal precorrection to sawtooth		6		%
	Deviation amplitude 50/60Hz			5	%

NOTES:
1. Typical value taken at starting level of AGC.
2. Signal with negative-going sync, maximum white level 10% of the maximum sync amplitude (see Figure 2).
3. Signal-to-noise ratio equals $20\log \frac{V_O(\text{black-to-white})}{V_{N(RMS)} \text{ at } B = 5MHz}$
4. Starting point tuner takeover NPN current 1.8mA;
5. $V_{I(RMS)} = 10mV$; see Figure 1; Q-factor = 36.

FUNCTIONAL DESCRIPTION

IF Amplifier, Demodulator, and AFC

The IF amplifier has a symmetrical input (Pins 8 and 9), the input impedance of which is suitable for SAW filtering to be used. The synchronous demodulator and the AFC circuit share an external reference tuned circuit (Pins 20 and 21). An internal RC network provides the necessary phase-shifting for AFC operation. The AFC circuit provides a control voltage output with a swing greater than 9V from Pin 18.

AGC Circuit

Gating of the AGC detector is performed to reduce sensitivity of the IF amplifier to external electrical noise. The AGC time constant is provided by an RC circuit connected to Pin 19. Tuner AGC voltage is supplied from Pin 5 and is suitable for tuners with PNP or NPN RF stages. The sense of the AGC (to increase in a positive or negative direction) and the point of tuner take-over are preset by the voltage level at Pin 1.

Video Amplifier

The signal through the video amplifier comprises video and sound information; therefore, no gating of the video amplifier is performed during flyback periods.

Sound Circuit and Horizontal Oscillator Starting Function

The input to the sound IF amplifier is obtained by a bandpass filter coupling from the video output (Pin 17). The sound is demodulated and passed via a dual-function volume control stage to the audio output amplifier. The volume control function is obtained by connecting a variable resistor (10kΩ) between Pin 11 and ground, or by supplying Pin 11 with a variable voltage. Sound output is suppressed by an internal mute signal when no input signal is present.

The horizontal oscillator starting function is obtained by supplying Pin 11 with a current of 6mA during the switching-on period. The IC then uses this current to generate drive pulses for the horizontal deflection. For this application, the main supply voltage for the IC can be obtained from the horizontal deflection circuit.

Vertical Divider System

A triggered divider system is used to synchronize the vertical drive waveforms, adjusting automatically to 50 or 60Hz working. A large window (search window) is opened between counts of 488 and 722; when a separated vertical sync pulse occurs before count 576, the system works in the 60Hz mode; otherwise, 50Hz working is chosen.

A narrow window is opened when 15 approved sync pulses have been detected. Divider ratio between 522 and 528 switches to 60Hz mode; between 622 and 628 switches to 50Hz mode.

The vertical blanking pulse is also generated via the divider system by adding the anti-topflutter pulse and the blanking pulse.

Line Phase Detector

The circuit has three operating conditions:

a. Strong input signal and synchronized.

b. Weak signal and synchronized.

c. Non-synchronized (weak and strong) signal.

The input signal condition is obtained from the AGC circuit.

DC Volume Control/Horizontal Oscillator Start

The operation depends on the application. When during switch-on no current is supplied, Pin 11 will act as volume control. When a current of 6mA is applied, the volume control is set to maximum and the circuit will generate drive pulses for the horizontal deflection.

Small-Signal Subsystem IC for Color TV

TDA4501

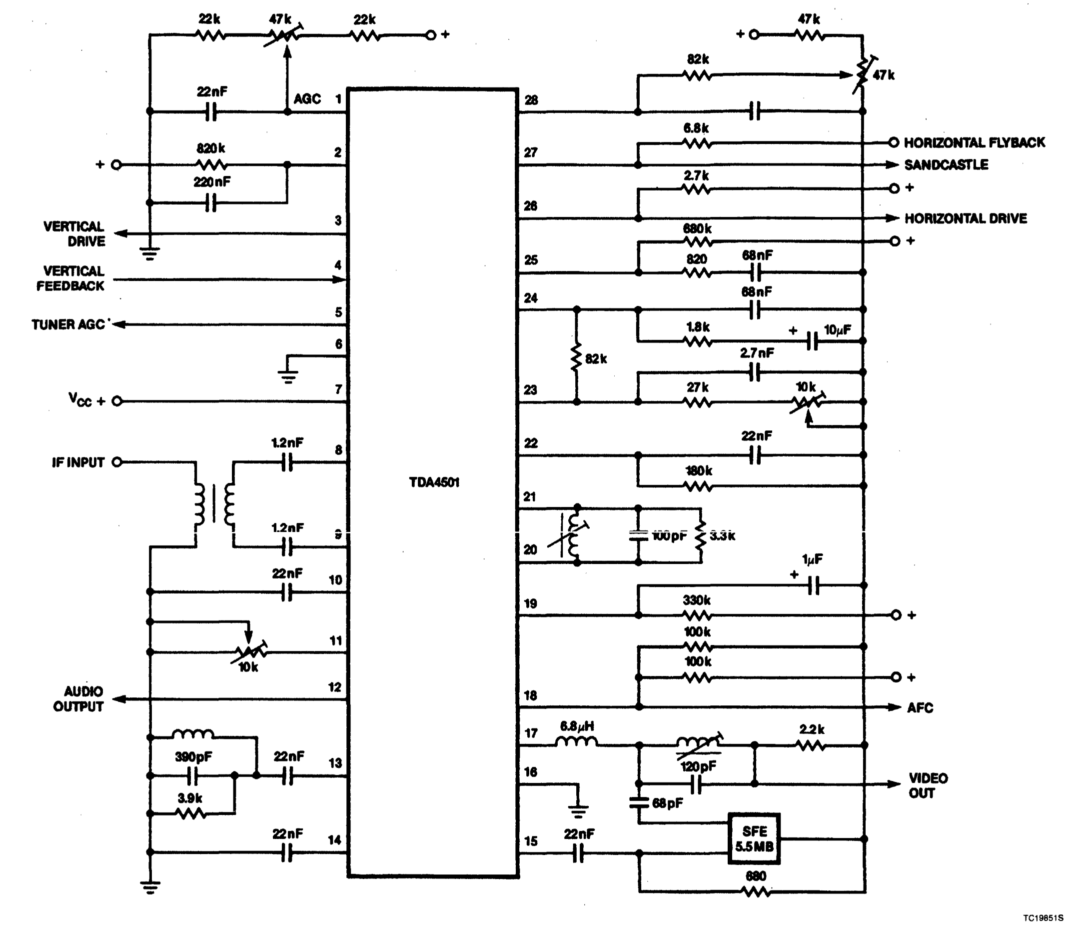


Figure 1. Application Diagram

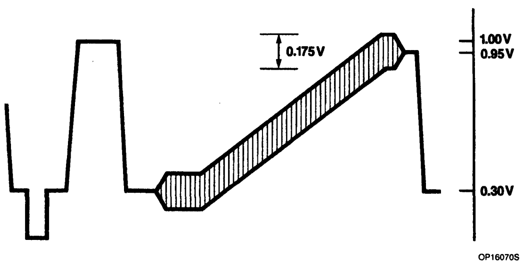


Figure 2. Video Output Signal

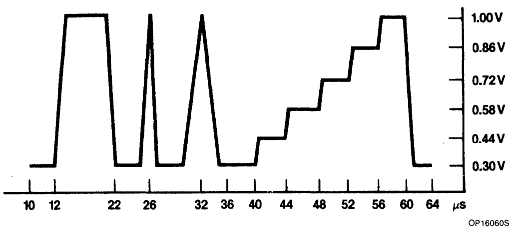


Figure 3. EBU Test Signal Waveform (Line 330)

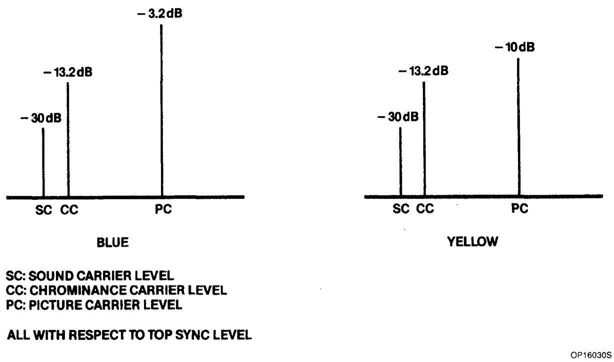


Figure 4. Input Signal Conditions

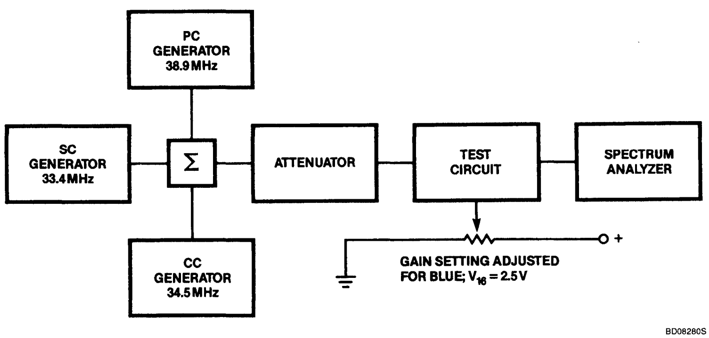


Figure 5. Test Setup Intermodulation

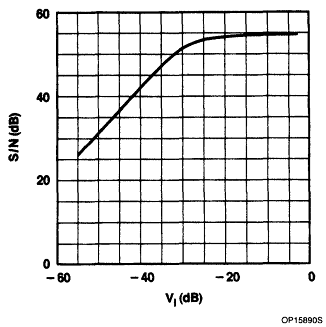


Figure 6. S/N Ratio as a Function of the Input Voltage

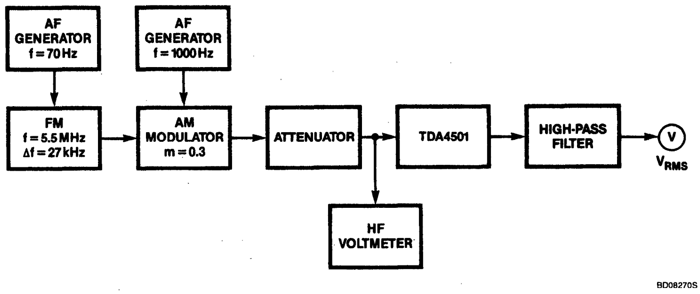


Figure 7. Test Setup AM Suppression

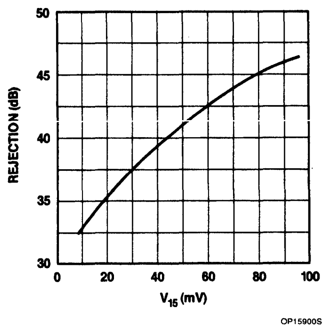


Figure 8. AM Rejection

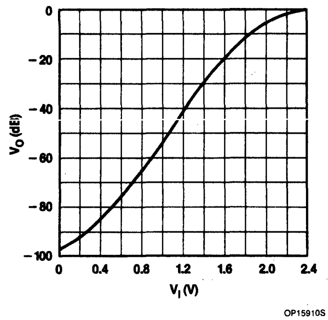


Figure 9. Volume Control Characteristics

Signetics

TDA4502
Small-Signal Subsystem IC for Color TV With Video Switch

Objective Specification

Linear Products

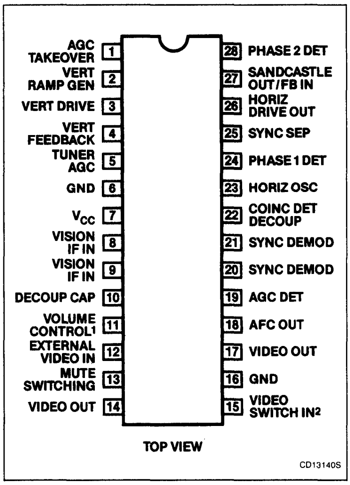
DESCRIPTION

The TDA4502 is a TV subsystem circuit intended to be used in color TV receivers. It is similar to the TDA4505, with the exception that it has no sound IF circuit or audio preamplifiers. Instead, it has a video switching input circuit for switching an external video signal.

FEATURES

- **Vision IF amplifier with synchronous demodulator**
- **AGC detector suited for negative modulation**
- **Tuner AGC**
- **AFC circuit with on/off switch**
- **Video preamplifier**
- **Video switch for an external video signal**
- **Horizontal synchronization circuit with two control loops**
- **Vertical synchronization (divider system) and sawtooth generation**
- **Sandcastle pulse generation**

PIN CONFIGURATION



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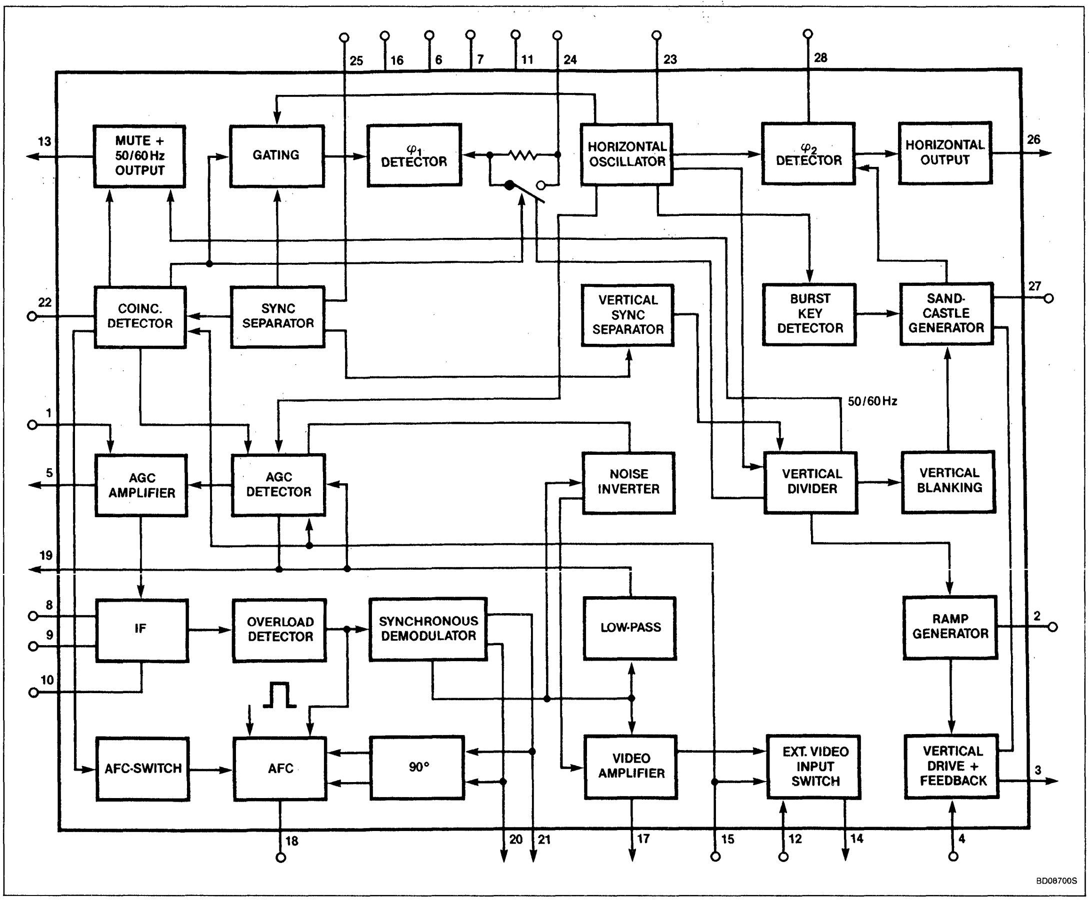
Small-Signal Subsystem IC for Color TV With Video Switch TDA4502

BLOCK DIAGRAM



Signetics

TDA4503
Small-Signal Subsystem for Monochrome TV

Product Specification

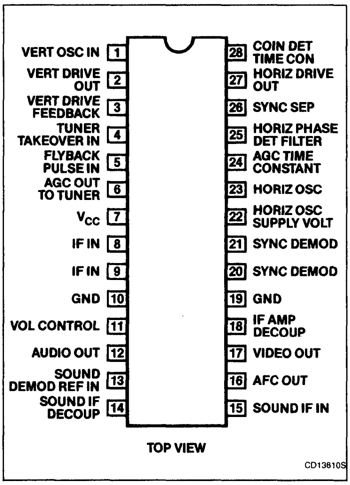
Linear Products

DESCRIPTION
The TDA4503 combines all small-signal functions (except the tuner) which are required for monochrome TV receivers. For a complete monochrome TV receiver only power output stages are required to be added for horizontal and vertical deflection, video and sound. This part is designed to work with the TDA3561, Vertical Output IC.

The TDA4503 can also be used in low cost color television receivers.

FEATURES
- **Vertical sync separator and oscillator**
- **Video preamplifier**
- **AGC detector**
- **Sync separator**
- **Horizontal synchronization**
- **Vision IF amplifier and synchronous demodulator**
- **Tuner AGC**
- **AFC circuit**
- **Sound IF amplifier and demodulator**
- **Audio preamplifier with DC volume control**
- **Gate pulse generator**

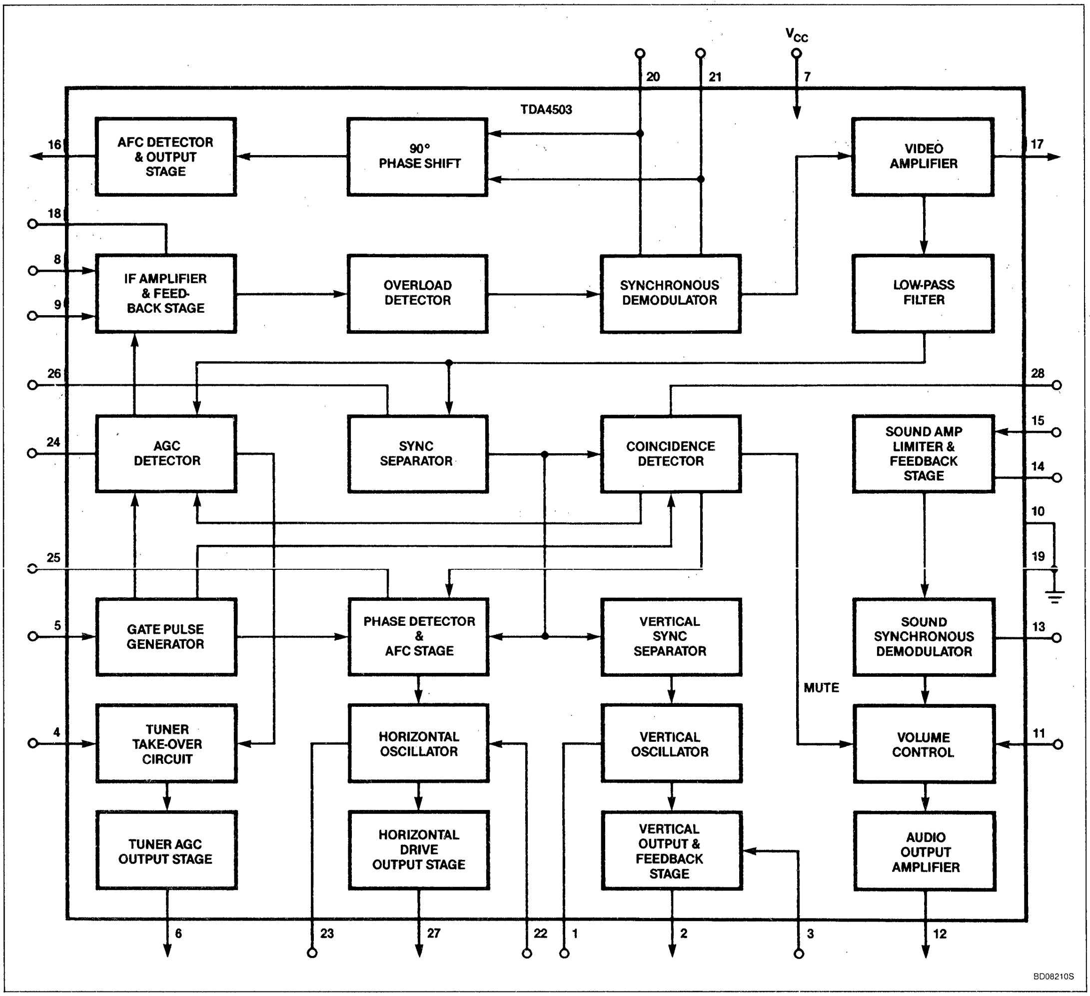
APPLICATIONS
- **Television receiver**
- **CATV converter**

PIN CONFIGURATION


ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
28-Pin Plastic DIP (SOT-117)	-25°C to +65°C	TDA4503N

BLOCK DIAGRAM



Small-Signal Subsystem for Monochrome TV TDA4503

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
$V_{CC} = V_{7-10}$	Supply voltage (Pin 7)	13.2	V
P_{TOT}	Total power dissipation	1.7	W
T_A	Operating ambient temperature range	−25 to +65	°C
T_{STG}	Storage temperature range	−65 to +150	°C

DC AND AC ELECTRICAL CHARACTERISTICS $V_{7-10} = 10.5V$; $V_{22-10} = 10.5V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supplies					
V_{7-10}	Supply voltage (Pin 7)	9.5	10.5	13.2	V
I_7	Supply current (Pin 7)		82	100	mA
V_{22-10}	Supply voltage (Pin 22)	9.5	10.5	13.2	V
I_{22}	Supply current (Pin 22)[1]		5	6.5	mA
P_{TOT}	Total power dissipation		920	1150	mW
Vision IF amplifier (Pins 8 and 9)					
V_{8-9}	Input sensitivity at 38.9 MHz[2]	40	80	120	μV
V_{8-9}	Input sensitivity at 45.75 MHz[2]		90		μV
R_{8-9}	Differential input resistance (Pin 8 to 9)		1.3		kΩ
C_{8-9}	Differential input capacitance (Pin 8 to 9)		5		pF
	AGC range		59		dB
V_{8-9}	Maximum input signal	50	70		mV
ΔV_{17-10}	Expansion of output signal (Pin 17) for 50dB variation of input signal (Pins 8 and 9)[3]		0.5	1.0	dB
Video amplifier[4]					
V_{17-10}	Output level for zero signal input (zero point of switched demodulator)	4.2	4.5	4.8	V
V_{17-10}	Output signal top sync level[5]	1.25	1.45	1.65	V
$V_{17-10(P-P)}$	Amplitude of video output signal (peak-to-peak value)	2.4	2.7	3.0	V
$I_{17(INT)}$	Internal bias current of output transistor (NPN emitter-follower)	1.4	2.0		mA
BW	Bandwidth of demodulated output signal		5		MHz
G_{17}	Differential gain[6] (Figure 5)		6		%
	Differential phase[6] (Figure 5)		4		%
	Video non-linearity over total video amplitude (peak white to black)			10	%
	Intermodulation (Figures 6 and 7) at gain control = 45dB f = 1.1MHz; blue f = 1.1MHz; yellow f = 3.3MHz; blue f = 3.3MHz; yellow	 55 50 60 55	 60 54 66 59		 dB dB dB dB
 S/N S/N	Signal-to-noise ratio[7] at $V_I = 10mV$ at end of AGC range	 50 50	 54 56		 dB dB
S/N	as a function of input signal	see Figure 8			
	Residual AM of intercarrier output signal[8]		5	10	%
	Residual carrier signal		7	30	mV
	Residual 2nd harmonic of carrier signal		3	30	mV

Small-Signal Subsystem for Monochrome TV TDA4503

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) V_{7-10} = 10.5V; V_{22-10} = 10.5V; T_A = 25°C, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS Min	Typ	Max	UNIT
Tuner AGC[9]					
V_{4-10}	Takeover voltage (Pin 4) for positive-going tuner AGC (NPN tuner)		3.5		V
$V_{8-9(RMS)}$	Starting point takeover at V_{4-10} = 5V (RMS value)		0.4	2.0	mV
$V_{8-9(RMS)}$	Starting point takeover at V_{4-10} = 1.2V (RMS value)	50	70		mV
V_{4-10}	Takeover voltage (Pin 1) for negative-going tuner AGC (PNP tuner)		8		V
$V_{8-9(RMS)}$	Starting point takeover at V_{4-10} = 9.5V (RMS value)		0.3	2.0	mV
$V_{8-9(RMS)}$	Starting point takeover at V_{4-10} = 5.6V (RMS value)	50	70		mV
I_{6MAX}	Maximum tuner AGC output swing	2	3		mA
$V_{6-10(SAT)}$	Output saturation voltage at I_6 = 2mA			300	mV
I_6	Leakage current at Pin 6			1	μA
ΔV_{8-9}	Input signal variation required for complete tuner control	0.5	2	4	dB
AFC circuit (Pin 16)[10]					
$V_{16-10(P-P)}$	AFC output voltage swing (peak-to-peak value)	9		10	V
$\pm I_{16}$	Available output current		1		mA
	Control steepness at 100% picture carrier 10% picture carrier	 20 	 40 15	 80 	 mV/kHz mV/kHz
V_{16-10}	Output voltage at nominal tuning of the reference-tuned circuit		5.25		V
V_{16-10}	Output voltage without input signal	2.7	6.0	8.5	V
Sound circuit					
V_{15LIM}	Input limiting voltage[11] (RMS value) at $V_O = V_{O\ MAX}$−3dB		2		mV
R_{15-10}	Input resistance at $V_{I(RMS)}$ = 1mV		2.6		kΩ
C_{15-10}	Input capacitance at $V_{I(RMS)}$ = 1mV		6		pF
 AMR AMR	AM rejection (Figures 7 and 8) at V_I = 10mV V_I = 50mV		 35 43		 dB dB
$V_{12-6(RMS)}$	AF output signal[12] (RMS value)	220	320		mV
Z_{12-10}	AF output impedance		150		Ω
THD	Total harmonic distortion[12]		1		%
 RR RR	Ripple rejection at f_K = 100Hz, volume control 20dB when muted		 22 26		 dB dB
V_{12-10}	Output voltage in mute condition		2.6		V
S/N	Signal-to-noise-ratio; weighted noise (CCIR 468)		47		dB
Volume control					
V_{11-10}	Voltage (Pin 11 disconnected)		6.9		V
I_{11}	Current (Pin 11 connected to ground)		1		mA
R_{11-10}	External control resistor[13]		5		kΩ
	Suppression of output signal during mute condition		66		dB

Small-Signal Subsystem for Monochrome TV TDA4503

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) V_{7-10} = 10.5V; V_{22-10} = 10.5V; T_A = 25°C, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS Min	Typ	Max	UNIT
Horizontal synchronization					
	Slicing level sync separator[14]		30		%
	Phase-locked loop holding range	±800	±1100	±1500	Hz
	Phase-locked loop catching range	±600	1000		Hz
	Control sensitivity video to flyback[15]		2.3		kHz/μs
	Delay between leading edge of sync pulse and zero cross-over of sawtooth (Pin 5)		3		μs
Horizontal oscillator (Pin 23)					
f_{FR}	Free-running frequency; R = 35kΩ; C = 2.7nF		15,626		Hz
	Spread with fixed external components			4	%
Δf_{FR}	Frequency variation due to change of supply voltage from 8 to 12V		0	0.5	%
TC	Temperature coefficient			1×10^{-4}	°C^{-1}
Δf_{FR}	Maximum frequency shift			10	%
Δf_{FR}	Maximum frequency deviation (V_{7-10} = 8V)			10	%
Horizontal output (Pin 27)					
I_{27}	Output current	5			mA
R_{27}	Output impedance		200		Ω
V_{27-10} V_{27-22}	Output voltage at I_{27} = 5mA		1.4 2.5		V V
a	Duty factor of horizontal output signal[16]	0.35	0.40	0.45	%
t_R, t_F	Rise and fall times of output pulse		400		ns
Flyback input (Pin 5)					
V_5	Amplitude of input pulse	2	4	6	V
V_5	Voltage at which gate pulse generator changes state[17]		0		V
Coincidence detector mute output (Pin 28)[18]					
V_{28-10}	Voltage for in-sync condition		9.5		V
V_{28-10}	Voltage for no-sync condition (no input signal)		1.0	1.5	V
V_{28-10}	Voltage level for phase detector to switch from slow to fast	3.7	4.1	4.5	V
	Fast-to-slow hysteresis		1		V
V_{28-10}	Voltage level to activate mute function (transmitter identification)	2.25	2.5	2.75	V
$I_{22(P-P)}$	Output current for in-sync condition (peak-to-peak value)	0.7	1.0		mA

6

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) V_{7-10} = 10.5V; V_{22-10} = 10.5V; T_A = 25°C, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Vertical oscillator (Pin 1)					
f_{FR}	Free-running frequency at C = 220nF; R = 560kΩ		47.5		Hz
	Spread with fixed external components			4	%
	Holding range at nominal frequency	52.5			Hz
TC	Temperature coefficient			2×10^{-4}	$°C^{-1}$
Δf_{FR}	Frequency variation due to change of supply voltage from 9.5 to 12V		3	5	%
I_1	Leakage current at Pin 1			1.6	μA
Vertical output (Pin 2)					
I_2	Output current	1	1.3		mA
R_2	Output resistance		2		kΩ
Feedback input (Pin 3)					
V_{3-10} $V_{3-10(P-P)}$	Input voltage DC component AC component (peak-to-peak value)	4.0	5.0 1.2	5.5	V V
I_3	Input current			12	μA
ΔI_3	Non-linearity of deflector current at V_{7-10} = 10.5V			2.5	%
	Delay between leading edge of vertical sync and start of vertical oscillator flyback	6		10	μs

NOTES:

1. The horizontal oscillator can be started by supplying a current of 6mA to Pin 22. Taking this current from the mains rectifier allows the positive supply voltage to Pin 7 to be derived from the horizontal output stage (the load current of Pin 27 is additional to the 6mA quoted).
2. At start of AGC.
3. Measured with 0dB = 200μV.
4. Measured at 10mV (RMS) top sync output signal.
5. Signal with negative-going sync; top white = 10% of the top sync amplitude.
6. Measured with test line as shown in Figure 3. The differential gain is expressed as a percentage of the difference in peak amplitudes between the largest and smallest values relative to the subcarrier amplitude at blanking level. The differential phase is defined as the difference in degrees between the largest and smallest phase angles.
7. Measured with a source impedance of 75Ω.

   $$\text{Signal-to-noise ratio} = 20\log \frac{V_O \text{ black-to-white}}{V_{I(RMS)} \text{ at } B = 5\text{MHz}}$$

8. Measured with a sawtooth-modulated input signal: m = 90%; $V_{I(RMS)}$ = 10mV;

   $$\text{Amplitude modulation} = \frac{V_O \text{ SC at top sync} - V_O \text{ SC at white}}{V_O \text{ SC at top sync} + V_O \text{ SC at white}} \times 100\%.$$

   (SC = sound carrier)
9. Starting point of tuner take-over for an NPN tuner is when I_6 = 1.8mA, and for a PNP tuner is when I_6 = 0.2mA.
10. Measured at $V_{8\text{-}9(RMS)}$ = 10mV and Pin 16 loaded with 2 × 100kΩ between V_7 and ground. Reference tuned circuit Q-factor = 36.
11. Reference tuned circuit Q-factor = 16; audio frequency = 1kHz; carrier frequency = 5.5 MHz.
12. The demodulator tuned circuit must be tuned for minimum distortion; output signal is measured at Δf = 7.5kHz; other measurements are at Δf = 27.5kHz.
13. Volume control can be realized by a variable resistor (5kΩ) connected between Pin 11 and ground, or by a variable voltage direct to Pin 11 (the low value of input impedance to Pin 11 must be taken into account).
14. The sync separator is noise-gated; the slicing level is referred to the top sync level and is independent of the video signal. The value stated is a percentage of the sync pulse amplitude, the level being dependent on external resistors connected to Pin 26.
15. The phase detector current is increased by a factor of seven during catching and when the phase detector is switched to 'fast' via Pin 28, thus ensuring a wide catching range and a high dynamic loop gain.
16. The negative going edge initiates switching-off of the line output transistor (simultaneous driver).
17. The circuit requires an integrated flyback pulse. Gate pulses for AGC and coincidence detectors are obtained from the sawtooth waveform.
18. The functions of in-sync, out-of-sync, and transmitter identification are combined on Pin 28. For the reception of VCR signals, V_{28} must be fixed between 3V and 4.5V so that the time constant is fast and sound information is preserved.

FUNCTIONAL DESCRIPTION

IF Amplifier, Demodulator, and AFC

The IF amplifier operates with symmetrical inputs at Pins 8 and 9 and has an input impedance suitable for SAW filter application. The amplifier sensitivity gives a peak-to-peak output voltage of 3V for an RMS input of 70μV. The demodulator and the AFC circuit share an external reference tuned circuit (Pins 20 and 21) and an internal RC network provides the phase-shifting necessary for AFC operation. The AFC circuit provides a control voltage output with a (typical) swing of 9V from Pin 16 (V_{CC} = 10.5V).

AGC Circuit

Gating of the AGC detector is performed to reduce sensitivity of the IF amplifier to external electrical noise. The AGC time constant is provided by an RC network connected to Pin 24. The typical gain control range of the IF amplifier is 60dB. Tuner AGC voltage is supplied from Pin 6 and is suitable for tuners with PNP or NPN RF stages. The sense of the AGC (to increase in a positive or negative direction) and the point of tuner takeover are preset by the voltage level at Pin 4 (V_4 = 3.5V (typ.) for positive AGC; V_4 = 8V (typ.) for negative AGC).

Video Amplifier

The video signal output from Pin 17 has a peak-to-peak value of 3V (top sync level = 1.5V) and carries negative-going sync. In order to retain sound information at Pin 17, the video signal is not blanked during flyback periods.

Sound Circuit

The sound IF signal present at the video output (Pin 17) is coupled to the sound circuit by a bandpass filter to Pin 15. The sound circuit has an amplifier-limiter stage, a synchronous demodulator with reference tuned circuit at Pin 13, a volume control stage, and an output amplifier. The volume control has a range of approximately 80dB and the audio output signal at maximum volume and with Δf = 7.5kHz is 320mV (RMS value). The sound output signal is suppressed when no input signal is detected.

Synchronization Circuits

The sync separator slicing level is determined by an external resistor network at Pin 26. The slicing level is referred to the top sync level and the recommended value for slicing is 30%. Internal protection from electrical noise is included.

A gated phase detector compares the phase of the separated sync pulses with a sawtooth waveform obtained from the flyback pulse at Pin 5. In sync and out-of-sync conditions are detected by the coincidence detector at Pin 28 (this circuit also gives transmitter identification). During the out-of-sync condition, gating of the phase detector is switched off and the output current from the phase detector increases to give the detector a short time-constant and thus a fast response. This condition can be imposed by clamping the voltage at Pin 28 to 3.5V for the reception of VCR signals.

The horizontal oscillator frequency is controlled by the output voltage of the phase detector circuit. The horizontal drive output from Pin 27 has a duty factor of 40%.

Vertical sync pulses are separated by an internal integrating network and are used to trigger the vertical oscillator. A comparator circuit compares the vertical sawtooth waveform, generated by the vertical oscillator, with feedback from the deflection coils, and supplies the drive voltage for the output stage at Pin 2.

Power Supplies

The main supply is to Pin 7 (positive supply) and Pin 10 (ground). The horizontal oscillator is supplied from Pin 22 to facilitate starting of the oscillator from a high-voltage rail. A special ground connection at Pin 19 is used by critical voltage dividers in the feedback loops of the vision and sound IF circuits.

Small-Signal Subsystem for Monochrome TV TDA4503

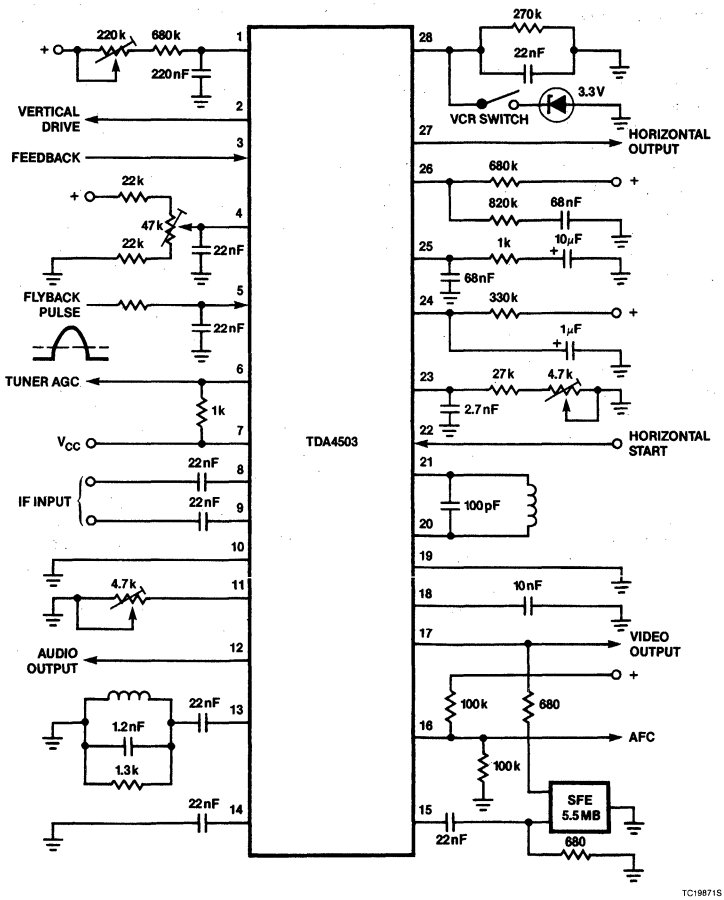


Figure 1. Application Circuit Diagram

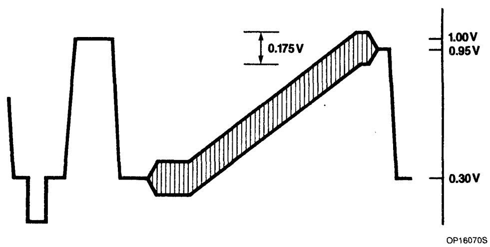


Figure 2. Video Output Signal

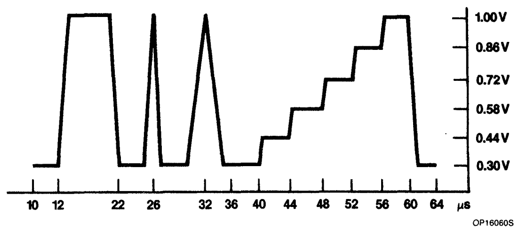


Figure 3. EBU Test Signal Line 330

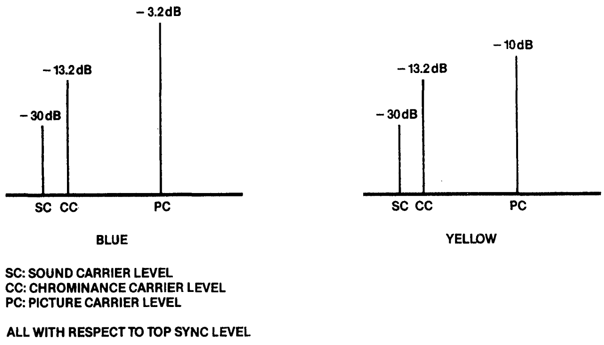


Figure 4. Input Signal Conditions for Intermodulation Test

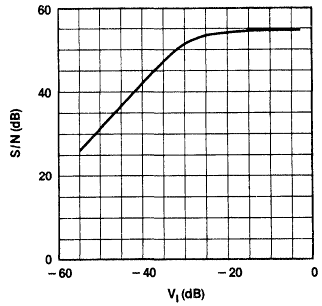
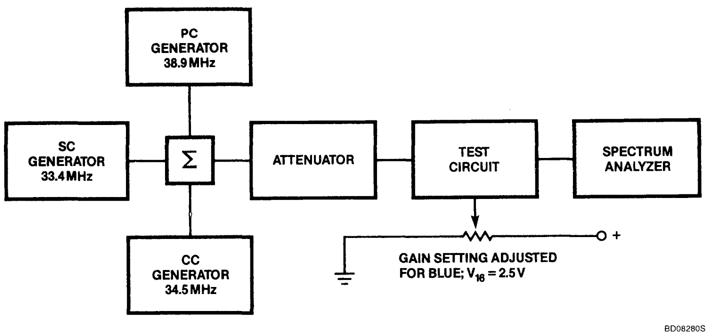


Figure 6. Signal-to-Noise Ratio as a Function of Input Voltage



NOTE:

$$\text{Value at 1.1MHz} = 20\log\frac{V_O \text{ at } 4.4\text{MHz}}{V_O \text{ at } 1.1\text{MHz}} + 3.6\text{dB};$$

$$\text{Value at 3.3MHz} = 20\log\frac{V_O \text{ at } 4.4\text{MHz}}{V_O \text{ at } 3.3\text{MHz}}.$$

Figure 5. Circuit for Intermodulation Test

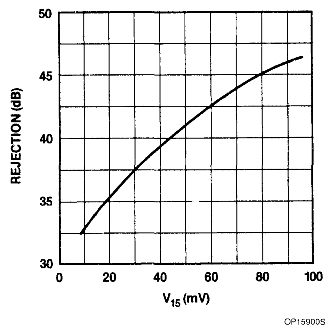


Figure 8. Typical Amplitude Modulation Rejection Curve

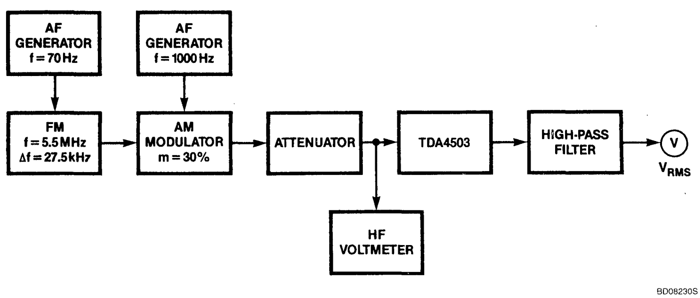


Figure 7. Circuit for Amplitude Modulation Rejection Test

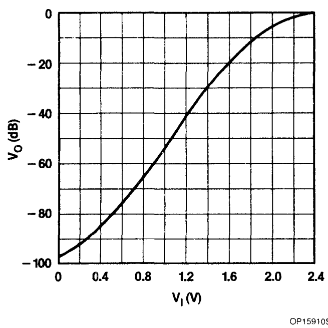


Figure 9. Volume Control Characteristic

Signetics

TDA4505
Small-Signal Subsystem IC for Color TV

Preliminary Specification

Linear Products

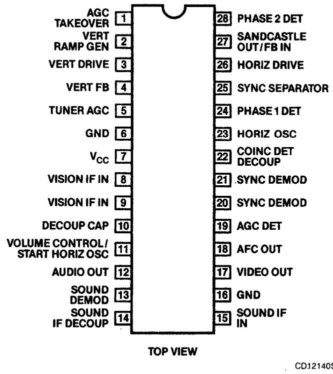
DESCRIPTION
The TDA4505 is a TV subsystem circuit intended to be used for base-band demodulation applications. This circuit consists of all small-signal functions (except the tuner) required for a quality color television receiver. The only additional circuits needed to complete a receiver are a tuner, the deflection output stages, and a color decoder. The TDA3563 or 67, NTSC color decoder, and the TDA3654 vertical output, are ideal complements for the TDA4505.

FEATURES
- **Vision IF amplifier with synchronous demodulator**
- **Tuner AGC (negative-going control voltage with increasing signal)**
- **AGC detector for negative modulation**
- **AFC circuit**
- **Video preamplifier**
- **Sound IF amplifier, demodulator and preamplifier**
- **DC volume control**
- **Horizontal synchronization circuit with two control loops**
- **Extra time constant switches in the horizontal phase detector**
- **Vertical synchronization (divider system) and sawtooth generation with automatic amplitude adjustment for 50 or 60Hz**
- **Three-level sandcastle pulse generation**

APPLICATIONS
- **Color television receiver**
- **CATV converters**
- **Base-band processing**

PIN CONFIGURATION

N Package



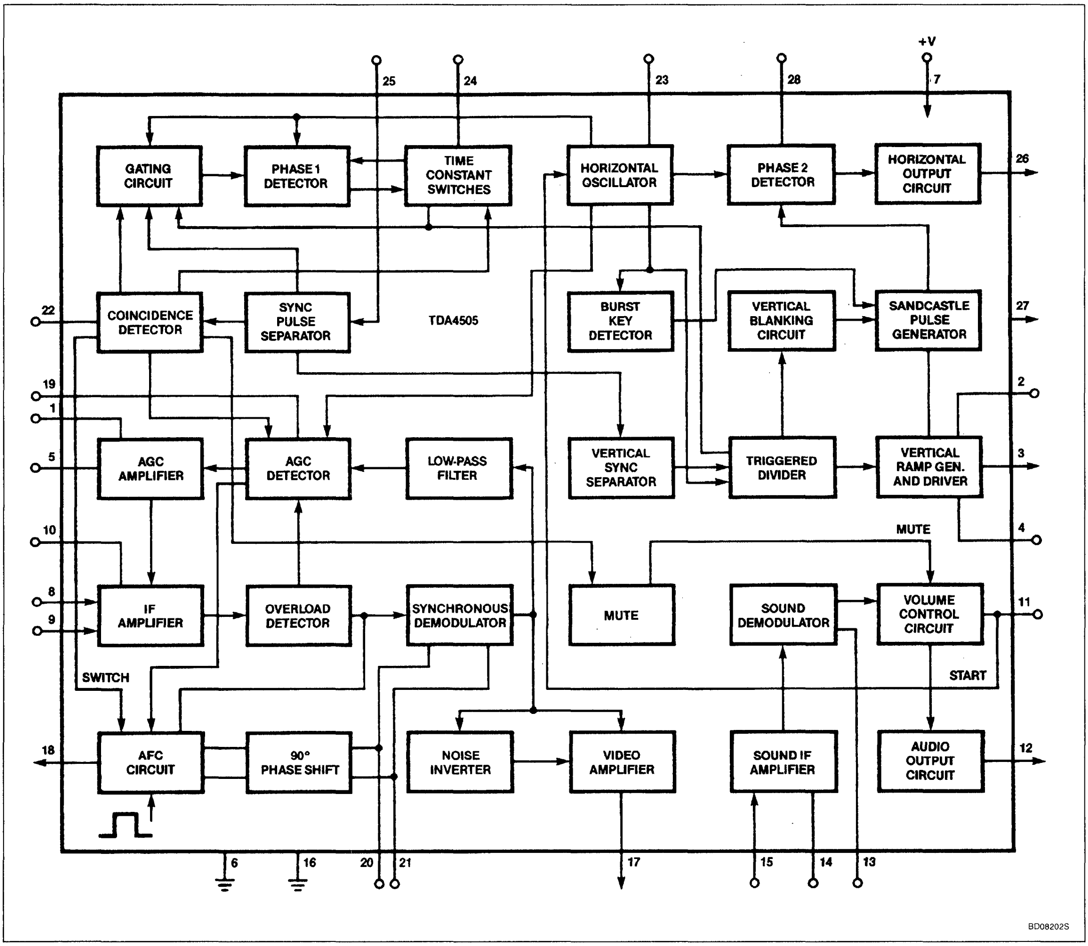
ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
28-Pin Plastic DIP (SOT-117)	−25°C to +65°C	TDA4505N
28-Pin Plastic DIP (SOT-117)	−25°C to +65°C	TDA4505AN
28-Pin Plastic DIP (SOT-117)	−25°C to +65°C	TDA4505BN

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{CC}	Supply voltage (Pin 7)	13.2	V
P_{TOT}	Total power dissipation	2.3	W
T_A	Operating ambient temperature range	-25 to +65	°C
T_{STG}	Storage temperature range	-65 to +150	°C

BLOCK DIAGRAM



DC AND AC ELECTRICAL CHARACTERISTICS $V_{CC} = V_{7-6} = 12V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supplies					
V_{7-6}	Supply voltage (Pin 7)	9.5	12	13.2	V
I_7	Supply current (Pin 7)		135		mA
V_{11-6}	Supply voltage (Pin 11)[1]		8.6		V
I_{11}	Supply current (Pin 11) for horizontal oscillator start		6	8	mA
Vision IF amplifier (Pins 8 and 9)					
V_{8-9}	Input sensitivity 38.9MHz on set AGC	60	100	140	μV
V_{8-9}	45.75MHz on set AGC		120		μV
R_{8-9}	Differential input resistance (Pin 8 to 9)	800	1300	1800	Ω
C_{8-9}	Differential input capacitance (Pin 8 to 9)		5		pF
G_{8-9}	Gain control range	56	60		dB
V_{8-9}	Maximum input signal	50	100		mV
ΔV_{17-6}	Expansion of output signal for 50dB variation of input signal with V_{8-9} at 150μV (0dB)		1		dB
Video amplifier measured at top sync input signal voltage (RMS value) of 10mV					
V_{17-6}	Output level for zero signal input (zero point of switched demodulator)		5.8		V
V_{17-6}	Output signal top sync level[2]	2.7	2.9	3.1	V
$V_{17-6(P-P)}$	Amplitude of video output signal (peak-to-peak value)		2.6		V
$I_{17(INT)}$	Internal bias current of output transistor (NPN emitter-follower)	1.4	2.0		mA
BW	Bandwidth of demodulated output signal	5			MHz
G_{17}	Differential gain (Figure 3)[3]		4	10	%
φ	Differential phase (Figure 3)[3]		3	10	deg.
	Video non-linearity[4] complete video signal amplitude			10	%
	Intermodulation (Figure 4) at gain control = 45dB f = 1.1MHz; blue f = 1.1MHz; yellow f = 3.3MHz; blue f = 3.3MHz; yellow	 55 50 60 55	 60 54 66 59		 dB dB dB dB
 S/N S/N	Signal-to-noise ratio[5] $Z_S = 75Ω$; $V_I = 10mV$ end of gain control range	 50 50	 54 56		 dB dB
	Residual carrier signal		7	30	mV
	Residual 2nd harmonic of carrier signal		24	30	mV

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $V_{CC} = V_{7-6} = 12V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Tuner AGC[13]					
$V_{1-6(RMS)}$	Minimum starting point take-over			0.5	mV
$V_{1-6(RMS)}$	Maximum starting point take-over	50	100		mV
I_{5MAX}	Maximum output swing	6	8		mA
$V_{5-6(SAT)}$	Output saturation voltage I = 2mA			300	mV
I_5	Leakage current			1	μA
ΔV_I	Input signal variation complete tuner control ($\Delta I_5 = 2mA$)	0.5	2	5	dB
AFC circuit (Pin 18)[6]					
$V_{18-6(P-P)}$	AFC output voltage swing	9.5	10.35	11	V
$\pm I_{18}$	Available output current		2.6		mA
	Control steepness		70		mV/kHz
V_{18-6}	Output voltage at nom. tuning of the reference-tuned circuit		6		V
I_{18}	Offset current AFC output (Pins 20 and 21 short-circuited)		TBD		μA
Sound circuit					
V_{15LIM}	Input limiting voltage $V_O = V_{O\ MAX} - 3dB$; $Q_L = 16$; $f_{AF} = 1kHz$; $f_C = 5.5MHz$		400	800	μV
R_{15-6}	Input resistance $V_{I(RMS)} = 1mV$		2.6		kΩ
C_{15-6}	Input capacitance $V_{I(RMS)} = 1mV$		6		pF
AMR AMR	AM rejection (Figures 7 and 8) $V_I = 10mV$ $V_I = 50mV$		46 50		dB dB
$V_{12-6(RMS)}$	AF output signal $\Delta f = 7.5kHz$; minimum distortion	400	600	800	mV
$V_{12-6(RMS)}$	AF output signal; $\Delta f = 50kHz$ Pin 11 used as starting pin	300	700	1200	mV
Z_{12-6}	AF output impedance		25	100	Ω
THD	Total harmonic distortion volume control 20dB, $\Delta f = 27.5kHz$; weighted acc. CCIR 468		1	3	%
RR RR	Ripple rejection $f_k = 100Hz$, volume control 20dB when muted		35 30		dB dB
V_{12-6}	Output voltage in Mute condition		3.0		V
S/N	Signal-to-noise ratio; $\Delta f = 27.5kHz$ weighted noise (CCIR 468)		45		dB
Volume control (Figure 8)					
V_{11-6}	Voltage (Pin 11 disconnected)		5.0		V
I_{11}	Circuit (Pin 11 short circuited)		0.9		mA
R_{11-6}	External control resistor		5		kΩ
OSS	Suppression output signal during mute condition		66		dB

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $V_{CC} = V_{7-6} = 12V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS Min	Typ	Max	UNIT
Sync separator and first control loop					
$V_{25-6(P-P)}$	Required sync pulse amplitude; $R_{17-25} = 2k\Omega$[7]	200	800		mV
 I_{25} I_{25}	Input current $V_{25-6} > 5V$ $V_{25-6} = 0V$		 10 TBD		 μA mA
$\pm\Delta f$	Holding range PLL		1100	1500	Hz
$\pm\Delta f$	Catching range PLL	600	1000		Hz
	Control sensitivity[8] video to oscillator; at weak signal at strong signal during scan during vertical retrace and catching		 2.5 3.75 7.5		 kHz/μs kHz/μs kHz/μs
Second control loop (positive edge)					
$\Delta t_D/\Delta t_O$	Control sensitivity R_{28-6} = see Figure 1		50		
t_D	Control range		25		μs
Phase adjustment (via second control loop)					
	Control sensitivity		25		μA/μs
α	Maximum allowed phase shift		±2		μs
Horizontal oscillator (Pin 23)					
f_{FR}	Free-running frequency $R = 34k\Omega$; $C = 2.7nF$		15,625		Hz
Δf	Spread with fixed external components		0.4	4	%
Δf_{FR}	Frequency variation due to change of supply voltage from 9.5 to 13.2V		0	0.5	%
TC	Frequency variation with temperature			1×10^{-4}	$°C^{-1}$
Δf_{FR}	Maximum frequency shift			10	%
Δf_{FR}	Maximum frequency deviation at start H-out		8	10	%
Horizontal output (Pin 26)					
V_{26-6}	Output voltage high level			13.2	V
V_{26-6}	Output voltage at which protection commences			15.8	V
V_{26-6}	Output voltage low at $I_{26} = 10mA$		0.15	0.5	V
d	Duty cycle of horizontal output signal at $t_P = 10\mu s$		0.45		
t_R	Rise time of output pulse		260		ns
t_F	Fall time of output pulse		100		ns

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $V_{CC} = V_{7-6} = 12V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS Min	Typ	Max	UNIT
Flyback input and sandcastle output[9]					
I_{27}	Input current required during flyback pulse	0.1		2	mA
V_{27-6}	Output voltage during burst key pulse	8	9.0		V
V_{27-6}	Output voltage during horizontal blanking	4	4.35	5	V
V_{27-6}	Output voltage during vertical blanking	2.1	2.5	2.9	V
t_W	Width of burst key pulse (60Hz)	3.1	3.5	3.9	μs
t_W	Width of burst key pulse (50Hz)	3.6	4.0	4.4	μs
	Width of horizontal blanking pulse	flyback pulse width			
	Width of vertical blanking pulse 50Hz divider in search window 60Hz divider in search window 50Hz divider in narrow window 60Hz divider in narrow window		 21 17 25 21		 lines lines lines lines
	Delay between start of sync pulse at video output and rising edge of burst key pulse		5.2		μs
Coincidence detector mute output[10]					
V_{22-6}	Voltage for in-sync condition		10.3		V
V_{22-6}	Voltage for no-sync condition no signal		1.5		V
V_{22-6}	Switching level to switch off the AFC		6.4		V
V_{22-6}	Hysteresis AFC switch		0.4		V
V_{22-6}	Switching level to activate mute function (transmitter identification)		2.4		V
V_{22-6}	Hysteresis Mute function		0.5		V
$I_{22(P-P)}$	Charge current in sync condition 4.7μs	0.7	1.0		mA
$I_{22(P-P)}$	Discharge current in sync condition 1.3μs		0.5		mA
Vertical ramp generator[11]					
I_2	Input current during scan		0.5	2	μA
I_2	Discharge current during retrace		0.4		mA
$V_{2-6(P-P)}$	Sawtooth amplitude		0.8	1.1	V
Vertical output (Pin 3)					
I_3	Output current			7	mA
V_{3-6}	Maximum output voltage		5.7		V
Feedback input (Pin 4)					
 V_{4-6} $V_{4-6(P-P)}$	Input voltage DC component AC component (peak-to-peak value)		 3.3 1.2		 V V
I_4	Input current			12	μA
Δt_P	Internal precorrection to sawtooth		5		%
	Deviation amplitude 50/60Hz		0	2	%
Vertical guard[12]					
  ΔV_{4-6} ΔV_{4-6}	Active at a deviation with respect to the DC feedback level; $V_{27-6} = 2.5V$; at switching level low at switching level high		 1.3 1.9		 V V

NOTES:

1. Pin 11 has a double function. When during switch-on a current of 6mA is supplied to this pin, this current is used to start the horizontal oscillator. The main supply can then be obtained from the horizontal deflection stage. When no current is supplied to this pin it can be used as volume control. The indicated maximum value is the current at which all ICs will start. Higher currents are allowed: the excess current is bypassed to ground.
2. Signal with negative-going sync top white 10% of the top sync amplitude (Figure 2).
3. Measured according to the test line given in Figure 3.
   — The differential gain is expressed as a percentage of the difference in peak amplitudes between the largest and smallest value relative to the subcarrier amplitude at blanking level.
   — The differential phase is defined as the difference in degrees between the largest and smallest phase angle.
4. This figure is valid for the complete video signal amplitude (peak white to black).
5. The $S/N = 20 \log \frac{V_{OUT\ BLACK\text{-}TO\text{-}WHITE}}{V_{N(RMS)} \text{ at } B = 5MHz}$.
6. The AFC control voltage is obtained by multiplying the IF-output signal (which is also used to drive the synchronous demodulator) with a reference carrier. This reference carrier is obtained from the demodulator tuned circuit via a 90° phase shift network. The IF-output signal has an asymmetrical frequency spectrum with respect to the carrier frequency. To avoid problems due to this asymmetrical signal, the AFC circuit is gated by means of an internally generated gating pulse. As a result the detector is operative only during black level at a constant carrier amplitude which contains no additional side bands. As a result the AFC output voltage contains no video information.
   At very weak input signals, the driver signal for the AFC circuit will contain a lot of noise. This noise signal has again an asymmetrical frequency spectrum and this will cause an offset of the AFC output voltage. To avoid problems due to this effect, the AFC is switched off when the AGC is controlled to maximum gain.
   The measured figures are obtained at an input sign RMS voltage of 10mV and the AFC output loaded with 2 times 220kΩ between $+V_S$ and ground. The unloaded Q-factor of the reference tuned circuit is 70. The AFC is switched off when no signal is detected by the coincidence detector or when the voltage at Pin 22 is between 1.2V and 6.4V. This can be realized by a resistor of 68kΩ connected between Pin 22 and ground.
7. The slicing level can be varied by changing the value of R_{17-25}. A higher resistor value results in a larger value of the minimum sync pulse amplitude. The slicing level is independent of the video information.
8. Frequency control is obtained by supplying a correction current to the oscillator RC-network via a resistor, connected between the phase 1 detector output and the oscillator network. The oscillator can be adjusted to the right frequency in one of the two following ways:
   a) Interrupt R_{23-24}.
   b) Short circuit the sync separator bias network (Pin 25) to $+V_{CC}$.
   To avoid the need of a VCR switch, the time constant of phase detector at strong input signal is sufficient short to get a stable picture during VCR playback. During the vertical retrace period, the time constant is even shorter so that the head errors of the VCR are compensated at the beginning of the scan. Only at weak signal conditions (information derived from the AGC circuit) is the time constant increased to obtain a good noise immunity.
9. The flyback input and sandcastle output have been combined on one pin.
   The flyback pulse is clamped to a level of 4.5V. The minimum current to drive the second control loop is 0.1mA.
10. The functions in-sync/out-of-sync and transmitter identification have been combined on this pin. The capacitor is charged during the sync pulse and discharged during the time difference between gating and sync pulse.
11. The vertical scan is synchronized by means of a divider system. Therefore no adjustment is required for the ramp generator. The divider detects whether the incoming signal has a vertical frequency of 50 or 60Hz and corrects the vertical amplitude.
12. To avoid screenburn due to a collapse of the vertical deflection, a continuous blanking level is inserted into the sandcastle pulse when the feedback voltage of the vertical deflection is not within the specified limits.
13. Starting point tuner takeover at 1 = 0.2mA. Takeover to be adjusted with a potentiometer of 47kΩ.

FUNCTIONAL DESCRIPTION

IF Amplifier, Demodulator, and AFC

The IF amplifier has a symmetrical input (Pins 8 and 9). The synchronous demodulator and the AFC circuit share an external reference tuned circuit (Pins 20 and 21). An internal RC-network provides the necessary phase-shifting for AFC operation. The AFC circuit is gated by means of an internally generated gating pulse. As a result, the AFC output voltage contains no video information. The AFC circuit provides a control voltage output with a swing greater than 10V from Pin 18.

AGC Circuit

Gating of the AGC detector is performed to reduce sensitivity of the IF amplifier to external electrical noise. The AGC time constant is provided by an RC circuit connected to Pin 19. The point of tuner take-over is preset by the voltage level at Pin 1.

Video Amplifier

The signal through the video amplifier comprises video and sound information.

Sound Circuit and Horizontal Oscillator Starting Function

The input to the sound IF amplifier is obtained by a band-pass filter coupling from the video output (Pin 17). The sound is demodulated and passed via a dual-function volume control stage to the audio output amplifier. The volume control function is obtained by connecting a variable resistor (5kΩ) between Pin 11 and ground, or by supplying Pin 11 with a variable voltage. Sound output is suppressed by an internal mute signal when no TV signal is identified.

DC Volume Control/Horizontal Oscillator Start

The circuit can be used with a DC volume control or with a starting possibility of the horizontal oscillator. The operation depends on the application. When during switch-on no current is supplied to Pin 11, this pin will act as volume control. When a current of 6mA is supplied to Pin 11, the volume control is set to a fixed output signal and the IC will generate drive pulses for the horizontal deflection. The main supply of the IC can then be derived from the horizontal deflection.

Horizontal Synchronization

The video input signal (positive video) is connected to Pin 25.

The horizontal synchronization has two control loops. This has been introduced because a sandcastle pulse had to be generated. An accurate timing of the burstkey pulse can be made in an easy way when the oscillator sawtooth is used. Therefore, the phase of this sawtooth must have a fixed relation with respect to the sync pulse. That can only be realized when a second loop is used.

Horizontal Phase Detector

The circuit has the following operating conditions:

a. Strong input signal, synchronized or not synchronized. (The input signal condition is obtained from the AGC-circuit, the in-sync/out-of-sync from the coincidence detector). In this condition the time constant is optimal for VCR playback; i.e., fast time constant during the vertical retrace (to be able to correct head-errors of the VCR) and such a time constant during scan that fluctuations of the sync are corrected. In this condition the phase detector is not gated.

b. Weak signal. In this condition the time constant is doubled compared with the previous condition. Furthermore, the phase detector is gated when the oscillator is synchronized. This ensures a stable display which is not disturbed by the noise in the video signal.

c. Not synchronized (weak signal). In this condition the time constant during scan and vertical retrace are the same as during scan in condition a.

Vertical Sync Pulse

The vertical sync pulse integrator will not be disturbed when the vertical sync pulses have a width of only 10μs with a separation of 22μs. This type of vertical sync pulses are generated by certain video tapes with anti-copy guard.

Vertical Ramp Generator

To avoid problems during VCR-playback in the so-called feature modes (fast or slow), the vertical ramp generator is not coupled to the horizontal oscillator when such signals are received. For normal signals the coupling between vertical ramp generator and horizontal oscillator is maintained. This ensures a reliable interface.

Vertical Divider System

The IC embodies a synchronized divider system for generating the vertical sawtooth at Pin 2. The divider system has an internal frequency doubling circuit, so the horizontal oscillator is working at its normal line frequency; one line period equals 2 clock pulses.

Due to the divider system no vertical frequency adjustment is needed. The divider has a discriminator window for automatically switching over from the 60Hz to 50Hz system. When the trigger pulse comes before line 576 the system works in the 60Hz mode, otherwise 50Hz mode is chosen. The divider system operates with 2 different divider reset windows for maximum interference/disturbance protection.

The windows are activated via an up/down counter.

The counter increases its counter value with 1 for each time the separated vertical sync. pulse is within the search window. When it is not, the counter value is lowered with 1.

The different working modes of the divider system are specified below.

a. Large (search) window: divider ratio between 488 and 722.

This mode is valid for the following conditions:

1. Divider is locking for a new transmitter.
2. Divider ratio found, not within the narrow window limits.
3. Non-standard TV signal condition detected while a double or enlarged vertical sync pulse is still found after the internally-generated anti-topflutter pulse has ended. This means a vertical sync pulse width larger than 10 clock pulses (50Hz) viz. 12 clock pulses (60Hz).

   In general this mode is activated for video tape recorders operating in the feature trick mode. When the wide vertical sync. pulses are detected, the vertical ramp generator is decoupled from the horizontal oscillator. As a consequence, the retrace time of this ramp generator is now determined by the external capacitor and the discharge current. This decoupling prevents instability of the picture due to irregular incoming signals (variable number of lines per field).
4. Up/down counter value of the divider system operating in the narrow window mode drops below count 6.

b. Narrow window: divider ratio between 522 – 528 (60Hz) or 622 – 628 (50Hz).

The divider system switches over to this mode when the up/down counter has reached its maximum value of 15 approved vertical sync pulses. When the divider operates in this mode and a vertical sync pulse is missing within the window, the divider is reset at the end of the window and the counter value is lowered with 1. At a counter value below 6, the divider system switches over the large window mode. The divider system also generates the so-called anti-topflutter pulse which inhibits the phase 1 detector during the vertical sync pulse. The width of this pulse depends on the divider mode. For the divider mode a the start is generated at the reset of the divider. In mode b the anti-topflutter pulse starts at the beginning of the first equalizing pulse.

The anti-topflutter pulse ends at count 10 for 50Hz and count 12 for 60Hz. The vertical blanking pulse is also generated via the divider system. The start is at the reset of the divider while the blanking pulse width is 34 (17 lines) for 60Hz and at count 42 (21 lines) for 50Hz systems.

The vertical blanking pulse generated at the sandcastle output Pin 27 is made by adding the anti-topflutter pulse and the blanking pulse. In this way the vertical blanking pulse starts at the beginning of the first equalizing pulse when the divider operates in the b mode. The total length of the vertical blanking in this condition is 21 lines in the 60Hz mode and 25 lines in the 50Hz mode.

Application When External Video Signals Have to Be Synchronized

The input of the sync separator is externally available. For the normal application, the video output signal (Pin 17) is AC-coupled to this input (see Figure 2). It is possible to interrupt this connection and to drive the sync separator from another source; e.g., a teletext decoder in serial mode or a signal coming from the PT-plug. When a teletext decoder is applied, the IF-amplifier and synchronization circuit are running in the same phase so that the various connections between the two parts (like AGC gating) can remain active. When external signals are applied to the sync separator, the connections between the two parts must be interrupted. This can be obtained by connecting Pin 22 to ground.

This results in the following condition:

- AGC detector is not gated.
- AFC circuit is active.
- Mute circuit not active so that the sound channel remains switched-on.
- The first phase detector has an optimal time constant for external video sources.

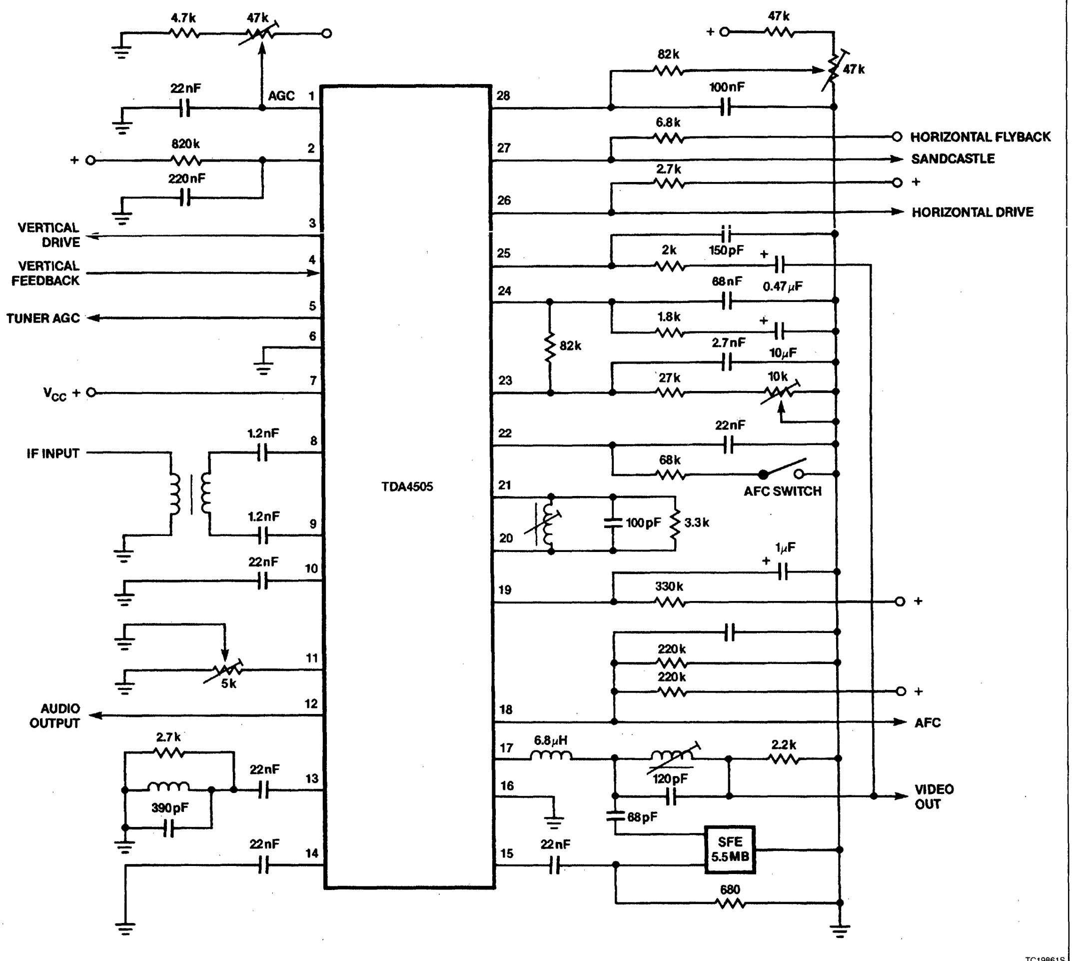


Figure 1. Application Diagram

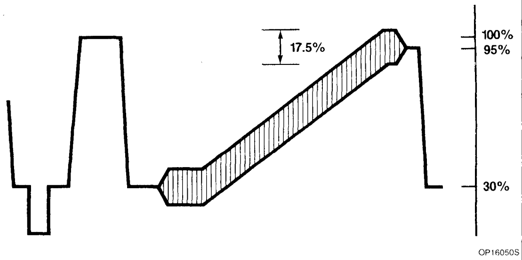


Figure 2. Video Output Signal

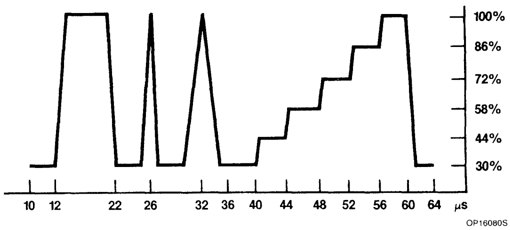
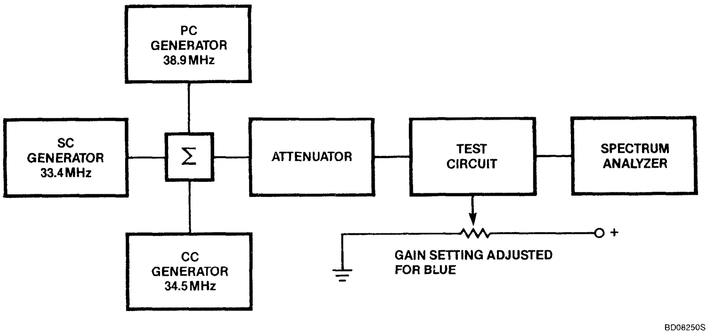


Figure 3. E.B.U. Test Signal Waveform (Line 330)



NOTES:
Value at 1.1MHz; $20\log\frac{V_O \text{ at } 4.4\text{MHz}}{V_O \text{ at } 1.1\text{MHz}} + 3.6\text{dB}$.
Value at 3.3MHz; $20\log\frac{V_O \text{ at } 4.4\text{ MHz}}{V_O \text{ at } 3.3\text{MHz}}$.

Figure 4. Test Setup Intermodulation

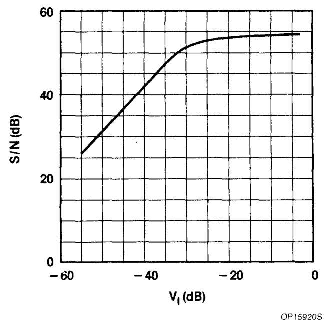


Figure 5. S/N Ratio as a Function of the Input Voltage

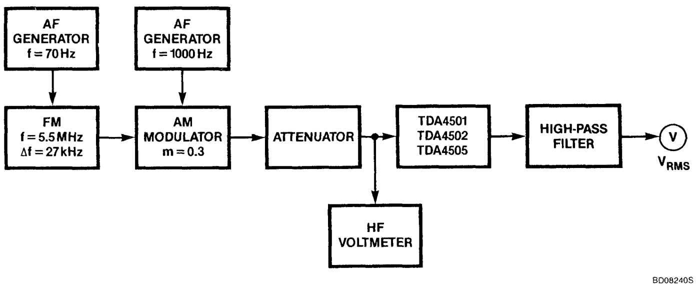


Figure 6. Test Setup AM Suppression

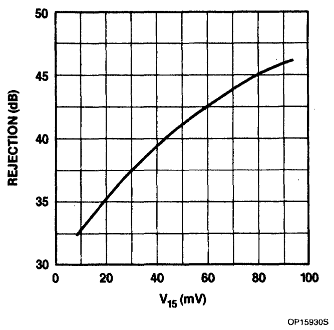


Figure 7. AM Rejection

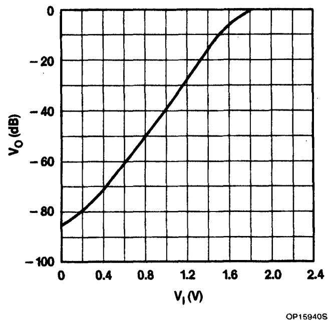


Figure 8. Volume Control Characteristics

Signetics

Section 7
Video/IF

Linear Products

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Signetics

TDA8340, TDA8341
Television IF Amplifier and Demodulator

Linear Products

DESCRIPTION

The TDA8340 and TDA8341 are integrated IF amplifier and demodulator circuits for color or black and white television receivers. The TDA8340 is for application with NPN tuners and the TDA8341 for PNP tuners.

FEATURES

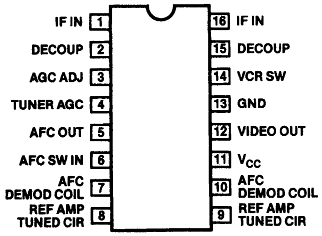
- **Full range gain-controlled wideband IF amplifier**
- **Linear synchronous demodulator with excellent intermodulation performance**
- **White spot inverter**
- **Wideband video amplifier with noise protection**
- **AFC circuit with AFC on/off switching and sample and hold function**
- **Low impedance AFC output**
- **AGC circuit with noise gating**
- **Tuner AGC output for NPN tuners (TDA8340) or PNP tuners (TDA8341)**
- **External video switch for switching off the video output**
- **Reduced sensitivity for high sound carriers**
- **Integrated filter to limit second harmonic IF signals**
- **Wide supply voltage range**
- **Requires few external components**

APPLICATIONS

- **Black/white and color TV receivers**
- **Video casette recorders (VCR's)**
- **CATV converters**

PIN CONFIGURATION

N Package



TOP VIEW

CD12100S

PIN NO.	DESCRIPTION
1	Balanced IF inputs
2	IF Amplifier Decoupling
3	Tuner AGC Starting Point Adjustment
4	Tuner AGC Output
5	AFC Output
6	AFC On/Off Switch and Sample and Hold Capacitor
7	Reference Carrier $\pi/2$ rad. Phase Shift
8	IF Picture Carrier Passive Regeneration
9	IF Picture Carrier Passive Regeneration
10	Reference Carrier $\pi/2$ rad. Phase Shift
11	Positive Supply Voltage (V_{CC})
12	Video Output
13	Ground (V_{EE})
14	IF AGC Capacitor and VCR Switch
15	IF Amplifier Decoupling
16	Balanced IF Inputs

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
16-Pin Plastic DIP (SOT-38)	-25 to +60°C	TDA8340N
16-Pin Plastic DIP (SOT-38)	-25 to +60°C	TDA8341N

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{CC}	Supply voltage	13.2	V
$V_{4\text{-}13}$	Tuner AGC voltage	12	V
P_{TOT}	Total power dissipation	1.2	mW
T_{STG}	Storage temperature range	-65 to +150	°C
T_A	Operating ambient temperature range	-25 to +70	°C

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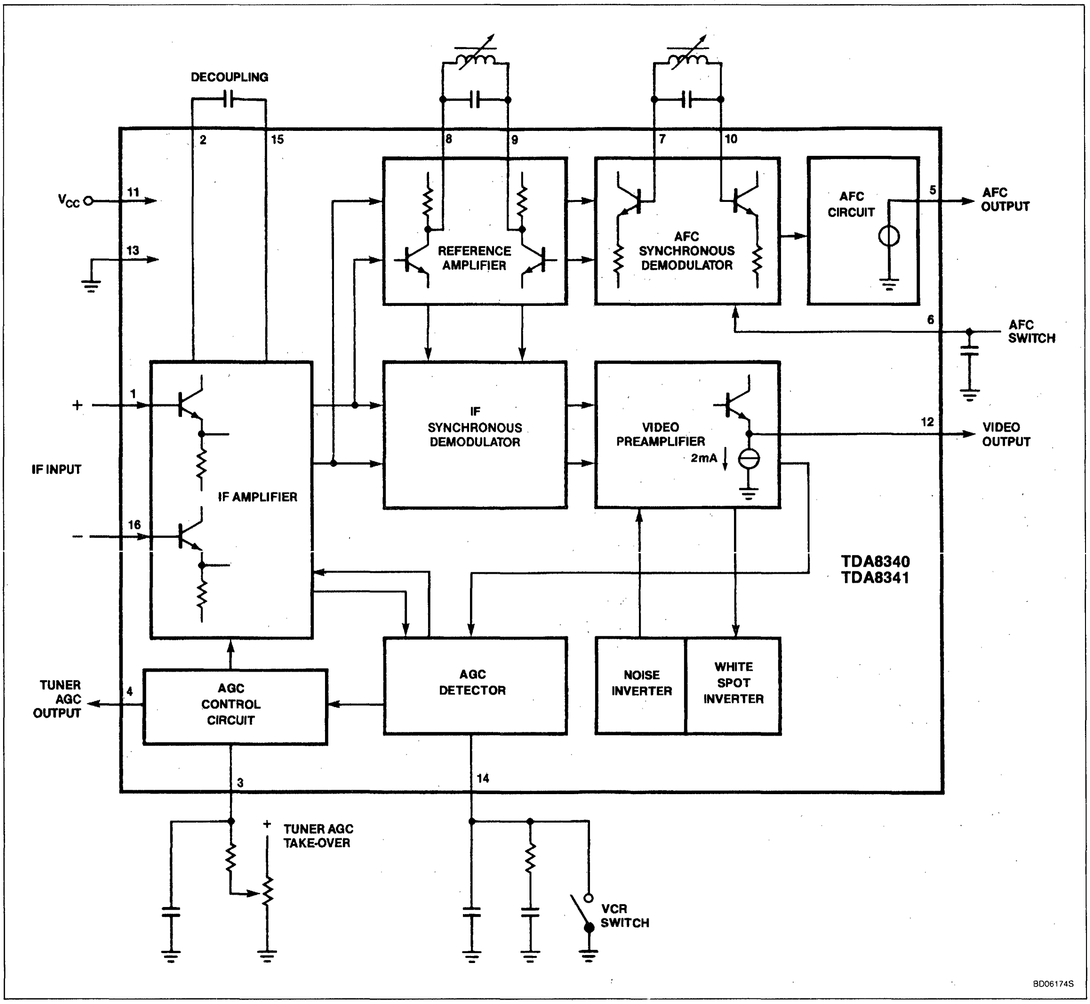
Television IF Amplifier and Demodulator — TDA8340, TDA8341

BLOCK DIAGRAM



Television IF Amplifier and Demodulator TDA8340, TDA8341

DC ELECTRICAL CHARACTERISTICS Measured in circuit of Figure 2; V_{CC} = 12V; T_A = 25°C, unless otherwise specified.

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNIT
$V_{CC} = V_{11-13}$	Supply voltage (Pin 11)		9.4	12	13.2	V
I_{11}	Supply current	No input signal	30	42	55	mA
V_{1-16}	**IF Amplifier**[1] Input sensitivity	At onset of AGC	20	40	80	μV
R_{1-16}	Differential input resistance			2		kΩ
C_{1-16}	Differential input capacitance			3		pF
G_V	Gain control range			67		dB
V_{12-13}	Input signal variation[2]				0.5	dB
V_{1-16}	Maximum input signal		100			mV
V_{1-16}	**Tuner AGC**[1] Tuner AGC starting point[3]	R_{3-11} = 39kΩ			3.0	mV
V_{1-16}	Tuner AGC starting point[3]	R_{3-13} = 39kΩ	70			mV
I_4	Maximum current swing of Tuner AGC output		10			mA
V_{1-16}	Input signal variation[4]	I_4 = 1 to 9mA			3.0	dB
V_{4-13}	Output saturation voltage	I_4 = 7mA		200	300	mV
I_4	Leakage current	V_4 = 12V			1	μA
V_{12-13}	**Video Output**[4] Zero-signal output level[5]		5.7	6.0	6.3	V
V_{12-13}	Top sync output level		2.8	3.0	3.2	V
$V_{12-13(P-P)}$	Video output voltage (Peak-to-peak value)	White signal; 10% top sync	2.4	2.7	3.0	V
	Internal bias current of emitter follower output transistor		1.4	2.2	3.0	mA
Z_{12}	Output impedance			100		Ω
B	Bandwidth of demodulated output signal		7.5	10.0		MHz
G_d	Differential gain[6]			2.0	5.0	%
d	Differential phase[6]			2.0	5.0	deg
	Luminance non-linearity[7]			2.0	5.0	%
$V_{12-13(RMS)}$	Residual carrier signal[8] (RMS value)			2.0	10	mV
$V_{12-13(RMS)}$	Residual 2nd harmonic of carrier signal (RMS value)[8]			2.0	10	mV
$\frac{\Delta V_{12-13(P-P)}}{\Delta V_{11-13}}$	Variation of video voltage for ΔV_{CC} = 1V		0.1	0.2	0.3	mV
α	Intermodulation[8, 9]	1.1MHz, blue		-65	-60	dB
α		1.1MHz, yellow		-60	-56	dB
α		3.3MHz			-68	dB
S/(S + N)	Signal-to-noise ratio[10]	V_I = 10mV Maximum gain	50 54	58 61		dB dB
V_{12-13}	**Spot Inverter**[11] Threshold level		6.3	6.8	7.3	V
V_{12-13}	Insertion level		4.2	4.5	4.8	V

DC ELECTRICAL CHARACTERISTICS (Continued) Measured in circuit of Figure 2; V_{CC} = 12V; T_A = 25°C, unless otherwise specified.

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNIT
V_{12-13}	**Noise Inverter**[11] Threshold level		1.6	1.8	2.0	V
V_{12-13}	Insertion level		3.5	3.8	4.1	V
V_{14-13}	**VCR Switch** Level below which video output switches off		1.8	2.2	2.6	V
$-I_{14}$	Switch current	V_{12-13} = 0.7V	40	60	100	μA
	AFC Circuit[12] Output voltage swing					
$V_{5-13(P-P)}$	(Peak-to-peak value)			10		V
Δf	Change of frequency for an AFC output voltage swing of 10V			60	120	kHz
V_{5-13}	AFC output voltage	At f = 38.9MHz		6		V
V_{5-13}		No input signal	4	6	8	V
V_{5-13}		During AFC off	5	6	7	V
R_{5-13}	AFC output resistance			500		Ω
V_{6-13}	AFC switch: Level below which AFC output switches off		1.4	2.0	2.8	V
I_6	AFC switch current	During AFC on		200	500	μA
I_6	Max. AFC switch current	During AFC off; V_{6-13} = 0V			5	mA

NOTES:
1. All input signals are measured RMS at top sync and 38.9MHz.
2. Measured with 0dB = 200μV.
3. Tuner AGC starting point is defined as "level of input signal when tuner AGC current = 1mA".
4. Measured with Pin 3 connected via 39kΩ resistor to V_{CC} (Pin 11), with an RMS voltage of 10mV top sync input signal and with Pin 12 not loaded.
5. At the "projected zero point", i.e., with switched demodulator.
6. Measured in the circuit of Figure 6: The differential gain is expressed as a percentage of the difference in peak amplitudes between the largest and smallest value relative to the subcarrier amplitude at blanking level; The differential phase is defined as "the difference (in degrees) between the largest and smallest phase angles".
7. Measured according to the test line shown in Figure 8: The non-linearity is expressed as a percentage of the maximum deviation of a luminance step from the mean step, with respect to the mean step; The mean step is (white level-black level) divided by the number of steps.
8. Measured up to 45dB gain control.
9. Test setup and input conditions for intermodulation measurements as in Figures 5 and 6.
10. Measured with a 75Ω source:

$$S/(S+N) = 20 \log \frac{V_{out\ black\ to\ white}}{V_{n(RMS)}\ at\ B = 5\ MHz}.$$

11. Video output waveform showing white spot and noise inverter threshold levels.
12. Measured with input signal V_{1-16} = 10mV and with no load at AFC output.

Television IF Amplifier and Demodulator TDA8340, TDA8341

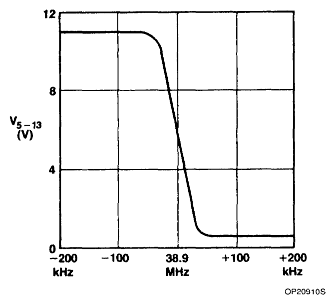


Figure 1. AFC Output Voltage as a Function of Frequency

IF Amplifier

This is a 3-stage, gain-controlled IF amplifier with a wide dynamic range. On-chip capacitors in the DC feedback loop of the amplifier maintain stability at maximum gain. Internal stablization of the supply voltage ensures the desired sensitivity and gain control range over the whole supply voltage range and also gives very good power supply ripple rejection in this part of the circuit.

Demodulator

The redesigned IF demodulator is a quasi-synchronous circuit that employs passive carrier regeneration and logarithmic clamping to give improved signal handling. The demodulator input is AC-coupled to the IF amplifier to reduce DC offsets and to thus minimize residual IF carrier in the output signal.

Video Amplifier

The linearity and bandwidth of the video amplifier are sufficient to meet all wideband requirements, i.e., for teletext transmissions. Second harmonics of the IF carrier are effectively reduced by a Sallen-Key low pass interstage filter between the demodulator output and the video amplifier input. An integrated filter in the noise inverter reduces the sensitivity of the video amplifier for high sound carriers.

White spot protection comprises a white spot clamp system combined with a delayed-action inverter which is also highly resistant to high sound carriers.

To prevent radiated video output at the input pins, connect at 6.8μH inductor in series with Pin 12 and place as closely as possible to the IC body. Use short leads.

AGE Detector

A Bessel low-pass filter between the video output and the AGC detector improves the detector function in the presence of high sound carriers. No hang-up occurs in the detector after Pin 14 has been short-circuited to ground (VCR switch operated). The detector also generates the sample and hold pulse for the AFC system.

AGC Control Circuit

This converts the AGC detector voltage (Pin 14) into a current signal which controls the gain of the IF amplifier. It also provides a tuner AGC control output from Pin 4; current limiting is incorporated to prevent internal damage. The AGC starting point is adjusted via Pin 3.

AFC Circuit

The AFC circuit provides a voltage output which controls the IF frequency of the tuner. Video information on the AFC output (Pin 5) is eliminated by a sample and hold circuit (external capacitor at Pin 6). Coupling between the AFC and reference tuned circuits is via two small capacitos (or parasitic capacitance) between the respective tracks of the printed circuit board. If the capacitance is less than 1pF, the steepness of the AFC charasteristic is reduced.

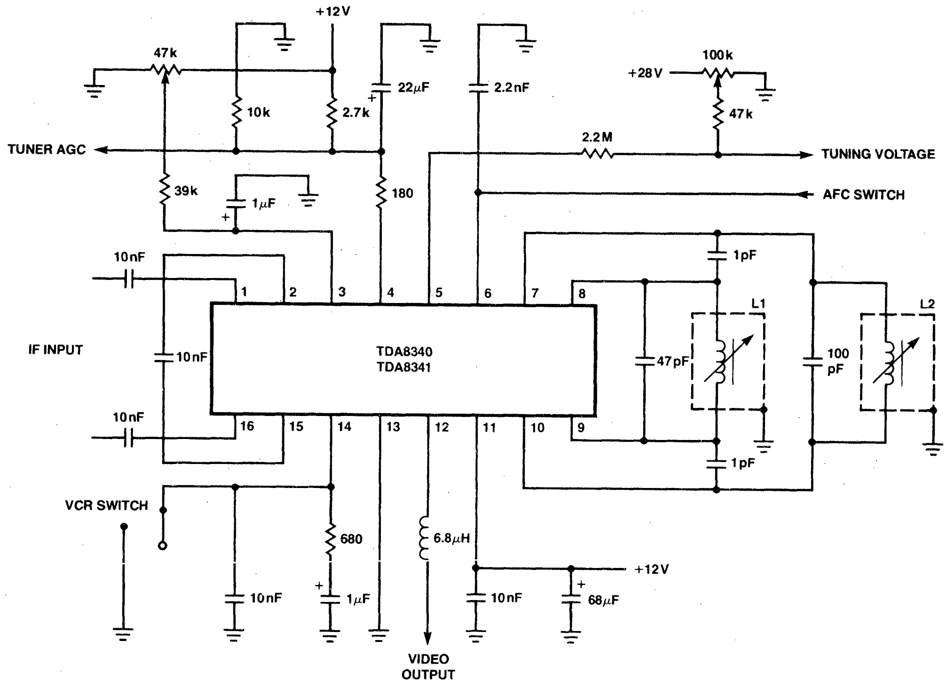


Figure 2. Typical application circuit diagram; Q of L1 and L2 = 80; f_o = 38.9 MHz

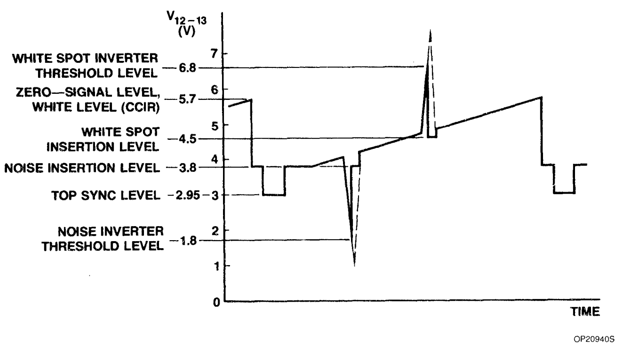


Figure 3. Video Output Waveform Showing White Spot and Noise Inverter Threshold Levels

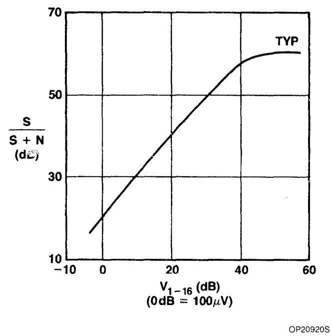
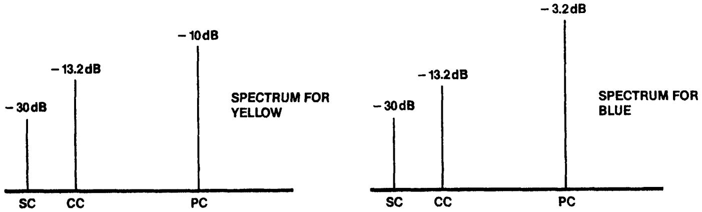


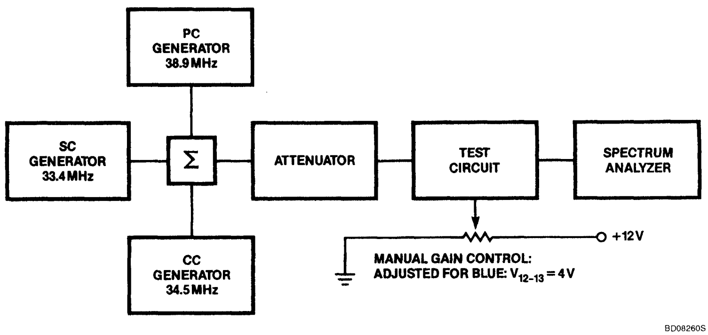
Figure 4. Signal-To-Noise Ratio as a Function of Input Voltage

Intermodulation at 1,1 MHz: blue*	> 46 dB
	typ. 60 dB
yellow*	> 46 dB
	typ. 50 dB
at 3,3MHz**	> 46 dB
	typ. 54 dB



S.C.: sound carrier level
C.C.: chrominance carrier level
P.C.: picture carrier level
} with respect to top sync level

Figure 5. Input Conditions for Intermodulation Measurements; Standard Colour Bar with 75% Contrast



$$*20 \log \frac{V_O \text{ at } 4{,}4 \text{ MHz}}{V_O \text{ at } 1{,}1 \text{ MHz}} + 3{,}6\text{dB.} \qquad **20 \log \frac{V_O \text{ at } 4{,}4 \text{ MHz}}{V_O \text{ at } 3{,}3 \text{ MHz}}.$$

Figure 6. Test Setup for Intermodulation

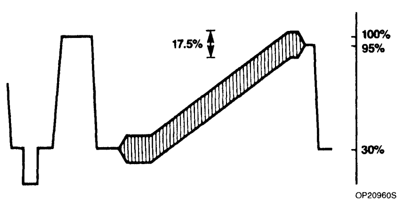


Figure 7. Video Output Signal

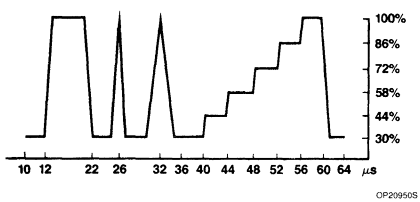


Figure 8. E.B.U. Test Signal Waveform (Line 330)

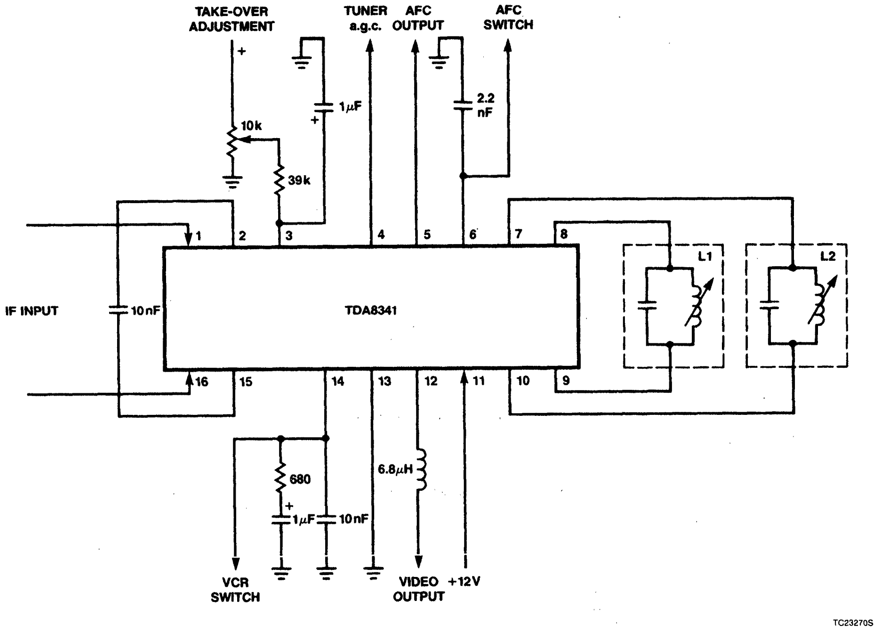


Figure 9. Application Diagram

Signetics

Section 8
Sound IF and Special Audio Decoding

Linear Products

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8

Signetics

TDA2545A
Quasi-Split-Sound Circuit

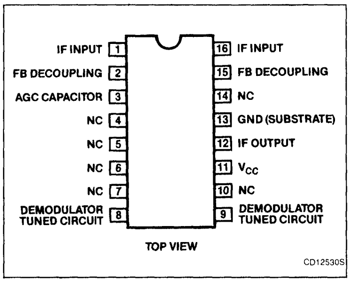
Product Specification

Linear Products

DESCRIPTION
The TDA2545A is a monolithic integrated circuit for quasi-split-sound processing in television receivers.

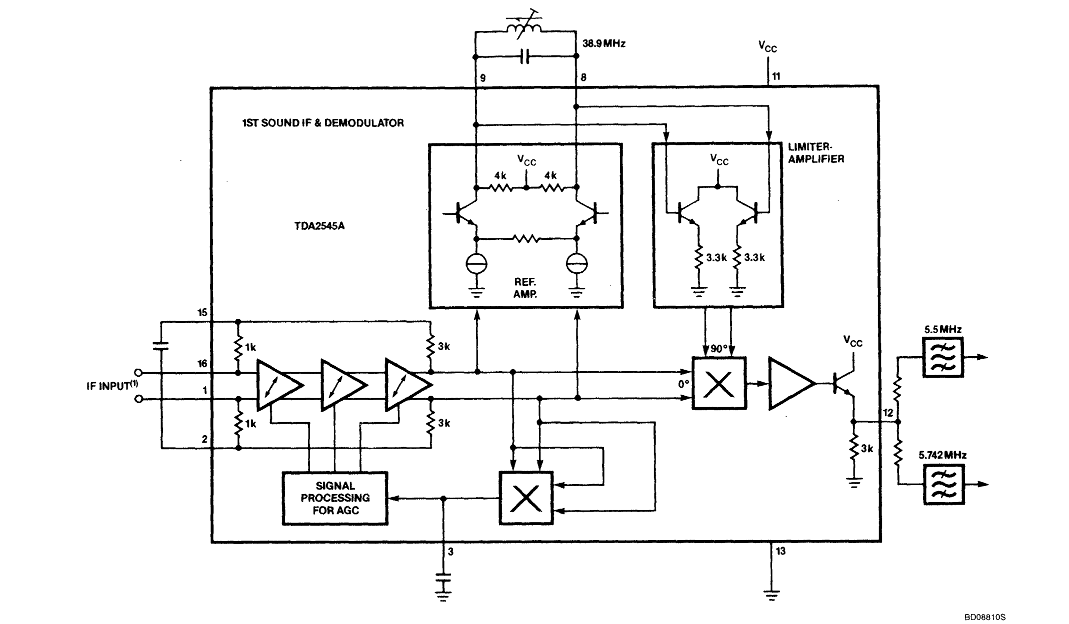
FEATURES
- **3-stage gain-controlled IF amplifier**
- **AGC circuit**
- **Reference amplifier and limiter amplifier for vision carrier processing**
- **Linear multiplier for quadrature demodulation**

APPLICATIONS
- **Stereo MTS television receiver**
- **Video cassette recorder with MTS**
- **CATV converters**

PIN CONFIGURATION


ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
16-Pin Plastic DIP (SOT-38)	0 to +70°C	TDA2545AN

BLOCK DIAGRAM


NOTE:
1. IF signal: vision carrier (VC) and sound carrier (SC).

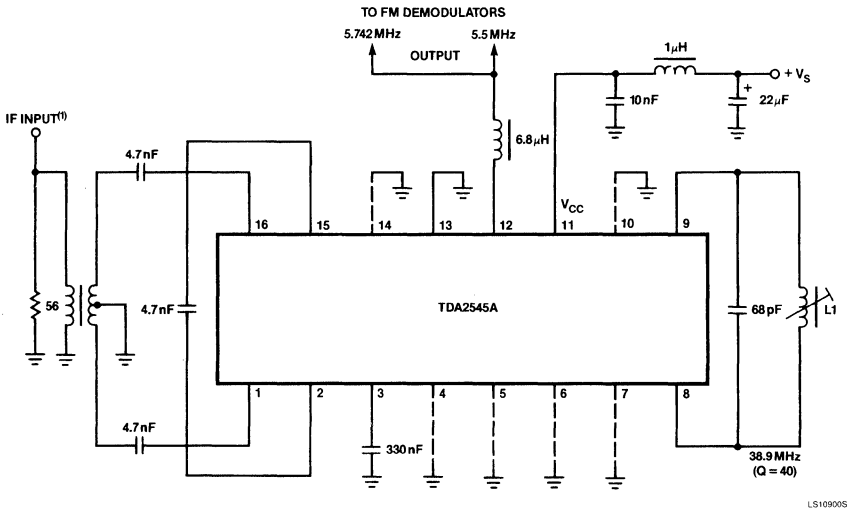
ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{CC}	Supply voltage (Pin 11)	13.2	V
T_{STG}	Storage temperature range	−65 to +150	°C
T_A	Operating ambient temperature range	0 to +70	°C

DC ELECTRICAL CHARACTERISTICS

V_{CC} = 12V; T_A = 25°C; measured at f_{VC} = 38.9MHz, f_{SC1} = 33.4MHz, f_{SC2} = 33.158MHz:
Vision carrier (VC) modulated with 2T/20T pulses, line-for-line alternating with white bars; modulation depth 100% (proportional to 10% residual carrier).
Sound carriers (SC1, SC2) modulated with f = 1kHz and Δf = ±30kHz.
Vision-to-sound carrier ratios are VCSC1 = 13dB and VCSC2 = 20dB.
Vision carrier amplitude (RMS value) is V_{VC} = 10mV.
For measuring circuit see Figure 1, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supply (Pin 11)					
V_{CC}	Supply voltage	10.8	12	13.2	V
$I_{CC} = I_{11}$	Supply current		42		mA
IF amplifier					
$V_{VC1-16(RMS)}$	Minimum input voltage (RMS value) (intercarrier signals −3dB)		50		μV
$V_{VC1-16(RMS)}$	Maximum input voltage (RMS value) (intercarrier signals +1dB		100		mV
ΔG_V	IF control range	66			dB
V_{3-13}	Control voltage range	4		9	V
R_{1-16}	Input resistance		2		kΩ
C_{1-16}	Input capacitance		2		pF
Intercarrier generation					
$V_{12-13(RMS)}$	Output voltage; 5.5MHz (RMS value)		100		mV
$V_{12-13(RMS)}$	Output voltage; 5.742MHz (RMS value)		45		mV
V_{12-13}	DC output voltage		5.9		V
R_{12-13}	Allowable load resistance at the output	7			kΩ
$-I_{12}$	Allowable output current			1	mA
Intercarrier signal-to-noise (measured behind the FM demodulators)					
	Signal-to-weighted-noise ratio according to CCIR 468-2, quasi-peak				
S + W/W	at 5.5MHz	53			dB
S + W/W	at 5.742MHz	51			dB
	with black level (vision carrier modulated with sync pulses only)				
S + W/W	at 5.5MHz	60			dB
S + W/W	at 5.742MHz	58			dB



NOTES:
Pins 4, 5, 6, 7, 10 and 14 not connected.
1. IF signal: vision carrier (VC) and sound carrier (SC)

Figure 1. Measuring Circuit for TDA2545A

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Signetics

TDA2546A
Quasi-Split-Sound IF With Sound Demodulator

Product Specification

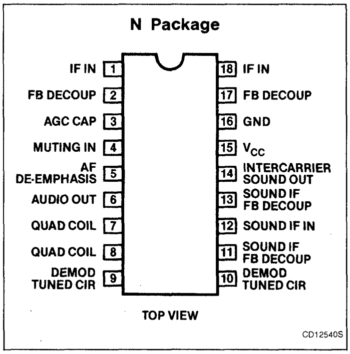
Linear Products

DESCRIPTION
The TDA2546A is a monolithic integrated circuit for quasi-split-sound processing, including 5.5MHz demodulation, in television receivers.

FEATURES
First IF (VC: vision carrier plus SC: sound carrier)
- **3-stage, gain-controlled IF amplifier**
- **AGC circuit**
- **Reference amplifier and limiter amplifier for vision carrier (VC) processing**
- **Linear multiplier for quadrature demodulation**

Second IF (5.5MHz signal)
- **8-stage limiter amplifier**
- **Quadrature demodulator**
- **AF amplifier with de-emphasis**
- **AV switch**

APPLICATIONS
- **Television stereo MTS receiver**
- **Video cassette recorder with MTS stereo**

PIN CONFIGURATION


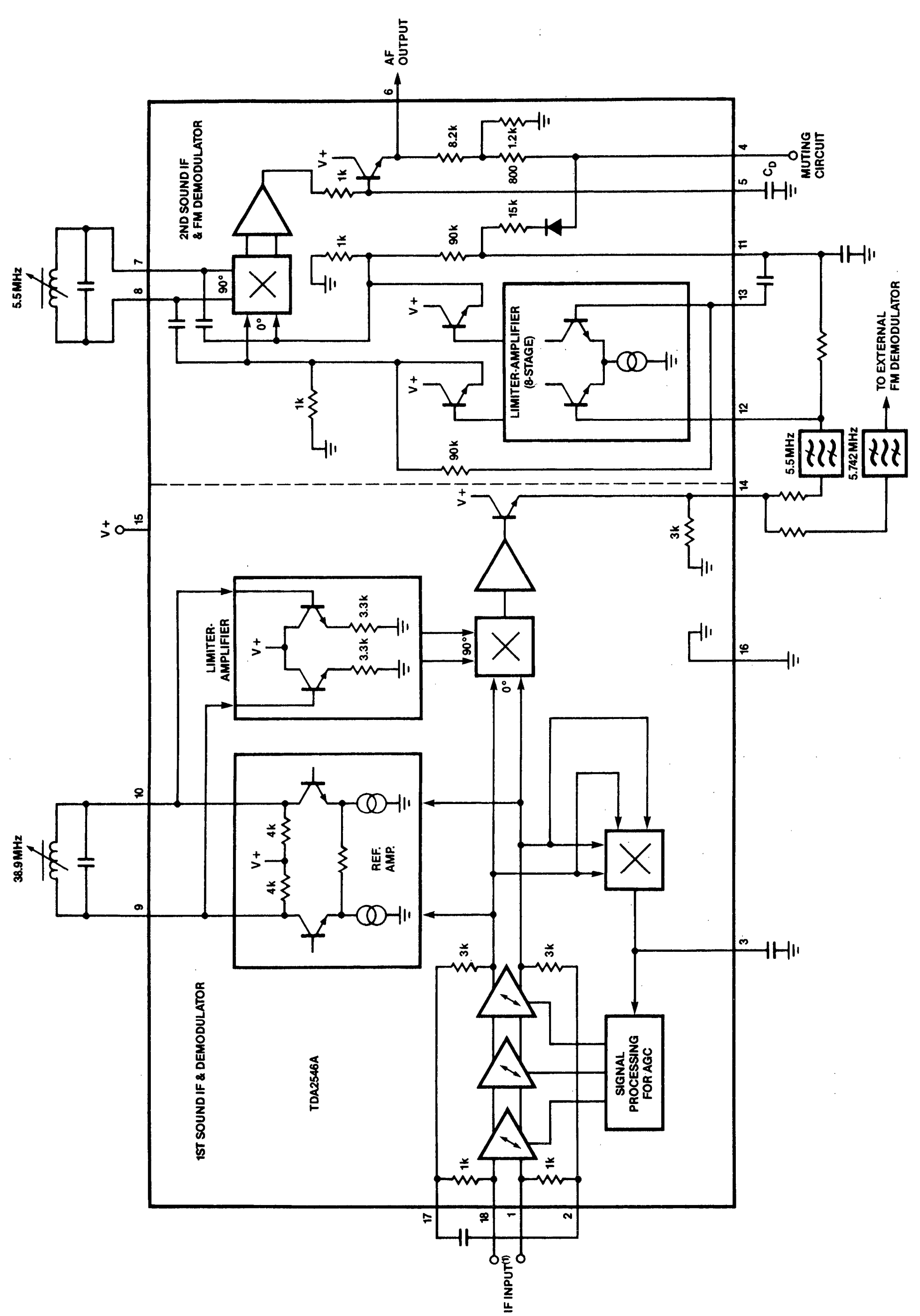
ORDERING INFORMATION
DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
18-Pin Plastic DIP (SOT-102CS)	0 to +70°C	TDA2546AN

ABSOLUTE MAXIMUM RATINGS
SYMBOL	PARAMETER	RATING	UNIT
V_{CC}	Supply voltage (Pin 15)	13.2	V
I_{IN}	Input current (Pin 4)	5	mA
T_{STG}	Storage temperature range	-25 to +150	°C
T_A	Operating ambient temperature range	0 to +70	°C

Quasi-Split-Sound IF With Sound Demodulator

TDA2546A

BLOCK DIAGRAM



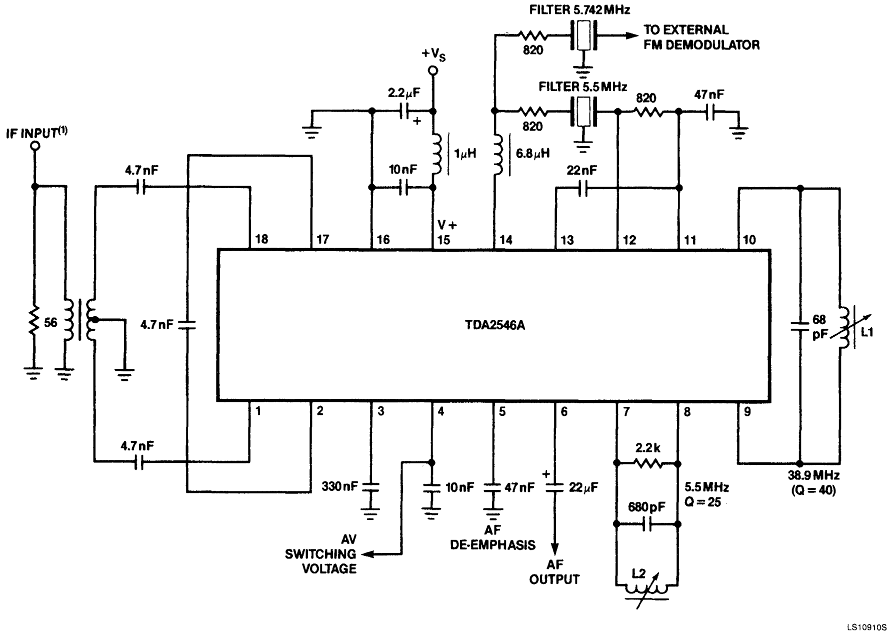
NOTE:
1. IF signal: Vision Carrier (VC) and Sound Carrier (SC).

Quasi-Split-Sound IF With Sound Demodulator — TDA2546A

DC ELECTRICAL CHARACTERISTICS

$V_{CC} = V_{15-16} = 12V$; $T_A = 25°C$; measured at $f_{VC} = 38.9MHz$, $f_{SC1} = 33.4MHz$, $f_{SC2} = 33.158MHz$:
Vision carrier (VC) modulated with 2T/20T pulses, line-for-line alternating with white bars; modulation depth 100% (proportional to 10% residual carrier).
Sound carriers (SC1, SC2) modulated with f = 1kHz and $\Delta f = \pm 30kHz$.
Vision-to-sound carrier ratios are VC/SC1 = 13dB and VC/SC2 = 20dB.
Vision carrier amplitude (RMS value) is $V_{VC} = 10mV$.
For measuring circuit see Figure 1, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supply (Pin 15)					
$V_{CC} = V_{15-16}$	Supply voltage	10.8	12	13.2	V
$I_{CC} = I_{15}$	Supply current		54		mA
IF amplifier					
$V_{VC1-18(RMS)}$	Minimum input voltage (RMS value) (intercarrier signals −3dB)		50		μV
$V_{VC1-18(RMS)}$	Maximum input voltage (RMS value) (intercarrier signals +1dB)		100		mV
ΔG_V	IF control range	66			dB
V_{3-16}	Control voltage range	4		9	V
R_{1-18}	Input resistance		2		kΩ
C_{1-18}	Input capacitance		2		pF
Intercarrier generation					
$V_{14-16(RMS)}$	Output voltage; 5.5MHz (RMS value)		100		mV
$V_{14-16(RMS)}$	Output voltage; 5.742MHz (RMS value)		45		mV
V_{14-16}	DC output voltage		5.9		V
R_{14-16}	Allowable load resistance at the output	7			kΩ
$-I_{14}$	Allowable output current			1	mA
Frequency demodulator (measured at f = 5.5MHz)					
$V_{12-16(RMS)}$	Input voltage for start of limiting (RMS value)			100	μV
$V_{12-16(RMS)}$	Maximum input voltage (RMS value)		200		mV
$V_{11,\ 12,\ 13-16}$	DC output voltage		2.2		V
$V_{6-16(RMS)}$	AF output voltage (RMS value)		600		mV
V_{6-16}	DC output voltage		4		V
R_{6-16}	Allowable load resistance at the output	27			kΩ
THD	Total harmonic distortion			1	%
R_{15-16}	Internal de-emphasis resistance		1		kΩ
	Switching voltage (Pin 4)				
V_{4-16}	for mute	9			V
V_{4-16}	for AF on			2.5	V
Intercarrier signal-to-noise (measured behind the FM demodulators)					
	Signal-to-weighted-noise ratio according to CCIR 468-2, quasi-peak				
S + W/W	at 5.5MHz	53			dB
S + W/W	at 5.742MHz	51			dB
	with black level (vision carrier modulated with sync pulses only)				
S + W/W	at 5.5MHz	60			dB
S + W/W	at 5.742MHz	58			dB



NOTE:
1. IF signal: Vision Carrier (VC) and Sound Carrier (SC).

Figure 1. Measuring Circuit for TDA2546A

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Signetics

Section 9 SYNC Processing and Generation

Linear Products

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9

Signetics

TDA2577A
Sync Circuit With Vertical Oscillator and Driver

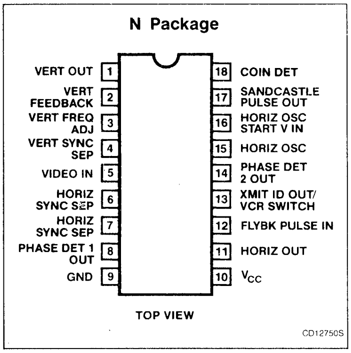
Product Specification

Linear Products

DESCRIPTION
The TDA2577A separates the vertical and horizontal sync pulses from the composite TV video signal and uses them to synchronize horizontal and vertical oscillators.

FEATURES
- **Horizontal sync separator and noise inverter**
- **Horizontal oscillator**
- **Horizontal output stage**
- **Horizontal phase detector (sync to oscillator)**
- **Time constant switch for phase detector (fast time constant during catching)**
- **Slow time constant for noise-only conditions**
- **Time constant externally switchable (e.g., fast for VCR)**
- **Inhibit of horizontal phase detector and video transmitter identification circuit during vertical oscillator flyback**
- **Second phase detector (φ2) for storage compensation of horizontal deflection stage**
- **Sandcastle pulse generator (3 levels)**
- **Video transmitter identification circuit**
- **Stabilizer and supply circuit for starting the horizontal oscillator and output stage directly from the supply voltage**
- **Duty factor of horizontal output pulse is 50% when flyback pulse is absent**
- **Vertical sync separator**
- **Bandgap 6.5V reference voltage for vertical oscillator and comparator**
- **Synchronized vertical oscillator/sawtooth generator (synchronization inhibited when no video transmitter is detected)**
- **Internal circuit for 3% parabolic precorrection of the oscillator/sawtooth generator. Comparator supplied with precorrected sawtooth and external feedback input**
- **Vertical comparator with internal 3% precorrection circuit for vertical oscillator/sawtooth generator**
- **Vertical driver stage**
- **Vertical blanking pulse generator with external adjustment of pulse duration (50Hz: 21 lines; 60Hz: 17 lines)**
- **Vertical guard circuit**

APPLICATIONS
- **Video monitors**
- **TV receivers**
- **Video processing**

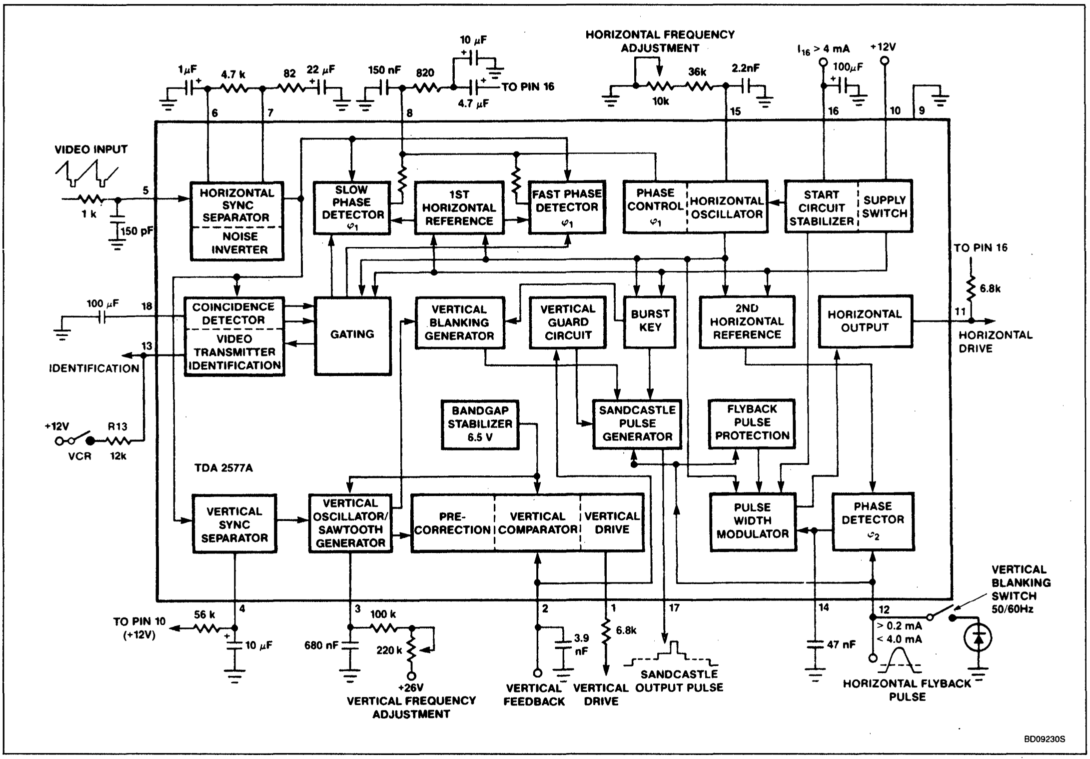
PIN CONFIGURATION


ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
18-Pin Plastic DIP (SOT-102HE)	−25°C to +65°C	TDA2577AN

Sync Circuit With Vertical Oscillator and Driver
TDA2577A

BLOCK DIAGRAM



Sync Circuit With Vertical Oscillator and Driver

TDA2577A

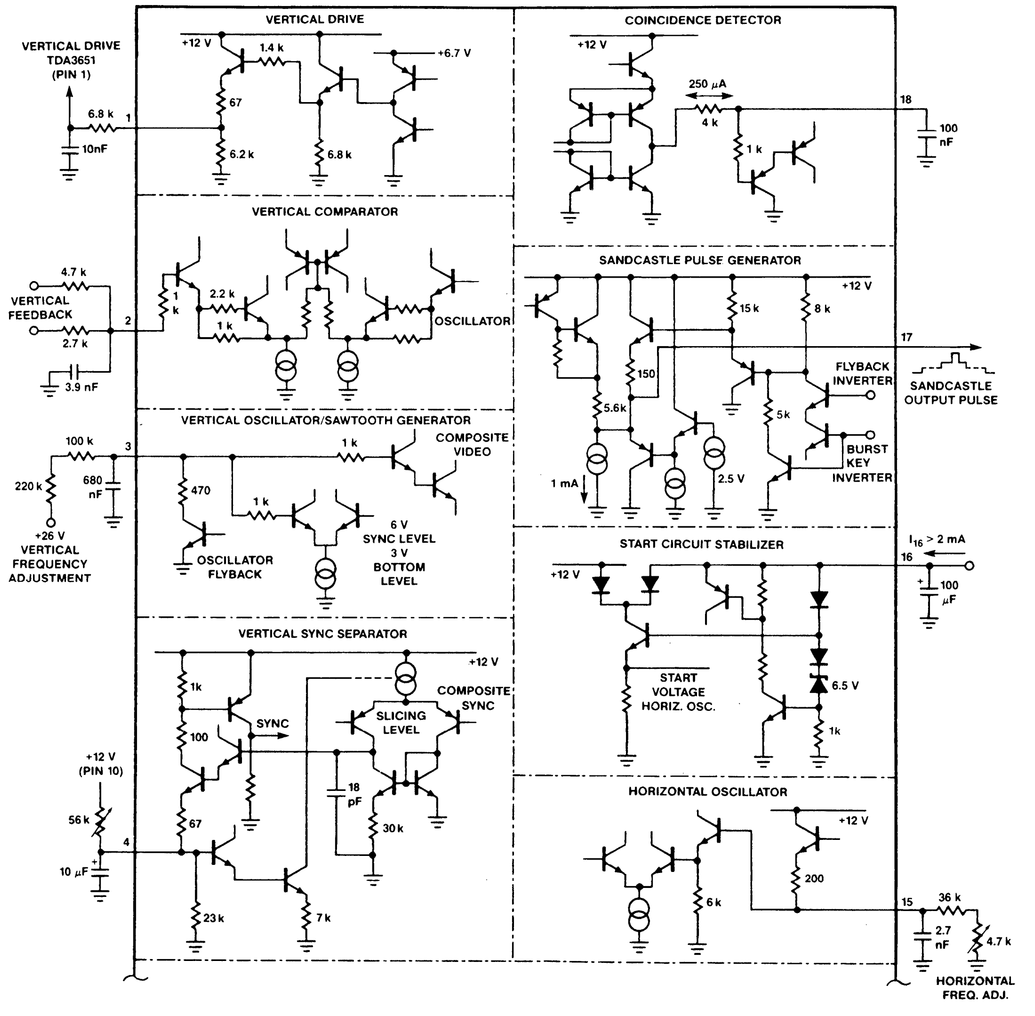


Figure 1. TDA2577A Circuit Diagram

Sync Circuit With Vertical Oscillator and Driver — TDA2577A

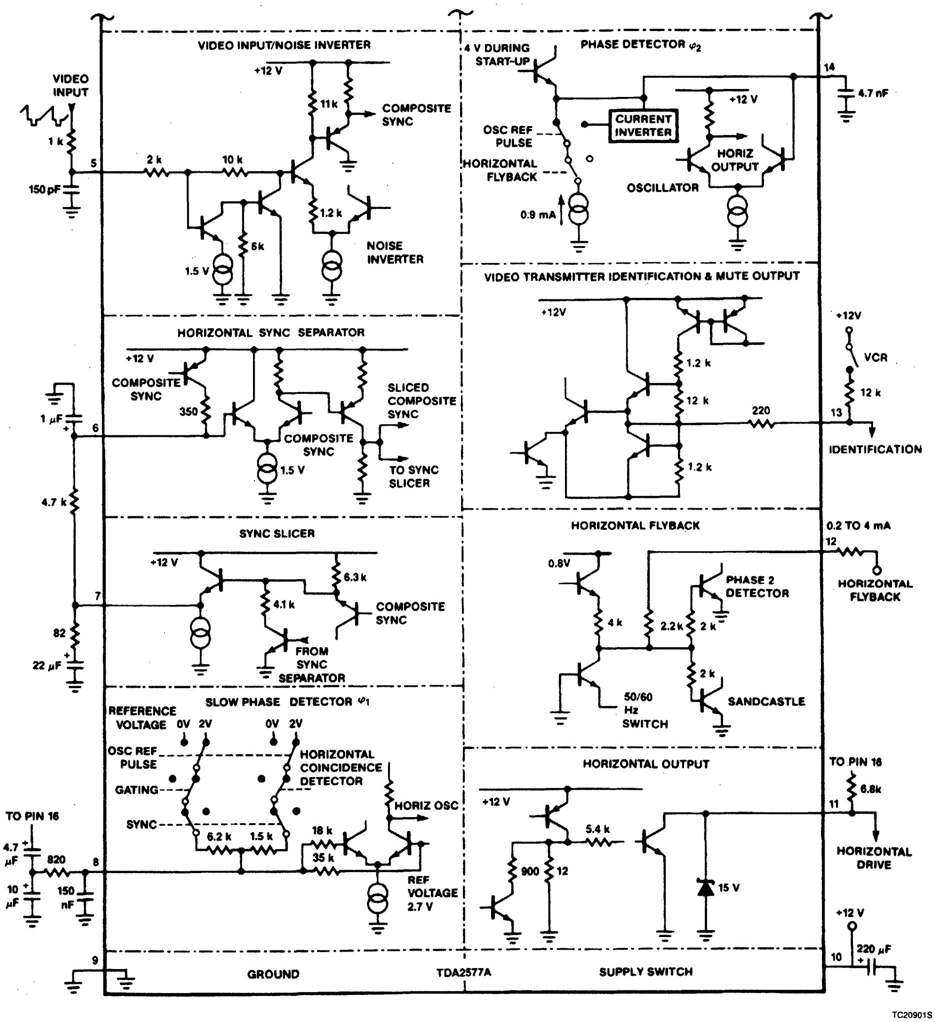


Figure 1. TDA2577A Circuit Diagram (Continued)

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
I_{16}	Start current (Pin 16)	8	mA
$V_{CC} = V_{10-9}$	Supply voltage (Pin 10)	13.2	V
P_{TOT}	Total power dissipation	1.1	W
T_{STG}	Storage temperature range	−65 to +150	°C
T_A	Operating ambient temperature range	−25 to +65	°C
θ_{JA}	Thermal resistance from junction to ambient in free air	50	°C/W

DC ELECTRICAL CHARACTERISTICS $I_{16} = 5mA$; $V_{CC} = 12V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supply					
I_{16}	Supply current at Pin 16	4		8	mA
V_{16-9}	Stabilized supply voltage (Pin 16)	8.0	8.7	9.5	V
I_{10}	Supply current (Pin 10)		55	70	mA
$V_{CC} = V_{10-9}$	Supply voltage (Pin 10)	10	12	13.2	V
Video input (Pin 5)					
V_{5-9}	Top-sync level	1.5	3.1	3.75	V
$V_{5-9(P-P)}$	Sync pulse amplitude (peak-to-peak value)[1]	0.15	0.6	1	V
	Slicing level	35	50	65	%
t_1	Delay between video input and detector output		0.35		μs
Noise gate (Pin 5)					
V_{5-9}	Switching level		0.7	1	V
First control loop (sync to oscillator; Pin 8)					
Δf	Holding range		±800		Hz
Δf	Catching range	±600	800	1100	Hz
	Control sensitivity video with respect to oscillator, burst key, and flyback pulse for slow time constant for fast time constant		 1 275		 kHz/μs kHz/μs
Second control loop (horizontal output to flyback; Pin 14)					
$\Delta t_D/\Delta t_O$	Control sensitivity; static[2]		400		μs/μs
t_D	Control range	1		50	μs
	Controlled edge		negative		
Phase adjustment (via 2nd control loop; Pin 14)					
	Control sensitivity		25		μA/μs
$\pm I_{14}$	Maximum permissible control current		0	50	μA
Horizontal oscillator (Pin 15)					
f_{OSC}	Frequency (no sync)		15625		Hz
Δf_{OSC}	Frequency spread ($C_{OSC} = 2.2nF$; $R_{OSC} = 40k\Omega$)			4	%
Δf_{OSC}	Frequency deviation between starting point of output signal and stabilized condition		6	8	%
T_C	Temperature coefficient		1×10^{-4}		°C

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Sync Circuit With Vertical Oscillator and Driver — TDA2577A

DC ELECTRICAL CHARACTERISTICS (Continued) $I_{16} = 5mA$; $V_{CC} = 12V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS Min	Typ	Max	UNIT
Horizontal output (Pin 11)					
V_{11-9}	Output voltage; high level			13.2	V
V_{11-9}	Voltage at which protection starts	13		15.8	V
V_{11-9}	Output voltage; low level start condition at $I_{11} = 10mA$		0.3	0.5	V
V_{11-9}	normal condition at $I_{11} = 40mA$		0.3	0.5	V
δ	Duty factor of output signal during starting (no phase shift; voltage at Pin 11 Low)		65		%
δ	Duty factor of output signal without flyback pulse	45	50	55	%
	Controlled edge	negative			
	Duration of output pulse (see Figure 2)	$t_D + t_O + 2.5$			μs
Sandcastle output pulse (Pin 17)					
	Output voltage during:				
V_{17-9}	burst key	10			V
V_{17-9}	horizontal blanking	4.2	4.6	5	V
V_{17-9}	vertical blanking	2	2.5	3	V
t_P	Pulse duration				
	burst key	3.6	4	4.4	μs
	horizontal blanking	flyback pulse[3]			
	vertical blanking				
	for 50Hz application ($-I_{12}$: 0 to 0.1mA)			21	lines
	for 60Hz application ($-I_{12}$: typ. 0.2mA)			17	lines
t_2	Delay between the start of the sync at the video input and the rising edge of the burst key pulse	4.8	5.2	5.6	μs
Coincidence detector; video transmitter identification circuit; time constant switches (Pin 18); see also Figure 1					
$\pm I_{18}$	Detector output current		300		μA
V_{18-9}	Voltage during noise[4]		0.3		V
V_{18-9}	Voltage level for in-sync condition		7.5		V
V_{18-9}	Switching level slow-to-fast	3.2	3.5	3.8	V
	Switching level				
V_{18-9}	mute function active; φ_1 fast-to-slow	1.0	1.2	1.4	V
	vertical period counter				
V_{18-9}	3 periods fast	0.08	0.12	0.16	V
V_{18-9}	Switching level slow-to-fast (locking) mute function inactive	1.5	1.7	1.9	V
V_{18-9}	Switching level fast-to-slow (locking)	4.7	5.0	5.3	V
V_{18-9}	Switching level for VCR (fast time constant) without mute function	8.2	8.6	9	V

DC ELECTRICAL CHARACTERISTICS (Continued) $I_{16} = 5mA$; $V_{CC} = 12V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Video transmitter identification output (Pin 13)					
V_{13-9}	Output voltage active (no sync) at $I_{13} = 1mA$	10	11		V
V_{13-9}	Output voltage active (no sync) at $I_{13} = 5mA$	7	10		V
V_{13-9}	Output voltage inactive		0.1	0.5	V
VCR switching (Pin 13)					
I_{13}	Input current for fast time constant phase detector φ_1, with mute function active	0.4	0.6	0.8	mA
Flyback input pulse (Pin 12)					
V_{12-9}	Switching level		1		V
I_{12}	Input current	0.2		4	mA
$V_{12-9(P-P)}$	Input pulse amplitude (peak-to-peak value)			12	V
R_{12-9}	Input resistance		2.7		kΩ
t_O	Delay time of sync pulse (measured in φ_1) to flyback at switching level; $t_{FL} = 12\mu s^2$ (see also Figure 3)		1.3		μs
Duration of vertical blanking pulse (Pin 12)					
 $-I_{12}$ $-I_{12}$	Required input current (negative) for 50Hz application; 21 lines blanking for 60Hz application; 17 lines blanking	 0.15	 0.2	 0.3 0.1	mA mA mA
$-I_{12}$	Maximum allowed input current			0.4	mA
Vertical sawtooth generator (Pin 3)					
f_S	Vertical frequency (no sync)		46		Hz
Δf_S	Frequency spread ($C_{OSC} = 680nF$; $R_{OSC} = 180k\Omega$; at +26V)			4	%
	Synchronization range		22		%
I_3	Input current at $V_{3-9} = 6V$			2	μA
Δf_S	Frequency shift for $V_{CC} = 10$ to 13V			0.2	%
T_C	Temperature coefficient		1×10^{-4}		$°C^{-1}$
Comparator (Pin 2)					
V_{2-9} $V_{2-9(P-P)}$	Input voltage DC level AC level (peak-to-peak value)	4.0	4.4 1.6	4.8	V V
I_2	Input current at $V_{2-9} = 6V$			2	μA
	Sawtooth internal precorrection (parabolic convex)		3		%
Vertical output stage; emitter-follower (Pin 1)					
V_{1-9}	Output voltage at $I_1 = 10mA$	3.2	3.6	5	V
I_1	Output current			20	mA
Vertical guard circuit					
 V_{2-9} V_{2-9}	Activating voltage levels (vertical blanking level is 2.5V) switching level Low switching level High	 2.7 5.4	 3 5.8	 3.3 6.3	 V V

NOTES:
1. Up to $1V_{P-P}$ the slicing level is constant; at amplitudes exceeding $1V_{P-P}$, the slicing level will increase.
2. t_D = delay between negative transient of horizontal output pulse and the rising edge of the flyback pulse. t_O = delay between the rising edge of the flyback pulse and the start of the current in $\varphi 1$ (Pin 8).
3. The duration of the flyback pulse is measured at the input switching level, which is about $1V(t_{FL})$.
4. Depends on DC level at Pin 5; value given applicable for $V_{5-9} \approx 5V$.

APPLICATION INFORMATION

The TDA2577A generates the signal for driving the horizontal deflection output circuit. It also contains a synchronized vertical sawtooth generator for direct drive of the vertical deflection output stage.

The horizontal oscillator and output stage can start operating on a very low supply current ($I_{16} \geqslant 4mA$), which can be taken directly from the supply line. Therefore, it is possible to derive the main supply (Pin 10) from the horizontal deflection output stage. The duty factor of the horizontal output signal is about 65% during the starting-up procedure. After starting up, the second phase detector ($\varphi 2$) is activated to control the timing of the negative-going edge of the horizontal output signal.

A bandgap reference voltage (6.5V) is provided for supply and reference of the vertical oscillator and comparator stage.

The slicing level of the horizontal sync separator is independent of the amplitude of the sync pulse at the input. The resistor between Pins 6 and 7 determines its value. A 4.7kΩ resistor gives a slicing level at the middle of the sync pulse. The nominal top sync level at the input is 3.1V. The amplitude selective noise inverter is activated at a level of 0.7V.

Good stability is obtained by means of the two control loops. In the first loop, the phase of the horizontal sync signal is compared to a waveform with its rising edge refering to the top of the horizontal oscillator signal. In the second loop, the phase of the flyback pulse is compared to another reference waveform, the timing of which is such that the top of the flyback pulse is situated symmetrically on the horizontal blanking interval of the video signal. Therefore, the first loop can be designed for a good noise immunity, whereas the second loop can be as fast as desired for compensation of switch-off delays in the horizontal output stage.

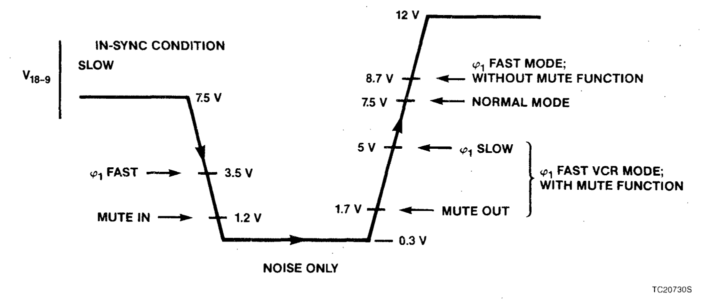


Figure 2. Voltage Levels at Pin 18 (V_{18-9})

The first phase detector is gated with a pulse derived from the horizontal oscillator signal. This gating (slow time constant) is switched off during catching. Also, the output current of the phase detector is increased fivefold during the catching time and VCR conditions (fast time constant). The first phase detector is inhibited during the retrace time of the vertical oscillator.

The in-sync, out-of-sync, or no-video condition is detected by the video transmitter identification/coincidence detector circuit (Pin 18). The voltage on Pin 18 defines the time constant and gating of the first phase detector. The relationship between this voltage and the various switching levels is shown in Figure 2. The complete survey of the switching actions is given in Table 1.

Table 1. Switching Levels at Pin 18

VOLTAGE AT PIN 18	FIRST PHASE DETECTOR φ_1				MUTE OUTPUT AT PIN 13		RECEIVING CONDITIONS
	Time Constant		Gating		On	Off	
	Slow	Fast	On	Off			
7.5V	X		X			X	Video signal detected
7.5 to 3.5V	X		X			X	Video signal detected
3.5 to 1.2V		X		X		X	Video signal detected
1.2 to 0.1V	X		X		X		Noise only
0.1 to 1.7V	X	*	X	*	X		New video signal detected
1.7 to 5.0V		X		X		X	Horizontal oscillator locked
							VCR playback with mute function
5.0 to 7.5V	X		X			X	Horizontal oscillator locked
8.7V		X		X		X	VCR playback without mute function

Where: * = 3 vertical periods.

The stability of displayed video information (e.g., channel number) during noise-only conditions is improved by the first phase detector time constant being set to slow.

The average voltage level of the video input on Pin 5 during noise-only conditions should not exceed 5.5V. Otherwise, the time constant switch may be set to fast due to the average voltage level on Pin 18 dropping below 0.1V. When the voltage on Pin 18 drops below 100mV, a counter is activated which sets the time constant switch to fast, and not gated for 3 vertical periods. This condition occurs when a new video signal is present at Pin 5. When the horizontal oscillator is locked, the voltage on Pin 18 increases. Nominally, a level of 5V is reached within 15ms (1 vertical period). The mute switching level of 1.2V is reached within 5ms (C_{18} = 47nF). If the video transmitter identification circuit is required to operate under VCR playback conditions, the first phase detector can be set to fast by connecting a resistor of 180kΩ between Pin 18 and ground. Also, a current of 0.6mA into Pin 13 sets the first phase detector to fast without affecting the mute output function (active High with no video signal detected). For VCR playback without mute function, the first phase detector can be set to fast by connecting a resistor of 1kΩ to the supply (Pin 10).

The supply for the horizontal oscillator (Pin 15) and horizontal output stage (Pin 11) is derived from the voltage at Pin 16 during the start condition. The horizontal output signal starts at a nominal supply current into Pin 16 of 3.5mA, which will result in a supply voltage of about 5.5V (for guaranteed operation of all devices $I_{16} > 4mA$). It is possible that the main supply voltage at Pin 10 is 0V during starting, so the main supply of the IC can be taken from the horizontal deflection output stage. The start of the other IC functions depends on the value of the main supply voltage at Pin 10. At 5.5V, all IC functions start operating except the second phase detector (oscillator to flyback pulse). The output voltage of the second phase detector at Pin 14 is clamped by means of an internally-loaded NPN emitter-follower. This ensures that the duty factor of the horizontal output signal (Pin 11) remains at about 65%. The second phase detector will close if the supply voltage at Pin 10 reaches 8.8V. At this value, the supply current for the horizontal oscillator and output stage is delivered by Pin 10, which also causes the voltage at Pin 16 to change to a stabilized 8.7V. This change switches off the NPN emitter-follower at Pin 14 and activates the second phase detector. The supply voltage for the horizontal oscillator will, however, still be referred to the stabilized voltage at Pin 16, and the duty factor of the output signal at Pin 12 is at the value required by the delay at the horizontal deflection stage. Thus, switch-off delays in the horizontal output stage are compensated. When no horizontal flyback signal is detected, the duty factor of the horizontal output signal is 50%.

Horizontal picture shift is possible by externally charging or discharging the 47nF capacitor connected to Pin 14.

The IC also contains a synchronized vertical oscillator/sawtooth generator. The oscillator signal is connected to the internal comparator (the other side of which is connected to Pin 2) via an inverter and amplitude divider stage. The output of the comparator drives an emitter-follower output stage at Pin 1. For a linear sawtooth in the oscillator, the load resistor at Pin 3 should be connected to a voltage source of 26V or higher. The sawtooth amplitude is not influenced by the main supply at Pin 10. The feedback signal is applied to Pin 2 and compared to the sawtooth signal at Pin 3. For an economical feedback circuit with less picture bounce, the sawtooth signal is internally precorrected by 3% (convex) referred to Pin 2. The linearity of the vertical deflection current depends upon the oscillator signal at Pin 3 and the feedback signal at Pin 2.

Synchronization of the vertical oscillator is inhibited when the mute output is present at Pin 13.

To minimize the influence of the horizontal part on the vertical part, a 6.5V bandgap reference source is provided for supply and reference of the vertical oscillator and comparator.

The sandcastle pulse, generated at Pin 17, has three different voltage levels. The highest level (11V) can be used for burst gating and black level clamping. The second level (4.6V) is obtained from the horizontal flyback pulse at Pin 12 and used for horizontal blanking. The third level (2.5V) is used for vertical blanking and is derived by counting the horizontal frequency pulses. For 50Hz, the blanking pulse duration is 21 lines and for 60Hz it is 17 lines. The blanking pulse duration is set by the negative voltage value of the horizontal flyback pulse at Pin 12.

The IC also incorporates a vertical guard circuit which monitors the vertical feedback signal at Pin 2. If this level is below 3V or higher than 5.8V, the guard circuit will insert a continuous level of 2.5V into the sandcastle output signal. This will result in complete blanking of the screen if the sandcastle pulse is used for blanking in the TV set.

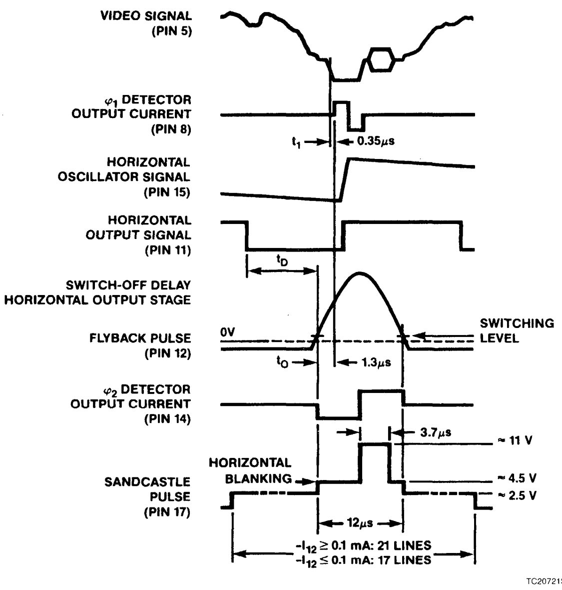


Figure 3. Timing Diagram of the TDA2577A

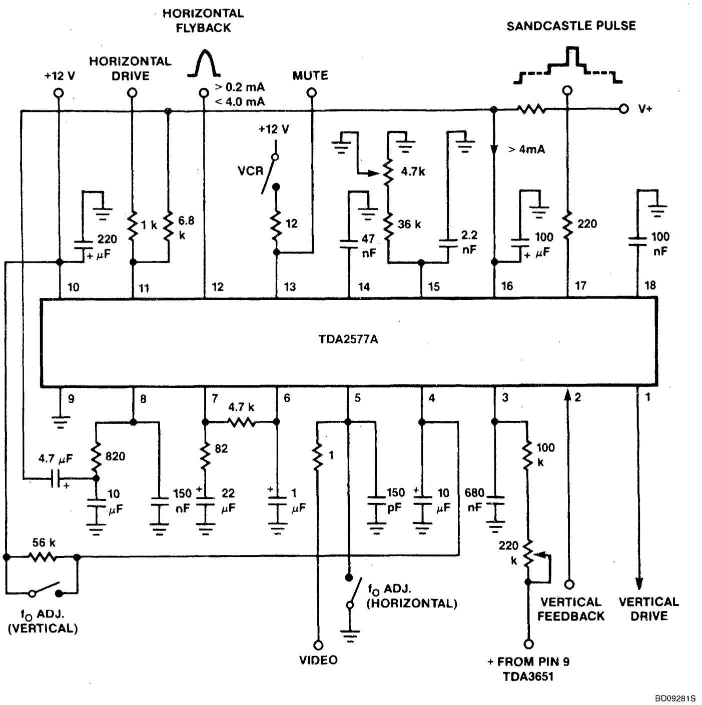


Figure 4. Typical Application Circuit Diagram; for Combination of the TDA2577A with the TDA3651 (see Figure 6)

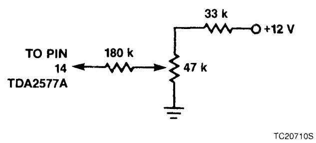


Figure 5. Circuit Configuration at Pin 14 for Phase Adjustment

Sync Circuit With Vertical Oscillator and Driver TDA2577A

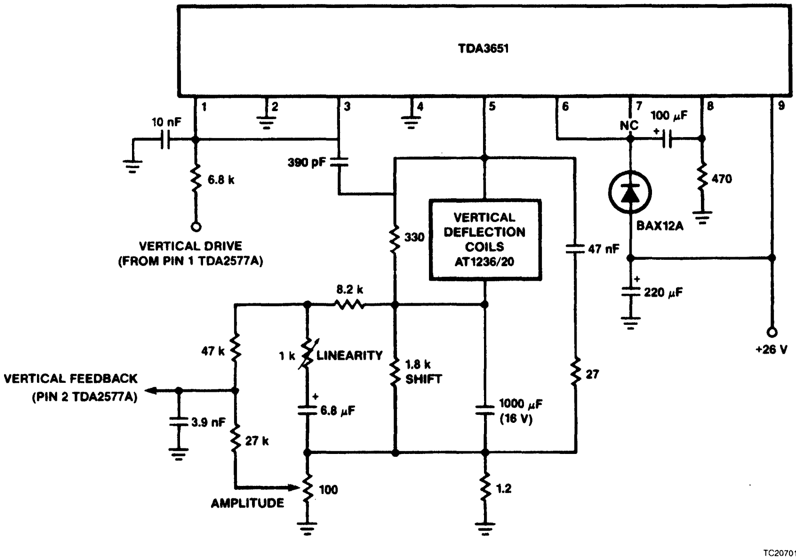


Figure 6. Typical Application Circuit Diagram of the TDA3651 (Vertical Output) When Used in Combination With the TDA2577A (90°C Application)

Signetics

Linear Products

TDA2578A
Sync Circuit With Vertical Oscillator and Driver

Product Specification

DESCRIPTION

The TDA2578A separates the vertical and horizontal sync pulses from the composite TV video signal and uses them to synchronize horizontal and vertical oscillators.

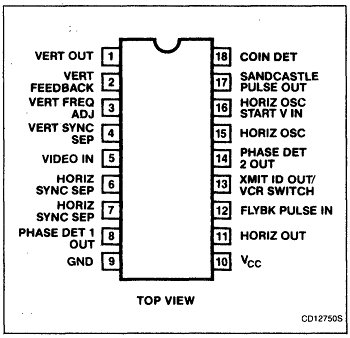
FEATURES

- **Horizontal sync separator and noise inverter**
- **Horizontal oscillator**
- **Horizontal output stage**
- **Horizontal phase detector (sync-to-oscillator)**
- **Time constant switch for phase detector (fast time constant during catching)**
- **Slow time constant for noise-only conditions**
- **Time constant externally switchable (e.g., fast for VCR)**
- **Inhibit of horizontal phase detector and video transmitter identification circuit during vertical oscillator flyback**
- **Second phase detector (φ2) for storage compensation of horizontal deflection stage**
- **Sandcastle pulse generator (3 levels)**
- **Video transmitter identification circuit**
- **Stabilizer and supply circuit for starting the horizontal oscillator and output stage directly from the power line rectifier**
- **Duty factor of horizontal output pulse is 50% when flyback pulse is absent**
- **Vertical sync separator**
- **Bandgap 6.5V reference voltage for vertical oscillator and comparator**
- **Synchronized vertical oscillator/sawtooth generator (synchronization inhibited when no video transmitter is detected)**
- **Internal circuit for 6% parabolic pre-correction of the oscillator/sawtooth generator. Comparator supplied with pre-corrected sawtooth and external feedback input**
- **Vertical driver stage**
- **Vertical blanking pulse generator**
- **50/60Hz detector**
- **50/60Hz identification output**
- **Automatic amplitude adjustment for 60Hz**
- **Automatic adjustment of blanking pulse duration (50Hz: 21 lines; 60Hz: 17 lines)**
- **Vertical guard circuit**

APPLICATIONS

- **Video terminals**
- **Television**

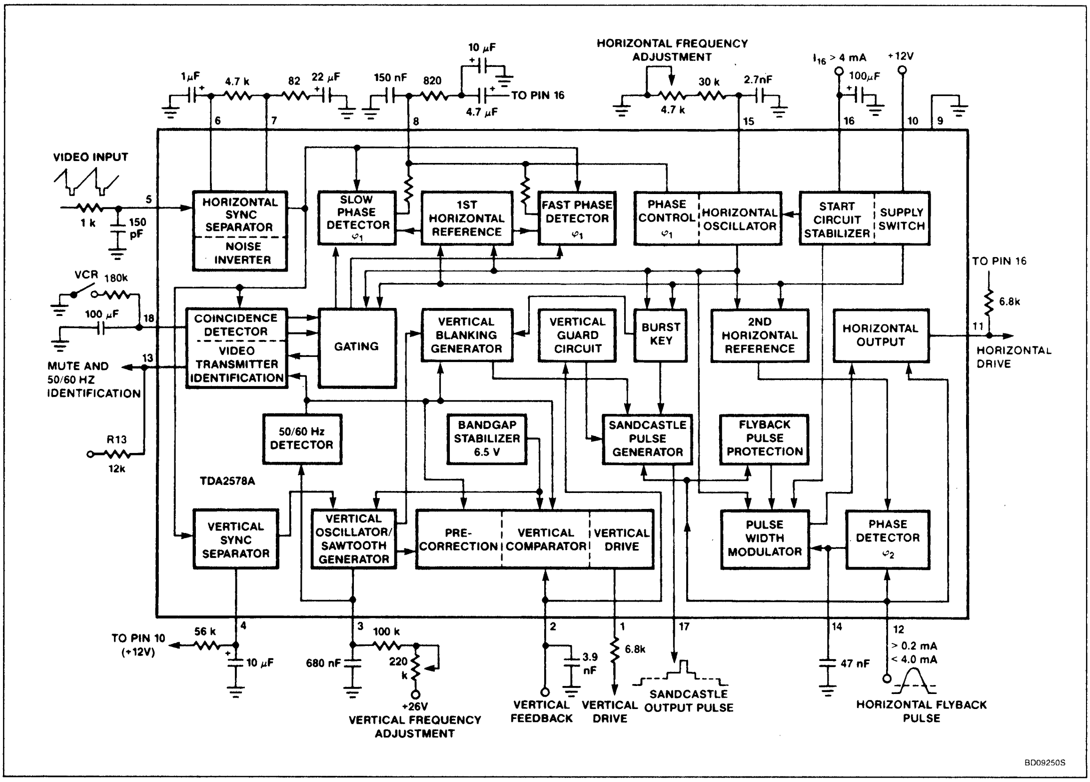
PIN CONFIGURATION



ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
18-Pin Plastic DIP (SOT-102HE)	−25°C to +65°C	TDA2578A

BLOCK DIAGRAM



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Sync Circuit With Vertical Oscillator and Driver — TDA2578A

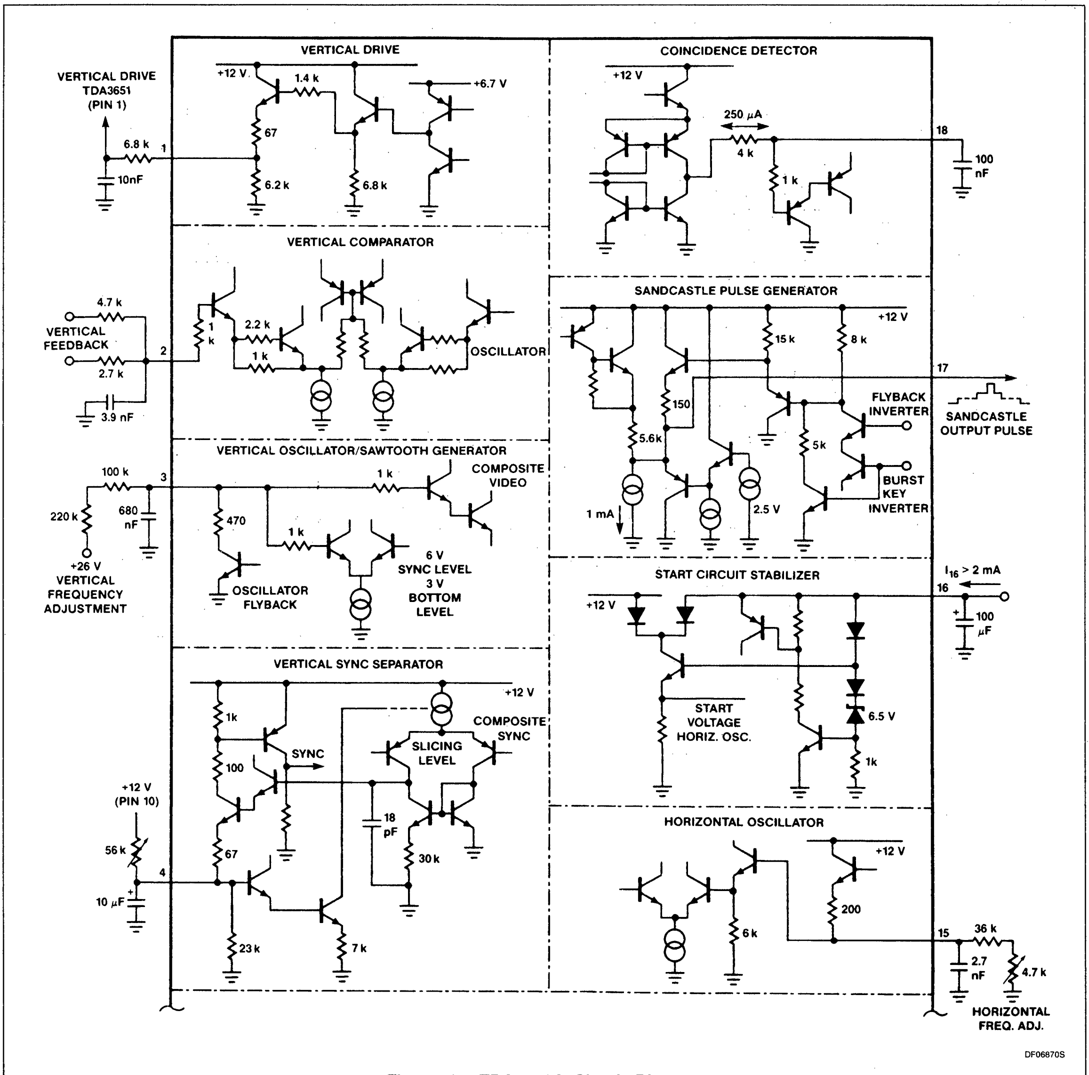


Figure 1a. TDA2578A Circuit Diagram

Sync Circuit With Vertical Oscillator and Driver TDA2578A

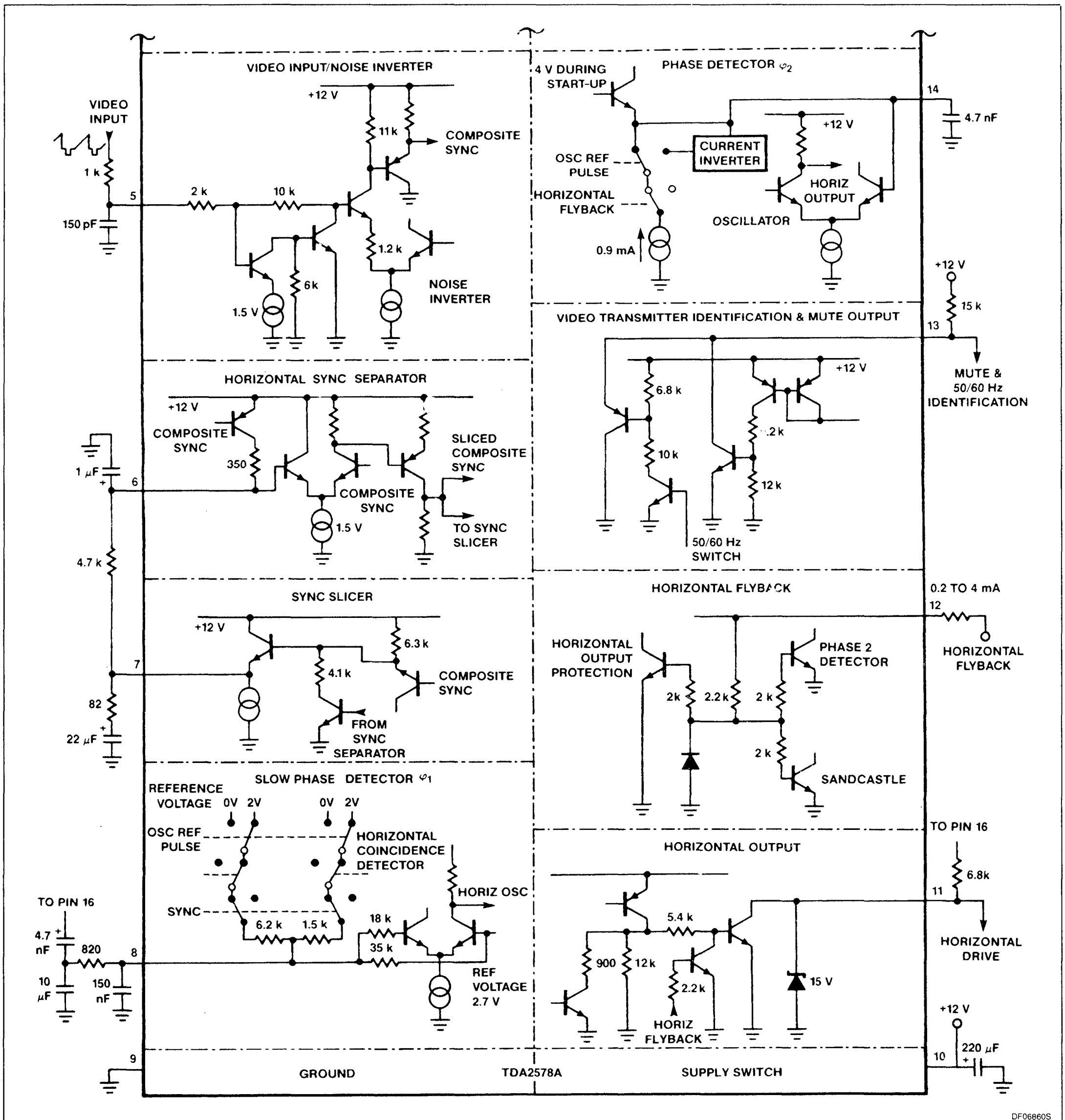


Figure 1b. TDA2578A Circuit Diagram

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
I_{16}	Start current (Pin 16)	8	mA
$V_{CC} = V_{10-9}$	Supply voltage (Pin 10)	13.2	V
P_{TOT}	Total power dissipation	1.1	W
T_{STG}	Storage temperature range	-65 to +150	°C
T_A	Operating ambient temperature range	-25 to +65	°C
θ_{JA}	Thermal resistance from junction to ambient in free air	50	°C

DC AND AC ELECTRICAL CHARACTERISTICS $I_{16} = 5mA$; $V_{CC} = 12V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supply					
I_{16}	Supply current at Pin 16	4		8	mA
V_{16-9}	Stabilized supply voltage (Pin 16)	8	8.7	9.5	V
I_{10}	Supply current (Pin 10)		55	70	mA
$V_{CC} = V_{10-9}$	Supply voltage (Pin 10)	10	12	13.2	V
Video input (Pin 5)					
V_{5-9}	Top-sync level	1.5	3.1	3.75	V
$V_{5-9(P-P)}$	Sync pulse amplitude (peak-to-peak value)[1]	0.15	0.6	1	V
	Slicing level	35	50	65	%
t_1	Delay between video input and detector output		0.35		µs
Noise gate (Pin 5)					
V_{5-9}	Switching level		0.7	1	V
First control loop (sync to oscillator; Pin 8)					
Δf	Holding range		±800		
Δf	Catching range	600	800	1100	Hz
	Control sensitivity video with respect to oscillator, burst key, and flyback pulse for slow time constant for fast time constant		 1 2.75		 kHz/µs kHz/µs
Second control loop (horizontal output to flyback; Pin 14)					
$\Delta t_D/\Delta t_O$	Control sensitivity; static[2]		400		µs/µs
t_D	Control range	1		45	µs
	Controlled edge (positive)				
Phase adjustment (via 2nd control loop; Pin 14)					
	Control sensitivity		25		µA
$\pm I_{14}$	Maximum permissible control current			50	µA
Horizontal oscillator (Pin 15)					
f_{OSC}	Frequency (no sync)		15625		Hz
Δf_{OSC}	Frequency spread ($C_{OSC} = 2.7nF$; $R_{OSC} = 33k\Omega$; no sync)			4	%
Δf_{OSC}	Frequency deviation between starting point of output signal and stabilized condition	6		8	%
TC	Temperature coefficient		10^{-4}		°C

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) I_{16} = 5mA; V_{CC} = 12V; T_A = 25°C, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS Min	LIMITS Typ	LIMITS Max	UNIT
Horizontal output (Pin 11)					
V_{11-9}	Output voltage; high level			13.2	V
V_{11-9}	Voltage at which protection starts	13		15.8	V
 V_{11-9} V_{11-9}	Output voltage; low level start condition at I_{11} = 10mA normal condition at I_{11} = 40mA		 0.3 0.3	 0.5 0.5	 V V
δ	Duty factor of output signal during starting (no phase shift) I_{16} = 4mA (voltage at Pin 11 low)		65		%
δ	Duty factor of output signal without flyback pulse	45	50	55	%
	Controlled edge (positive)				
	Duration of output pulse (see Figure 3)	t_D+ horizontal flyback pulse			
Sandcastle output pulse (Pin 17)					
 V_{17-9} V_{17-9} V_{17-9}	Output voltage during: burst key horizontal blanking vertical blanking	 4.2 2	 4.6 2.5	 10 5 3	 V V V
t_P	Pulse duration burst key	3.6	4	4.4	µs
	horizontal blanking (flyback pulse)[3] vertical blanking at 50Hz at 60Hz	 21 lines 17 lines			
t_2	Delay between the start of the sync at the video input and the rising edge of the burst key pulse	4.5			
Coincidence detector; video transmitter identification circuit; time constant switches (Pin 18) (see also Figure 2)					
$\pm I_{18}$	Detector output current		300		µA
V_{18-9}	Voltage during noise[4]		0.3		V
V_{18-9}	Voltage level for in-sync condition		7.5		V
V_{18-9}	Switching level slow to fast	3.2	3.5	3.8	V
 V_{18-9} V_{18-9}	Switching level mute function active; φ_1 fast to slow vertical period counter; 3 periods fast	 1 0.08	 1.2 0.12	 1.4 0.16	 V V
V_{18-9}	Switching level slow-to-fast (locking) mute function inactive	1.5	1.7	1.9	V
V_{18-9}	Switching level fast-to-slow (locking)	4.7	5	5.3	V
V_{18-9}	Switching level for VCR (fast time constant) without mute function	8.2	8.6	9	V
Video transmitter identification output (Pin 13)					
V_{13-9}	Output voltage active (no sync) at I_{13} = 1mA		0.3	0.5	V
I_{13}	Sink current active (no sync)		5		mA
I_{13}	Output current inactive (sync: 50Hz)			1	µA
50/60Hz identification (Pin 13)					
 V_{13-9} V_{13-9}	R13 = 15kΩ to +12V[5] at f = 50Hz (in sync condition) at f = 60Hz (in sync condition)	 7.2	 V_{10-9} 7.6	 8	 V V

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) I_{16} = 5mA; V_{CC} = 12V; T_A = 25°C, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS Min	LIMITS Typ	LIMITS Max	UNIT
Flyback input pulse (Pin 12)					
V_{12-9}	Switching level		1		V
I_{12}	Input current	0.2		4	mA
$V_{12-9(P-P)}$	Input pulse amplitude (peak-to-peak value)			12	V
R_{12-9}	Input resistance		2.7		kΩ
t_O	Delay time of sync pulse (measured in φ_1) to flyback at switching level; t_{FL} = 12μs[2] (see also Figure 3)		1.3		μs
Vertical sawtooth generator (Pin 3)					
f_S	Vertical frequency (no sync)		46		Hz
Δf_S	Frequency spread (C_{OSC} = 680nF; R_{OSC} = 180kΩ; at +26V)			4	%
	Synchronization range[6]		33		%
I_3	Input current at V_{3-9} = 6V			3	μA
Δf_S	Frequency shift for V_{CC} = 10 to 13V			0.2	%
TC	Temperature coefficient		10^{-4}		°C
Comparator (Pin 2)					
V_{2-9} $V_{2-9(P-P)}$	Input voltage; DC level AC level (peak-to-peak value)	4	4.4 0.8	4.8	V V
I_2	Input current at V_{2-9} = 6V			2	μA
	Sawtooth internal precorrection (parabolic convex)		6		%
Vertical output stage; emitter-follower (Pin 1)					
V_{1-9}	Output voltage at I_1 = 10mA	3.2		5	V
I_1	Output current			20	mA
Vertical guard circuit					
 V_{2-9} V_{2-9}	Activating voltage levels (vertical blanking level is 2.5V) switching level LOW switching level HIGH	 3 4.75	 3.35 5.15	 3.7 5.55	 V V

NOTES:
1. Up to $1V_{P-P}$ the slicing level is constant; at amplitudes exceeding $1V_{P-P}$ the slicing level will increase.
2. t_D = delay between positive transient of horizontal output pulse and the rising edge of the flyback pulse.
   t_O = delay between the rising edge of the flyback pulse and the start of the current in φ_1 (Pin 8).
3. The duration of the flyback pulse is measured at the input switching level, which is about 1V(t_{FL}).
4. Depends on DC level at Pin 5; value given applicable for $V_{5-9} \approx$ 5V.
5. For 60Hz, a PNP emitter clamp is activated.
6. When f_O = 46Hz, the 50/60Hz detector switches over to 60Hz; video input signal at Pin 5 ≈ 55Hz.

Sync Circuit With Vertical Oscillator and Driver

TDA2578A

Table 1. Switching Levels at Pin 18

VOLTAGE AT PIN 18	FIRST PHASE DETECTOR φ_1				MUTE OUTPUT AT PIN 13		RECEIVING CONDITIONS
	Time Constant		Gating				
	Slow	Fast	On	Off	On	Off	
7.5V	X		X			X	Video signal detected
7.5 to 3.5V	X		X			X	Video signal detected
3.5 to 1.2V		X		X		X	Video signal detected
1.2 to 0.1V	X		X		X		Noise only
0.1 to 1.7V	X	*	X	*	X		New video signal detected
1.7 to 5.0V		X		X		X	Horizontal oscillator locked VCR playback with mute function
5.0 to 7.5V	X		X			X	Horizontal oscillator locked
8.7V		X		X		X	VCR playback without mute function

Where: * = 3 vertical periods.

APPLICATION INFORMATION

The TDA2578A generates the signal for driving the horizontal deflection output circuit. It also contains a synchronized vertical sawtooth generator for direct drive of the vertical deflection output stage.

The horizontal oscillator and output stage can start operating on a very low supply current ($I_{16} \geqslant 4mA$), which can be taken directly from the power line rectifier. Therefore, it is possible to derive the main supply (Pin 10) from the horizontal deflection output stage. The duty factor of the horizontal output signal is about 65% during the starting-up procedure. After starting up, the second phase detector ($\varphi 2$) is activated to control the timing of the positive-going edge of the horizontal output signal.

A bandgap reference voltage (6.5V) is provided for supply and reference of the vertical oscillator and comparator stage.

The slicing level of the horizontal sync separator is independent of the amplitude of the sync pulse at the input. The resistor between Pins 6 and 7 determines its value. A 4.7kΩ resistor gives a slicing level at the middle of the sync pulse. The nominal top sync level at the input is 3.1V. The amplitude selective noise inverter is activated at a level of 0.7V.

Good stability is obtained by means of the two control loops. In the first loop, the phase of the horizontal sync signal is compared to a waveform with its rising edge refering to the top of the horizontal oscillator signal. In the second loop, the phase of the flyback pulse is compared to another reference waveform, the timing of which is such that the top of the flyback pulse is situated symmetrically on the horizontal blanking interval of the video signal. Therefore the first loop can be designed for a good noise immunity, whereas the second loop can be as fast as desired for compensation of switch-off delays in the horizontal output stage.

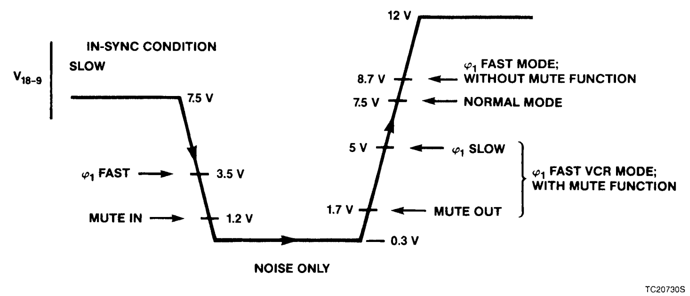


Figure 2. Voltage Levels at Pin 18 (V_{18-9})

The first phase detector is gated with a pulse derived from the horizontal oscillator signal. This gating (slow time constant) is switched off during catching. Also, the output current of the phase detector is increased fivefold, during the catching time and VCR conditions (fast time constant). The first phase detector is inhibited during the retrace time of the vertical oscillator.

The in-sync, out-of-sync, or no-video condition is detected by the video transmitter identification/coincidence detector circuit (Pin 18). The voltage on Pin 18 defines the time constant and gating of the first phase detector. The relationship between this voltage and the various switching levels is shown in Figure 2. The complete survey of the switching actions is given in Table 1.
The stability of displayed video information (e.g., channel number) during noise-only conditions is improved by the first phase detector time constant being set to slow.

The average voltage level of the video input on Pin 5 during noise-only conditions should not exceed 5.5V. Otherwise, the time constant switch may be set to fast due to the average voltage level on Pin 18 dropping below 0.1V. When the voltage on Pin 18 drops below 100mV, a counter is activated which sets the time constant switch to fast, and not gated for 3 vertical periods. This condition occurs when a new video signal is present at Pin 5. When the horizontal oscillator is locked, the voltage on Pin 18 increases. Nominally a level of 5V is reached within 15ms (1 vertical period). The mute switching level of 1.2V is reached within 5ms (C_{18} = 47nF). If the video transmitter identification circuit is required to operate under VCR playback conditions, the first phase detector can be set to fast by connecting a resistor of 180kΩ between Pin 18 and ground (see Figure 6).

The supply for the horizontal oscillator (Pin 15) and horizontal output stage (Pin 11) is derived from the voltage at Pin 16 during the start condition. The horizontal output signal starts at a nominal supply current into Pin 16 of 3.6mA, which will result in a supply voltage of about 5.5V (for guaranteed operation of all devices $I_{16} > 4mA$). It is possible that the main supply voltage at Pin 10 is 0V during starting, so the main supply of the IC can be taken from the horizontal deflection output stage. The start of the other IC functions depends on the value of the main supply voltage at Pin 10. At 5.5V, all IC functions

Sync Circuit With Vertical Oscillator and Driver | TDA2578A

start operating except the second phase detector (oscillator to flyback pulse). The output voltage of the second phase detector at Pin 14 is clamped by means of an internally-loaded NPN emitter-follower. This ensures that the duty factor of the horizontal output signal (Pin 11) remains at about 65%. The second phase detector will close if the supply voltage at Pin 10 reaches 8.8V. At this value, the supply current for the horizontal oscillator and output stage is delivered by Pin 10, which also causes the voltage at Pin 16 to change to a stabilized 8.7V. This change switches off the NPN emitter-follower at Pin 14 and activates the second phase detector. The supply voltage for the horizontal oscillator will, however, still be referred to the stabilized voltage at Pin 16, and the duty factor of the output signal at Pin 12 is at the value required by the delay at the horizontal deflection stage. Thus, switch-off delays in the horizontal output stage are compensated. When no horizontal flyback signal is detected, the duty factor of the horizontal output signal is 50%.

Horizontal picture shift is possible by externally charging or discharging the 47nF capacitor connected to Pin 14.

The IC also contains a synchronized vertical oscillator/sawtooth generator. The oscillator signal is connected to the internal comparator (the other side of which is connected to Pin 2), via an inverter and amplitude divider stage. The output of the comparator drives an emitter-follower output stage at Pin 1. For a linear sawtooth in the oscillator, the load resistor at Pin 3 should be connected to a voltage source of 26V or higher. The sawtooth amplitude is not influenced by the main supply at Pin 10. The feedback signal is applied to Pin 2 and compared to the sawtooth signal at Pin 3. For an economical feedback circuit with less picture bounce, the sawtooth signal is internally pre-corrected by 6% (convex) referred to Pin 2. The linearity of the vertical deflection current depends upon the oscillator signal at Pin 3 and the feedback signal at Pin 2.

Synchronization of the vertical oscillator is inhibited when the mute output is present at Pin 13.

To minimize the influence of the horizontal part on the vertical part, a 6.7V bandgap reference source is provided for supply and reference of the vertical oscillator and comparator.

The sandcastle pulse, generated at Pin 17, has three different voltage levels. The highest level (11V) can be used for burst gating and black level clamping. The second level (4.6V) is obtained from the horizontal flyback pulse at Pin 12 and used for horizontal blanking. The third level (2.5V) is used for vertical blanking and is derived by counting the horizontal frequency pulses. For 50Hz the blanking pulse duration is 21 lines, and for 60Hz it is 17 lines. The blanking pulse duration and sawtooth amplitude is automatically adjusted via the 50/60Hz detector.

The IC also incorporates a vertical guard circuit which monitors the vertical feedback signal at Pin 2. If this level is below 3.35V or higher than 5.15V, the guard circuit will insert a continuous level of 2.5V into the sandcastle output signal. This will result in complete blanking of the screen if the sandcastle pulse is used for blanking in the TV set.

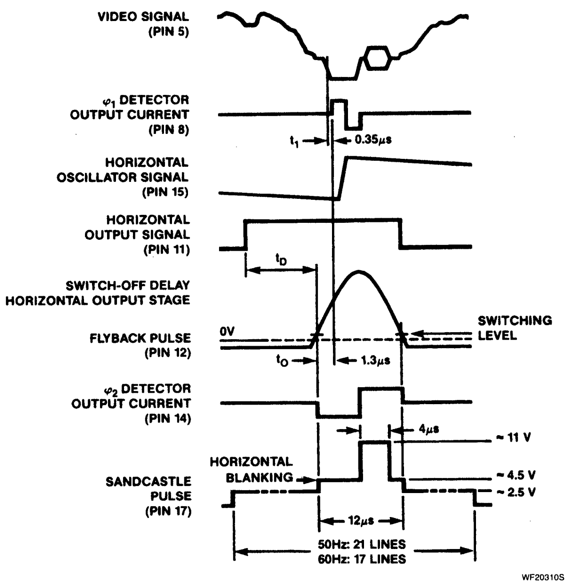
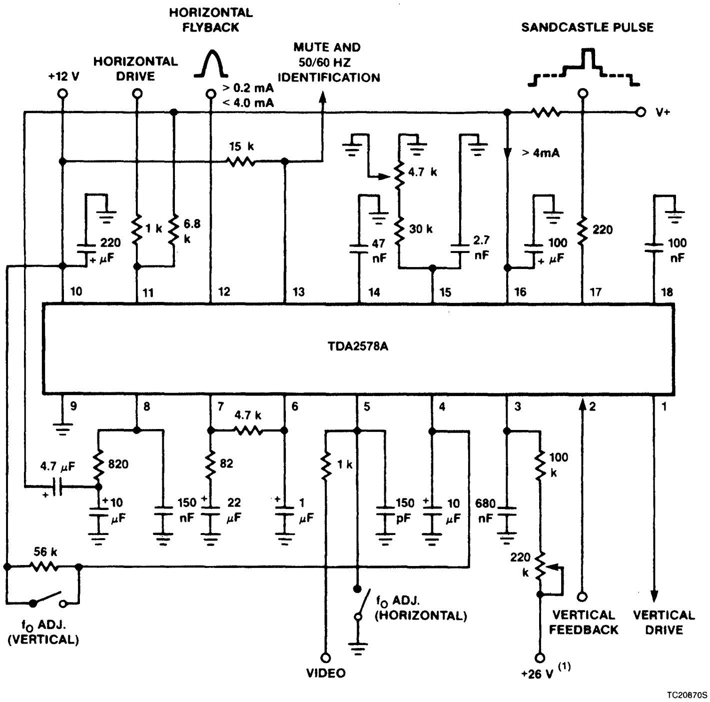


Figure 3. Timing Diagram of the TDA2578A

APPLICATION INFORMATION (Continued)



NOTE:
1. ⩾ 26V for linear scan.

Figure 4. Typical Application Circuit Diagram; for Application of the TDA2578A With the TDA3651 — See Figure 7

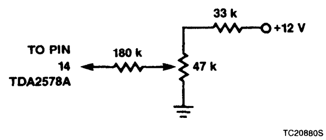
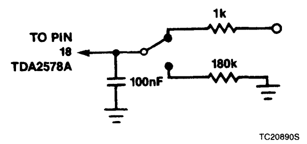


Figure 5. Circuit Configuration at Pin 14 for Phase Adjustment



NOTES:
1kΩ resistor between Pin 18 and +12V: without mute function.
180kΩ between Pin 18 and ground: with mute function.

Figure 6. Circuit Configuration at Pin 18 for VCR Mode

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APPLICATION INFORMATION (Continued)

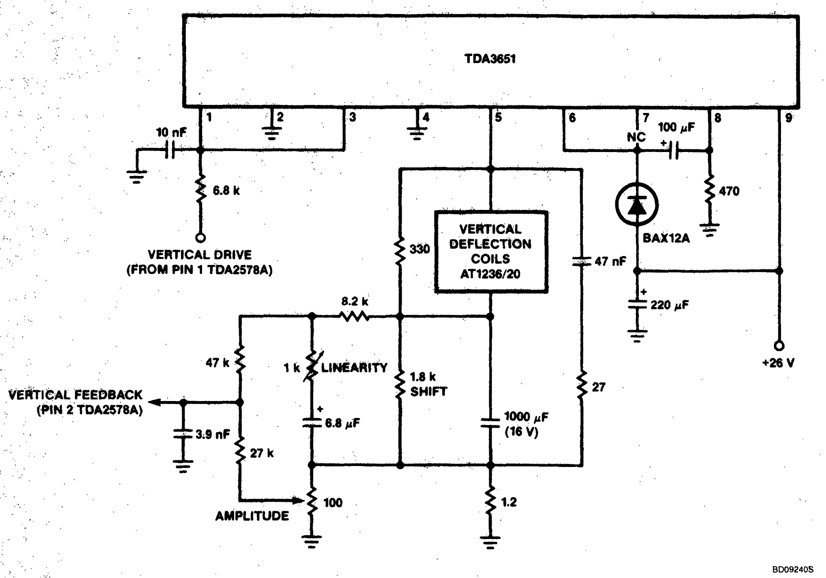

Figure 7. Typical Application Circuit Diagram of the TDA3651 (Vertical Output) When Used in Combination With the TDA2578A, (90° Application)

Signetics

Linear Products

AN162
A Versatile High-Resolution Monochrome Data and Graphics Display Unit

Application Note

INTRODUCTION

The Data and Graphics Display (DGD) unit, (also referred to as a Video Display Unit), is built for wide ranging applications. It cons ists of a very high resolution CRT paired with precision deflection coils and all the associated display circuitry, as shown in Figure 1. Using the same printed circuit board and components, it can easily be adapted to operate over a wide range of line and field frequencies with different flyback times in either horizontal (landscape) or vertical (portrait) format.

The possible applications of this unit range from video games to high-resolution displays. However, it is as a computer terminal display device that the DGD will be most useful. Normally, it is the logic design that determines all the parameters to be specified in a computer system, and it is only when the logic circuitry has been finalized that a suitable display is sought. Consequently, the display must be tailormade for the application. There are no signs of any standardization in the future. For this reason the DGD has been designed to allow different dedicated display units to be built up very simply from one basic design.

The DGD is a straightforward and efficient design which will operate with line frequencies of between 15 and 70kHz and field frequencies of 50 to 100Hz, interlaced or non-interlaced. All the design features combine to provide the resolution required for very high density displays (up to 1.5 million picture elements per page). They also ensure a sharp picture right to the screen corners, and allow operation at high horizontal line frequencies without undue temperature rise. A diode-split transformer provides combined line scan and EHT and it is this component which allows changes in line frequency and flyback time to be accomplished very easily.

NOTE:
EHT stands for *extreme haute-tension*, or extreme high voltage.

GENERAL DESCRIPTION

Figure 2 shows a block diagram of the DGD unit and its auxiliary circuits. (The unit is to the right of the broken line, with the auxiliary circuits to the left.) The circuit diagram is shown in Figure 3.

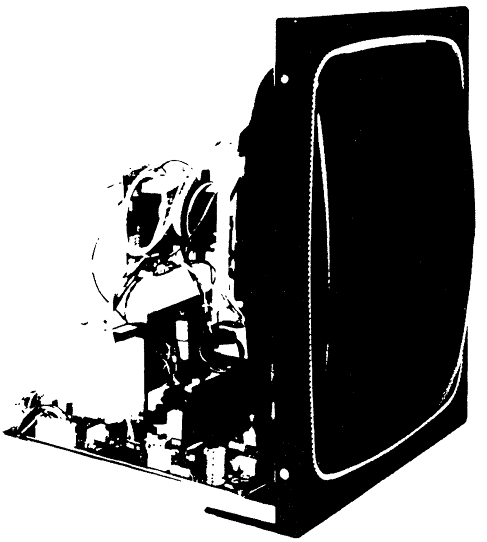

Figure 1. DGD Unit

The normal DGD requirements of good raster geometry and minimal loss of display quality between the screen center and corners are even more important in high-definition systems. To ensure a display offering the best possible resolution over the whole line frequency range, the unit uses high-quality purpose-designed deflection coils type AT1039. These are paired with either the 12 in (M31-326) or 15 in (M38-328) picture tubes. These coils have been designed using recently developed techniques to give good deflection performance and raster geometry suitable for correction by built-in magnets. For the 12 in tube, type AT1039/03 deflection coils are used. Two types of coil are available for the 15 in tube, the AT1039/00 which has been optimized for portrait (vertical) formats and the AT1039/01 for landscape (horizontal) displays. Terminations to each coil are brought out separately to allow for both series and parallel connections.

Both line scanning and EHT are provided by a purpose-built diode-split transformer. It is the flexibility of this device which produces the extreme versatility of the DGD unit as a whole and allows operation of the wide range of line frequencies and flyback times. In addition, all auxiliary power supply requirements are obtained from the same transformer. The primary is provided with several taps, each of which corresponds to a different peak voltage and hence flyback time. By careful positioning of these transformer primary taps, and by utilizing both parallel and series connection of the line deflection coils, a wide variety of flyback times can be accomodated in steps. Each step allows sensible values of flyback ratio for the different line frequencies. Apart from the selection of the correct transformer tap, the only other components that may need to be changed in order to use a different line frequency are the oscillator timing capacitor C6, S-correction capacitor C22, base drive resistor R52, linearity control L1, and heater resistor R84 (see Figure 3).

Although deflection defocusing has been minimized by careful design of the line deflection coils, there is still some focusing action in the deflection process. Also, there is a difference between the electron beam path lengths for axial beams and those deflected to the tube corners. These effects combine to produce a change in focus requirements from the center to the edges of the picture tube. To overcome this, dynamic focus is employed. The active dynamic focus circuit applies parabolic cor-

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A Versatile High-Resolution Monochrome Data and Graphics Display Unit

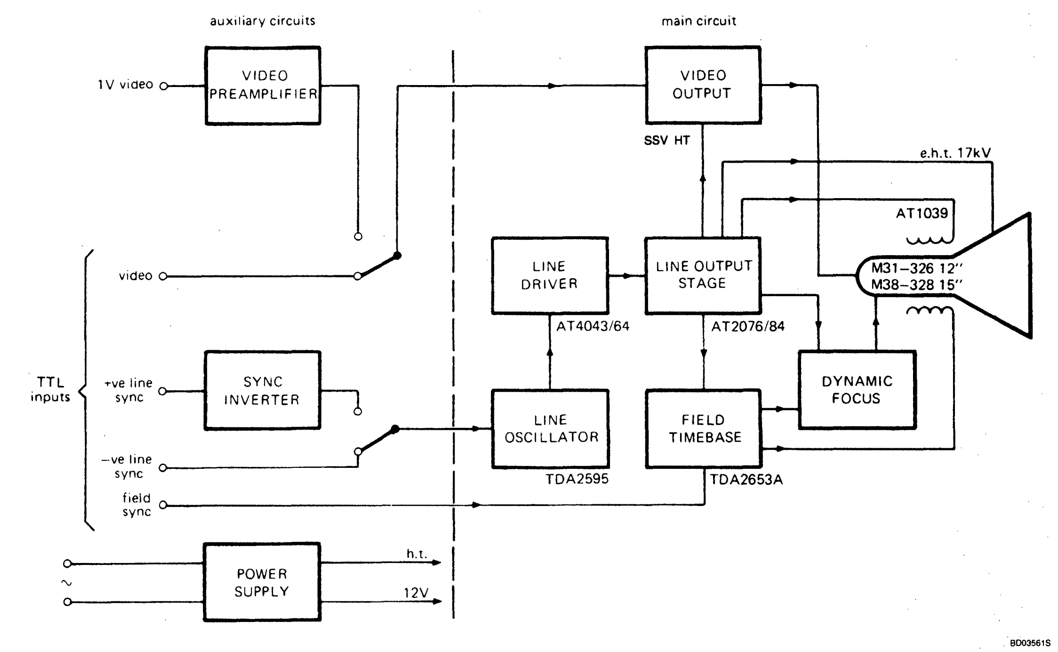


Figure 2. DGD Unit Block Diagram

rection in both the line and field directions to give precise focus over the whole raster. Because the electron gun is a unipotential type, the tube has a fairly flat focus characteristic. The amplitude of the dynamic focus can therefore be preset and adjustment is unnecessary.

Width control is accomplished with a series-parallel inductance arrangement which does not affect the flyback time or EHT. Adjustable picture shift is supplied in both the line and field directions by passing DC through the appropriate deflection coils.

The TDA2595 line oscillator combination IC provides the correct waveforms to drive the line output transistor via a transformer-coupled driver stage. This IC includes both the line oscillator and coincidence detector, a line flyback pulse, obtained from the collector of the line output transistor TR2, is required for phase detection. A protection circuit which turns off the output drive if the voltage at Pin 8 is either below 4 or above 8V is used to provide overvoltage protection for the line output stage.

All the field timebase functions are converted by the TDA2653A IC. It takes a positive-going field sync input at TTL level and drives the impedance-matched AT-1039 deflection coils in series connection. A field blanking pulse, which may be used for screen burn protection, is available from Pin 2. The IC is contained in a 13-lead DIP plastic power encapsulation type SOT-141, which offers straightforward heatsinking.

An emitter-driven video output stage is used with output transistor TR6 and driver TR7. The collector load resistors R87 and R88 with peaking coil L5 and some compensation in the emitter circuit ensure a bandwidth of 60MHz at 35V, measured at the cathode. In order to minimize stray capacitance, the video amplifier is placed on the tube-base printed circuit board close to the cathode pin of the tube. The 55V HT (High Tension) line is provided from the line output stage.

The unit will accept video input at TTL level with positive-going field sync and negative-going line sync. However, inputs at other levels and polarities may be accepted by using the auxiliary circuits, as shown in Figure 2.

The main HT line input will depend upon the line frequency and varies from about 30 to 150V. If lower values of HT are preferred, a floating tap will accommodate a series boosted circuit arrangement.

A 12V supply is required at all frequencies. The total power consumption of the unit is about 40W.

Standard measures are taken to protect the circuitry in the event of a picture tube flashover. Spark gaps for all picture tube pins are provided and all are returned to a single point which is, in turn, connected to the outside aquadag layer of the tube and the common earth point.

To achieve a satisfactory stable display with good linearity and one that is free from undesirable modulation, well recognized procedures should be adopted with regard to printed circuit board layout. It is essential that each individual circuit block has its own grounding system connected to a central point on the main printed circuit board which is, in turn, connected to the chassis. Circuit layout within the individual blocks may also be critical.

A Versatile High-Resolution Monochrome Data and Graphics Display Unit

AN162

Table 1. DGD Unit Specifications

Picture tube	12 in M31-326 series 15 in M38-328 series
Deflection coils	AT1039 series
Line output transformer	AT2076/84
Character display	Up to 1.5×10^6 pixels
Line frequency landscape format portrait format	 15 to 50kHz 15 to 70kHz
Field frequency non-interlaced or interlaced	 50 to 100Hz
EHT	17kV
Line linearity	Better than 3%
Field linearity	Better than 3%
Raster breathing (0 to 100μA)	 Better than 2%
Line flyback time	3 to 9μs
Field flyback time	0.6ms
Video bandwidth (at 35V output measured at the cathode)	 60MHz
Input signals	Positive field sync at TTL level, negative line sync at TTL level, video input at TTL level
Power input	40W total 30 to 150V 36W 12V 4W

Originally published as "Technical Publication 115," ELCOMA, The Netherlands, 1983.

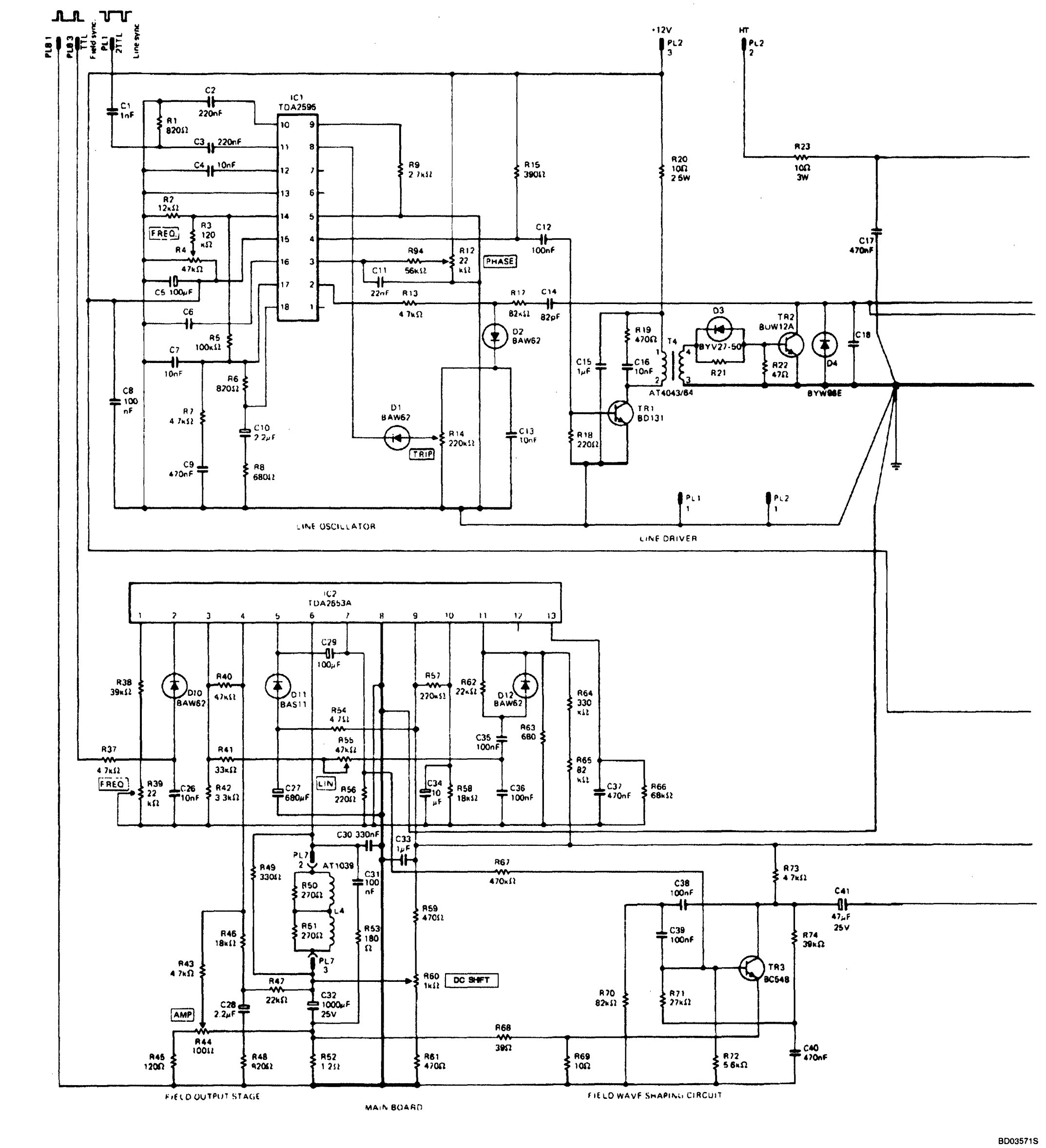


Figure 3. Data and Graphics Display Unit Circuit Diagram

A Versatile High-Resolution Monochrome Data and Graphics Display Unit

AN162

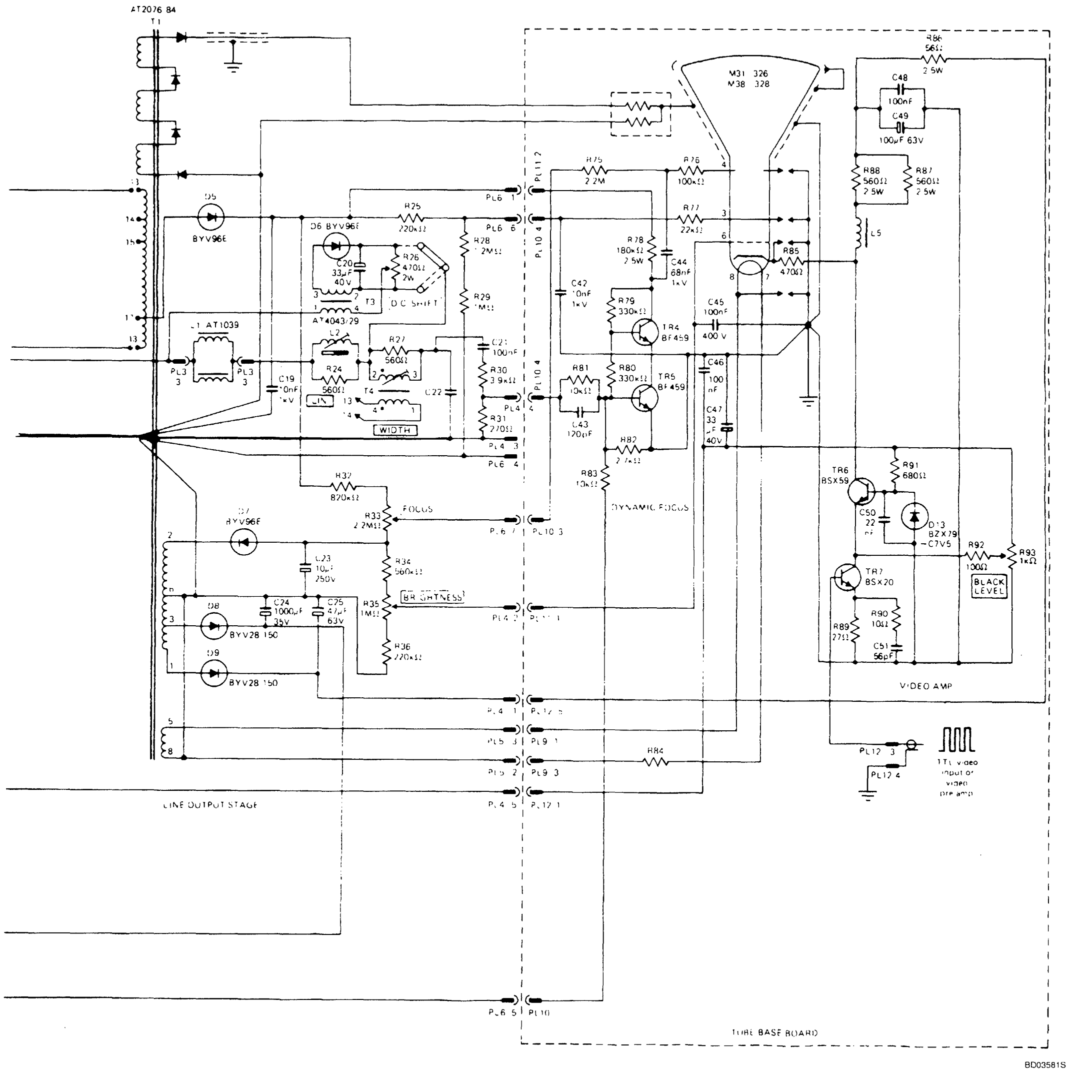


Figure 3. Data and Graphics Display Unit Circuit Diagram (Continued)

Signetics

AN1621
TDA2578A/TDA3651 PCB Layout Directives

Application Note

Linear Products

The TDA2578A is a sync separator and horizontal/vertical synchronization circuit while the TDA3651 is a vertical deflection output driver.

This application note covers general directives for the circuit and PCB layout to achieve stable horizontal time stability and correct vertical interface.

The TDA2578A combines both a horizontal oscillator/PLL and a vertical oscillator/PLL. When used in conjunction with a TDA3651 vertical driver, high system loop gains are involved. This requires careful attention to ground points and consideration to magnetic fields within the receiver/monitor design.

GENERAL PCB LAYOUT DIRECTIVES

- Each IC and discrete component should be surrounded by a good ground plane (See Figure 1).
- The ground plane should not be a complete closed-loop. This is to avoid ground plane-induced currents created by magnetic fields.
- All circuit peripheral components should be connected to the ground plane.
- All high current points should be grounded on another ground plane (double-sided PCB).
- Each IC circuit should have its own common "solid" ground point and should be connected to the other circuitry so that no "strange" ground plane currents are injected.
- Input leads should be short and direct to avoid cross-coupling by both electrostatic and electromagnetic fields.
- A small value resistor in series with input leads can decrease flashover IC failure problems
- Position components with respect to leakage fields of the horizontal line output transformer.

TDA2578A PCB CONSIDERATION

- Grounding point of vertical oscillator timing capacitor (Pin 3 & ground) should be connected to the Pin 9 ground pin, *not* via a PCB trace which carries either large horizontal line currents or video information.
- The vertical feedback voltage input (Pin 2) decoupling capacitor should be connected to the same PCB trace as the vertical oscillator timing capacitor.
- The vertical feedback input (Pin 2) has a very high input impedance; therefore, the scaling resistors should be situated close to Pin 2 to prevent parasitic capacitive horizontal line cross-coupling.
- The vertical integrator capacitor (Pin 4) can carry high peak currents up to 30mA during vertical interval. Therefore it should be firmly grounded to Pin 9, *not,* however, by the same ground PCB trace as used by the vertical oscillator timing capacitor.
- The TDA2578A horizontal output (Pin 11) to drive the base of the horizontal output transistor should be restricted to 30mA peak. This prevents disturbing voltage drops on the TDA2578A ground lead which can result in an offset voltage to the vertical comparator.
- Special attention is required when capacitive coupling is used to drive the horizontal output transistor.
- Vertical interlace is strongly influenced by parasitic signals when coincidence occurs between the vertical oscillator flyback and the horizontal blanking interval. Coincidence is determined by slicing in the vertical integrator and the pre-adjustment of the vertical oscillator.
- Decoupling of the supply voltages (Pins 10 and 16) should be kept as short and direct to the ground pin (Pin 9) as possible. Ripple on the supplies should be less than 1%.

TDA3651 PCB LAYOUT CONSIDERATIONS

- The vertical deflection current loop should be short and be of low impedance, i.e., ample PCB traces on Pin 5 deflection coil, coupling capacitor, and connection to the feedback resistor on Pin 4.
- Damping components and horizontal line suppression across the yoke deflection coil should be located as close as possible to the deflection coil connector.
- Horizontal line information modulated on the vertical waveform at Pin 5 should not exceed $1V_{P-P}$. This is usually caused by:
  1. Inductive & capacitive coupling across the yoke coils.
  2. Capacitive coupling within vertical control loop.
  3. Inductive magnetic coupling.
  4. Supply voltage variations.
- Vertical input (Pin 1) requires a bypass capacitor of 10pF to ground (Pin 2) to suppress the IC current noise.
- Feedback capacitance of 220pF from Pin 1 (input) and Pin 5 (output) improves loop stability.
- Supply voltage decoupling (Pin 9) should be connected directly to ground (Pin 4).
- The supply to both the TDA2578A and the TDA3651 should be decoupled at the source to remove any extraneous noise.

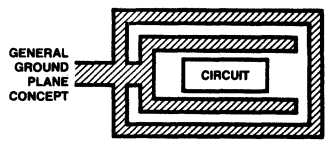


Figure 1. General Ground Plane Concept

Signetics

TDA2579
Synchronization Circuit

Product Specification

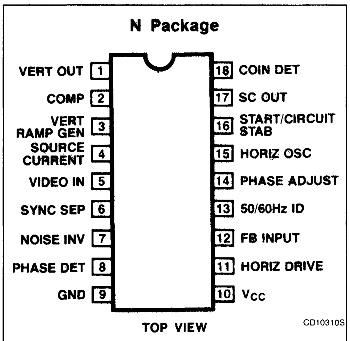
Linear Products

DESCRIPTION
The TDA2579 generates and synchronizes horizontal and vertical signals. The device has a 3-level sandcastle output, a transmitter identification signal and also 50/60Hz identification.

FEATURES
- **Horizontal phase detector, (sync to osc), sync separator and noise inverter**
- **Triple current source in the phase detector with automatic selection**
- **Inhibit of horizontal phase detector and video transmitter identification**
- **Second phase detector for storage compensation of the horizontal output stage**
- **Stabilized direct starting of the horizontal oscillator and output stage**
- **Horizontal output pulse with constant duty cycle value of 29µs**
- **Duty factor of the horizontal output pulse is 50% when horizontal flyback pulse is absent**
- **Internal vertical sync separator and two integration selection times**
- **Divider system with three different reset enable windows**
- **Synchronization is set to 628 divider ratio when no vertical sync pulses and no video transmitter is identified**
- **Vertical comparator with a low DC feedback signal**
- **50/60Hz identification output combined with mute function**
- **Automatic amplitude adjustment for 50 and 60Hz and blanking pulse duration**

APPLICATIONS
- **Video terminals**
- **Television**
- **Video tape recorder**

PIN CONFIGURATION



ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
16-Pin Plastic DIP (SOT-102HE)	0 to +70°C	TDA2579N

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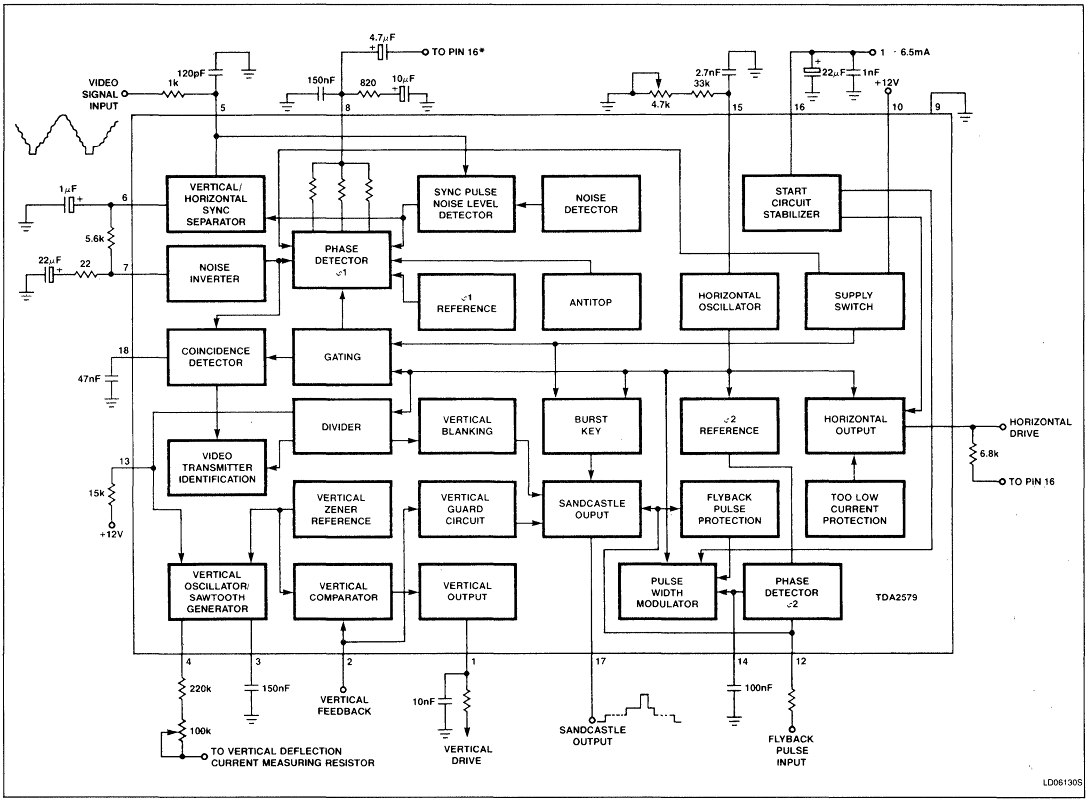
Synchronization Circuit TDA2579

BLOCK DIAGRAM



ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
I_{16}	Start current	10	mA
V_{10}	Supply voltage	13.2	V
P_{TOT}	Power dissipation	1.2	W
T_{STG}	Storage temperature	-65 to +150	°C
T_A	Operating ambient temperature	-25 to +65	°C
θ_{JA}	Thermal resistance from junction to ambient in free air	50	°C/W

Synchronization Circuit | TDA2579

DC AND AC ELECTRICAL CHARACTERISTICS

$T_A = 25°C$; $I_{16} = 6.5mA$; $V_{10} = 12V$, unless otherwise specified. Voltage measurements are taken with respect to Pin 9 (ground).

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supply					
I_{16}	Supply current, Pin 16 $V_{10} = 0V$	6.5		10	mA
I_{16}	Supply current, Pin 16 $V_{10} = 9.5V$	2.5		10	mA
V_{16}	Stabilized voltage, Pin 16	8.1	8.7	9.3	V
I_{10}	Current consumption, Pin 10		68	85	mA
V_{CC}	Supply voltage range, Pin 10	9.5	12	13.2	V
Video input (Pin 5)					
V_5	Top sync. level	1.5	3.1	3.75	V
V_5	Sync. pulse amplitude[1]	0.1	0.6	1	V_{CC}
	Slicing level[2]	35	50	65	%
	Delay between video input and det. output (see also Figure 2)	0.2	0.3	0.5	μs
	Sync. pulse noise level detector circuit active		600		mV_{TT}
Sync. Pulse					
	Noise level detector circuit hysteresis		3		dB
Noise gate (Pin 5)					
V_5	Switching level		+0.7	+1	V
First control loop (Pin 8) (Horizontal osc. to sync.)					
Δf	Holding range		±800		Hz
Δf	Catching range	±600	±800	±1100	Hz
	Control sensitivity video with respect to burstkey and flyback pulse				
	Slow time constant		2.5		kHz/μs
	Normal time constant		10		kHz/μs
	Fast time constant		5		kHz/μs
	Phase modulation due to hum on the supply line Pin 10[3]		0.2		$\mu s/V_{TT}$
	Phase modulation due to hum on input current Pin 16[3]		0.08		$\mu s/mA_{TT}$
Second control loop (Pin 14) (Horizontal flyback to horizontal oscillator)					
$\Delta t_d/\Delta t_o$	Control sensitivity $t_D = 10\mu s$	200	300	600	μs
t_D	Control range	1		> 45	μs
t_D	Control range for constant duty cycle horizontal output	1	29 (–t flyback pulse)		
	Controlled edge of horizontal output signal Pin 11		positive		
Phase adjustment (Pin 14) (via second control loop)					
	Control sensitivity $t_D = 10\mu s$		25		μA/μs
I_{14}	Maximum allowed control current			±60	μA

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DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $T_A = 25°C$; $I_{16} = 6.5mA$; $V_{10} = 12V$, unless otherwise specified. Voltage measurements are taken with respect to Pin 9 (ground).

SYMBOL	PARAMETER	LIMITS Min	Typ	Max	UNIT
Horizontal oscillator (Pin 15) (C = 2.7nF; $R_{OSC} = 33k\Omega$					
f	Frequency (no sync.)		15625		Hz
Δf	Spread (fixed external component, no sync.)			±4	%
Δf	Frequency deviation between starting point output signal and stabilized condition		+5	+8	%
TC	Temperature coefficient		10		°C
Horizontal output (Pin 11) (Open-collector)					
V_{11}	Output voltage high			13.2	V
V_{11}	Start voltage protection (internal zener diode)	13		15.8	V
I_{16}	Low input current Pin 16 protection output enabled		5.5	6.5	mA
V_{11}	Output voltage low start condition ($I_{11} = 10mA$)		0.1	0.5	V
	Duty cycle output current during starting $I_{16} = 6.5mA$	55	65	75	%
V_{11}	Output voltage low normal condition ($I_{11} = 25mA$)		0.3	0.5	V
	Duty cycle output current without flyback pulse Pin 12	45	50	55	%
	Duration of the output pulse high $t_D = 8\mu s$	27	29	31	μs
	Controlled edge		positive		
	Temperature coefficient horizontal output pulse		−0.05		μs/°C
Sandcastle output signal (Pin 17) ($I_{LOAD} = 1mA$)					
 V_{17} V_{17} V_{17}	Output voltage during: burstkey horizontal blanking vertical blanking	 9.75 4.1 2	 10.6 4.5 2.5	 4.9 3	 V V V
V_{17}	Zero level output voltage $I_{SINK} = 0.5mA$			0.7	V
 t_P V_{12}	Pulse width: burstkey horizontal blanking	 3.45	 3.75 1	 4.1	 μs V
	Phase position burstkey Time between middle synchronization pulse at Pin 5 and start burst at Pin 17	2.3	2.7	3.1	μs
	Time between start sync. pulse and end of burst pulse, Pin 17			9.2	μs

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) T_A = 25°C; I_{16} = 6.5mA; V_{10} = 12V, unless otherwise specified. Voltage measurements are taken with respect to Pin 9 (ground).

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Coincidence detector, video transmitter identification circuit and time constant switching levels (see also Figure 1)					
I_{18}	Detector output current		0.25		mA
V_{18}	Voltage level for in sync. condition (φ1 normal)		6.5		V
V_{18}	Voltage for noisy sync. pulse (φ1 slow and gated)	9	10		V
V_{18}	Voltage level for noise only[5]		0.3		V
V_{18}	Switching level normal-to-fast	3.2	3.5	3.8	V
V_{18}	Switching level Mute output active and fast-to-slow	1.0	1.2	1.4	V
V_{18}	Switching level frame period counter (3 periods fast)	0.08	0.12	0.16	V
V_{18}	Switching level Slow-to-fast (locking) Mute output inactive	1.5	1.7	1.9	V
V_{18}	Switching level fast-to-normal (locking)	4.7	5.0	5.3	V
V_{18}	Switching level normal-to-slow (gated sync. pulse)	7.4	7.8	8.2	V
Video transmitter identification output (Pin 13)					
V_{13}	Output voltage active (no sync., I_{13} = 2mA)		0.15	0.32	V
I_{13}	Sink current active (no sync.), $V_{13} < 1V$			5	mA
I_{13}	Output current inactive (sync. 50Hz)			1	μA
50/60Hz identification (Pin 13) (R_{13} positive supply 15kΩ)					
	Emitter-follower, PNP				
V_{13}	60Hz: $\frac{2 \times fH}{fV}$ < 576 voltage	7.2	7.65	8.1	V
V_{13}	50Hz: $\frac{2 \times fH}{fV}$ > 576 voltage		V_{10}		V
Flyback input pulse (Pin 12)					
V_{12}	Switching level		+1		V
I_{12}	Input current	+0.2		+4	mA
V_{12}	Input pulse			12	V_{CC}
R_{IN}	Input resistance		3		kΩ
	Phase position without shift				
t_D	Time between the middle of the sync. pulse at Pin 5 and the middle of the horizontal blanking pulse of Pin 17		2.5		μs
Vertical ramp generator (Pin 3)					
	Pulse width charge current		26		clock pulses
I_3	Charge current		3		mA
	Top level ramp signal voltage				
V_3	Divider in 50Hz mode[6]	5.1	5.5	5.9	V
V_3	Divider in 60Hz mode[6]	4.35	4.7	5.05	V
	Ramp amplitude C_3 = 150nF, R_4 = 330kΩ, 50Hz[6] R_4 = 330kΩ, 60Hz[6]		 3.1 2.5		 V_{CC} V_{CC}

Synchronization Circuit | TDA2579

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $T_A = 25°C$; $I_{16} = 6.5mA$; $V_{10} = 12V$, unless otherwise specified. Voltage measurements are taken with respect to Pin 9 (ground).

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Current source (Pin 4)					
$V_{4.9}$	Output voltage $I_4 = 20\mu A$	6.6	7.1	7.6	V
I_4	Allowed current range	10		55	μA
	Temperature coefficient output voltage				
TC	$I_4 = 20\mu A$		+50		$10^{-6}/°C$
TC	$I_4 = 40\mu A$		+20		$10^{-6}/°C$
TC	$I_4 = 50\mu A$		-40		$10^{6}/°C$
Comparator (Pin 2) $C_3 = 150nF$; $R_4 = 330k\Omega$					
	Input voltage				
V_{2-9}	DC level[6]	0.9	1	1.1	V
V_{2-9}	AC level		0.8		V_{CC}
	Deviation amplitude 50/60Hz			2.5	%
	Vertical output stage, Pin 1 (NPN) emitter follower				
V_{1-9}	Output voltage I_O Pin 1 = +1.5mA	4.8	5.2	5.6	V
R_S	Sync. separator resistor		160		Ω
	Continuous sink current		0.25		mA
Vertical guard circuit (Pin 2) Active ($V_{17} = 2.5V$)					
V_2	Switching level low[6]	> 1.7	1.9	2.1	V
V_2	Switching level high[6]	< 0.3	0.4	0.5	V

NOTES:

1. Up to $1V_{P-P}$ the slicing level is constant, at amplitudes exceeding $1V_{P-P}$ the slicing level will increase.
2. The slicing level is fixed by the formula:

$$P = \frac{R_S}{5.3 + R_S} \times 100\% \quad (R_S \text{ value in k}\Omega)$$

3. Measured between Pin 5 and sandcastle output Pin 17.
4. Divider in search (large) mode:
   start: reset divider = start vertical sync. plus 1 clock pulse
   stop:
   $$n = \frac{2 \times fH}{fV} > 576 \text{ clock pulse } 42$$
   $$n = \frac{2 \times fH}{fV} < 576 \text{ clock pulse } 34$$
   Divider in small window mode:
   start: clock pulse 517 (60Hz) clock pulse 619 (50Hz)
   stop: clock pulse 34 (60Hz) clock pulse 42 (50Hz)
5. Depends on DC level of Pin 5, given value is valid for $V_5 \approx 5V$.
6. Value related to internal zener diode reference voltage source spread includes the complete spread of reference voltage.

FUNCTIONAL DESCRIPTION

Vertical Part (Pins 1, 2, 3, 4)

The IC embodies a synchronized divider system for generating the vertical sawtooth at Pin 3. The divider system has an internal frequency doubling circuit, so the horizontal oscillator is working at its normal line frequency and one line period equals 2 clock pulses. Due to the divider system, no vertical frequency adjustment is needed. The divider has a discriminator window for automatically switching over from the 60Hz to 50Hz system. The divider system operates with 3 different divider reset windows for maximum interference/disturbance protection.

The windows are activated via an up/down counter. The counter increases its counter value by 1 for each time the separated vertical sync. pulse is within the searched window. The count is reduced by 1 when the vertical sync. pulse is not present.

Large (Search) Window: Divider Ratio Between 488 and 722

This mode is valid for the following conditions:

1. Divider is looking for a new transmitter.
2. Divider ratio found, not within the narrow window limits.
3. Non-standard TV-signal condition detected while a double or enlarged vertical sync. pulse is still found after the internally-generated antitop flutter pulse has ended. This means a vertical sync. pulse width larger than 8 clock pulses (50Hz), that is, 10 clock pulses (60Hz). In general this mode is activated for video tape recorders operating in the feature/trick mode.
4. Up/down counter value of the divider system operating in the narrow window mode drops below count 1.
5. Externally setting. This can be reached by loading Pin 18 with a resistor of 180kΩ to earth or connecting a 3.6V diode stabistor between Pin 18 and ground.

Narrow Window: Divider Ratio Between 522 – 528 (60Hz) or 622 – 628 (50Hz).

The divider system switches over to this mode when the up/down counter has reached its maximum value of 12 approved vertical sync. pulses. When the divider operates in this mode and a vertical sync. pulse is missing within the window, the divider is reset at the end of the window and the counter value is lowered by 1. At a counter value below count 1 the divider system switches over to the large window mode.

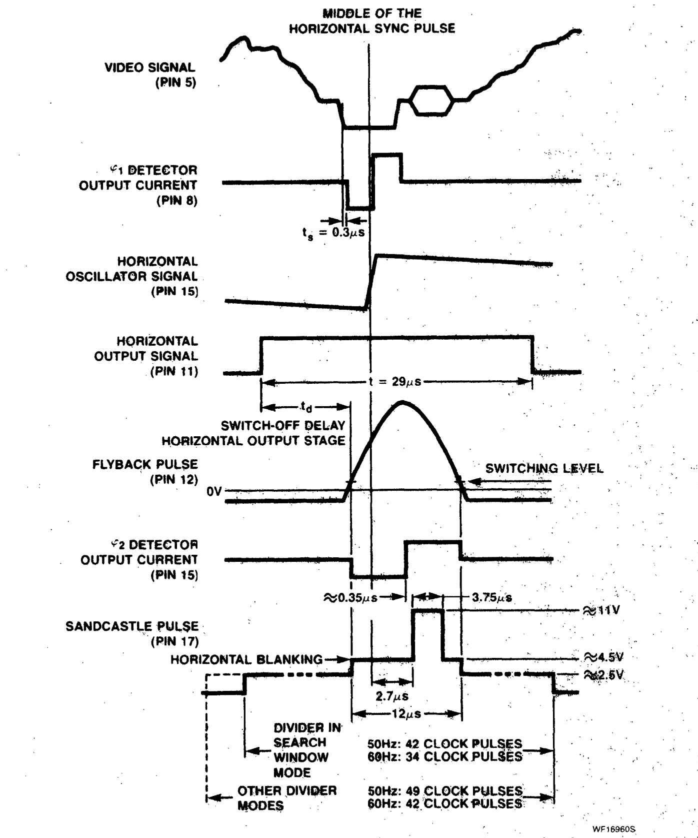


Figure 1. Timing Diagram of the TDA2579

Standard TV Norm

When the up/down counter has reached its maximum value of 12 in the narrow window mode, the information applied to the up/down counter is changed such that the standard divider ratio value is tested. When the counter has reached a value of 14, the divider system is changed over to the standard divider ratio mode. In this mode the divider is always reset at the standard value even if the vertical sync. pulse is missing. A missed vertical sync. pulse lowers the counter value by 1. When the counter reaches the value of 10, the divider system is switched over to the large window mode. The standard TV norm condition gives maximum protection for video recorders playing tapes with anti-copy guards.

No TV Transmitter Found: (Pin 18 < 1.2V)

In this condition, only noise is present, the divider is reset to count 628. In this way a stable picture display at normal height is achieved.

Video Tape Recorders in Feature Mode

It should be noted that some VTRs operating in the feature modes, such as picture search, generate such distorted pictures that the no TV transmitter detection circuit can be activated as Pin V_{18} drops below 1.2V. This would imply a rollowing picture (condition d). In general, VTR machines use a reinserted vertical sync. pulse in the feature mode. Therefore, the divider system has been made such that the automatic reset of the divider at count 628 when V_{18} is below 1.2V is inhibited when a vertical sync. pulse is detected.

The divider system also generates the antitop flutter pulse which inhibits the phase 1 detector during the vertical sync. pulse. The width of this pulse depends on the divider mode. For the divider mode a, the start is generated at the reset of the divider. In modes b and c, the anti-top flutter pulse starts at the beginning of the first equalizing pulse. The anti-top flutter pulse ends at count 8 for 50Hz and count 10 for 60Hz. The vertical blanking pulse is also generated via the divider system. The start is at the reset of the divider while the blanking pulse ends at count 34 (17 lines for 60Hz, and at count 42 (21 lines) for 50Hz systems. The vertical blanking pulse generated at the sandcastle output Pin 17 is made by adding the anti-top flutter pulse and the blank pulse. In this way the vertical blanking pulse starts at the beginning of the first equalizing pulse when the divider operates in the b or c mode. For generating a vertical linear sawtooth voltage a capacitor should be connected to Pin 3. The recommended value is 150nF to 330nF (see Block Diagram).

The capacitor is charged via an internal current source starting at the reset of the divider system. The voltage on the capacitor is monitored by a comparator which is activated also at reset. When the capacitor has reached a voltage value of 5.5V for the 50Hz system or 4.7V for the 60Hz system the voltage is kept constant until the charging period ends. The charge period width is 26 clock pulses. At clock pulse 26 the comparator is switched off and the capacitor is discharged by an NPN transistor current source, the value of which can be set by an external resistor between Pin 4 and ground (Pin 9). Pin 4 is connected to a PNP transistor current source which determines the current of the NPN current source. The PNP current source on Pin 4 is connected to an internal zener diode reference voltage which has a typical voltage of ≈ 7.1V. The recommended operating current range is 10 to 50μA. The resistance at pin R_4 should be 140 to 700kΩ. By using a double current mirror concept the vertical sawtooth pre-correction can be set on the desired value by means of external components between Pin 4 and Pin 3, or by connecting the Pin 4 resistor to the vertical current measuring resistor of the vertical output stage. The vertical amplitude is set by the current of Pin 4. The vertical feedback voltage of the output stage has to be applied to Pin 2. For the normal amplitude adjustment the values are DC = 1V and AC = 0.8V. Due to the automatic system adaption both values are valid for 50Hz and 60Hz.

The low DC-voltage value improves the picture bounce behaviour as less parabola compensation is necessary. Even a fully DC-coupled feedback circuit is possible.

Vertical Guard

The IC also contains a vertical guard circuit. This circuit monitors the vertical feedback signal on Pin 2. When the level on Pin 2 is below 0.4V or higher than 1.9V, the guard circuit inserts a continuous level of 2.5V in the sandcastle output signal of Pin 17. This results in the blanking of the picture displayed, thus preventing a burnt-in horizontal line. The guard levels specified refer to the zener diode reference voltage source level.

Driver Output

The driver output is at Pin 1, it can deliver a drive current of 1.5mA at 5V output. The internal impedance is about 150Ω. The output pin is also connected to an internal current source with a sinking current of 0.25mA.

Sync. Separator, Phase Detector and TV Station Identification, (Pins 5, 6, 7, 8, and 18)

The video input signal is connected to Pin 5. The sync. separator is designed such that the slicing level is independent of the amplitude of the sync. pulse. The black level is measured and stored in the capacitor at Pin 7. The slicing level value is stored in the capacitor at Pin 6. The slicing level value can be chosen by the value of the external resistor between Pins 6 and 7. The value is given by the formula:

$$P = \frac{R_S \times 100}{5.3 + R_S} \quad (R_S \text{ value in } k\Omega)$$

Where R_S is the resistor between Pins 6 and 7 and top sync. level equals 100%. The recommended resistor value is 5.6kΩ.

Black Level Detector

A gating signal is used for the black level detector. This signal is composed of an internal horizontal reference pulse with a duty cycle of 50% and the flyback pulse at Pin 12. In this way the TV transmitter identification operates also for all DC conditions at input Pin 5 (no video modulation, plain carrier only).

During the frame interval the slicing level detector is inhibited by a signal which starts with the anti-top flutter pulse and ends with the reset vertical divider circuit. In this way shift of the slicing level due to the vertical sync. signal is reduced and separation of the vertical sync. pulse is improved.

Noise Inverter

An internal noise inverter is activated when the video level at Pin 5 drops below 0.7V. The IC embodies also a built-in sync. pulse noise level detection circuit. This circuit is directly connected to Pin 5 and measures the noise level at the middle of the horizontal sync. pulse. When a noise level of $600mV_{P-P}$ is detected, a counter circuit is activated. A video input signal is processed as "acceptable noise-free" when 12 out of 16 sync. pulses have a noise level below 600mV for two succeeding frame periods. The sync. pulses are processed during a 16 line width gating period generated by the divider system. The measuring circuit has a built-in noise level hysteresis of about 150mV (≈ 3dB).

When the "acceptable noise-free" condition is found, the phase detector of Pin 8 is switched to not-gated and normal time constant. When a higher sync. pulse noise level is found, the phase detector is switched over to slow time constant and gated sync. pulse phase detection. At the same time the integration time of the vertical sync. pulse separator is adapted.

Phase Detector

The phase detector circuit is connected to Pin 8. This circuit consists of 3 separate phase detectors which are activated depending on the voltage of Pin 18 and the state of the sync. pulse noise detection circuit.

All three phase detectors are activated during the vertical blanking period, this with the exception of the anti-top flutter pulse period, and the separated vertical sync. pulse time.

As a result, phase jumps in the video signal related to video head takeover of video recorders are quickly restored within the vertical blanking period. At the end of the blanking period, the phase detector time constant is lowered by 2.5 times. In this way no need for external VTR time constant switching exists, so all station numbers are suitable for signals from VTR, video games or home computers.

For quick locking of a new TV station starting from a noise-only signal condition (normal time constant), a special circuit is incorporated. A new TV station which is not locked to the horizontal oscillator will result in a voltage drop below 0.1V at Pin 18. This will activate a frame period counter which switches the phase detector to fast for 3 frame periods.

Horizontal Oscillator

The horizontal oscillator will now lock to the new TV station and as a result, the voltage on Pin 18 will increase to about 6.5V. When Pin 18 reaches a level of 1.8V the mute output transistor of Pin 13 is switched off and the divider is set to the large window. In general the mute signal is switched off within 5ms (pin C_{18} = 47nF) after reception of a new TV signal. When the voltage on Pin 18 reaches a level of 5V, usually within 15ms, the frame counter is switched off and the time constant is switched from fast to normal.

If the new TV station is weak, the sync. noise detector is activated. This will result in a changeover of Pin 18 voltage from 7V to ≈ 10V. When Pin 18 exceeds the level of 7.8V the phase detector is switched to slow time constant and gated sync. pulse condition.

When desired, most conditions of the phase detector can also be set by external means in the following way:

a. Fast time constant TV transmitter identification circuit not active, connect Pin 18 to earth (Pin 9).

b. Fast time constant TV transmitter identification circuit active, connect a resistor of 180kΩ between Pin 18 and ground. This condition can also be set by using a 3.6V stabistor diode instead of a resistor.

c. Slow time constant, (with exception of frame blanking period), connect Pin 18 via a resistor of 10kΩ to +12V, Pin 10. In this condition the transmitter identification circuit is not active.

d. No switching to slow time constant desired (transmitter identification circuit active), connect a 6.8V zener diode between Pin 18 and ground.

Figure 2 illustrates the operation of the 3 phase detector circuits.

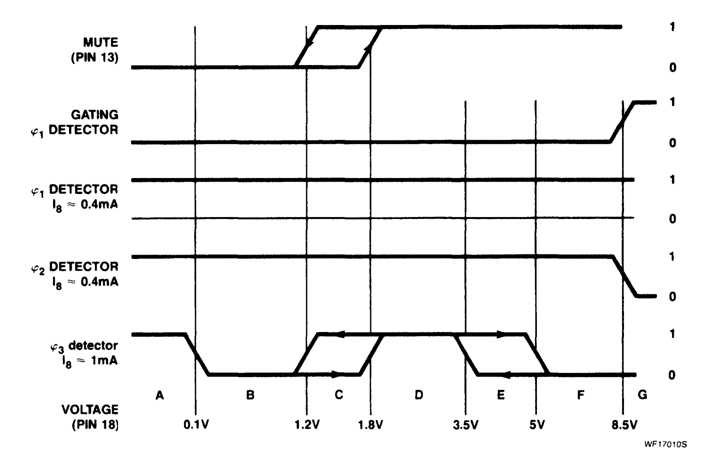


Figure 2. Timing Diagram, Phase Detectors.

Supply (Pins 9, 10 and 16)

The IC has been designed such that the horizontal oscillator and output stage can start operating by application of a very low supply current into Pin 16.

The horizontal oscillator starts at a supply current of about 4.5mA. The horizontal output stage is forced into the non-conducting stage until the supply current has a typical value of 5.5mA. The circuit has been designed so that after starting the horizontal output function a current drop of ≈ 1mA is allowed. The starting circuit gives the possibility to derive the main supply (Pin 10), from the horizontal output stage. The horizontal output signal can also be used as the oscillator signal for synchronized switch-mode power supplies. The maximum allowed starting current is 10mA. The main supply should be connected to Pin 10, and Pin 9 should be used as ground. When the voltage on Pin 10 increases from zero to its final value (typically 12V) a part of the supply current of the starting circuit is taken from Pin 10 via internal diodes, and the voltage on Pin 16 will stabilize to a typical value of 8.7V.

In stabilized condition (Pin $V_{10} > 9.5V$) the minimum required supply current to Pin 16 is ≈ 2.5mA. All other IC functions are switched on via the main supply voltage on Pin 10. When the voltage on Pin 10 reaches a value of ≈ 7V the horizontal phase detector circuit is activated and the vertical ramp on Pin 3 is started. The second phase detector circuit and burst pulse circuit are started when the voltage on Pin 10 reaches the stabilized voltage value of Pin 16 which is typically 8.7V.

For closing the second phase detector loop, a flyback pulse must be applied to Pin 12. When no flyback is detected, the duty cycle of the horizontal output stage is 50%.

For remote switch-off Pin 16 can be connected to ground (via an NPN transistor with a series resistor of ≈ 500Ω) which switches off the horizontal output.

Horizontal Oscillator, Horizontal Output Transistor, and Second Phase Detector (Pins 11, 12, 14 and 15)

The horizontal oscillator is connected to Pin 15. The frequency is set by an external RC combination between Pin 15 and ground, Pin 9. The open collector horizontal output stage is connected to Pin 11. An internal zener diode configuration limits the open voltage of Pin 11 to ≈ 14.5V.

The horizontal output transistor at Pin 11 is blocked until the current into Pin 16 reaches a value of ≈ 5.5mA.

A higher current results in a horizontal output signal at Pin 11, which starts with a duty cycle of ≈ 35% HIGH.

The duty cycle is set by an internal current source-loaded NPN emitter-follower stage connected to Pin 14 during starting. When Pin 16 changes over to voltage stabilization, the NPN emitter-follower and current source load at Pin 14 are switched off and the second phase detector circuit is activated, provided a horizontal flyback pulse is present at Pin 12. When no flyback pulse is detected at Pin 12 the duty cycle of the horizontal output stage is set to 50%.

The phase detector circuit at Pin 14 compensates for storage time in the horizontal deflection output stage. The horizontal output pulse duration in 29μs HIGH for storage times between 1μs and 17μs (29μs flyback pulse of 12μs). A higher storage time increases the HIGH time. Horizontal picture shift is possible by forcing an external charge or discharge current into the capacitor of Pin 14.

Mute Output and 50/60Hz Identification (Pin 13)

The collector of an NPN transistor is connected to Pin 13. When the voltage on Pin 18 drops below 1.2V (no TV transmitter) the NPN transistor is switched ON.

When the voltage on Pin 18 increases to a level of ≈ 1.8V (new TV transmitter found) the NPN transistor is switched OFF.

Pin 13 has also the possibility for 50/60Hz identification. This function is available when Pin 13 is connected to Pin 10 (+12V) via an external pull-up resistor of 10 – 20kΩ. When no TV transmitter is identified, the voltage on Pin 13 will be LOW (< 0.5V). When a TV transmitter with a divider ratio > 576 (50Hz) is detected the output voltage of Pin 13 is HIGH (+12).

When a TV transmitter with a divider ratio < 576 (60Hz) is found an internal PNP transistor with its emitter connected to Pin 13 will force this pin output voltage down to ≈ 7.5V.

Sandcastle Output (Pin 17)

The sandcastle output pulse generated at Pin 17, has three different voltage levels. The highest level, (11V), can be used for burst gating and black level clamping. The second level, (4.5V), is obtained from the horizontal flyback pulse at Pin 12, and is used for horizontal blanking. The third level, (2.5V), is used for vertical blanking and is derived via the vertical divider system. For 50Hz the blanking pulse duration is 42 clock pulses and for 60Hz it is 34 clock pulses started from the vertical divider reset. For TV signals which have a divider ratio between 622 and 628 or 522 and 528 the blanking pulse is started at the first equalizing pulse.

TYPICAL APPLICATION

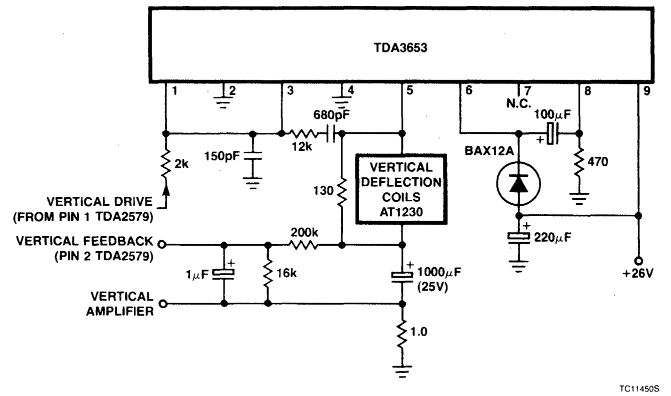


Figure 3

Signetics

TDA2593
Horizontal Combination

Product Specification

Linear Products

DESCRIPTION

The TDA2593 is a monolithic integrated circuit intended for use in color television receivers in combination with TDA2510, TDA2520, TDA2560 as well as with TDA3505, TDA3510, and TDA3520.

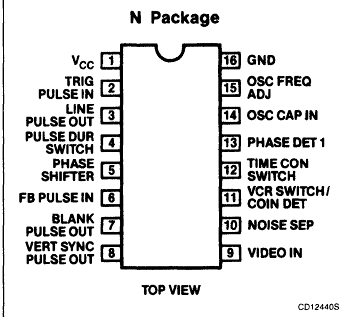
FEATURES

- **Horizontal oscillator based on the threshold switching principle**
- **Phase comparison between sync pulse and oscillator voltage (φ_1)**
- **Internal key pulse for phase detector (φ_1) (additional noise limiting)**
- **Phase comparison between line flyback pulse and oscillator voltage (φ_2)**
- **Larger catching range obtained by coincidence detector (φ_3; between sync and key pulse)**
- **Switch for changing the filter characteristic and the gate circuit (VCR operation)**
- **Sync separator**
- **Noise separator**
- **Vertical sync separator and output stage**
- **Color burst keying and line flyback blanking pulse generator**
- **Phase shifter for the output pulse**
- **Output pulse duration switching**
- **Output stage with separate supply voltage for direct drive of thyristor deflection circuits**
- **Low supply voltage protection**

APPLICATIONS

- **Video monitors**
- **TV receivers**

PIN CONFIGURATION



ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
16-Pin Plastic DIP (SOT-38)	-20°C to +70°C	TDA2593N

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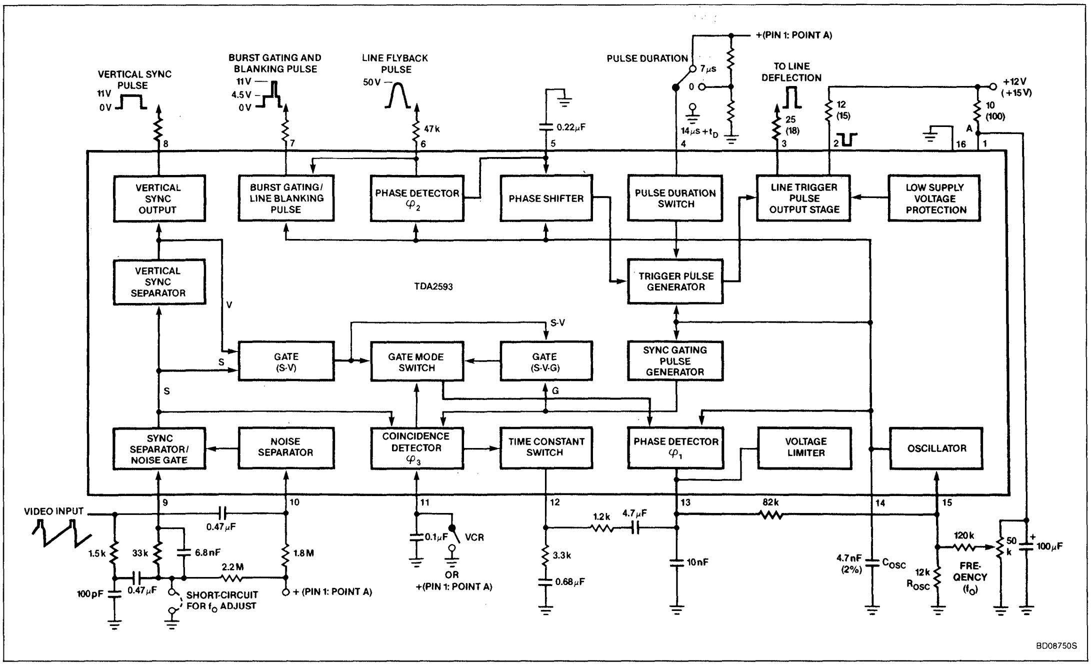
Horizontal Combination | TDA2593

BLOCK DIAGRAM



ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
	Supply voltage		
V_{1-16}	at Pin 1 (voltage source)	13.2	V
V_{2-16}	at Pin 2	18	V
	Voltages		
V_{4-16}	Pin 4	13.2	V
$\pm V_{9-16}$	Pin 9	6	V
$\pm V_{10-16}$	Pin 10	6	V
V_{11-16}	Pin 11	13.2	V
	Currents		
$I_{2M}, -I_{3M}$	Pins 2 and 3 (thyristor driving) (peak value)	650	mA
$I_{2M}, -I_{3M}$	Pins 2 and 3 (transistor driving) (peak value)	400	mA
I_4	Pin 4	1	mA
$\pm I_6$	Pin 6	10	mA
$-I_7$	Pin 7	10	mA
I_{11}	Pin 11	2	mA
P_{TOT}	Total power dissipation	800	mW
T_{STG}	Storage temperature range	-25 to +125	°C
T_A	Operating ambient temperature range	-20 to +70	°C

DC AND AC ELECTRICAL CHARACTERISTICS at $V_{CC} = 12V$; $T_A = 25°C$; measured in Block Diagram.

SYMBOL	PARAMETER	LIMITS Min	Typ	Max	UNIT
Sync separator					
V_{9-16}	Input switching voltage		0.8		V
I_9	Input keying current	5		100	μA
I_9	Input leakage current at $V_{9-16} = -5V$			1	μA
I_9	Input switching current			5	μA
I_9	Switch off current	100	150		μA
$V_{9-16(P-P)}$	Input signal (peak-to-peak value)	3		4	V^1
Noise separator					
V_{10-16}	Input switching voltage		1.4		V
I_{10}	Input keying current	5		100	μA
I_{10}	Input switching current	100	150		μA
I_{10}	Input leakage current at $V_{10-16} = -5V$			1	μA
$V_{10-16(P-P)}$	Input signal (peak-to-peak value)	3		4	V^1
$V_{10-16(P-P)}$	Permissible superimposed noise signal (peak-to-peak value)			7	V
Line flyback pulse					
I_6	Input current	0.02	1	2	mA
V_{6-16}	Input switching voltage		1.4		V
V_{6-16}	Input limiting voltage	−0.7		+1.4	V
Switching on VCR					
V_{11-16} V_{11-16}	Input voltage		0 to 2.5 9 to V_{1-16}		V V
$-I_{11}$ I_{11}	Input current			200 2	μA mA
Pulse duration switch for $t = 7\mu s$ (thyristor driving)					
V_{4-16}	Input voltage		9.4 to V_{1-16}		V
I_4	Input current	200			μA
Pulse duration switch for $t = 14\mu s + t_D$ (transistor driving)					
V_{4-16}	Input voltage	0		3.5	V
$-I_4$	Input current	200			μA
Pulse duration switch for $t = 0$; $V_{3-16} = 0$ or input Pin 4 open					
V_{4-16}	Input voltage	5.4		6.6	V
I_4	Input current		0	0	μA
Vertical sync pulse (positive-going)					
$V_{8-16(P-P)}$	Output voltage (peak-to-peak value)	10	11		V
R_8	Output resistance		2		$k\Omega$
t_{ON}	Delay between leading edge of input and output signal		15		μs
t_{OFF}	Delay between trailing edge of input and output signal		t_{on}		μs

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) at V_{CC} = 12V; T_A = 25°C; measured in Block Diagram.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Burst gating pulse (positive-going)					
$V_{7-16(P-P)}$	Output voltage (peak-to-peak value)	10	11		V
R_7	Output resistance		70		Ω
t_P	Pulse duration; V_{7-16} = 7V	 3.7	4 4.3		μs μs
t	Phase relation between middle of sync pulse at the input and the leading edge of the burst gating pulse; V_{7-16} = 7V	2.15	2.65	3.15	μs
I_7	Output trailing edge current		2		mA
Line flyback-blanking pulse (positive-going)					
$V_{7-16(P-P)}$	Output voltage (peak-to-peak value)	4	5		V
R_7	Output resistance		70		Ω
I_7	Output trailing edge current		2		mA
Line drive pulse (positive-going)					
$V_{3-16(P-P)}$	Output voltage (peak-to-peak value)		10.5		V
 R_3 R_3	Output resistance for leading edge of line pulse for trailing edge of line pulse		 2.5 20		 Ω Ω
t_P	Pulse duration (thyristor driving) V_{4-16} = 9.4 to V_{1-16} V	5.5	7	8.5	μs
t_P	Pulse duration (transistor driving) V_{4-16} = 0 to 4V; t_{FP} = 12μs		14 + t_D		μs[2]
V_{1-16}	Supply voltage for switching off the output pulse		4		V
Overall phase relation					
t	Phase relation between middle of sync pulse and the middle of the flyback pulse		2.6		μs[3]
$\|\Delta t\|$	Tolerance of phase relation			0.7	μs
$\Delta I_5/\Delta t$	The adjustment of the overall phase relation and consequently the leading edge of the line drive occurs automatically by phase control φ_2. If additional adjustment is applied it can be arranged by current supply at Pin 5		30		μA/μs
Oscillator					
V_{14-16}	Threshold voltage low level		4.4		V
V_{14-16}	Threshold voltage high level		7.6		V
$\pm I_{14}$	Discharge current		0.47		mA
f_O	Frequency; free running (C_{OSC} = 4.7nF; R_{OSC} = 12kΩ)		15.625		kHz
$\Delta f_O/f_O$	Spread of frequency		< ±5		%[4]
$\Delta f_O/\Delta I_{15}$	Frequency control sensitivity		31		Hz/μA
$\Delta f_O/f_O$	Adjustment range of network in circuit (see Block Diagram)		±10		%
$\frac{\Delta f_O/f_O}{\Delta V/V_{NOM}}$	Influence of supply voltage on frequency		< ±0.05		%[4]
Δf_O	Change of frequency when V_{1-16} drops to 5V		< ±10		%[4]
	Temperature coefficient of oscillator frequency		< ±10^{-4}		Hz/°C[4]

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) at $V_{CC} = 12V$; $T_A = 25°C$; measured in Block Diagram.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Phase comparison φ_1					
V_{13-16}	Control voltage range	3.8	8.2		V
$\pm I_{13M}$	Control current (peak value)	1.9	2.3		mA
I_{13}	Output leakage current at $V_{13-16} = 4$ to 8V			1	μA
 R_{13} R_{13}	Output resistance at $V_{13-16} = 4$ to 8V[5] at $V_{13-16} < 3.8V$ or $> 8.2V$[6]		 high ohmic low ohmic		
	Control sensitivity		2		kHz/μs
Δf	Catching and holding range (82kΩ between Pins 13 and 15)		± 780		Hz
$\Delta(\Delta f)$	Spread of catching and holding range		± 10		%[4]
Phase comparison φ_2 and phase shifter					
V_{5-16}	Control voltage range	5.4		7.6	V
$\pm I_{5M}$	Control current (peak value)		1		mA
  R_5	Output resistance at $V_{5-16} = 5.4$ to 7.6V[7] at $V_{5-16} < 5.4$ or $> 7.6V$		 high ohmic 8		 kΩ
I_5	Input leakage current $V_{5-16} = 5.4$ to 7.6V			5	μA
t_D	Permissible delay between leading edge of output pulse and leading edge of flyback pulse ($t_{FP} = 12\mu s$)			15	μs
$\Delta t/\Delta t_D$	Static control error			0.2	%
Coincidence detector φ_3					
V_{11-16}	Output voltage	0.5		6	V
 I_{11M} $-I_{11M}$	Output current (peak value) without coincidence with coincidence		 0.1 0.5		 mA mA
Time constant switch					
V_{12-16}	Output voltage		6		V
$\pm I_{12}$	Output current (limited)			1	mA
 R_{12} R_{12}	Output resistance at $V_{11-16} = 2.5$ to 7V at $V_{11-16} < 1.5V$ or $> 9V$		 0.1 60		 kΩ kΩ
Internal gating pulse					
t_P	Pulse duration		7.5		μs

NOTES:
1. Permissible range 1 to 7V.
2. t_D = switch-off delay of line output stage.
3. Line flyback pulse duration $t_{FP} = 12\mu s$.
4. Excluding external component tolerances.
5. Current source.
6. Emitter-follower.
7. Current source.

9

Signetics

TDA2594
Horizontal Combination

Product Specification

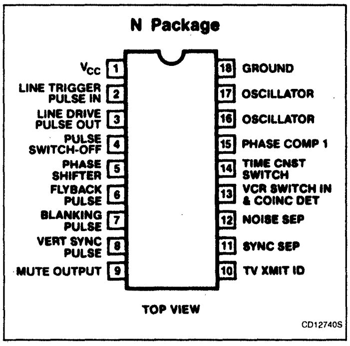
Linear Products

DESCRIPTION
The TDA2594 is a monolithic integrated circuit intended for use in color television receivers.

FEATURES
- **Horizontal oscillator based on the threshold switching principle**
- **Phase comparison between sync pulse and oscillator voltage (φ_1)**
- **Internal key pulse for phase detector (φ1) (additional noise limiting)**
- **Phase comparison between line flyback pulse and oscillator voltage (φ_2)**
- **Larger catching range obtained by coincidence detector (φ_3 between sync and key pulse)**
- **Switch for changing the filter characteristic and the gate circuit (VCR operation)**
- **Sync separator**
- **Noise separator**
- **Vertical sync separator and output stage**
- **Color burst keying and line flyback blanking pulse generator and clamp circuit for vertical blanking**
- **Phase shifter for the output pulse**
- **Output pulse duration for transistor reflection systems**
- **External switching off of the line trigger pulse**
- **Output stage with separate supply voltage**
- **Low supply voltage protection**
- **Transmitter identification and muting circuit, and vertical sync switch-off**

APPLICATIONS
- **Video processing**
- **Television receivers**
- **Video monitors**
- **Sync separator**

PIN CONFIGURATION

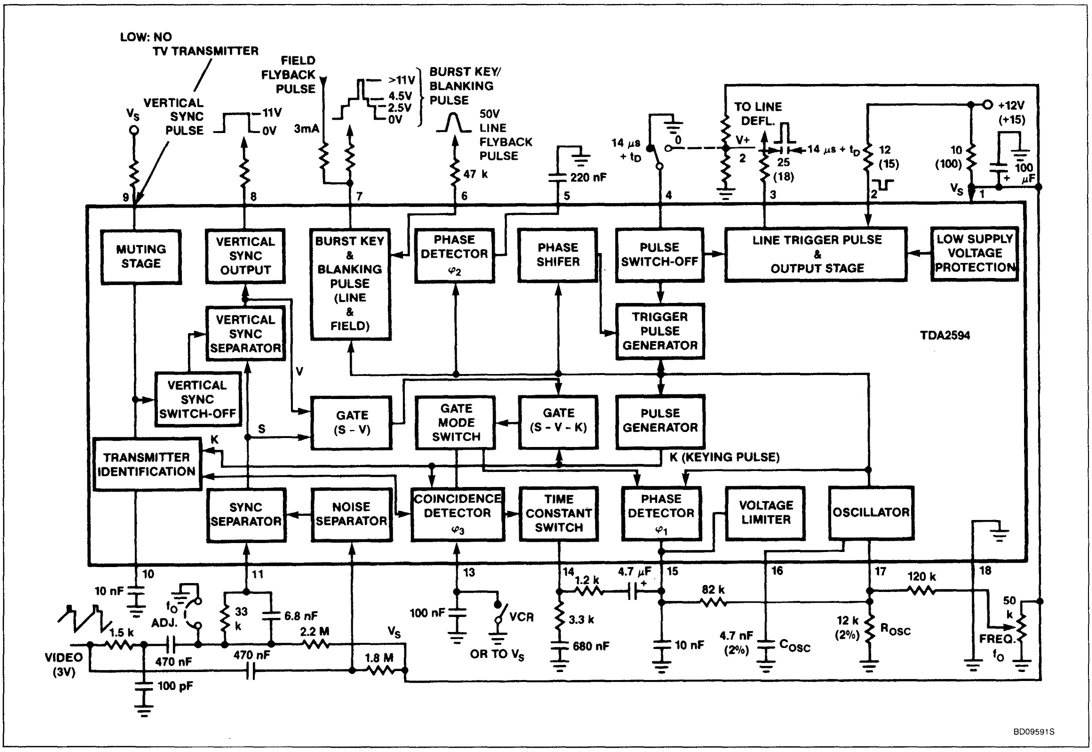


ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
18-Pin Plastic DIP (SOT-102DS)	−20°C to +70°C	TDA2594N

Horizontal Combination TDA2594

BLOCK DIAGRAM



ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
	Supply voltage		
$V_{1-18} = V_S$	at Pin 1 (voltage source)	13.2	V
V_{2-18}	at Pin 2	18	V
	Voltages		
V_{4-18}	Pin 4	13.2	V
V_{9-18}	Pin 9	18	V
$-V_{9-18}$		0.5	V
$\pm V_{11-18}$	Pin 11	6	V
$\pm V_{12-18}$	Pin 12	6	V
V_{13-18}	Pin 13	13.2	V
	Currents		
I_{2M}, $-I_{3M}$	Pins 2 and 3 (transistor driving) (peak value)	400	mA
I_4	Pin 4	1	mA
$\pm I_6$	Pin 6	10	mA
$-I_7$	Pin 7	5	mA
I_9	Pin 9	10	mA
I_{13}	Pin 13	2	mA
P_{TOT}	Total power dissipation	800	mW
T_{STG}	Storage temperature range	−25 to +125	°C
T_A	Operating ambient temperature range	−20 to +70	°C

Horizontal Combination TDA2594

DC AND AC ELECTRICAL CHARACTERISTICS at V_{1-18} = 12V; T_A = 25°C; measured in Block Diagram.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Sync separator (Pin 11)					
V_{11-18}	Input switching voltage		0.8		V
I_{11}	Input keying current	5		100	μA
I_{11}	Input leakage current at V_{11-18} = −5V			1	μA
I_{11}	Input switching current			5	μA
I_{11}	Switch off current	100	150		μA
$V_{11-18(P-P)}$	Input signal (peak-to-peak value)	3		4	V[1]
Noise separator (Pin 12)					
V_{12-18}	Input switching voltage		1.4		V
I_{12}	Input keying current	5		100	μA
I_{12}	Input switching current	100	150		μA
I_{12}	Input leakage current at V_{12-18} = −5V			1	μA
$V_{12-18(P-P)}$	Input signal (peak-to-peak value)	3		4	V[1]
$V_{12-18(P-P)}$	Permissible superimposed noise signal (peak-to-peak value)			7	V
Line flyback pulse (Pin 6)					
I_6	Input current	0.02	1		mA
V_{6-18}	Input switching voltage		1.4		V
V_{6-18}	Input limiting voltage	−0.7		+1.4	V
Switching on VCR (Pin 13)					
V_{13-18}	Input voltage	0		2.5 9 to V_S	V V
$-I_{13}$ or: I_{13}	Input current			200 2	μA mA
Pulse switching off (Pin 4) For t = 0; input Pin 4 open or V_{3-18} = 0					
V_{4-18}	Input voltage	5.4		6.6	V
I_4	Input current		0		μA
Vertical sync pulse (Pin 8) (positive-going)					
$V_{8-18(P-P)}$	Output voltage (peak-to-peak value)	10	11		V
R_8	Output resistance		2		kΩ
t_{ON}	Delay between leading edge of input and output signal		15		μs
t_{OFF}	Delay between trailing edge of input and output signal	t_{ON}			μs
V_{10-18}	Switching off the vertical sync pulse			3	V
Burst key pulse (Pin 7) (positive-going)					
V_{7-18}	Output voltage	10	11		V
R_7	Output resistance		70		Ω
t_P	Pulse duration; V_{7-18} = 7V	3.7	4	4.3	μs
t	Phase relation between middle of sync pulse at the input and the leading edge of the burst key pulse; V_{7-18} = 7V	2.15	2.65	3.15	μs
I_7	Output trailing edge current		2	2	mA
V_{7-18}	Saturation voltage during line scan			1	V

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) at $V_{1-18} = 12V$; $T_A = 25°C$; measured in Block Diagram.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Line flyback-blanking pulse (Pin 7) (positive-going)					
V_{7-18}	Output voltage	4.1		4.9	V
R_7	Output resistance		70		Ω
I_7	Output trailing edge current		2		mA
Field flyback/blanking pulse (Pin 7)					
V_{7-18}	Output voltage with externally forced in current $I_7 = 2.4$ to 3.6mA	2		3	V
R_7	Output resistance at $I_7 = 3mA$		70		Ω
TV transmitter identification output (Pin 9) (open-collector)					
V_{9-18}	Output voltage at $I_9 = 3mA$; no TV transmitter			0.5	V
R_9	Output resistance at $I_9 = 3mA$; no TV transmitter			100	Ω
I_9	Output current at $V_{10-18} \geqslant 3V$; TV transmitter identified			5	μA
TV transmitter identification (Pin 10)					
	When receiving a TV signal, the voltage V_{10-18} will change from $\leqslant 1V$ to $\geqslant 7V$				
Line drive pulse (positive-going)					
$V_{3-18(P-P)}$	Output voltage (peak-to-peak value)		10		V
R_3	Output resistance for leading edge of line pulse for trailing edge of line pulse		 2.5 20		 Ω Ω
t_P	Pulse duration (transistor driving) $V_{4-18} = 0$ to 3.5V; $-I_4 \geqslant 200\mu A$; $t_{FP} = 12\mu s$			$14 + t_D$	μs^2
V_{1-18}	Supply voltage for switching off the output pulse		4		V
Overall phase relation					
Δt	Phase relation between middle of sync pulse and the middle of the flyback pulse		2.6		μs^3
	The adjustment of the overall phase relation and consequently the leading edge of the line drive pulse occurs automatically by phase control φ_2.				
$\Delta I/\Delta t$	If additional adjustment is applied, it can be arranged by current supply at Pin 5, such that: supplying current		30		μA/μs

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DC AND AC ELECTRICAL CHARACTERISTICS (Continued) at V_{1-18} = 12V; T_A = 25°C; measured in Block Diagram.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Oscillator (Pins 16 and 17)					
V_{16-18}	Threshold voltage low level		4.4		V
V_{16-18}	Threshold voltage high level		7.6		V
$\pm I_{16}$	Charging current		0.47		mA
f_O	Frequency; free running (C_{OSC} = 4.7nF; R_{OSC} = 12kΩ)		15.625		kHz
Δf_O	Spread of frequency			±5	%[6]
$\Delta f_O/\Delta_{17}$	Frequency control sensitivity		31		Hz/μA
Δf_O	Adjustment range of network in circuit (Block Diagram)		±10		%
$\frac{\Delta f_O/f_O}{\Delta V/V_{NOM}}$	Influence of supply voltage on frequency; reference at V_S = 12V			±0.05	%[6]
Δf_O	Change of frequency when V_S drops to 5V; reference at V_S = 12V			±10	%[6]
TC	Temperature coefficient of oscillator frequency			$\pm 10^{-4}$	K^{-1} [6]
Phase comparison φ_1 (Pin 15)					
V_{15-18}	Control voltage range	4.1		7.9	V
$\pm I_{15M}$	Control current (peak value)	1.8		2.2	mA
I_{15}	Output leakage current at V_{15-18} = 4.3 to 7.7V			1	μA
	Output resistance				
R_{13}	at V_{15-18} = 4.3 to 7.7V[4]		high ohmic		
R_{13}	at $V_{15-18} \leqslant$ 4.1V or $\geqslant$ 7.9V[5]		low ohmic		
	Control sensitivity		2		kHz/μs
Δf	Catching and holding range (82kΩ between Pins 15 and 17)		±680		Hz
$\Delta(\Delta f)$	Spread of catching and holding range		±12		%[6]
Phase comparison φ_2 and phase shifter (Pin 5)					
V_{5-18}	Control voltage range	5.4		7.6	V
$\pm I_{5M}$	Control current (peak value)		1		mA
R_5	Output resistance at V_{5-18} = 5.4 to 7.6V[4]		high ohmic		
I_5	Input leakage current at V_{5-18} = 5.4 to 7.6V			5	μA
t_D	Permissible delay between leading edge of output pulse and leading edge of flyback pulse (t_{FP} = 12μs)			15.5	μs
$\Delta t/\Delta t_D$	Static control error			0.2	%
Coincidence detector φ_3 (Pin 13)					
V_{13-18}	Output voltage	0.5		6	V
	Output current (peak value)				
I_{13M}	without coincidence		0.1		mA
$-I_{13M}$	with coincidence		0.5		mA

NOTES:
1. Permissible range 1 to 7V.
2. t_D = switch-off delay of line output stage.
3. Line flyback pulse duration t_{FP} = 12μs.
4. Current source.
5. Emitter-follower.
6. Excluding external component tolerances.

Signetics

TDA2595
Horizontal Combination

Product Specification

Linear Products

DESCRIPTION

The TDA2595 is a monolithic integrated circuit intended for use in color television receivers.

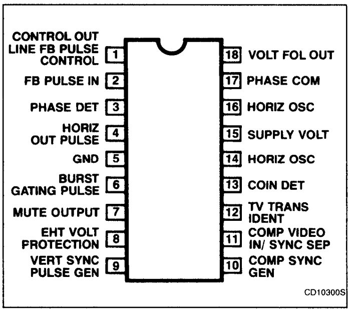
FEATURES

- **Positive video input; capacitively coupled (source impedance $< 200\Omega$)**
- **Adaptive sync separator; slicing level at 50% of sync amplitude**
- **Internal vertical pulse separator with double slope integrator**
- **Output stage for vertical sync pulse or composite sync depending on the load; both are switched off at muting**
- **φ_1 phase control between horizontal sync and oscillator**
- **Coincidence detector φ_3 for automatic time constant switching; overruled by the VCR switch**
- **Time constant switch between two external time constants for loop gain; both controlled by the coincidence detector φ_3**
- **φ_1 gating pulse controlled by coincidence detector φ_3**
- **Mute circuit depending on TV transmitter identification**
- **φ_2 phase control between line flyback and oscillator; the slicing levels for φ_2 control and horizontal blanking can be set separately**
- **Burst keying and horizontal blanking pulse generation, in combination with clamping of the vertical blanking pulse (three-level sandcastle)**
- **Horizontal drive output with constant duty cycle inhibited by the protection circuit or the supply voltage sensor**
- **Detector for too low supply voltage**
- **Protection circuit for switching off the horizontal drive output continuously if the input voltage is below 4V or higher than 8V**
- **Line flyback control causing the horizontal blanking level at the sandcastle output continuously in case of a missing flyback pulse**
- **Spot suppressor controlled by the line flyback control**

APPLICATIONS

- **Television receivers**
- **Video receivers**

PIN CONFIGURATION



ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
18-Pin Plastic DIP (SOT-102CS)	−20°C to +70°C	TDA2595N

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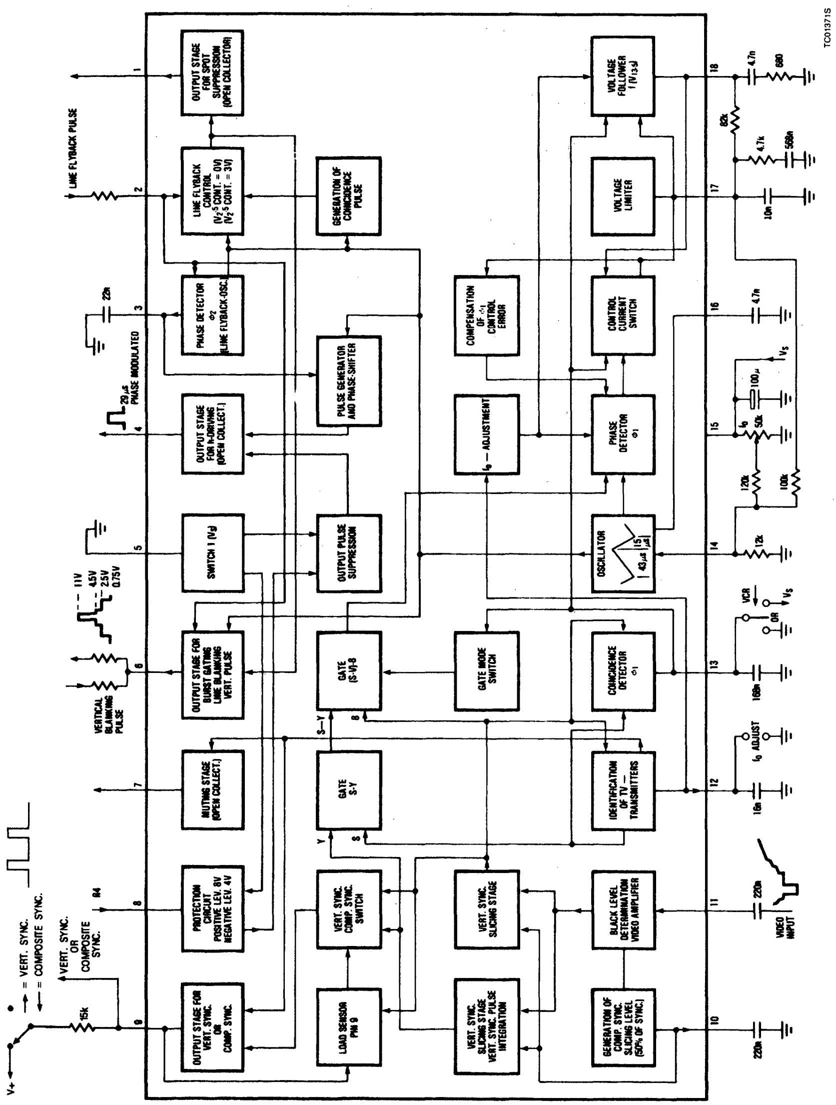
BLOCK DIAGRAM



ABSOLUTE MAXIMUM RATINGS

SYMBOL	DESCRIPTION	RATING	UNIT
$V_{15-5} = V_{CC}$	Supply voltage (Pin 15)	13.2	V
	Voltages at:		
$V_{1;4;7-5}$	Pins 1, 4 and 7	18	V
$V_{8;13;18-5}$	Pins 8, 13 and 18	V_{CC}	V
V_{11-5}	Pin 11 (range)	−0.5 to +6	V
	Currents at:		
I_1	Pin 1	10	mA
$\pm I_{2M}$	Pin 2 (peak value)	10	mA
I_4	Pin 4	100	mA
$\pm I_{6M}$	Pin 6 (peak value)	6	mA
I_7	Pin 7	10	mA
I_8	Pin 8 (range)	−5 to +1	mA
I_9	Pin 9 (range)	−10 to +3	mA
$\pm I_{18}$	Pin 18	10	mA
P_{TOT}	Total power dissipation	800	mW
T_{STG}	Storage temperature range	−65 to +150	°C
T_A	Operating ambient temperature range	−0 to +70	°C

DC AND AC ELECTRICAL CHARACTERISTICS V_{CC} = 12V; T_A = 25°C, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Composite video input and sync separator (Pin 11) (internal black level determination)					
$V_{11-5(P-P)}$	Input signal (positive video; standard signal; peak-to-peak value)	0.2	1	3	V
$V_{11-5(P-P)}$	Sync pulse amplitude (independent of video content)	50			mV
R_G	Generator resistance			200	Ω
	Input current during				
I_{11}	Video		5		μA
$-I_{11}$	Sync pulse		40		μA
$-I_{11}$	Black level		25		μA
Composite sync generation (Pin 10) horizontal slicing level at 50% of the sync pulse amplitude					
	Capacitor current during				
I_{10}	Video		16		μA
$-I_{10}$	Sync pulse		170		μA
Vertical sync pulse generation (Pin 9) slicing level at 30% (60% between black level and horizontal slicing level)					
V_{9-5}	Output voltage	10			V
t_P	Pulse duration		190		μs
t_D	Delay with respect to the vertical sync pulse (leading edge)		45		μs
	Pulse-mode control				
	Output current for vertical sync pulse (dual integrated)	No current applied at Pin 9			
	Output current for horizontal and vertical sync pulse (non-integrated separated signal)	Current applied via a resistor of 15kΩ from V_{CC} to Pin 9			

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DC AND AC ELECTRICAL CHARACTERISTICS (Continued) V_{CC} = 12V; T_A = 25°C, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Horizontal oscillator (Pins 14 and 16)					
f_{OSC}	Frequency; free-running		15.625		kHz
V_{14-5}	Reference voltage for f_{OSC}		6		V
$\Delta f_{OSC}/\Delta I_{14}$	Frequency control sensitivity		31		Hz/μA
Δf_{OSC}	Adjustment range of circuit Figure 1		±10		%
Δf_{OSC}	Spread of frequency			5	%
$\frac{\Delta f_{OSC}/f_{OSC}}{\Delta V_{15-5}/V_{15-5}}$ Δf_{OSC} TC	Frequency dependency (excluding tolerance of external components) with supply voltage (V_{CC} = 12V) with supply voltage drop of 5V with temperature		±0.05	 10 $\pm 10^{-4}$	% % °C^{-1}
 $-I_{16}$ I_{16}	Capacitor current during: Charging Discharging		 1024 313		 μA μA
 t_R t_F	Sawtooth voltage timing (Pin 14) Rise time Fall time		 49 15		 μs μs
Horizontal output pulse (Pin 4)					
V_{4-5}	Output voltage Low at I_4 = 30mA			0.5	V
t_P	Pulse duration (High)		29 ±1.5		μs
V_{CC}	Supply voltage for switching off the output pulse (Pin 15)		4		V
ΔV_P	Hysteresis for switching on the output pulse		250		mV
Phase comparison φ_1 (Pin 17)					
V_{17-5}	Control voltage range	3.55		8.3	V
I_{17}	Leakage current at V_{17-5} = 3.55 to 8.3V			1	μA
$\pm I_{17}$	Control current for external time constant switch	1.8	2	2.2	mA
$\pm I_{17}$	Control current at $V_{18-5} = V_{15-5}$ and $V_{13-5} < 2V$ or $V_{13-5} > 9.5V$		8		mA
$\pm I_{17}$	Control current at $V_{18-5} = V_{15-5}$ and V_{13-5} = 2 to 9.5V	1.8	2	2.2	mA
 S_φ Δf_{OSC} Δf_{OSC}	Horizontal oscillator control Control sensitivity Catching and holding range Spread of catching and holding range	 6	 ±680 ±10		 kHz/μs Hz %
t_P	Internal keying pulse at V_{13-5} = 2.9 to 9.5V		7.5		μs
 V_{13-5} V_{13-5}	Time constant switch Slow time constant Fast time constant	 9.5 2		 2 9.5	 V V
$\pm V_{17-18}$	Impedance converter offset voltage (slow time constant)			3	mV
 R_{18-5} R_{18-5}	Output resistance Slow time constant Fast time constant	 high-impedance		 10	 Ω
I_{18}	Leakage current			1	μA

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) V_{CC} = 12V; T_A = 25°C, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Coincidence detector φ_3 (Pin 13)					
	Output voltage				
V_{13-5}	without coincidence with composite video signal			1	V
V_{13-5}	without coincidence without composite video signal (noise)			2	V
V_{13-5}	With coincidence with composite video signal		6		V
	Output current				
I_{13}	without coincidence with composite video signal		50		μA
$-I_{13}$	with coincidence with composite video signal		300		μA
	Switching current				
I_{13}	at $V_{13-5} = V_{CC}-0.5V$			100	μA
$I_{13(av)}$	at V_{13-5} = 0.5V (average value)			100	μA
Phase comparison φ_2 (Pins 2 and 3)[1]					
Δt	Phase relation between middle of the horizontal sync pulse and the middle of the line flyback pulse at $t_{FP} = 12\mu s$[2]		2.6 ± 0.7		μs
$\Delta I/\Delta t$	If additional adjustment is required, it can be arranged by applying a current at Pin 3, such that for applied current:		30		μA/μs
Input for line flyback pulse (Pin 2)					
V_{2-5}	Switching level for φ_2 comparison		3		V
V_{2-5}	Switching level for horizontal blanking and flyback control		0.3		V
V_{2-5}	Input voltage limiting		−0.7 +4.5		V V
	Switching current				
I_2	at horizontal flyback	0.01	1		mA
I_2	at horizontal scan			2	μA
$-I_2$	Maximum negative input current			500	μA
Phase detector output (Pin 3)					
$\pm I_3$	Control current for φ_2		1		mA
$\Delta t_{\varphi 2}$	Control range		19		μs
$\Delta t/\Delta t_d$	Static control error			0.2	%
I_3	Leakage current			5	μA
Burst gating pulse (Pin 6)[3]					
V_{6-5}	Output voltage	10	11		V
t_P	Pulse duration	3.7	4	4.3	μs
$t_{\varphi 6}$	Phase relation between middle of sync pulse at the input and the leading edge of the burst gating pulse at V_{6-5} = 7V	2.15	2.65	3.15	μs
I_6	Output trailing edge current		2		mA

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DC AND AC ELECTRICAL CHARACTERISTICS (Continued) V_{CC} = 12V; T_A = 25°C, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Horizontal blanking pulse (Pin 6)[3]					
V_{6-5}	Output voltage	4.2	4.5	4.9	V
I_6	Output trailing edge current		2		mA
V_{6-5sat}	Saturation voltage at horizontal scan			0.5	V
Clamping circuit for vertical blanking pulse (Pin 6)[3]					
V_{6-5}	Output voltage at I_6 = 2.8mA	2.15	2.5	3	V
I_{6min}	Minimum output current at V_{6-5} > 2.15V		2.3		mA
I_{6max}	Maximum output current at V_{6-5} < 3V		3.3		mA
TV transmitter identification (Pin 12)					
	Output voltage				
V_{12-5}	no TV transmitter			1	V
V_{12-5}	TV transmitter identified	7			V
Mute output (Pin 7)					
V_{7-5}	Output voltage at I_7 = 3mA; no TV transmitter			0.5	V
R_{7-5}	Output resistance at I_7 = 3mA; no TV transmitter			100	Ω
I_7	Output leakage current at V_{12-5} > 3V; TV transmitter identified			5	μA
Protection circuit (beam current/EHT voltage protection) (Pin 8)					
V_{8-5}	No-load voltage for I_8 = 0 (operative condition)		6		V
V_{8-5}	Threshold at positive-going voltage		8 ±0.8		V
V_{8-5}	Threshold at negative-going voltage		4 ±0.4		V
$\pm I_8$	Current limiting for V_{8-5} = 1 to 8.5V		60		μA
R_{8-5}	Input resistance for V_{8-5} > 8.5V		3		kΩ
t_d	Response delay of threshold switch		10		μs
Control output of line flyback pulse control (Pin 1)					
V_{1-5sat}	Saturation voltage at standard operation; I_1 = 3mA			0.5	V
I_1	Output leakage current in case of break in transmission			5	μA

NOTES:
1. Phase comparison between horizontal oscillator and the line flyback pulse. Generation of a phase-modulated (φ_2) horizontal output pulse with constant duration.
2. t_{FP} is the line flyback pulse duration.
3. Three-level sandcastle pulse.

Signetics

Linear Products

AN158
Features of the TDA2595 Synchronization Processor

Application Note

FEATURES

- Positive video input, capacitive coupled (source impedance < 200Ω)
- Adaptive sync slicer at 50% of sync pulse amplitude
- Internal vertical pulse separator with double-slope integrator
- Outputstage for vertical sync pulse or composite sync depending on the load. Both are switched off by mute
- ϕ_1 phase control between H-sync and oscillator
- Coincidence detector ϕ_3 for automatic time-constant switching, overruled by the VCR-switch
- Time-constant switch between two external time-constants or loop-gain switch both controlled by coincidence detector ϕ_3
- ϕ_1 gating pulse controlled by coincidence detector ϕ_3
- Mute circuit depending on TV transmitter identification
- ϕ_2 phase control between line flyback and oscillator. The slicing levels for ϕ_2 control and line blanking can be set separately
- Burst keying and line blanking pulse generation, combined with clamping of field blanking pulse (triple-level sandcastle)
- H-drive output with constant duty cycle inhibited by the protection circuit or the supply voltage detector
- Detector for too low supply voltage
- Protection circuit switching off H-drive output continuously if input voltage is below 4V or higher than 8V
- Line flyback control causing line-blanking level at sandcastle output continuously in case of missing flyback pulse
- Spot-suppressor controlled by the line flyback control

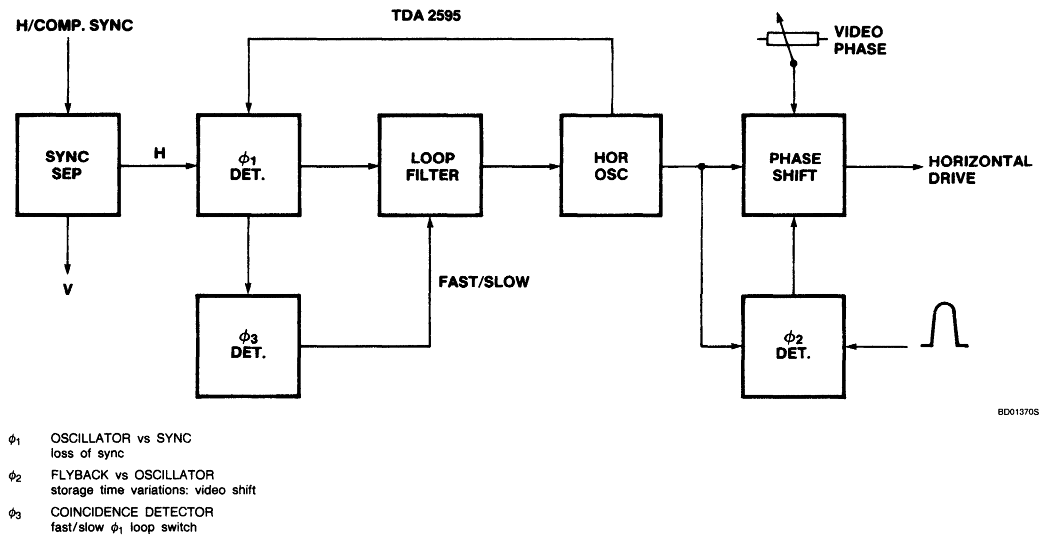

Figure 1

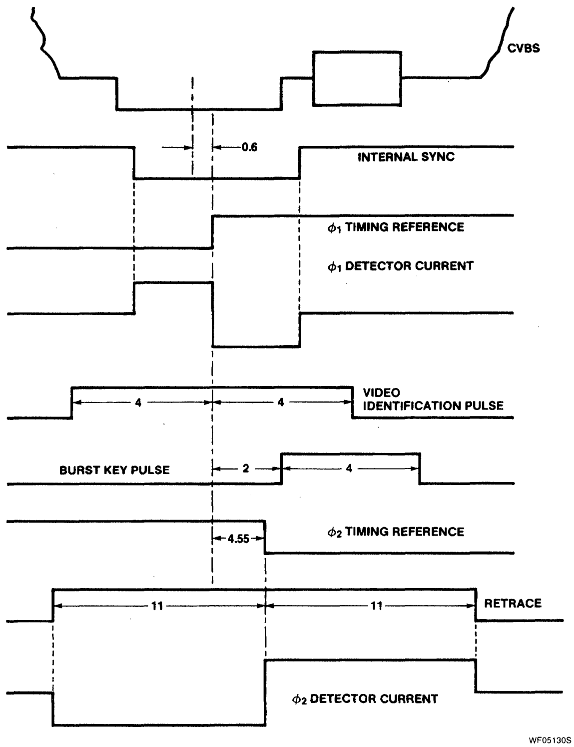

Figure 2. Timing Diagram

Features of the TDA2595 Synchronization Processor AN158

SYNC SEPARATOR

Adaptive sync separator to slice H-sync at 50% and V-sync at 25% independent on sync-amplitude. This is to insure immunity against deteriorated sync impulses. The black level is stored on a capacitor which is fed to the positive video-signal (source impedance 200Ω) into Pin 11. The slicing level is detected internally and stored in a capacitor at Pin 12.

The internal vertical integrator has a delay of 45μs and is of the double-slope type to avoid jitter and to improve noise immunity.

VERTICAL/COMPOSITE SYNC

The output stage at Pin 9 delivers a positive vertical pulse or a positive composite sync signal if the current drain is higher than 3mA.

If no TV transmitter is detected, the output is switched to ground. The source impedance is low-ohmic.

15kHz VCO

The VCO is a current controlled ramp oscillator with 49μs rise time and 15μs fall time. The timing capacitor is connected to Pin 16; the control current has to be fed into Pin 14.

While adjusting f_0, Pin 12 should be connected to ground.

The oscillator generates the following signals (see timing diagram Figure 2):

- timing reference for ϕ_1
- gating pulse for ϕ_1
- reference pulse for video identification circuit and coincidence detector ϕ_3
- burst keying pulse
- time reference for ϕ_2

ϕ_1 PHASE CONTROL

The phase control ϕ_1 compares the ϕ_1 timing reference of the VCO with the center of the H-sync signal and converts the time difference into a proportional current at Pin 17.

The external low-pass filter at Pin 17 determines the time constant and the catching and tracking range of the VCO.

If Pin 18 is connected to the V+, the loop gain is increased 4 times as long as the oscillator is not locked in or Pin 13 is connected to ground or V+ (VCR switch).

If Pin 18 is connected as shown in the circuit diagram, Pin 18 has the same voltage as Pin 17 as long as the oscillator is not locked in or Pin 13 is connected to ground. Due to this the "long" time constant connected from Pin 18 to ground, ground is electrically disconnected from Pin 17.

If the oscillator is locked in and Pin 13 not connected to ground, Pin 18 switches to high impedance and thus the loop filter to the "long" time-constant.

By switching loop gain or loop time-constant, the lock in condition of the oscillator is not disturbed. This enables a fast search tuning using the TV transmitter identification ($\overline{\text{mute}}$) as a search stop.

To increase noise immunity the phase detector is inhibited during horizontal retrace and vertical retrace if the oscillator is locked in and Pin 13 not connected to ground or V+.

COINCIDENCE DETECTOR ϕ_3

The coincidence circuit detects whether there is coincidence between the H-sync pulse and a 8μs impulse generated by the VCO. The capacitor at Pin 13 is discharged continuously by 8μs current pulses of 50μA. If there is coincidence, the capacitor is additionally charged by H-sync pulses of 350μA.

If the voltage at Pin 13 exceeds 3V, the loop gain is reduced and the loop time constant is switched to the "long" value.

If the voltage exceeds 4.5V, the phase detector ϕ_1 is gated to improve noise immunity.

MUTE CIRCUIT

The mute circuit detects whether there is coincidence between the H-sync impulse and a 8μs impulse generated by the VCO. The capacitor at Pin 12 is discharged during sync-pulses of 50μA and by 8μs current pulses of 50μA. If there is coincidence, the capacitor is additionally charged by H-sync pulses of 450μA.

If the voltage at Pin 12 exceeds 4V, mute is released and the mute output at Pin 7 is switched to high impedance. Although the coincidence detector ϕ_3 and the mute circuit act similarly, separate circuits have been chosen. This is to gain in design flexibility as far as the time constants are related and to keep the mute function alive independently on the VCR switch.

ϕ_2 PHASE CONTROL

The phase control ϕ_2 compares the center of the positive flyback pulse at Pin 2 at a threshold of 3V with the ϕ_2 timing reference. The time difference is converted into a proportional current at Pin 3. Loop gain and time-constant are influenced by the external components at Pin 3. The voltage at Pin 3 in turn controls the phase shift.

To achieve a small phase adjustment a small current may be injected into Pin 3.

The aim of having two different thresholds at the flyback input is to determine the performance of the ϕ_2 loop, e.g., a straight vertical center line, by the amplitude of the applied flyback pulse without affecting the blanking time.

SUPER SANDCASTLE

For burst keying and vertical and horizontal blanking there is a 3 level pulse at Pin 6.

The burst keying part is driven from the VCO and is 4μs wide. Due to its small tolerances in widths and phase it keys the burst very exactly and is suitable as black level clamping pulse.

The blanking part is derived from the line flyback pulse at Pin 2 at a threshold of 0.2V. If no flyback is applied to Pin 2, there will be continuous blanking level superimposed by the burst keying pulse.

The frame blanking part has to be fed in externally as a 2mA current.

HORIZONTAL DRIVE

The H-drive output is an open-collector output at Pin 4. The output pulse has a constant aspect ratio of 45.3% off and 54.7% on dependent upon the line frequency. An internal guard logic insures that there will be high level during flyback. The output is inhibited by the protection circuit also if the supply voltage is below 4V. In both cases the line flyback vanishes and by this the spot suppressor is activated.

SPOT SUPPRESSOR

The spot suppressor is an open collector output at Pin 1. If no flyback impulses are detected at Pin 2, the output switches to high impedance and remains there as long as the flyback pulses are missing even if the supply voltage vanishes during that time.

PROTECTION CIRCUIT

The protection circuit is activated if the voltage at Pin 8 exceeds 8V or decreases below 4V. One of both thresholds may be used (as indicated in Figures 4a and b) to have X-ray protection or overcurrent protection.

If activated, the H-drive is inhibited by this and the line flyback vanishes and in turn the spot suppressor is activated.

The protection circuit is reset if the supply voltage decreases below 4V, e.g., the set is switched off.

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Features of the TDA2595 Synchronization Processor AN158

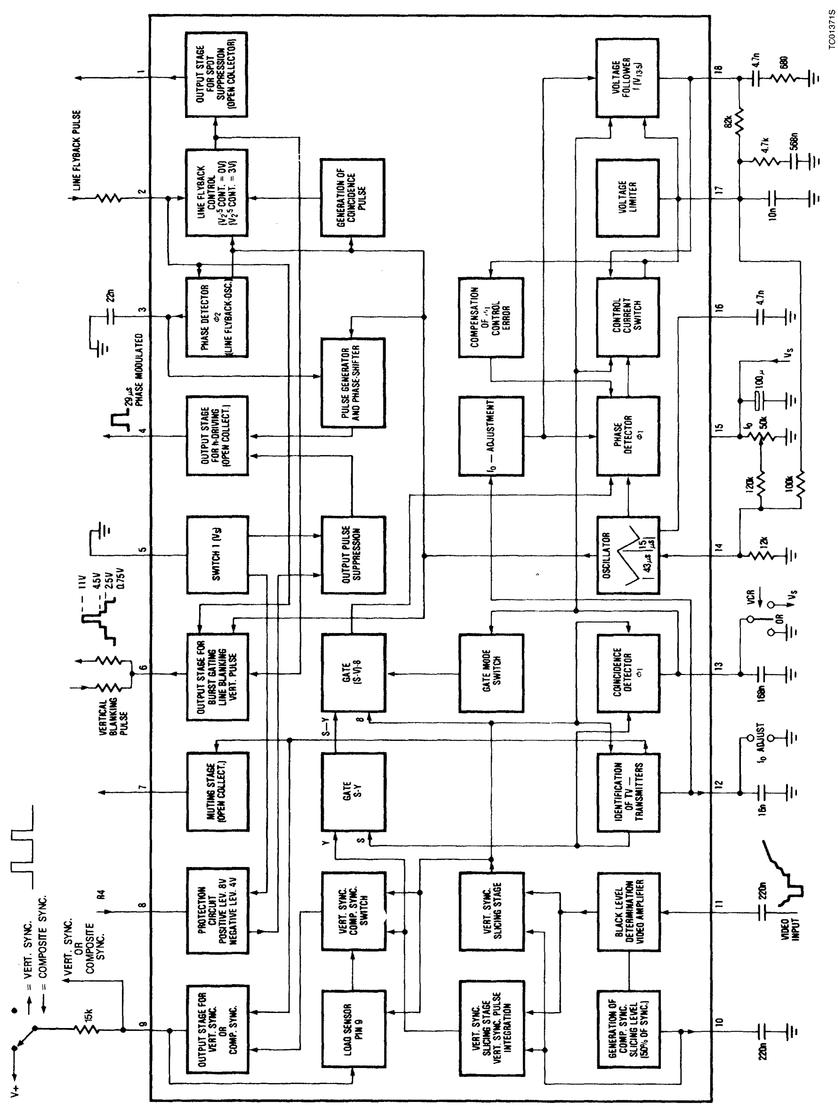
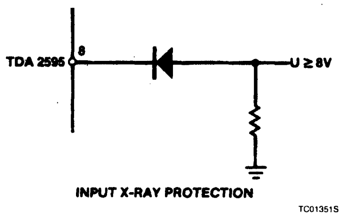
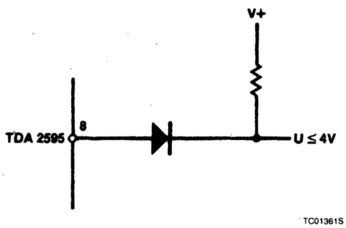


Figure 3



a. Input X-ray Protection



b. Input Over Current Protection

Figure 4

Signetics

Section 10
Color Decoding and Encoding

Linear Products

INDEX

Signetics

AN155A
Multi-Standard Color Decoder With Picture Improvement

Application Note

Linear Products

The decoder concept presented here comprises a multi-standard color decoder and a video combination. The concept can also be extended by means of a picture improvement circuit.

A brief overview will first be given to clarify this arrangement. Figure 1 shows the block diagram of a complete color decoder from the CVBS interface up to the picture tube. There are switchable filters for separation of the luminance and chrominance signals from one another. Only one IC is necessary for the demodulation of four color standards.

The output signals are the standard-independent color difference signals (B-Y) and (R-Y), i.e., U and V. The baseband signals (i.e., color difference signals and luminance signal Y) can either be directly supplied to the video combination or they can be supplied via a signal processor IC as shown here.

The video combination comprises all functions for advanced video signal processing. The RGB output signals of the IC can be fed to the video final stages directly.

The interface selected in this decoder concept, with the baseband signals as input signals of the video combination, also permits new circuit concepts to be introduced; e.g., the delay line which is required for PAL and SECAM can be realized with CCD lines. Picture improvement circuits with picture memories can also be added.

The Color Transient Improvement (CTI) IC which is incorporated in Figure 1 was also developed for this interface. Two functions are integrated in this circuit: a transient improvement for a better picture, and a Y delay line in gyrator technique to replace the previously-required wound line.

In the past, multi-standard color decoders (MSD) have been built up with a number of integrated circuits. Parallel working concepts are known, and also transcoder concepts specially for PAL and SECAM. The decoders of the various standards require circuit blocks of the same type; this applies in particular to the quadrature amplitude modulation standards (QAM standards) PAL and NTSC, but also to a large extent to the FM standard SECAM. Therefore, an obvious approach for the integration of a multi-standard decoder on one chip is to make use of as many circuit blocks as possible in common for the different standards in order to minimize the components and, also, the crystal area required. Under the condition of automatic standard identification, as is already the state of the art for present MSD concepts, multiple utilization of the circuit blocks can only be realized if automatic standard identification is effected by *sequential standard scanning*. A system of this kind gives the great advantage that the entire decoder, including the filters, can be designed in the optimum way for the individual standards.

The single-chip multi-standard decoder TDA4555/TDA4556 is examined fully in AN1551. Please refer to AN1551 for application information.

The Video Combination IC — TDA3505

The video combination IC incorporates all setting functions for color picture reproduction. A black current stabilizing circuit is provided. This saves three tuning operations and also automatically regulates operating-point changes due to warming up after switch-on and to aging.

RGB signal inputs are provided for signal supply from RGB sources via the audio/video plug, e.g., from cameras or from internal teletext decoders.

Figure 2 shows the block diagram of the input part of this IC. The two color difference signals −(R-Y) and −(B-Y) are fed in via capacitors and clamped in the input stages to reference values. After the saturation control stages, the −(G-Y) signal is generated with the (G-Y) matrix. These color difference signals, together with the Y-signal which is also clamped in the input stage, are converted to the R, G, and B signals in the R, G, and B matrix.

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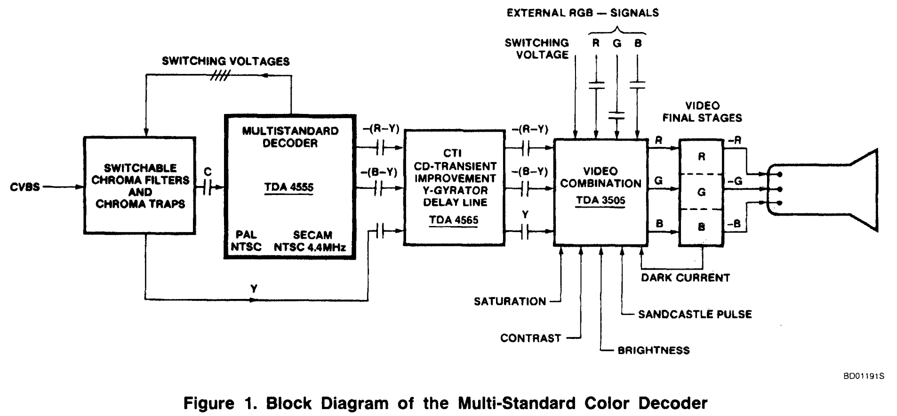

Figure 1. Block Diagram of the Multi-Standard Color Decoder

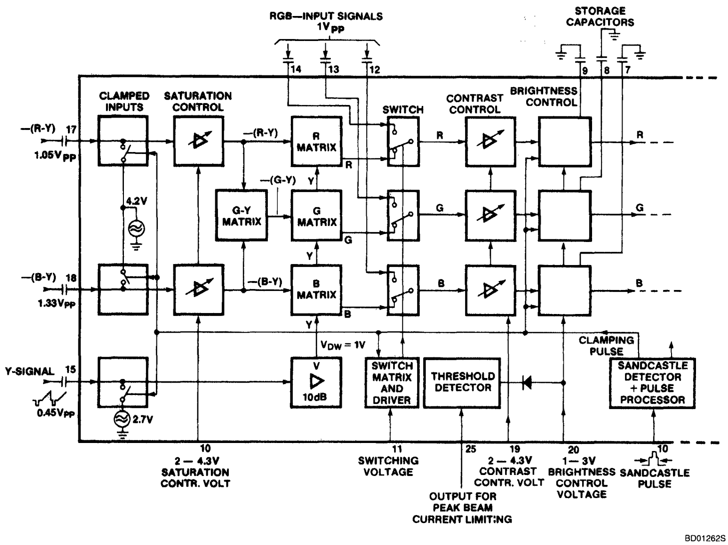

Figure 2. Front Part of the Video Combination TDA3505

Switching stages, together with a switching matrix and a driver stage for the switching, permit the choice between the picture signals from the color difference and Y inputs, or from the R, G, B inputs. When the R, G, B signals from the R, G, B inputs are selected, they are added to the black levels, which are simultaneously inserted. The switching times between blanking, insertion, and changeover are about 50ns and are so small that there are no visible errors in the picture. If the RGB inputs are constantly connected, synchronization with the other signals is not necessary. The signals also pass through the contrast- and brightness-control stages. A peak beam current limitation can be effected via an input to a threshold level switching circuit. The threshold level circuit then reduces the contrast-control voltage. Average beam current limitation is effected directly via the contrast-control voltage, whereby under certain circumstances the brightness control is also reduced via an internal diode.

All the pulses required in the IC, and especially for the black current stabilization which will be explained later, are derived from the sandcastle pulse.

Signal processing is effected in parallel in three R, G, B channels and, therefore, the description and explanation will continue to be limited to the R channel.

Figure 3 shows the functional block diagram of the black current stabilizer. The R signal is blanked out and a measuring pulse is inserted for the black current measurement. A subsequent limiter stage prevents overdriving of the video final stages. A control stage is provided for white-point adjustment, which can be effected by means of a DC setting voltage. There is an adding stage in which the voltage from the black current stabilization circuit is added to the R signal. The output stage of the IC can feed the video final stage directly. Its output voltage is supplied via a PNP measuring transistor to the cathode of the CRT. The collector circuit includes a measuring resistor at which voltage drops occur at the respective sequential measuring times; these are due on the one hand to any leakage currents which occur and on the other hand to dark current with leakage currents. These voltages are given to the IC. Following a buffer stage, the measurement voltage for the leakage currents is stored on the capacitor C_L. Switch S_L is only closed at the time when the signal is blanked and no signal current can flow. During the black level measurement time, a reference voltage of 0.5V is subtracted from the voltage to be measured and then compared in a comparator circuit with the stored voltage for the leakage currents. Switch S_d is only closed during the black measurement time and closes the control loop. Capacitor C_d stores the control voltage.

A dark current of 10μA is not too small for reliable evaluation and not too big, so that if it is in the right time position no disturbing effects are visible on the screen.

Insertion of the measurement pulses and their evaluation is sequential; this means that from the measuring resistor through the measurement input and leakage current storage up to and including the comparator circuit, these circuits only have to be realized once and are used for all three channels.

Figure 4 shows the time positions of the various measurement pulse insertions and evaluations. The measurement pulses are after the vertical flyback pulse and are thus above the upper picture edge in the overscan.

The R, G, B signals are blanked up to the inserted measurement pulses. The leakage current of all channels is measured in the line before the first measurement pulse. This is followed by the measurement pulses and their evaluation in the sequence red, green, blue.

A comprehensive application diagram with the video combination TDA3505 and the video final stages is shown in Figure 5.

For two sets of external RGB inputs and larger video input bandwidth, the TDA4580 can be used in place of the TDA3505 (see Figure 6).

The Color Transient Improvement IC — TDA4565

A complete multi-standard decoder can be built with the two ICs described above. A third IC, which can be interconnected in the color difference interface, can be used for color

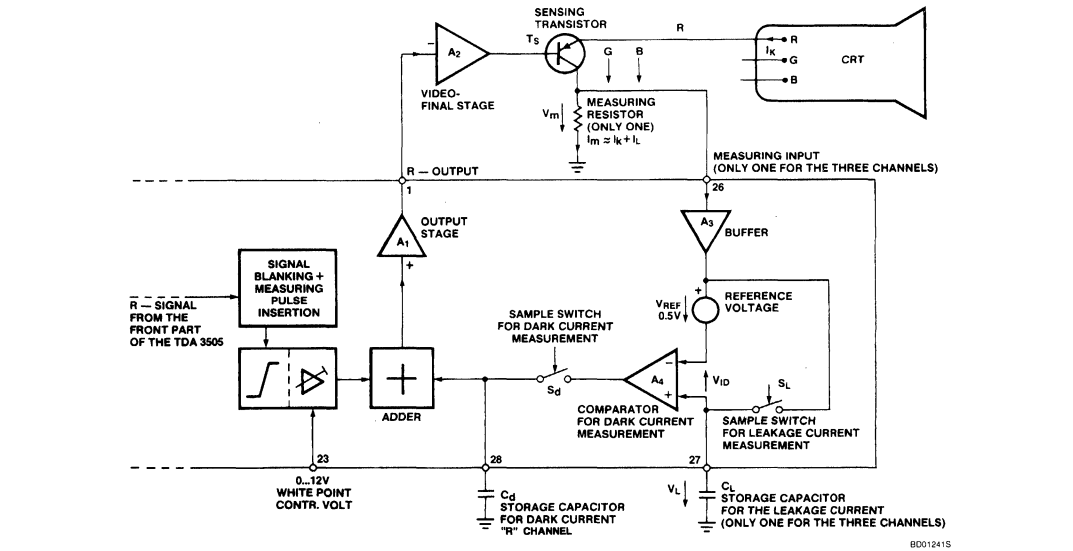

Figure 3. Functional Block Diagram for the Dark Current Stabilization With the Video Combination TDA3505 (R-channel)

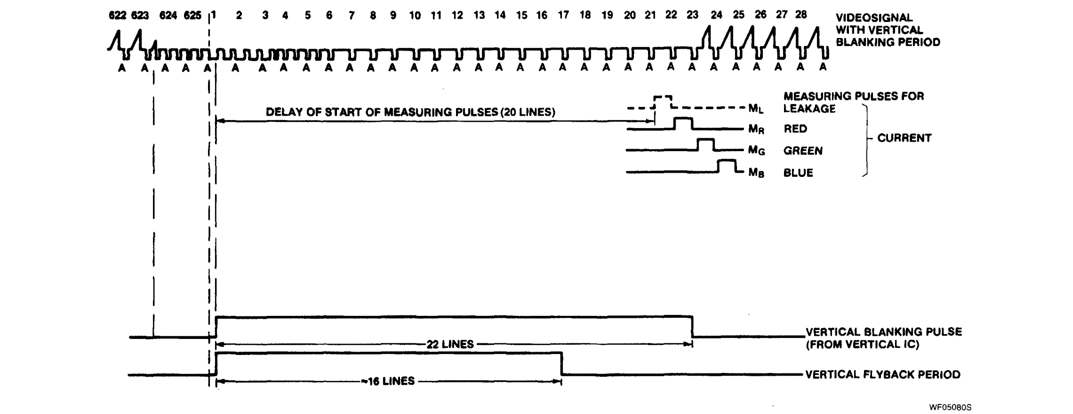

Figure 4. Position of the Measuring Lines of the Video Combination TDA3505

picture improvements by means of transient improvement of the color difference signals.

In Figure 7, the signal characteristics a) and b) show a transient in the Y and color difference signal. The rise time of the color difference signal is longer, corresponding to the smaller bandwidth. A delay line in the Y channel coordinates the centers of the transients as shown in Figure 7c.

In deviation from the previous signal processing, with the Color Transient Improvement IC, the color difference transient does not occur until the input signal transient is finished, but then occurs with a steepness corresponding to that of the Y signal. The characteristic of this color difference signal is shown in Figure 7d. It is now clear that – as shown in Figure 7e – a correspondingly longer delay is necessary for the Y signal in order to achieve coincidence of the transients.

Color signal transmissions, especially of test pictures coming via this CTI circuit, appear on the screen with the same color definition as RGB transmissions.

Figure 8 gives an explanation of the CTI function: the simplified circuits are shown on the left and the signals occurring at these are shown on the right. Part "A" shows a color difference input signal with a fast positive transient corresponding to the maximum bandwidth of the color difference signal.

The subsequent negative signal characteristics are slower. In this circuit, the input signal

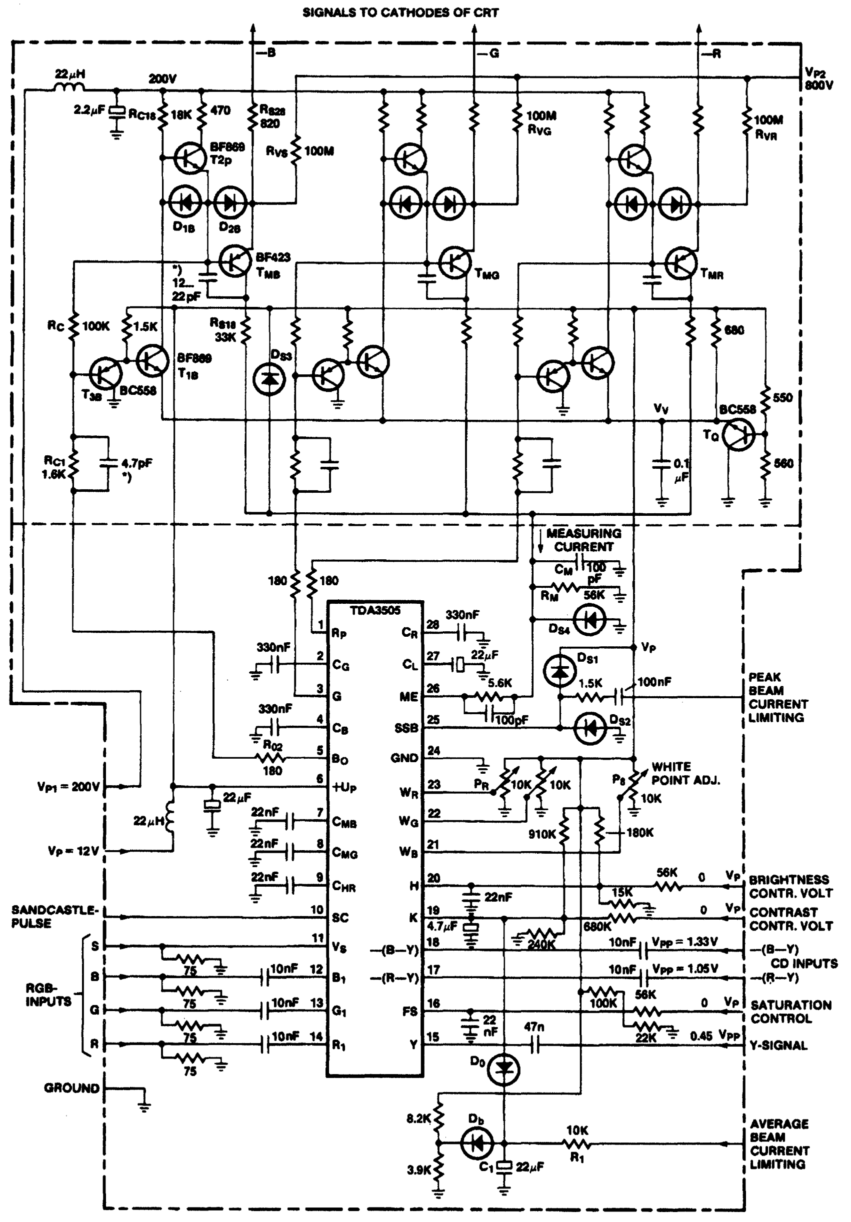

Figure 5. Application Diagram for the Video Combination TDA3505 and the Video Final Stages

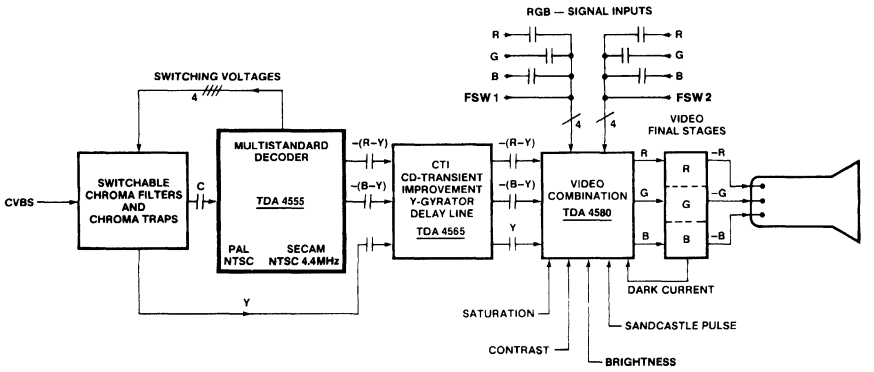

Figure 6. Multi-Standard Decoder Using the TDA4580

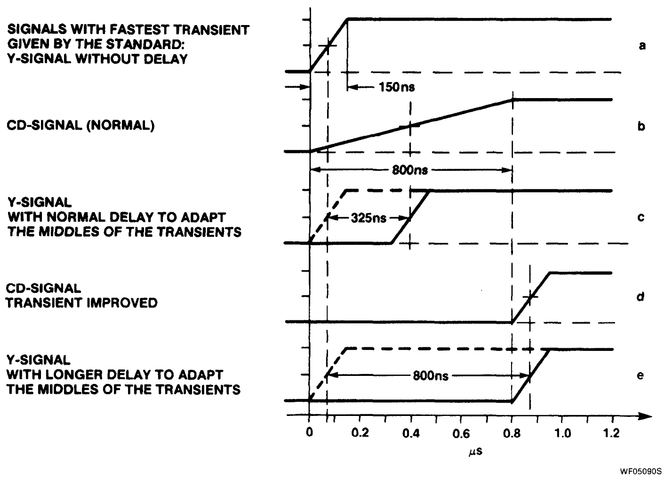

Figure 7. Y-delay Time for CD Signals With and Without Transient Improvement

is supplied after an impedance transformer via a switch and a further impedance transformer to the output. A storage capacitor is connected between the switch and the output impedance transformer, and is charged by the input impedance transformer in accordance with the signal characteristic.

Processing of the switching signal is affected by differentiation of the color difference signal, followed by full-wave rectification. Figure 8b shows the signals obtained in this way, which are supplied to a comparator via a high-pass filter. A diode at the high-pass filter reduces the charge reversal time and, thus, the dead time for generation of a switching signal for transients following in rapid succession. A comparator with threshold voltage generates a switching voltage as shown in Figure 8d from the signal of 8c when the threshold voltage is exceeded, and this triggers the switch. The switch is thus opened at the beginning of a transient and the voltage is maintained by the storage capacitor at the time before the transient. After completion of a fast transient, the switch is closed and the capacitor's charge is changed in approximately 150ns to the voltage after the transient. The effect of a slower transient characteristic is shown in the second part of the signal in Figure 8c. Only a small part is affected. For even slower characteristics, the differential quotient is so small that the threshold voltage is no longer exceeded and there is no effect on the signal. Thus, for the most part, only transients having a steepness approaching the system limit are improved, whereas slower signal characteristics remain unchanged.

Figure 10 shows the entire block diagram with external circuitry of the CTI IC.

The lower CTI section affects signal processing for the two color difference signals in parallel circuits, as already described. Only one switching signal forming stage is incorporated, and this is triggered by the differentiating stage of the two channels. Thus, the signal switches will always work in parallel, so that transient improvement is also parallel in the two channels.

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Multi-Standard Color Decoder With Picture Improvement AN155A

The transient-improved color difference signals require a longer Y signal delay line with a delay time of up to 1000ns, which is additionally realized in this IC in gyrator technique.

A selection capability has been incorporated for the delay time, by means of a switching voltage, since the total required delay time is dependent on the overall television receiver concept. The delay line comprises a total of 11 gyrator all-pass elements with a delay time of 90ns each, making a total of 990ns. The group delay and frequency behavior of the gyrator delay line is very good up to 5MHz.

A switching stage permits optional by-pass of one, two, or three of these elements, so that a minimum of 8 × 90ns = 720ns is effective. The transient improvement of the color difference signal makes coincidence errors with respect to the Y signal especially visible. A slight increase in delay time by 45ns has therefore been provided for fine tuning, working via an IC pin to be connected to ground.

A signal tapping is available before the last delay element for a further picture improvement capability by means of deflection modulation.

Figure 11 depicts the circuit diagram of the TDA4565.

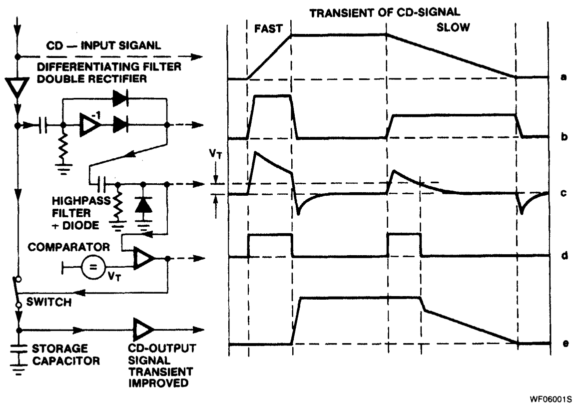
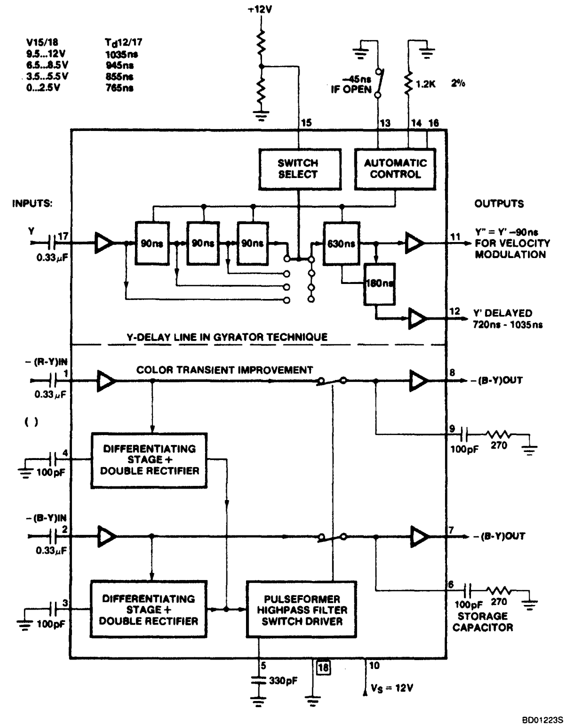


Figure 8. Function of CTI



NOTE:
* The TDA4565 is a high-performance TDA4560.

Figure 9. Block Diagram of Picture-Improvement IC TDA4565*

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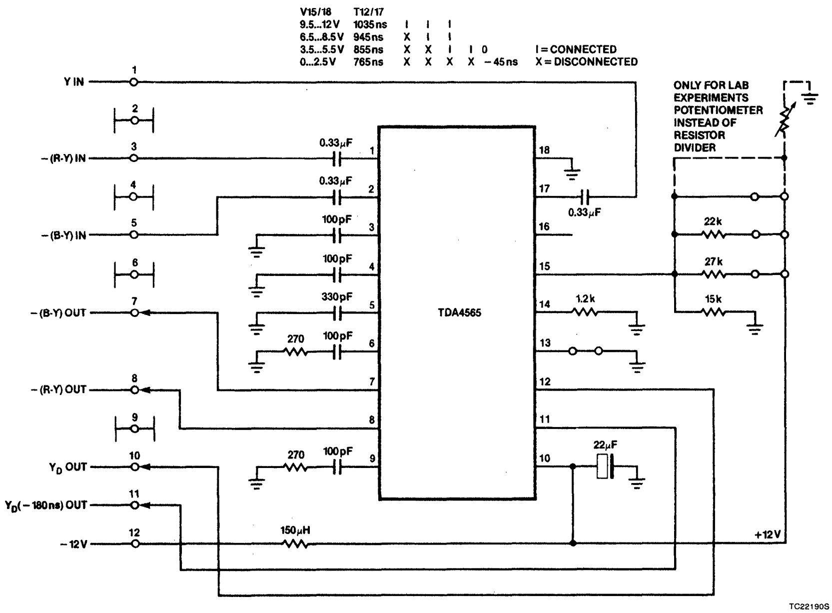


NOTE:
* If "Picture Signal Improvement" chip is *not* used, a 330ns delay line in Y channel, as shown in dotted lines, is required.

Figure 10. Various Combinations Used to Implement the Multi-Standard Color Decoder

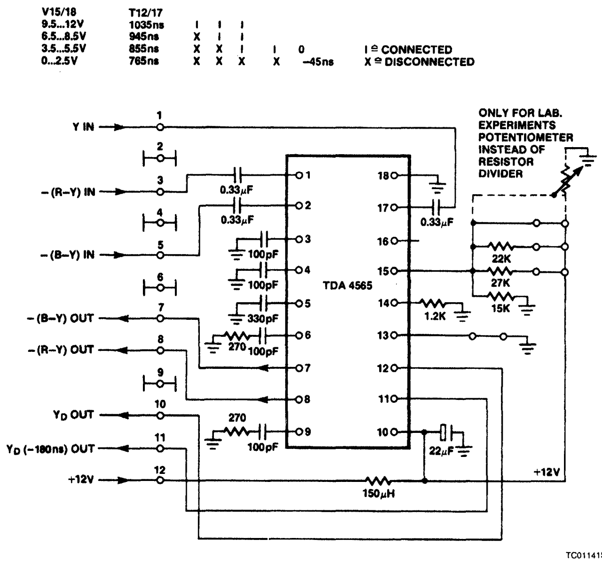


Figure 11. Circuit Diagram of TDA4565

Signetics

TDA3505
Chroma Control Circuit

Product Specification

Linear Products

DESCRIPTION

The TDA3505 performs the control functions in a PAL/SECAM decoder, which also comprises the TDA3510 (PAL decoder) and/or TDA3530 (SECAM decoder).

The required input signals are: luminance and color difference −(R-Y) and −(B-Y), while linear RGB signals can be inserted from external sources. RGB output signals are delivered for driving the video output stages. This circuit provides automatic cut-off control of the picture tube.

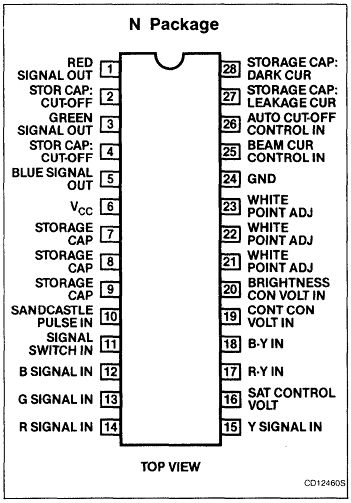
FEATURES

- Capacitive coupling of the color difference and luminance input signals with black level clamping in the input stages
- Linear saturation control in the color difference stages
- (G-Y) and RGB matrix
- Linear transmission of inserted signals
- Equal black levels for inserted and matrixed signals
- 3 identical channels for the RGB signals
- Linear contrast and brightness control, operating on both the inserted and matrixed RGB signals
- Peak beam current limiting input
- Horizontal and vertical blanking and clamping of the three input signals obtained via a 3-level sandcastle pulse
- DC gain controls for each of the RGB output signals (white point adjustment)
- Emitter-follower outputs for driving the RGB output stages
- Input for automatic cut-off control of the picture tube
- Compensation for leakage current of the picture tube

APPLICATIONS

- Video processing
- TV receivers

PIN CONFIGURATION



ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
28-Pin Plastic DIP (SOT-117)	−20°C to +80°C	TDA3505N

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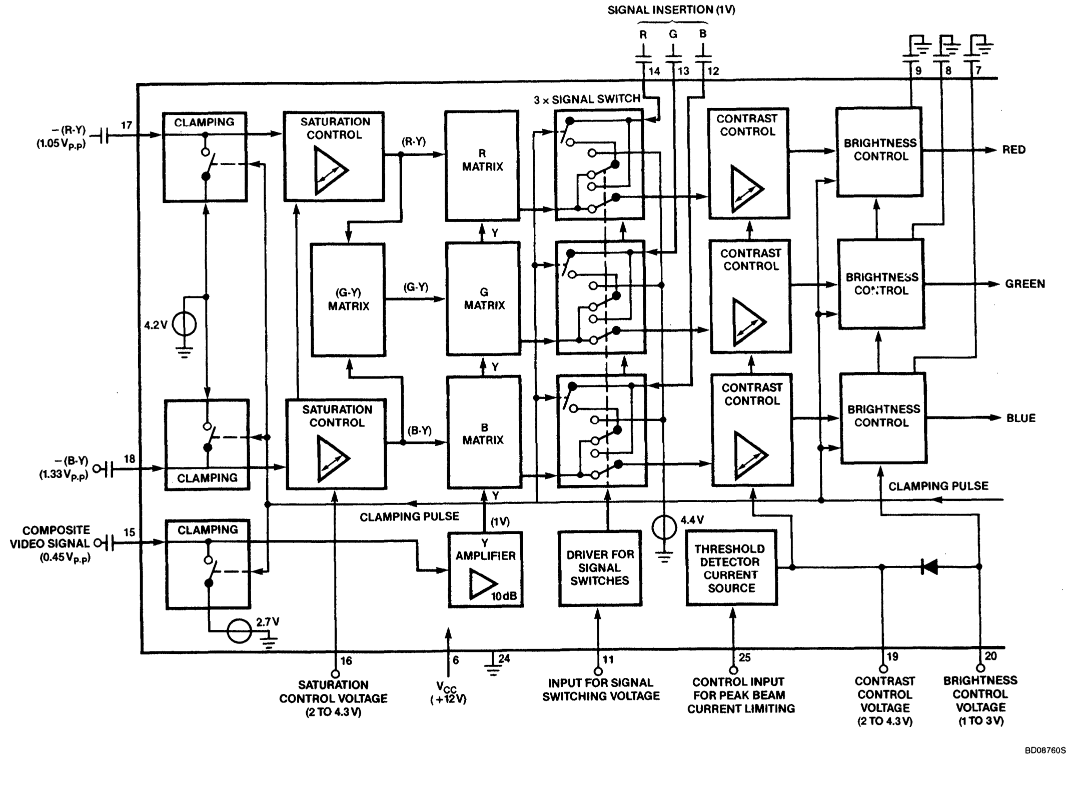
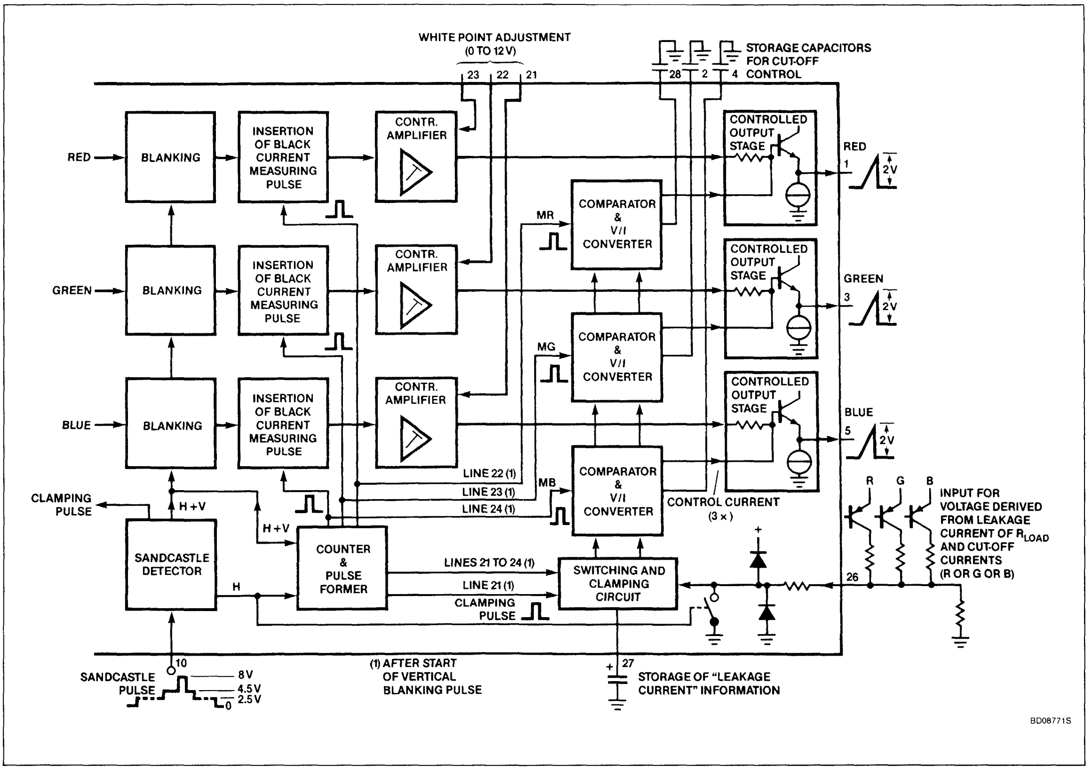
Chroma Control Circuit TDA3505

BLOCK DIAGRAM (PART A)



BLOCK DIAGRAM (PART B)



ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
$V_{CC} = V_{6-24}$	Supply voltage	13.2	V
	Voltages with respect to Pin 24		
V_{26-24}	Pin 26	V_{CC}	V
V_{25-24}	Pin 25	V_{CC}	V
V_{10-24}	Pin 10	V_{CC}	V
V_{11-24}	Pin 11	−0.5 to 3	V
$V_{16, 19, 20-24}$	Pins 16, 19, 20	0.5 V_{CC}	V
$V_{21, 22, 23-24}$	Pins 21, 22, 23	V_{CC}	V
No external DC voltage	Pins 1, 3, 5; 2, 4, 28; 7, 8, 9; 12, 13, 14; 15, 17, 18; 27		
	Currents		
$-I_{1, 3, 5}$	Pins 1, 3, 5	3	mA
I_{19}	Pin 19	10	mA
I_{20}	Pin 20	5	mA
$-I_{25}$	Pin 25	5	mA
P_{TOT}	Total power dissipation	1.7	W
T_{STG}	Storage temperature range	−65 to +150	°C
T_A	Operating ambient temperature range	−20 to +70	°C

Chroma Control Circuit | TDA3505

DC ELECTRICAL CHARACTERISTICS

The following characteristics are measured in a circuit similar to Figure 1; V_{CC} = 12V; T_A = 25°C; $V_{18-24(P-P)}$ = 1.33V; $V_{17-24(P-P)}$ = 1.05V; $V_{15-24(P-P)}$ = 0.45V; $V_{12,13,14-24(P-P)}$ = 1V, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS Min	Typ	Max	UNIT
$V_{CC} = V_{6-24}$	Supply voltage range	10.8		13.2	V
$I_6 = I_{CC}$	Supply current		85		mA
Color difference inputs					
$V_{18-24(P-P)}$	−(B-Y) input signal at Pin 18 (peak-to-peak value)		1.33		V
$V_{17-24(P-P)}$	−(R-Y) input signal at Pin 17 (peak-to-peak value)		1.05		V
$I_{17,\ 18}$	Input current during scanning			1	μA
$R_{17,\ 18-24}$	Input resistance	100			kΩ
$V_{17,\ 18-24}$	Internal DC voltage due to clamping		4.2		V
	Saturation control at Pin 16				
V_{16-24}	control voltage range for a change of saturation from −20dB to +6dB	2.1		4.3	V
V_{16-24}	control voltage for attenuation > 40dB			1.8	V
V_{16-24}	nominal saturation (6dB below maximum)		3.1		V
I_{16}	input current			20	μA
(G-Y) matrix					
$V_{(G-Y)} = -0.51\ V_{(R-Y)} -0.19\ V_{(B-Y)}$	Matrixed according to the equation				
Luminance amplifier (Pin 15)					
$V_{15-24(P-P)}$	Composite video input signal (peak-to-peak value)		0.45		V
R_{15-24}	Input resistance	100			kΩ
V_{15-24}	Internal DC voltage		2.7		V
I_{15}	Input current during scanning			1	μA
RGB channels					
	Signal switching input voltage for insertion (Pin 11)				
V_{11-24}	on level	0.9		3	V
V_{11-24}	off level			0.4	V
I_{11}	Input current	−100		+200	μA
	Signal insertion (Pin 12: blue; Pin 13: green; Pin 14: red)				
$V_{12,\ 13,\ 14-24(P-P)}$	external RGB input signal (black-to-white values)			1	V
$V_{12,\ 13,\ 14-24}$	internal DC voltage due to clamping[2]		4.4		V
$I_{12,\ 13,\ 14}$	input current during scanning			1	μA
	Contrast control (Pin 19)				
V_{19-24}	control voltage range for a change of contrast from −18dB to +3dB	2		4.3	V
V_{19-24}	nominal contrast (3dB below maximum)		3.6		V
V_{19-24}	control voltage for −6dB		2.8		V
I_{19}	input current at $V_{25-24} \geqslant 6V$			2	μA

DC ELECTRICAL CHARACTERISTICS (Continued) The following characteristics are measured in a circuit similar to Figure 1; $V_{CC} = 12V$; $T_A = 25°C$; $V_{18-24(P-P)} = 1.33V$; $V_{17-24(P-P)} = 1.05V$; $V_{15-24(P-P)} = 0.45V$; $V_{12,13,14-24(P-P)} = 1V$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
	Peak beam current limiting (Pin 25)				
V_{25-24}	internal DC bias voltage		5.5		V
R_{25-24}	input resistance		10		kΩ
I_{19}	input current at contrast control input at $V_{25-24} = 5.1V$		17		mA
V_{20-24}	Brightness control (Pin 20) control voltage range	1		3	V
$-I_{20}$	input current			10	μA
V_{20-24}	control voltage for nominal black level which equals the inserted artificial black level		2		V
ΔV_{20-24}	change of black level in the control range related to the nominal luminance signal (black-white)		50		%
	Internal signal limiting signal limiting for nominal luminance (black to white = 100%)				
	black		−25		%
	white		120		%
White point adjustment (Pin 21: blue; Pin 22: green; Pin 23: red)					
	AC voltage gain[3]				
	at $V_{21, 22, 23-24} = 5.5V$		100		%
	at $V_{21, 22, 23-24} = 0V$		60		%
	at $V_{21, 22, 23-24} = 12V$		140		%
$R_{21, 22, 23-24}$	Input resistance		20		kΩ
Emitter-follower outputs (Pin 1: red; Pin 3: green; Pin 5: blue)					
At nominal contrast, saturation, and white point adjustment					
$V_{1, 3, 5-24(P-P)}$	Output voltage (black-to-white signal, positive)		2		V
$V_{1, 3, 5-24}$	Black level without automatic cut-off control ($V_{28, 2, 4-24} = 10V$)		6.7		V
I_{SOURCE}	Internal current source		3		mA
$-\Delta V_{1, 3, 5-24}$	Cut-off current control range		4.6		V
Automatic cut-off control (Pin 26)					
The measurement occurs in the following lines after start of the vertical blanking pulse: line 21: measurement of leakage current; line 22: measurement of red cut-off current; line 23: measurement of green cut-off current; line 24: measurement of blue cut-off current					
V_{26-24}	Input voltage range	0		+6.5	V
ΔV_{26-24}	Voltage difference between cut-off current measurement and leakage current[4] measurement[5] Input 26 switches to ground during horizontal flyback		0.7		V

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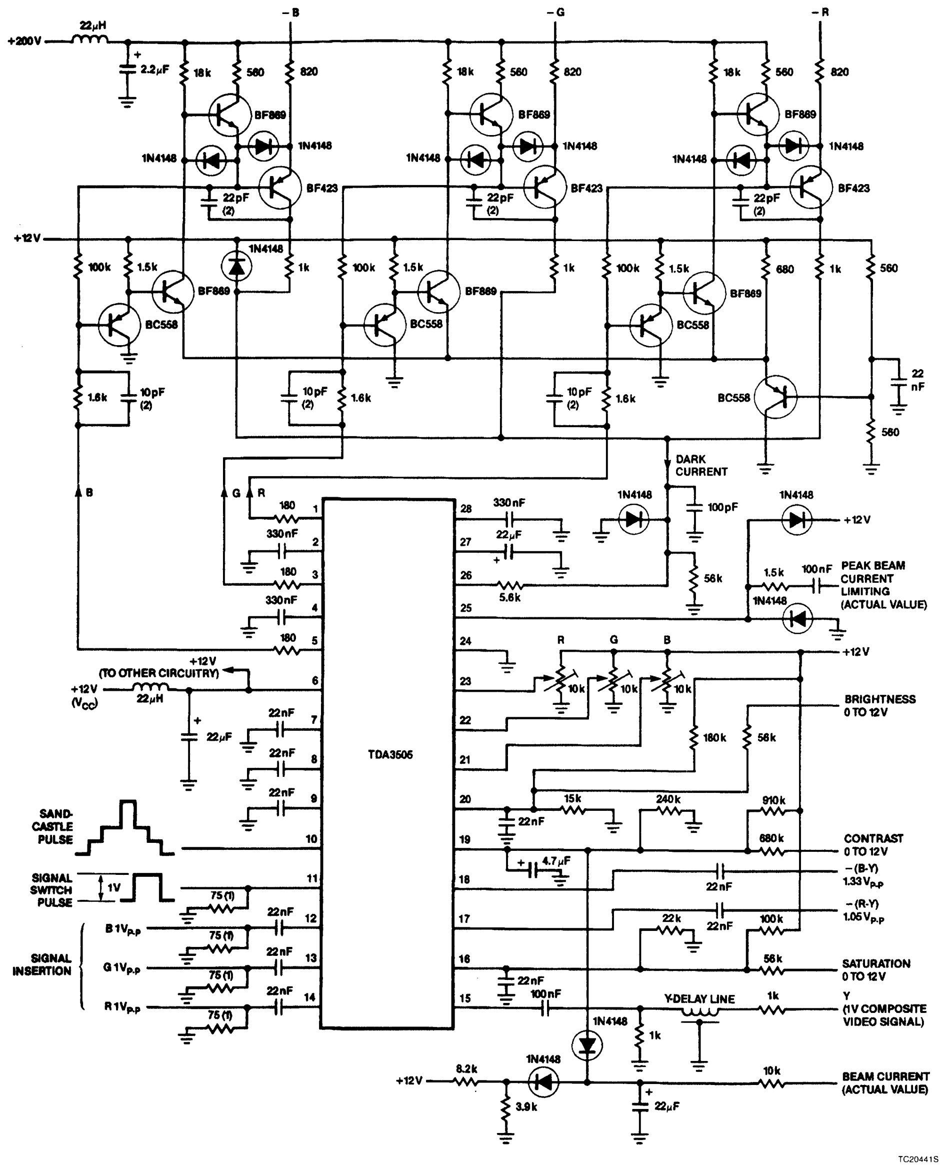
DC ELECTRICAL CHARACTERISTICS (Continued) The following characteristics are measured in a circuit similar to Figure 1; V_{CC} = 12V; T_A = 25°C; $V_{18-24(P-P)}$ = 1.33V; $V_{17-24(P-P)}$ = 1.05V; $V_{15-24(P-P)}$ = 0.45V; $V_{12,13,14-24(P-P)}$ = 1V, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Gain data					
At nominal contrast, saturation, and white point adjustment					
$G_{1,\ 3,\ 5-15}$	Voltage gain with respect to Y-input (Pin 15)		16		dB
$d_{1,\ 3,\ 5-15}$	Frequency response (0 to 5MHz)			3	dB
$G_{5-18} = G_{1-17}$	Voltage gain with respect to color difference inputs (Pins 17 and 18)		6		dB
$d_{5-18} = d_{1-17}$	Frequency response (0 to 2MHz)			3	dB
$G_{1-14} = G_{3-13} = G_{5-12}$	Voltage gain of inserted signals		6		dB
$d_{1-14} = d_{3-13} = d_{5-12}$	Frequency response (0 to 6MHz)			3	dB
Sandcastle detector (Pin 10)					
	There are 3 internal thresholds (proportional to V_{CC})[6]. The following amplitudes are required for separating the various pulses:				
V_{10-24}	horizontal and vertical blanking pulses[7]	2		3	V
V_{10-24}	horizontal pulse	4		5	V
V_{10-24}	clamping pulse[8]	7.5			V
V_{10-24}	DC voltage for artificial black level (scan and flyback)	7.5			V
V_{10-24}	no keying			1	V
$-I_{10}$	input current			110	μA

NOTES:
1. For saturated color bar with 75% of maximum amplitude.
2. $V_{11-24} < 0.4V$ during clamping time: the black levels of the inserted RGB signals are clamped on the black levels of the internal RGB signals.
   $V_{11-24} > 0.9V$ during clamping time: the black levels of the inserted signals are clamped on an internal DC voltage.
   Correct clamping of the external RGB signals is only possible when they are synchronous with the sandcastle pulse.
3. With input Pins 21, 22, and 23 not connected, an internal bias voltage of 5.5V is supplied.
4. Black level of measured channel is nominal; the other two channels are blanked to ultra-black.
5. All three channels blanked to ultra-black.
   The cut-off control cycle occurs when the vertical blanking part of the sandcastle pulse contains more than 3 line pulses.
   The internal signal blanking continues until the end of the last measurement line.
   The vertical blanking pulse is not allowed to contain more than 34 line pulses; otherwise, another control cycle begins.
6. The thresholds are for
   horizontal and vertical blanking: V_{10-24} = 1.5V
   horizontal pulse: V_{10-24} = 3.5V
   clamping pulse: V_{10-24} = 7.0V
7. Blanking to ultra-black (−25%).
8. Pulse duration ≥ 3.5μs.

Chroma Control Circuit TDA3505



NOTES:
1. When supplied via a 75Ω line.
2. Capacitor value depends on circuit layout.

Figure 1. Typical Application Circuit Diagram Using the TDA3505

Signetics

TDA3566
PAL/NTSC Decoder with RGB Inputs

Product Specification

Linear Products

DESCRIPTION

The TDA3566 is a monolithic, integrated decoder for the PAL and/or NTSC color television standards. It combines all functions required for the identification and demodulation of PAL/NTSC signals. Furthermore, it contains a luminance amplifier, and an RGB matrix and amplifier. These amplifiers supply output signals up to $4V_{P-P}$ (picture information) enabling direct drive of the discrete output stages. The circuit also contains separate inputs for data insertion, analog as well as digital, which can be used for text display systems (e.g., Teletext/broadcast antiope), channel number display, etc.

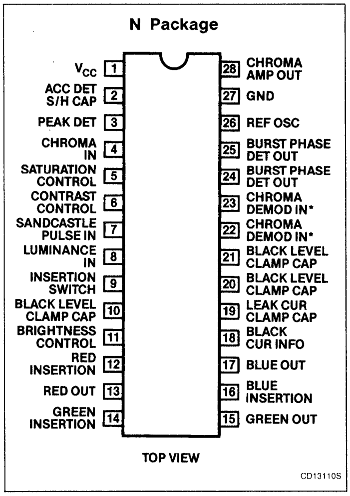
FEATURES

- **A black current stabilizer which controls the black currents of the three electron guns to a level low enough to omit the black level adjustment**
- **Contrast control of inserted RGB signals**
- **No black level disturbance when nonsynchronized external RGB signals are available on the inputs**
- **NTSC capability with hue control**
- **Single-chip chroma and luminance processor**
- **ACC with peak detector**
- **DC control settings**
- **External linear or digital RGB inputs**
- **High-level RGB outputs**
- **Luminance signal with clamp**
- **On-chip hue control for NTSC**

APPLICATIONS

- **Video monitors and displays**
- **Text display systems**
- **TV receivers**
- **Graphic systems**
- **Video processing**

PIN CONFIGURATION



ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
28-Pin Plastic DIP (SOT-117)	−25°C to +70°C	TDA3566N

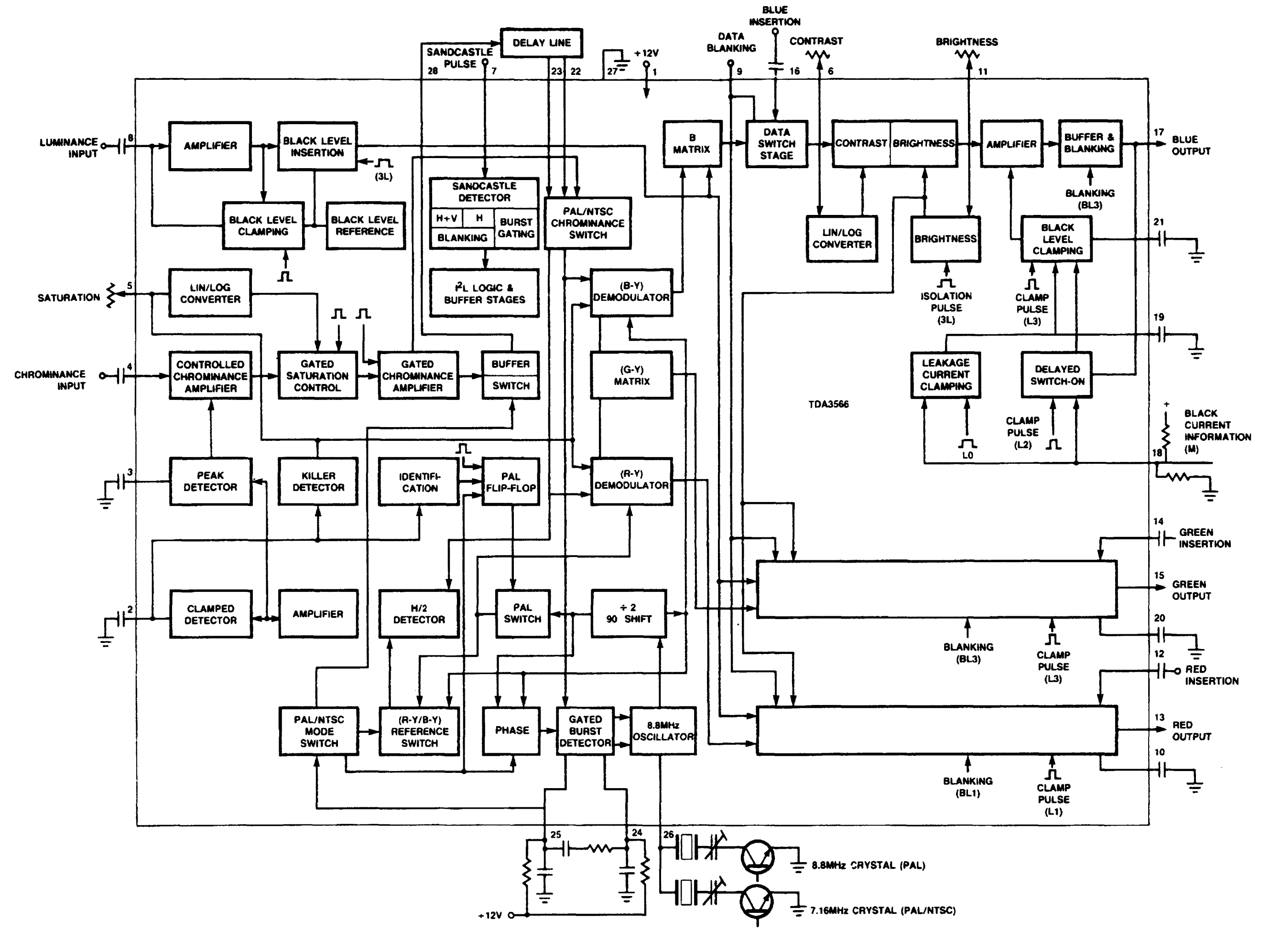
ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
$V_{CC} = V_{1-27}$	Supply voltage (Pin 1)	13.2	V
P_{TOT}	Total power dissipation	1.7	W
T_{STG}	Storage temperature range	−65 to +150	°C
T_A	Operating ambient temperature range	−25 to +70	°C
θ_{JA}	Thermal resistance from junction to ambient (in free air)	40	°C/W

PAL/NTSC Decoder with RGB Inputs

TDA3566

BLOCK DIAGRAM



NOTE:
For explanation of Pulse Mnemonics see Figure 5.

PAL/NTSC Decoder with RGB Inputs TDA3566

DC AND AC ELECTRICAL CHARACTERISTICS $V_{CC} = V_{1-27} = 12V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supply (Pin 1)					
$V_{CC} = V_{1-27}$	Supply voltage	10.8	12	13.2	V
$I_{CC} = I_1$	Supply current		80	110	mA
P_{TOT}	Total power dissipation		0.95	1.3	W
Luminance amplifier (Pin 8)					
$V_{8-27(P-P)}$	Input voltage[1] (peak-to-peak value)		0.45	0.63	V
V_{8-27}	Input level before clipping			1	V
I_8	Input current		0.1	1	μA
	Contrast control range (see Figure 1)	−15		+5	dB
I_7	Input current contrast control			15	μA
Chrominance amplifier (Pin 4)					
$V_{4-27(P-P)}$	Input voltage[2] (peak-to-peak value)	40	390	1100	mV
$\lvert Z_{4-27} \rvert$	Input impedance (Pin 4)		10		kΩ
C_{4-27}	Input capacitance			6.5	pF
	ACC control range	30			dB
ΔV	Change of the burst signal at the output over the whole control range			1	dB
A_V	Gain at nominal contrast/saturation Pin 4 to Pin 28[3]	34			dB
	Chrominance to burst ratio at nominal saturation at Pin 28[2, 3]		12		dB
$V_{28-27(P-P)}$	Maximum output voltage range (peak-to-peak value); $R_L = 2k\Omega$	4	5		V
d	Distortion of chrominance amplifier at $V_{28-27(P-P)} = 2V$ (output) up to $V_{4-27(P-P)} = 1V$ (input)			5	%
$\propto_{28-4}$	Frequency response between 0 and 5MHz			−2	dB
	Saturation control range (see Figure 2)	50			dB
I_5	Input current saturation control (Pin 5)			20	μA
	Cross-coupling between luminance and chrominance amplifier[4]			−46	dB
S/N	Signal-to-noise ratio at nominal input signal[5]	56			dB
$\Delta\varphi$	Phase shift between burst and chrominance at nominal contrast/saturation			±5	deg
$\lvert Z_{28-27} \rvert$	Output impedance of chrominance amplifier		10		Ω
I_{28}	Output current			15	mA

PAL/NTSC Decoder with RGB Inputs TDA3566

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $V_{CC} = V_{1-27} = 12V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS Min	Typ	Max	UNIT
Reference part					
	Phase-locked loop				
Δf	catching range[6]	500	700		Hz
$\Delta\varphi$	phase shift for ±400Hz deviation of f_{OSC}[6]			5	deg
	Oscillator				
TC_{OSC}	temperature coefficient of oscillator frequency[6]		−2	−3	Hz/°C
Δf_{OSC}	frequency variation when supply voltage increases from 10 to 13.2V[6]		40	100	Hz
R_{26-27}	input resistance (Pin 26)	280	400	520	Ω
C_{26-27}	input capacitance (Pin 26)			10	pF
	ACC generation (Pin 2)				
V_{2-27}	control voltage at nominal input signal		4.6		V
V_{2-27}	control voltage without chrominance input		2.6		V
V_{2-27}	color-off voltage		3.4		V
V_{2-27}	color-on voltage		3.6		V
V_{2-27}	identification-on voltage		2.0		V
	change in burst amplitude with temperature		0.1	0.25	%/°C
V_{3-27}	voltage at Pin 3 at nominal input signal		5.1		V
Demodulator part					
$V_{23-27(P-P)}$	Input burst signal amplitude[7] (peak-to-peak value) between Pins 23 and 27	68	80	95	mV
$\lvert Z_{22-27/23-27} \rvert$	Input impedance between Pins 22 or 23 and 27	0.7	1	1.3	kΩ
	Ratio of demodulated signals[8]				
$\frac{V_{17-27}}{V_{13-27}}$	(B-Y)/(R-Y)		1.78± 10%		
$\frac{V_{15-27}}{V_{13-27}}$	(G-Y)/(R-Y); no (B-Y) signal		−0.51± 10%		
$\frac{V_{15-27}}{V_{17-27}}$	(G-Y)/(B-Y); no (R-Y) signal		−0.19± 10%		
α_{17}	Frequency response between 0 and 1MHz			−3	dB
	Cross-talk between color difference signals	40			dB
$\Delta\varphi$	Phase difference between (R-Y) signal and (R-Y) reference signals			5	deg
$\Delta\varphi$	Phase difference between (R-Y) signal and (B-Y) reference signals	85	90	95	deg

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $V_{CC} = V_{1-27} = 12V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
RGB matrix and amplifiers					
$V_{13, 15, 17-27(P-P)}$	Output voltage (peak-to-peak value) at nominal luminance/contrast (black-to-white)[3]	3.5	4	4.5	V
$V_{13-27(P-P)}$	Output voltage at Pin 13 (peak-to-peak value) at nominal contrast/ saturation and no luminance signal to (R-Y)		4.2		V
$V_{13, 15, 17(m)}$	Maximum peak-white level	9.7	10	10.3	V
$I_{13, 15, 17}$	Available output current (Pins 13, 15, 17)	10			mA
$\Delta V_{13, 15, 17-27}$	Difference between black level and measuring level at the output for a brightness control voltage at Pin 11 of 2V[9]		0		V
ΔV	Difference in black level between the three channels without black current stabilization[10]			100	mV
	Control range of black-current stabilization $V_{CC1} = 3V$; $V_{11-17} = 2V$			±2	V
ΔV	Black level shift with vision contents			40	mV
	Brightness control voltage range		see Figure 2		
I_{11}	Brightness control input current			5	μA
$\Delta V/\Delta T$	Variation of black level with temperature		0		mV/°C
ΔV	Variation of black level with contrast*			100	mV
	Relative spread between the R, G, and B output signals			10	%
ΔV	Relative black-level variation between the three channels during variation of contrast, brightness, and supply voltage (± 10%)*		0	20	mV
ΔV	Differential black-level drift over a temperature range of 40°C		0	20	mV
V_{BL}	Blanking level at the RGB outputs		0.95	1.1	V
V_{BL}	Difference in blanking level of the three channels		0		mV
V_{BL}	Differential drift of the blanking levels over a temperature range of 40°C		0	10	mV
$\frac{\Delta V_{BL}}{V_{BL}} \times \frac{V_{CC}}{\Delta V_{CC}}$	Tracking of output black level with supply voltage	0.9	1	1.1	
	Tracking of contrast control between the three channels over a control range at 10dB			0.5	dB
V_O	Output signal during the clamp pulse (3L) after switch-on	7.5			V
S/N	Signal-to-noise ratio of output signals[5]	62			dB
$V_{R(P-P)}$	Residual 4.4MHz signal at RGB outputs (peak-to-peak value)			50	mV
$V_{R(P-P)}$	Residual 8.8MHz signal and higher harmonics at the RGB outputs (peak-to-peak value)			150	mV
$\|Z_{13, 15, 17-27}\|$	Output impedance of RGB outputs		50		Ω
α	Frequency response of total luminance and RGB amplifier circuits for f = 0 to 5MHz		−1	−3	dB
I_O	Current source of output stage	2	3		mA
ΔV	Difference of black level at the three outputs at nominal brightness*			10	mV
	Tracking of brightness control			2	%

NOTE:
*With respect to the measuring pulses.

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $V_{CC} = V_{1-27} = 12V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Signal insertion (Pins 12, 14, and 16)					
$V_{12, 14, 16-27(P-P)}$	Input signals (peak-to-peak value) for RGB output voltage of 4V (peak-to-peak) at nominal contrast	0.9	1	1.1	V
ΔV	Difference between the black levels of the RGB signals and the inserted signals at the output[11]			100	mV
t_R	Output rise time		50	80	ns
t_D	Differential delay time for the three channels		0	40	ns
$I_{12, 14, 16}$	Input current			10	μA
Data blanking (Pin 9)					
V_{9-27}	Input voltage for no data insertion			0.4	V
V_{9-27}	Input voltage for data insertion	0.9			V
$V_{9-27(m)}$	Maximum input voltage			3	V
t_D	Delay of data blanking			20	ns
R_{9-27}	Input resistance	7	10	13	kΩ
	Suppression of the internal RGB signals when $V_{9-27} > 0.9V$	46			dB
Sandcastle input (Pin 7)					
V_{7-27}	Level at which the RGB blanking is activated	1	1.5	2	V
V_{7-27}	Level at which the horizontal pulses are separated	3	3.5	4	V
V_{7-27}	Level at which burst gating and clamping pulse are separated	6.5	7.0	7.5	V
t_D	Delay between black level clamping and burst gating pulse		0.6		μs
 $-I_7$ I_7 I_7	Input current at $V_{7-27} = 0$ to 1V at $V_{7-27} = 1$ to 8.5V at $V_{7-27} = 8.5$ to 12V			 1 50 2	 mA μA mA
Black current stabilization (Pin 18)					
V_{18-27}	Bias voltage (DC)	3.5	5	7.0	V
ΔV	Difference between input voltage for 'black' current and leakage current	0.35	0.5	0.65	V
I_{18}	Input current during 'black' current			1	μA
I_{18}	Input current during scan			10	mA
V_{18-27}	Internal limiting at Pin 10	8.5	9	9.5	V
V_{18-27}	Switching threshold for 'black' current control ON	7.6	8	8.4	V
R_{18-27}	Input resistance during scan	1	1.5	2	kΩ
$I_{10, 20, 21}$	Input current during scan at Pins 10, 20, and 21 (DC)			TBD	nA
	Maximum charge/discharge current during measuring time		1		nA

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $V_{CC} = V_{1-27} = 12V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
NTSC					
V_{24-25}	Level at which the PAL/NTSC switch is activated (Pins 24 and 25)		8.8	9.2	V
$I_{24+25(AV)}$	Average output current[12]	75	90	105	μA
	Hue control	see Figure 4			

NOTES:
1. Signal with the negative-going sync; amplitude includes sync amplitude.
2. Indicated is a signal for a color bar with 75% saturation; chrominance to burst ratio is 2.2:1.
3. Nominal contrast is specified as the maximum contrast — 5dB and nominal saturation as the maximum saturation — 6dB.
4. Cross coupling is measured under the following condition: input signal nominal, contrast and saturation such that nominal output signals are obtained. The signals at the output at which no signal should be available must be compared with the nominal output signal at that output.
5. The signal-to-noise ratio is defined as peak-to-peak signal with respect to RMS noise.
6. All frequency variations are referred to 4.4MHz carrier frequency.
7. These signal amplitudes are determined by the ACC circuit of the reference part.
8. The demodulators are driven by a chrominance signal of equal amplitude for the (R-Y) and the (B-Y) components. The phase of the (R-Y) chrominance signal equals the phase of the (R-Y) reference signal. This also applies to the (B-Y) signals.
9. This value depends on the gain setting of the RGB output amplifiers and the drift of the picture tube guns. Higher black level values are possible (up to 5V), but in that application the amplitude of the output signal is reduced.
10. The variation of the black-level during brightness control in the three different channels is directly dependent on the gain of each channel. Discoloration during adjustment of contrast and brightness does not occur because amplitude and the black-level change with brightness control are directly related.
11. This difference occurs when the source impedance of the data signals is 150Ω and the black level clamp pulse width is 4μs (sandcastle pulse). For a lower impedance the difference will be lower.
12. The voltage at Pins 24 and 25 can be changed by connecting the load resistors (10kΩ in this application) to the slider bar of the hue control potentiometer (see Figure 7). When the transistor is switched on, the voltage at Pins 24 and 25 is reduced below 9V, and the circuit is switched to NTSC mode. The width of the burst gate is assumed to be 4μs typical.

FUNCTIONAL DESCRIPTION

The TDA3566 is a further development of the TDA3562A. It has the same pinning and almost the same application. The differences between the TDA3562A and the TDA3566 are as follows:

- The NTSC application has largely been simplified. In the case of NTSC, the chroma signal is now internally coupled to the demodulators, ACC, and phase detectors. The chroma output signal (Pin 28) is suppressed in this case. It follows that the external switches and filters which are needed for the TDA3562A are not needed for the TDA3566. Furthermore, there is no difference between the amplitude of the color output signals in the PAL or NTSC mode. The PAL/NTSC switch and the hue control of the TDA3566 and the TDA3562A are identical.
- The switch-on and the switch-off behavior of the TDA3566 has been improved. This has been obtained by suppressing the output signals during the switch-on and switch-off periods.
- The clamp capacitors connected to the Pins 10, 20, and 21 can be reduced to 100nF for the TDA3566. The clamp capacitors also receive a pre-bias voltage to avoid colored background during switch-on.
- The crystal oscillator circuit has been changed to prevent parasitic oscillations on the third overtone of the crystal. This has the consequence that optimal tuning capacitance must be reduced to 10pF.

Luminance Amplifier

The luminance amplifier is voltage driven and requires an input signal of 450mV peak-to-peak (positive video). The luminance delay line must be connected between the IF amplifier and the decoder. The input signal is AC coupled to the input (Pin 8). After amplification, the black level at the output of the preamplifier is clamped to a fixed DC level by the black clamping circuit. During three line periods after vertical blanking, the luminance signal is blanked out and the black level reference voltage is inserted by a switching circuit. This black level reference voltage is controlled via Pin 11 (brightness). At the same time, the RGB signals are clamped. Noise and residual signals have no influence during clamping; thus, simple internal clamping circuitry is used.

Chrominance Amplifiers

The chrominance amplifier has an asymmetrical input. The input signal must be AC coupled (Pin 4) and have a minimum amplitude of $40mV_{P-P}$. The gain control stage has a control range in excess of 30dB; the maximum input signal must not exceed $1.1V_{P-P}$ or clipping of the input signal will occur. From the gain-control stage, the chrominance signal is fed to the saturation control stage. Saturation is linear controlled via Pin 5. The control voltage range is 2 to 4V, the input impedance is high, and the saturation control range is in excess of 50dB. The burst signal is not affected by saturation control. The signal is then fed to a gated amplifier which has a 12dB higher gain during the chrominance signal. As a result, the signal at the output (Pin 28) has a burst-to-chrominance ratio which is 6dB lower than that of the input signal when the saturation control is set at −6dB. The chrominance output signal is fed to the delay line and, after matrixing, is applied to the demodulator input pins (Pins 22 and 23). These signals are fed to the burst phase detector. In the case of NTSC, the chroma signal is internally coupled to the demodulators, ACC, and phase detector.

Oscillator and Identification Circuit

The burst phase detector is gated with the narrow part of the sandcastle pulse (Pin 7). In the detector, the (R-Y) and (B-Y) signals are added to provide the composite burst signal again. This composite signal is compared to the oscillator signal divided-by-2 ((R-Y) reference signal). The control voltage is available at Pins 24 and 25, and is also applied to the 8.8MHz oscillator. The 4.4MHz signal is obtained via the divide-by-2 circuit, which generates both the (B-Y) and (R-Y) reference signals and provides a 90° phase shift between them.

The flip-flop is driven by pulses obtained from the sandcastle detector. For the identification of the phase at PAL mode, the (R-Y) reference signal coming from the PAL switch is compared to the vertical signal (R-Y) of the PAL delay line. This is carried out in the H/2 detector, which is gated during burst. When the phase is incorrect, the flip-flop gets a reset from the identification circuit. When the phase is correct, the output voltage of the H/2 detector is directly related to the burst amplitude so that this voltage can be used for the ACC. To avoid 'blooming-up' of the picture under weak input signal conditions, the ACC voltage is generated by peak detection of the H/2 detector output signal.

The killer and identification circuits get their information from a gated output signal of the H/2 detector. Killing is obtained via the saturation control stage and the demodulators to obtain good suppression. The time constant of the saturation control (Pin 5) provides a delayed switch-on after killing.

Adjustment of the oscillator is achieved by variation of the burst phase detector load resistance between Pins 24 and 25 (see Figure 6). With this application, the trimmer capacitor in series with the 8.8MHz crystal (Pin 26) can be replaced by a fixed value capacitor to compensate for imbalance of the phase detector.

Demodulator

The (R-Y) and (B-Y) demodulators are driven by the color difference signals from the delay-line matrix circuit and the reference signals from the 8.8MHz divider circuit. The (R-Y) reference signal is fed via the PAL-switch. The output signals are fed to the R and B matrix circuits and to the (G-Y) matrix to provide the (G-Y) signal which is applied to the G matrix. The demodulation circuits are killed and blanked by bypassing the input signals.

NTSC Mode

The NTSC mode is switched on when the voltage at the burst phase detector outputs (Pins 24 and 25) is adjusted below 9V. To ensure reliable application, the phase detector load resistors are external. When the TDA3566 is used only for PAL, these two 33kΩ resistors must be connected to +12V (see Figure 6). For PAL/NTSC application, the value of each resistor must be reduced to 10kΩ and connected to the slider of a potentiometer (see Figure 7). The switching transistor brings the voltage at Pins 24 and 25 below 9V, which switches the circuit to the NTSC mode. The position of the PAL flip-flop ensures that the correct phase of the (R-Y) reference signal is supplied to the (R-Y) demodulator. The drive to the H/2 detector is now provided by the (B-Y) reference signal. (In the PAL mode it is driven by the (R-Y) reference signal.)

Hue control is realized by changing the phase of the reference drive to the burst phase detector. This is achieved by varying the voltage at Pins 24 and 25 between 7.5V and 8.5V, nominal position 8.0V. The hue control characteristic is shown in Figure 4.

RGB Matrix and Amplifiers

The three matrix and amplifier circuits are identical and only one circuit will be described. The luminance and the color difference signals are added in the matrix circuit to obtain the color signal, which is then fed to the contrast control stage. The contrast control voltage is supplied to Pin 6 (high-input impedance). The control range is +3dB to −17dB nominal. The relationship between the control voltage and the gain is linear (see Figure 1).

During the 3-line period after blanking, a pulse is inserted at the output of the contrast control stage. The amplitude of this pulse is varied by a control voltage at Pin 11. This applies a variable offset to the normal black level, thus providing brightness control. The brightness control range is 1V to 3V.

While this offset level is present, the 'black-current' input impedance (Pin 18) is high and the internal clamp circuit is activated. The clamp circuit then compares the reference voltage at Pin 19 with the voltage developed across the external resistor network R_A and R_B (Pin 18) which is provided by picture tube beam current. The output of the comparator is stored in capacitors connected from Pins 10, 20, and 21 to ground, which controls the black level at the output. The reference voltage is composed by the resistor divider network and the leakage current of the picture tube into this bleeder. During vertical blanking, this voltage is stored in the capacitor connected to Pin 19, which ensures that the leakage current of the CRT does not influence the black current measurement.

The RGB output signals can never exceed a level of 10V. When the signal tends to exceed this level, the output signal is clipped. The black level at the outputs (Pins 13, 15, and 17) will be about 3V. This level depends on the spread of the guns of the picture tube. If a beam current stabilizer is not used, it is possible to stabilize the black levels at the outputs, which in this application must be connected to the black current measuring input (Pin 18) via a resistor network.

Data Insertion

Each color amplifier has a separate input for data insertion. A $1V_{P-P}$ input signal provides a $4V_{P-P}$ output signal. To avoid the 'black-level' of the inserted signal differing from the black level of the normal video signal, the data is clamped to the black level of the luminance signal. Therefore, AC coupling is required for the data inputs.

To avoid a disturbance of the blanking level due to the clamping circuit, the source impedance of the driver circuit must not exceed 150Ω.

The data insertion circuit is activated by the data blanking input (Pin 9). When the voltage at this pin exceeds a level of 0.9V, the RGB matrix circuits are switched off and the data amplifiers are switched on. To avoid colored edges, the data blanking switching time is short.

The amplitude of the data output signals is controlled by the contrast control at Pin 6. The black level is equal to the video black level and can be varied between 2 and 4V (nominal condition) by the brightness control voltage at Pin 11. Non-synchronized data signals do not disturb the black level of the internal signals.

Blanking of RGB and Data Signals

Both the RGB and data signals can be blanked via the sandcastle input (Pin 7). A slicing level of 1.5V is used for this blanking function, so that the wide part of the sandcastle pulse is separated from the remainder of the pulse. During blanking, a level of +1V is available at the output. To prevent parasitic oscillations on the third overtone of the crystal, the optimal tuning capacitance should be 10pF.

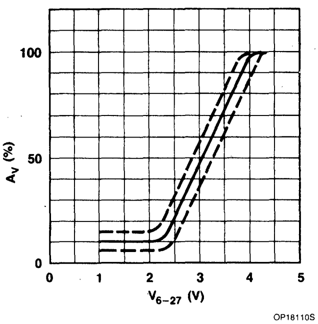


Figure 1. Contrast Control Voltage Range

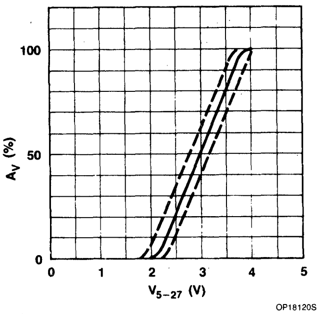


Figure 2. Saturation Control Voltage Range

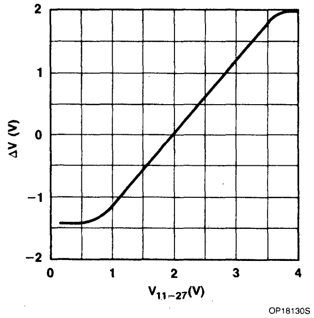


Figure 3. Difference Between Black Level and Measuring Level at the RGB Outputs (ΔV) as a Function of the Brightness Control Input Voltage (V_{11-27})

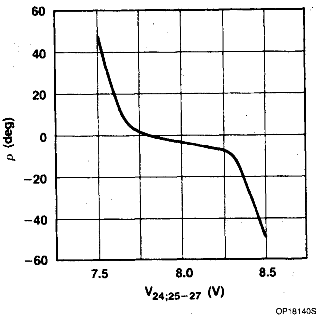


Figure 4. Hue Control Voltage Range

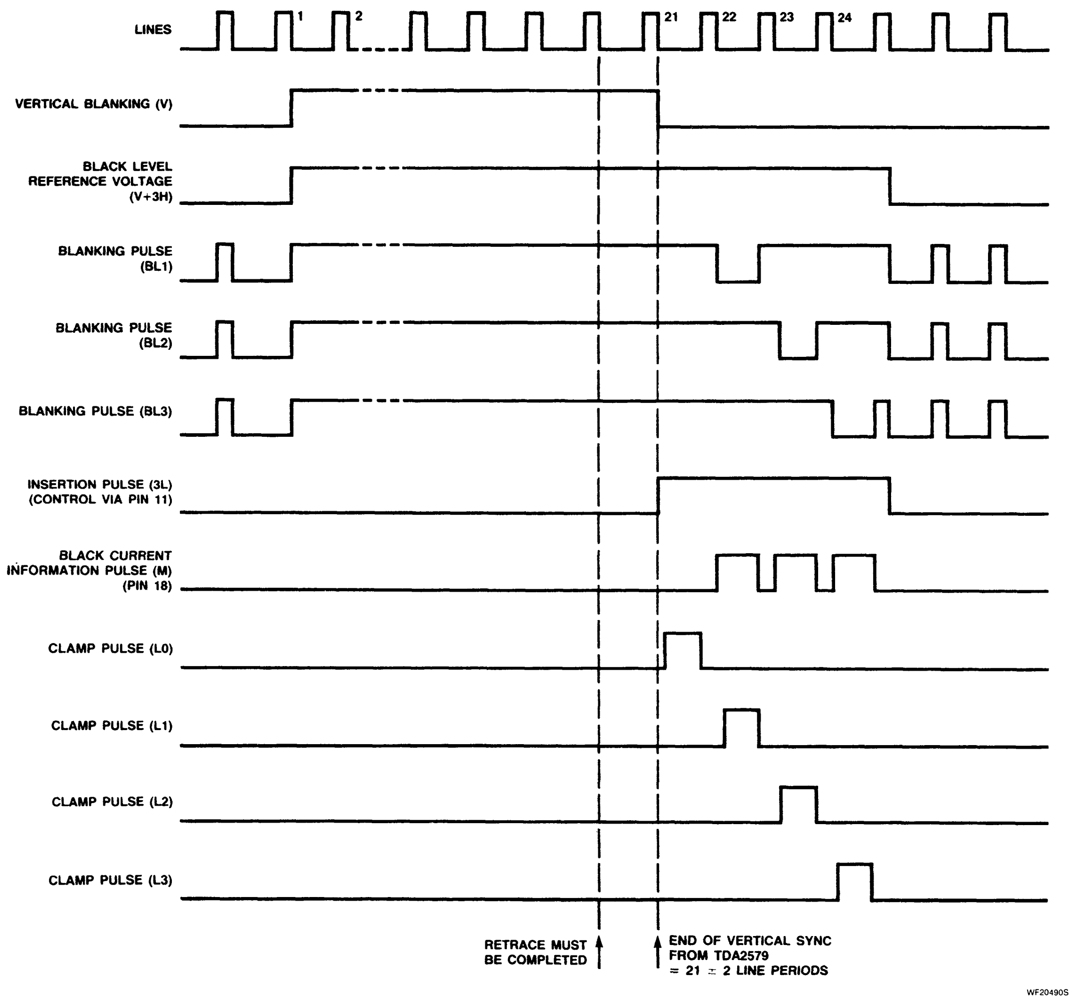
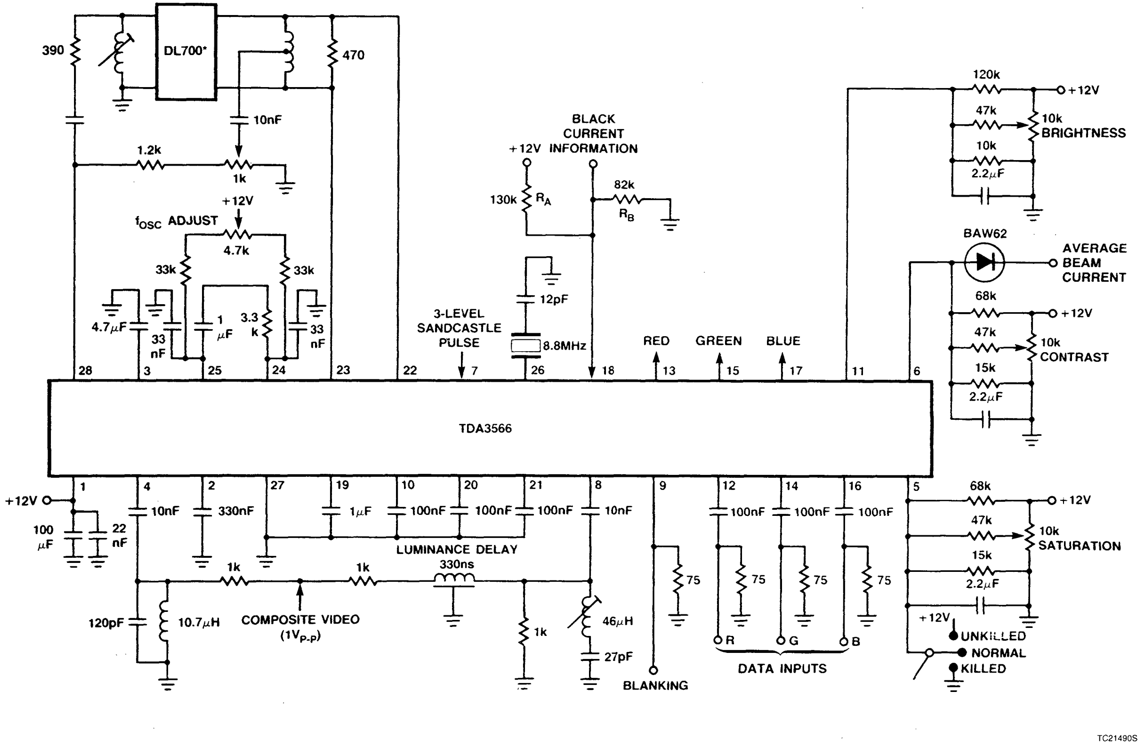


Figure 5. Timing Diagram for Black Current Stabilizing



NOTE:
*D1700 AMPEREX CORP.

Figure 6. Application Diagram Showing the TDA3566 for a PAL Decoder

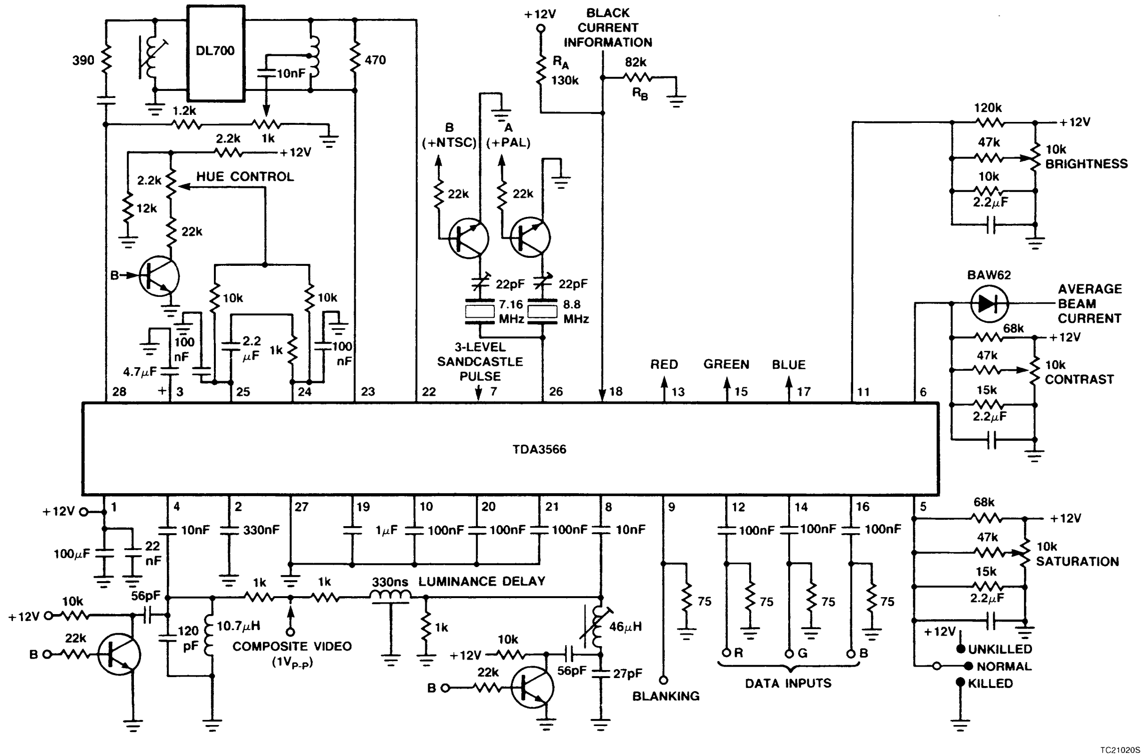
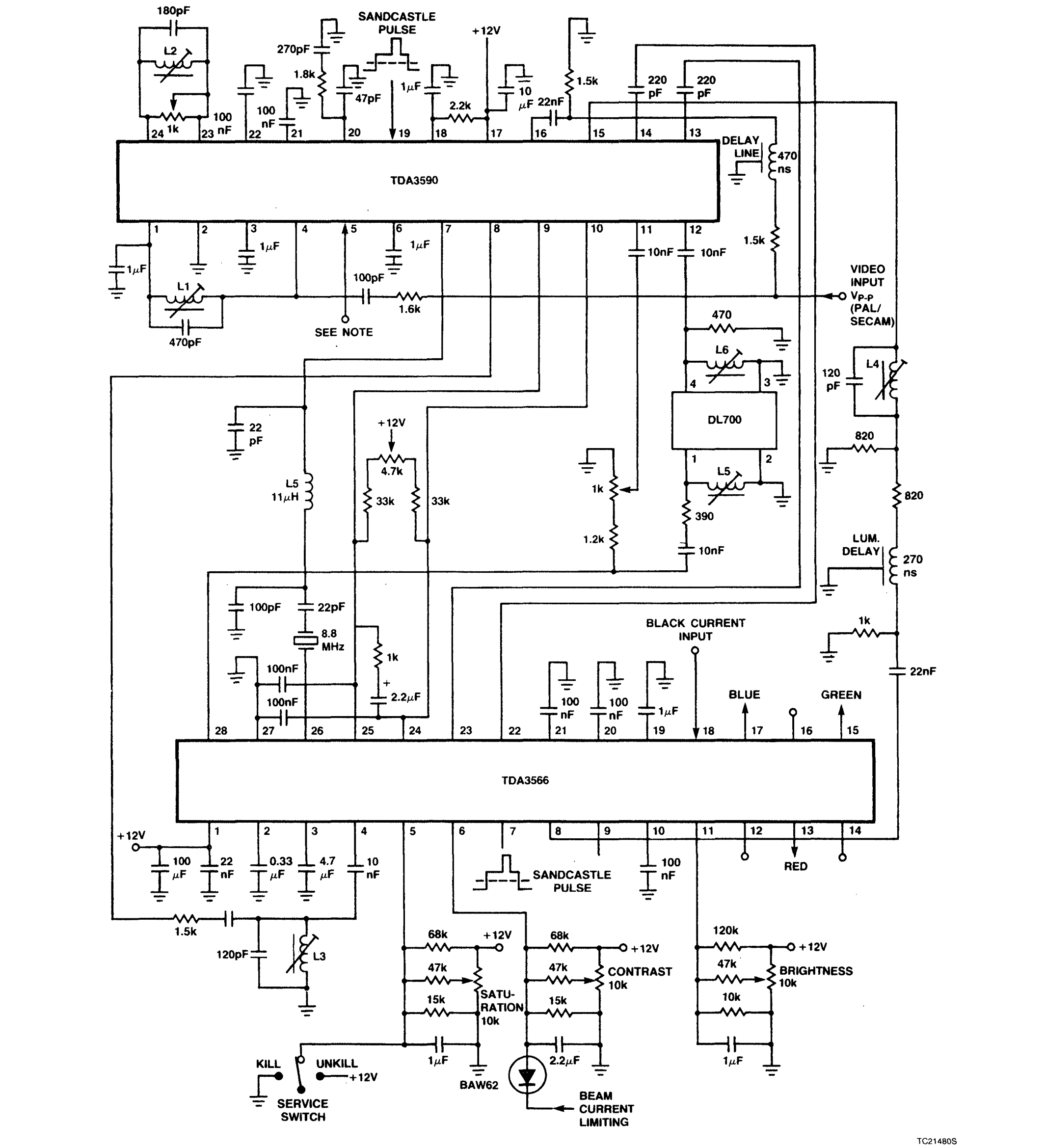


Figure 7. Application Diagram Showing the TDA3566 for a PAL/NTSC Decoder



NOTES:
Note to Pin 5 TDA3590:
$V_{5-2} < 1V$; horizontal identification and black level clamping.
$V_{5-2} > 11V$; vertical identification and artificial black level.
$V_{5-2} = 5$ to 7V; horizontal identification and artificial black level.

Figure 8. PAL/SECAM Application Circuit Diagram Using the TDA3590 and TDA3566

Signetics

TDA3567
NTSC Color Decoder

Product Specification

Linear Products

DESCRIPTION

The TDA3567 is a monolithic integrated decoder for the NTSC color television standards. It combines all functions required for the demodulation of NTSC signals. Furthermore, it contains a luminance amplifier, and an RGB-matrix and amplifier. These amplifiers supply output signals up to $5V_{P-P}$ (picture information) enabling direct drive of the discrete output stages.

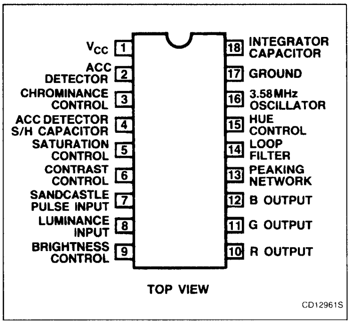
FEATURES

- **Single-chip chroma and luminance processor**
- **ACC with peak detector**
- **DC control settings**
- **High-level RGB outputs**
- **Luminance signal with clamp**
- **Requires few external components**
- **On-chip hue control circuit**

APPLICATIONS

- **Video monitors and displays**
- **TV receivers**
- **Video processing**

PIN CONFIGURATION



ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
18-Pin Plastic DIP (SOT-102HE)	−25°C to +65°C	TDA3567N

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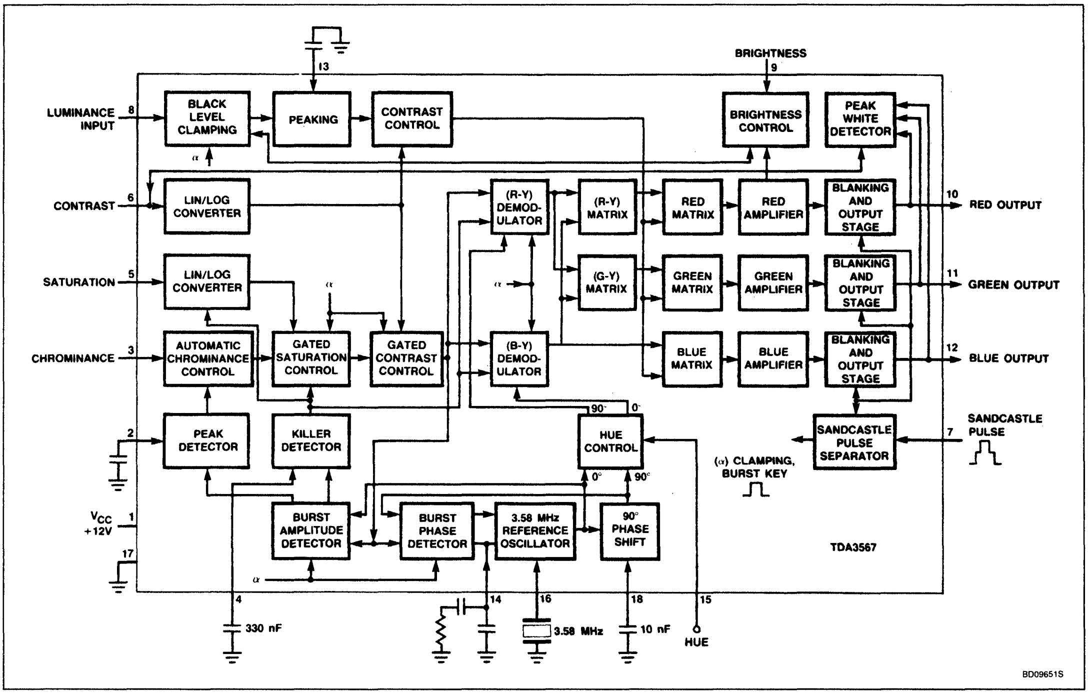
BLOCK DIAGRAM



ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
$V_{CC} = V_{1-17}$	Supply voltage	13.2	V
P_{TOT}	Total power dissipation	1.7	W
T_{STG}	Storage temperature range	−25 to +150	°C
T_A	Operating ambient temperature range	−25 to +65	°C
θ_{JA}	Thermal resistance from junction to ambient (in free-air)	50	°C/W

DC AND AC ELECTRICAL CHARACTERISTICS $V_{CC} = V_{1-17} = 12V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	CONDITIONS	LIMITS			UNIT
			Min	Typ	Max	
Supply						
$V_{CC} = V_{1-17}$	Supply voltage		9	12	13.2	V
$I_{CC} = I_1$	Supply current			65		mA
P_{TOT}	Total power dissipation			0.78		W
Luminance input signal						
$V_{8-17(P-P)}$	Input voltage[1] (peak-to-peak value)	Pin 8		450		mV
V_{8-17}	Input voltage level before clipping occurs in the input stage				1	V
I_8	Input current			0.15	1	μA
	Contrast control range	See Figure 1	−17		+3	dB
I_7	Input current contrast control	For $V_{6-17} < 6V$		0.5	15	μA
I_7	Input current when the peak white limiter is active	$V_{6-17} = 2.5V$		5.5		mA
R_{7-17}	Input resistance	$V_{6-17} > 6V$	1.4	2	2.6	kΩ
Peaking of luminance signal						
$\lvert Z_{13-17} \rvert$	Output impedance	Pin 13		200		Ω
	Ratio of internal/external current when Pin 13 is short-circuited			3		
Chrominance amplifier						
$V_{3-17(P-P)}$	Input signal amplitude[2] (peak-to-peak value)	Pin 3		550		mV
$V_{3-17(P-P)}$	Input signal amplitude before clipping occurs in the input stage (peak-to-peak value)				1.1	V
	Minimum burst signal amplitude within the ACC control range (peak-to-peak)		35			mV
	ACC control range		30			dB
ΔV	Change of the burst signal at the output for the complete control range				+1	dB
$\lvert Z_{3-17} \rvert$	Input impedance	Pin 3	6	8	10	kΩ
C_{3-17}	Input capacitance	Pin 3		4	6	pF
	Saturation control range	See Figure 3	50			dB
I_5	Input current saturation control	For $V_{5-17} > 6V$		1	20	μA
$\lvert Z_{5-17} \rvert$	Input impedance	$V_{5-17} = 6V$ to $10V$	1.4	2	2.6	kΩ
$\lvert Z_{5-17} \rvert$	Input impedance when the color killer is active		1.4	2	2.6	kΩ
$\lvert Z_{5-17} \rvert$	Input impedance	For $V_{5-17} > 10V$	0.7	1	1.3	kΩ
	Tracking between luminance and chrominance contrast	For 10dB of control		1	2	dB
	Cross-coupling between luminance and chrominance amplifier[4]			−50	−46	dB
Reference part phase-locked loop						
Δf	Catching range		±400	±500		Hz
Δ	Phase shift for 400Hz deviation of the carrier frequency				5	deg

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DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $V_{CC} = V_{1-17} = 12V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	CONDITIONS	LIMITS Min	LIMITS Typ	LIMITS Max	UNIT
Oscillator						
TC_{OSC}	Temperature coefficient of oscillator frequency			1.5	2.5	Hz/°C
Δf_{OSC9}	Frequency deviation	$\Delta V_{CC} = \pm 10\%$		150	250	Hz
R_{16-17}	Input resistance	Pin 16	260	360	460	Ω
C_{22-17}	Input capacitance	Pin 16			10	pF
ACC generation						
V_{4-17}	Voltage at Pin 4 nominal input signal			4		V
V_{4-17}	Voltage at Pin 4 without burst input			1.9		V
V_{4-17}	Color-off voltage			2.5		V
V_{4-17}	Color-on voltage			2.8		V
	Change in burst amplitude with temperature			0.1		%/°C
	Change in burst amplitude with 10% supply voltage change			0		%/V
V_{2-17}	Voltage at Pin 2 at nominal input signal			5		V
Hue control						
	Control voltage range		see Figure 4			
I_{14}	Input current	for $V_{15-17} < 5V$		0.5	20	μA
$\lvert Z_{14-17}\rvert$	Input impedance	for $V_{15-17} > 5V$	1.5	2.5	3.5	kΩ
Demodulation part						
	Ratio of demodulation signals (measured at the various outputs)[7]					
$\frac{V_{10-17}}{V_{12-17}}$	(R-Y)/(B-Y); no (R-Y) signal			−0.42		
$\frac{V_{10-17}}{V_{12-17}}$	(R-Y)/(B-Y); color bar signal			1.4		
$\frac{V_{11-17}}{V_{12-17}}$	(G-Y)/(R-Y); no (B-Y) signal			−0.25		
$\frac{V_{11-17}}{V_{12-17}}$	(G-Y)/(B-Y); no (R-Y) signal			−0.11		
	Frequency response	0 to 0.7MHz			−3	dB
RGB matrix and amplifier						
$V_{10,\ 11,\ 12-17(P-P)}$	Output signal amplitude[3]	at nominal luminance input signal and nominal contrast (peak-to-peak value) black-white	4	5	6	V
$V_{12-17(P-P)}$	Output signal amplitude of the "blue" channel	at nominal contrast and saturation control setting and no luminance signal to the input (B-Y) signal (peak-to-peak value)		3.8		V
$V_{10,\ 11,\ 12-7}$	Maximum peak-white level[6]		9	9.3	9.6	V

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $V_{CC} = V_{1-17} = 12V$; $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	CONDITIONS	LIMITS			UNIT
			Min	Typ	Max	
$I_{10,\ 11,\ 12-17}$	Maximum output current				10	mA
	Difference in the black level between the three channels				600	mV
	Black level shift with vision content			10	40	mV
	Brightness control voltage range		see Figure 3			
I_9	Brightness control input current				-50	μA
V/T	Black level variation with temperature			0.15	1	mV/°C
ΔV	Black level variation with contrast control			75	200	mV
	Relative spread between the three output signals				10	%
ΔV	Relative variation in black level between the three channels	during variations of contrast (10dB), brightness (± 1V), and supply voltage (± 10%)		0	20	mV
ΔV	Differential drift of black level over a temperature range of 40°C			0	20	mV
V_{B1}	Blanking level at the RGB outputs		1.95	2.15	2.35	V
$\frac{\Delta V_{B1}}{V_{B1}} \times \frac{V_{CC}}{\Delta V_{CC}}$	Tracking of output black levels with supply voltage		1	1.05	1.1	
S/N	Signal-to-noise ratio of output signals[5]		62			dB
$V_{R(P-P)}$	Residual 3.58MHz in RGB outputs (peak-to-peak value)			50	75	mV
$V_{R(P-P)}$	Residual 7.1MHz and higher harmonics in the RGB outputs (peak-to-peak value)			50	75	mV
$\lvert Z_{10,\ 11,\ 12-17} \rvert$	RGB output impedance				50	Ω
	Frequency response of total luminance and RGB amplifier circuits	0 to 5MHz			-3	dB
Sandcastle input						
V_{7-17}	Level at which the RGB blanking is activated		1	1.5	2	V
V_{7-17}	Level at which burst gate clamping pulses are separated		6.5	7	7.5	V
t_D	Delay between black level clamping and burst gating pulse		300	375	450	ns
I_7 I_7 I_7	Input currents	$V_{7-17} = 0$ to 1V $V_{7-17} = 1$ to 8.5V $V_{7-17} = 8.5$ to 12V		 -20 	-1 -40 2	mA μA mA

NOTES:
1. Signal with negative-going sync; amplitude includes sync pulse amplitude.
2. Indicated is a signal for color bar with 75% saturation, so the chrominance-to-burst ratio is 2.2:1.
3. Nominal contrast is specified as maximum contrast −3dB and nominal saturation as maximum saturation −10dB.
4. Cross-coupling is measured under the following conditions:
   - input signals nominal
   - contrast and saturation such that nominal output signals are obtained
   - the signals at the output at which no signal should be available must be compared with the nominal output signal at that output.
5. The signal-to-noise ratio is specified as peak-to-peak signal with respect to RMS noise.
6. When this level is exceeded, the amplifier of the output signal is reduced via a discharge of the capacitor on Pin 7 (contrast control). Discharge current is 5.5mA.
7. These matrixed values are found by measuring the ratio of the various output signals. The values are derived from the matrix equations given in the section 'FUNCTIONAL DESCRIPTION'.

FUNCTIONAL DESCRIPTION

Luminance Amplifier

The luminance amplifier is voltage driven and requires an input signal of $450mV_{P-P}$ [1]. The luminance delay line must be connected between the IF amplifier and the decoder. The input signal must be AC coupled to the input Pin 8.

The black level clamp circuit of the RGB amplifiers uses the coupling capacitor as a storage capacitor. After clamping, the signal is fed to a peaking stage. The RC network connected to Pin 13 is used to define the amount of overshoot.

The peaking stage is followed by a contrast control stage. The control voltage has to be supplied to Pin 6. The control voltage range is nominally −17 to +3dB. The linear curve of the contrast control voltage is shown in Figure 1.

Chrominance Amplifier

The chrominance amplifier has an asymmetrical input. The input signal at Pin 3 must be AC coupled, and must have an amplitude of $550mV_{P-P}$. The gain control stage has a control range in excess of 30dB, the maximum input signal should not exceed $1.1V_{P-P}$, otherwise clipping of the input signal will occur. From the gain control stage, the chrominance signal is fed to the saturation and contrast control stages. Chrominance and luminance control stages are directly coupled to obtain good tracking. The saturation is linearly controlled via Pin 5. The control voltage range is 2V to 4V. The impedance is high and the saturation control range is in excess of 50dB. The burst signal is not affected by contrast or saturation control. After the amplification and control stages, the chrominance signal is internally fed to the (R-Y) and (B-Y) demodulators, burst phase, and ACC detectors.

Oscillator and ACC Circuit

The 3.58MHz reference oscillator operates at the subcarrier frequency. The crystal must be connected between Pin 16 and ground. The oscillator does not require adjustment due to the small spreads of the IC. The free-running frequency of the oscillator can be checked by connecting the saturation control (Pin 5) to the positive supply line. Then the loop is opened so that the frequency can be measured. The oscillator has an internal gain-limiting stage which controls the gain to unity, so that internal signals are sinusoidal. This prevents the generation of higher harmonics of the subcarrier signals. The burst signal is compared to a 0° reference signal by the burst amplitude detector, and is then amplified and fed to a peak detector for ACC and to a sample-and-hold circuit which drives the color-killer circuit. The reference signal for the burst phase detector is provided by the 90° phase-shifted signal. An RC network is used to obtain the required catching range and noise immunity for the output voltage of the burst phase detector.

The hue control is obtained by mixing oscillator signals with a phase of 0° and 90° before they are fed to the (R-Y) and (B-Y) demodulators. The 90° phase-shifted signal is provided by a Miller integrator (biased by Pin 18). As the hue control is independent of the PLL, the control will react without time delay on the control voltage changes.

Demodulator Circuits

The demodulators are driven by the amplified and controlled chrominance signals; the reference signals are obtained from the hue control circuit. In nominal hue control position, the phase angle of (R-Y) reference signal is 0°; the phase angle of the (B-Y) reference signal is 90°.

For flesh-tone corrections, the demodulated (R-Y) signal is matrixed with the demodulated (B-Y) signal according to the following equations:

$$(R\text{-}Y)_{matrixed} = 1.61\ (R\text{-}Y)_{IN} - 0.42\ (B\text{-}Y)_{IN}$$

$$(G\text{-}Y)_{matrixed} = 0.43\ (R\text{-}Y)_{IN} - 0.11\ (B\text{-}Y)_{IN}$$

$$(B\text{-}Y)_{matrixed} = (B\text{-}Y)_{IN}$$

In these equations $(R\text{-}Y)_{IN}$ and $(B\text{-}Y)_{IN}$ indicate the color difference signal amplitudes when the chrominance signal is demodulated with a phase difference between the R-Y and B-Y demodulator of 90° and a gain ratio B-Y/R-Y = 1.78.

RGB Matrix Circuit and Amplifiers

The three matrix and amplifier circuits are identical. The luminance signal and the color difference signals are added in the matrix circuit to obtain the color signal.

Output signals are $5V_{P-P}$ (black-white) for the following nominal input signals and control settings:

- Luminance $450mV_{P-P}$
- Chrominance $550mV_{P-P}$ (burst-to-chrominance ratio of the input 1:2.2)
- Contrast −3dB (maximum)
- Saturation −10dB (maximum)

The maximum available output voltage is approximately $7V_{P-P}$. The black level of the red channel is compared to a variable external reference level (Pin 9), which provides the brightness control. The control loop is closed via the luminance input.

The luminance input is varied to control the black level control; therefore, the green and blue outputs will follow any variation of the red output. The output of the black control can be varied between 2V to 4V. The corresponding brightness control voltage is shown in Figure 3.

If the output signal surpasses the level of 9V, the peak white limiter circuit becomes active and reduces the output signal via the contrast control.

Blanking of RGB Signals

A slicing level of about 1.5V is used for this blanking function, so that the wide part of the sandcastle pulse is separated from the rest of the pulse. During blanking, a level of +2V is available at the output.

NOTE:
1. Signal with negative-going sync; amplitude includes sync pulse amplitude.

NTSC Color Decoder TDA3567

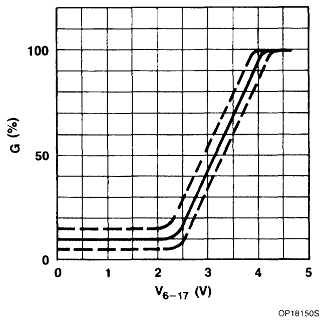

Figure. 1. Contrast Control Voltage Range

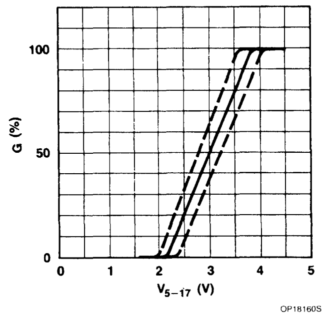

Figure 2. Saturation Control Voltage Range

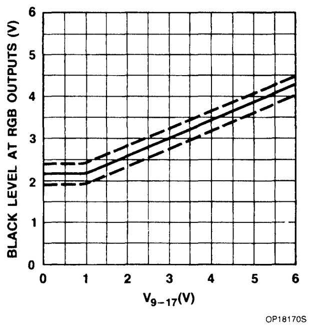

Figure 3. Brightness Control Voltage Range

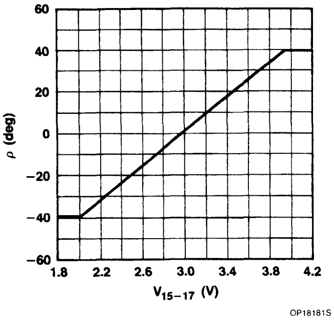

Figure 4. Hue Control Voltage Range

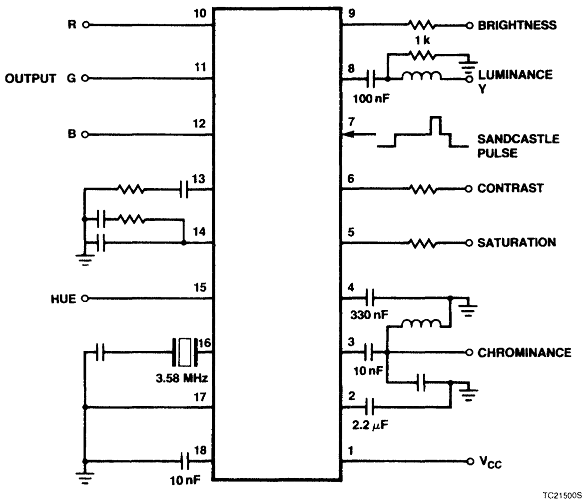

Figure 5. Application Diagram

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Signetics

TDA4555/56
Multistandard Color Decoder

Product Specification

Linear Products

DESCRIPTION

The TDA4555 and TDA4556 are monolithic, integrated, multistandard color decoders for the PAL®, SECAM, NTSC 3.58MHz and NTSC 4.43MHz standards. The difference between the TDA4555 and the TDA4556 is the polarity of the color difference output signals (B-Y) and (R-Y).

FEATURES

Chrominance Part

- **Gain-controlled chrominance amplifier for PAL, SECAM, and NTSC**
- **ACC rectifier circuits (PAL/NTSC, SECAM)**
- **Burst blanking (PAL) in front of 64μs glass delay line**
- **Chrominance output stage for driving the 64μs glass delay line (PAL, SECAM)**
- **Limiter stages for direct and delayed SECAM signal**
- **SECAM permutator**

Demodulator Part

- **Flyback blanking incorporated in the two synchronous demodulators (PAL, NTSC)**
- **PAL switch**
- **Internal PAL matrix**
- **Two quadrature demodulators with external reference-tuned circuits (SECAM)**
- **Internal filtering of residual carrier**
- **De-emphasis (SECAM)**
- **Insertion of reference voltages as achromatic value (SECAM) in the (B-Y) and (R-Y) color difference output stages (blanking)**

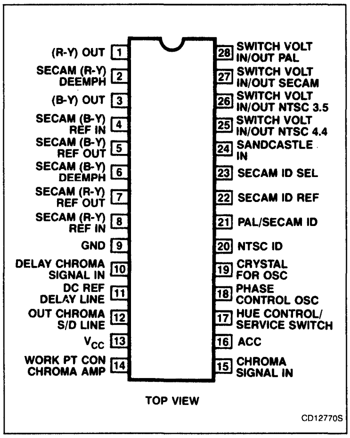
Identification Part

- **Automatic standard recognition by sequential inquiry**
- **Delay for color-on and scanning-on**
- **Reliable SECAM identification by PAL priority circuit**
- **Forced switch-on of a standard**
- **Four switching voltages for chrominance filters, traps, and crystals**
- **Two identification circuits for PAL/SECAM (H/2) and NTSC**
- **PAL/SECAM flip-flop**
- **SECAM identification mode switch (horizontal, vertical, or combined horizontal and vertical)**
- **Crystal oscillator with divider stages and PLL circuitry (PAL, NTSC) for double color subcarrier frequency**
- **HUE control (NTSC)**
- **Service switch**

APPLICATIONS

- **Video monitors**
- **Video processing**
- **TV receivers**

PIN CONFIGURATION

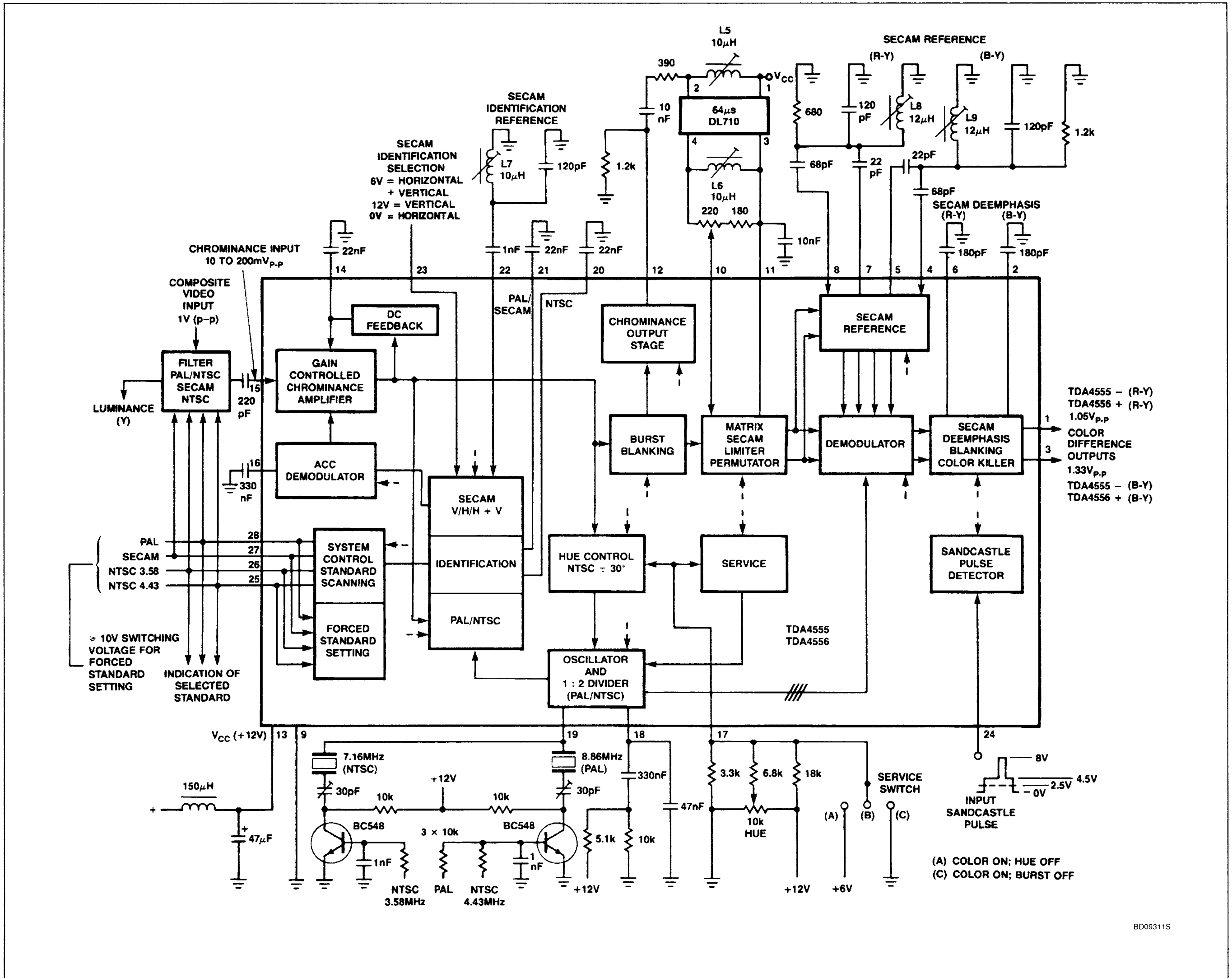


ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
28-Pin Plastic DIP (SOT-117)	0 to +70°C	TDA4555N

PAL® is a registered trademark of Monolithic Memories, Inc.

BLOCK DIAGRAM AND PERIPHERAL CIRCUITS



ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
$V_{CC} = V_{13-9}$	Supply voltage (Pin 13)	13.2	V
V_{n-9}	Voltage range at Pins 10, 11, 17, 23, 24, 25, 26, 27, 28, to Pin 9 (ground)	0 to V_{CC}	V
I_{12}	Current at Pin 12	8	mA
I_{12M}	Peak value	15	mA
P_{TOT}	Total power dissipation	1.4	W
T_{STG}	Storage temperature range	-65 to +150	°C
T_A	Operating ambient temperature range	0 to +70	°C

DC AND AC ELECTRICAL CHARACTERISTICS

$V_{CC} = V_{13-9} = 12V$; $T_A = 25°C$; measured in Block Diagram, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supply (Pin 13)					
$V_{CC} = V_{13-9}$	Supply voltage range	10.8		13.2	V
$I_{CC} = I_{13}$	Supply current		65		mA
Chrominance part					
	Chrominance input signal (Pin 15)				
$V_{15-9(P-P)}$	input voltage with 75% color bar signal (peak-to-peak value)	20	100	200	mV
$\lvert Z_{15-9} \rvert$	input impedance	2.3	3.3		kΩ
	Chrominance output signal (Pin 12)				
$V_{12-9(P-P)}$	output voltage (peak-to-peak value)		1.6		V
$\lvert Z_{12-9} \rvert$	output impedance (NPN emitter-follower)			20	Ω
V_{12-9}	DC output voltage		8.2		V
	Input for delayed signal (Pin 10)				
I_{10}	DC input current			10	μA
R_{10-9}	input resistance	10			kΩ
Demodulator part (PAL/NTSC)					
	Color difference output signals output voltage (proportional to V_{13-9}) (peak-to-peak value)				
	TDA4555				
$V_{1-9(P-P)}$	−(R-Y) signal (Pin 1)		1.05V ±2dB		V
$V_{3-9(P-P)}$	−(B-Y) signal (Pin 3)		1.33V ±2dB		V
	TDA4556				
$V_{1-9(P-P)}$	+(R-Y) signal (Pin 1)		1.05V ±2dB		V
$V_{3-9(P-P)}$	+(B-Y) signal (Pin 3)		1.33V ±2dB		V
$V_{1/3-9}$	Ratio of color difference output signals (R-Y)/(B-Y)		0.79 ±10%		
$V_{1,\,3-9(P-P)}$	Residual carrier (subcarrier frequency) (peak-to-peak value)			30	mV
$V_{1,\,3-9(P-P)}$	Residual carrier (PAL only) (peak-to-peak value)		10		mV
$V_{1-9(P-P)}$	H/2 ripple at (R-Y) output (Pin 1) (peak-to-peak value) without input signal			10	mV
	DC output voltage NPN emitter-follower with				
$V_{1,\,3-9}$	internal current source of 0.3mA		7.7		V
$\lvert Z_{1,\,3-9} \rvert$	output impedance			150	Ω

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $V_{CC} = V_{13-9} = 12V$; $T_A = 25°C$; measured in Block Diagram, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Demodulator part (SECAM)					
   $V_{1-9(P-P)}$ $V_{3-9(P-P)}$ $V_{1-9(P-P)}$ $V_{3-9(P-P)}$	Color difference signals[1] output voltage (proportional to V_{13-9}) (peak-to-peak value) TDA4555 −(R-Y) signal (Pin 1) −(B-Y) signal (Pin 3) TDA4556 +(R-Y) signal (Pin 1) +(B-Y) signal (Pin 3)		 1.05 1.33 1.05 1.33		 V V V V
$V_{1/3-9}$	Ratio of color difference output signals (R-Y)/(B-Y)		0.79[2] ±10%		
$V_{1,\ 3-9(P-P)}$	Residual carrier (4 to 5MHz) (peak-to-peak value)		20	30	mV
$V_{1,\ 3-9(P-P)}$	Residual carrier (8 to 10MHz) (peak-to-peak value)		20	30	mV
$V_{1,\ 3-9(P-P)}$	H/2 ripple at (R-Y) (B-Y) outputs (Pins 1 and 3) (peak-to-peak value) with f_O signals			20	mV
$V_{1,\ 3-9}$	DC output voltage		7.7		V
$\Delta V/\Delta T$(R-Y) $\Delta V/\Delta T$(B-Y)	Shift of inserted levels relative to levels of demodulated f_O frequencies (IC only)		−0.55 +0.25		mV/°C mV/°C
HUE control (NTSC)/service switch					
 $-\phi$ ϕ $+\phi$	Phase shift of reference carrier at $V_{17-9} = 2V$ at $V_{17-9} = 3V$ at $V_{17-9} = 4V$		 30[3] 0 30[3]		 deg deg deg
R_{17-9}	Input resistance		5		kΩ
Service position					
 V_{17-9} V_{17-9}	Switching voltage (Pin 17) burst OFF; color ON (for oscillator adjustment) Hue control OFF; color ON (for forced color ON)	 6		 0.5 	 V V
Crystal oscillator (Pin 19)					
 R_{19-9} Δf	For double color subcarrier frequency input resistance lock-in-range referred to subcarrier frequency	 ±400	 350 		 Ω Hz

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DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $V_{CC} = V_{13-9} = 12V$; $T_A = 25°C$; measured in Block Diagram, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS Min	Typ	Max	UNIT
Identification part					
	Switching voltages for chrominance filters and crystals at Pin 28 (PAL) at Pin 27 (SECAM) at Pin 26 (NTSC 3.58MHz) at Pin 25 (NTSC 4.43MHz)				
$V_{25, 26, 27, 28-9}$	Control voltage OFF state			0.5	V
 $V_{25, 26, 27, 28-9}$ $V_{25, 26, 27, 28-9}$	Control voltage ON state during scanning; color OFF color ON		 2.45 5.8		 V V
$-I_{25, 26, 27, 28-9}$	Output current			3	mA
 V_{28-9} V_{27-9} V_{26-9} V_{25-9}	Voltage for forced switching ON PAL SECAM NTSC 3.58MHz NTSC 4.43MHz	 9 9 9 9			 V V V V
 t_{DS} t_{DC1} t_{DC2}	Delay time for restart of scanning color ON color OFF	 2 to 3 vertical periods 2 to 3 vertical periods 0 to 1 vertical periods			
	SECAM identification (Pin 23)				
 V_{23-9} V_{23-9} V_{23-9}	Input voltage for horizontal identification (H) vertical identification (V) combined (H) and (V) identification	 10 	 6[2]	 2 	 V V V
	Sequence of standard inquiry PAL-SECAM-NTSC 3.58MHz NTSC 4.43MHz Reliable SECAM identification by PAL priority circuit				
t_S	Scanning time for each standard	4 vertical periods			
Sandcastle pulse detector[4]					
 V_{24-9} $V_{24-9(P-P)}$ V_{24-9} $V_{24-9(P-P)}$ V_{24-9} $V_{24-9(P-P)}$	Input voltage pulse levels (Pin 24) to separate vertical and horizontal blanking pulses required pulse amplitude to separate horizontal blanking pulse required pulse amplitude to separate burst gating pulse required pulse amplitude	 1.2 2.0 3.2 4.0 6.5 7.7		 2.0 3.0 4.0 5.0 7.7 V_{CC}	 V V V V V V
V_{24-9}	Input voltage during horizontal scanning			1.0	V
$-I_{24}$	Input current			100	μA

NOTES:
1. The signal amplitude of the color difference signals (R-Y) and (B-Y) is dependent on the characteristics of the external tuned circuits at Pins 7, 8 and 4, 5, respectively. Adjustment of the amplitude is achieved by varying the Q-factor of these tuned circuits. The resonant frequency must be adjusted such that the demodulated output frequency (f_O) provides the same output level as the internally inserted reference voltage (achromatic value).
2. Value measured without influence of external circuitry.
3. Relative to phase at $V_{17-9} = 3V$.
4. The sandcastle pulse is compared to three internal threshold levels, which are proportional to the supply voltage.

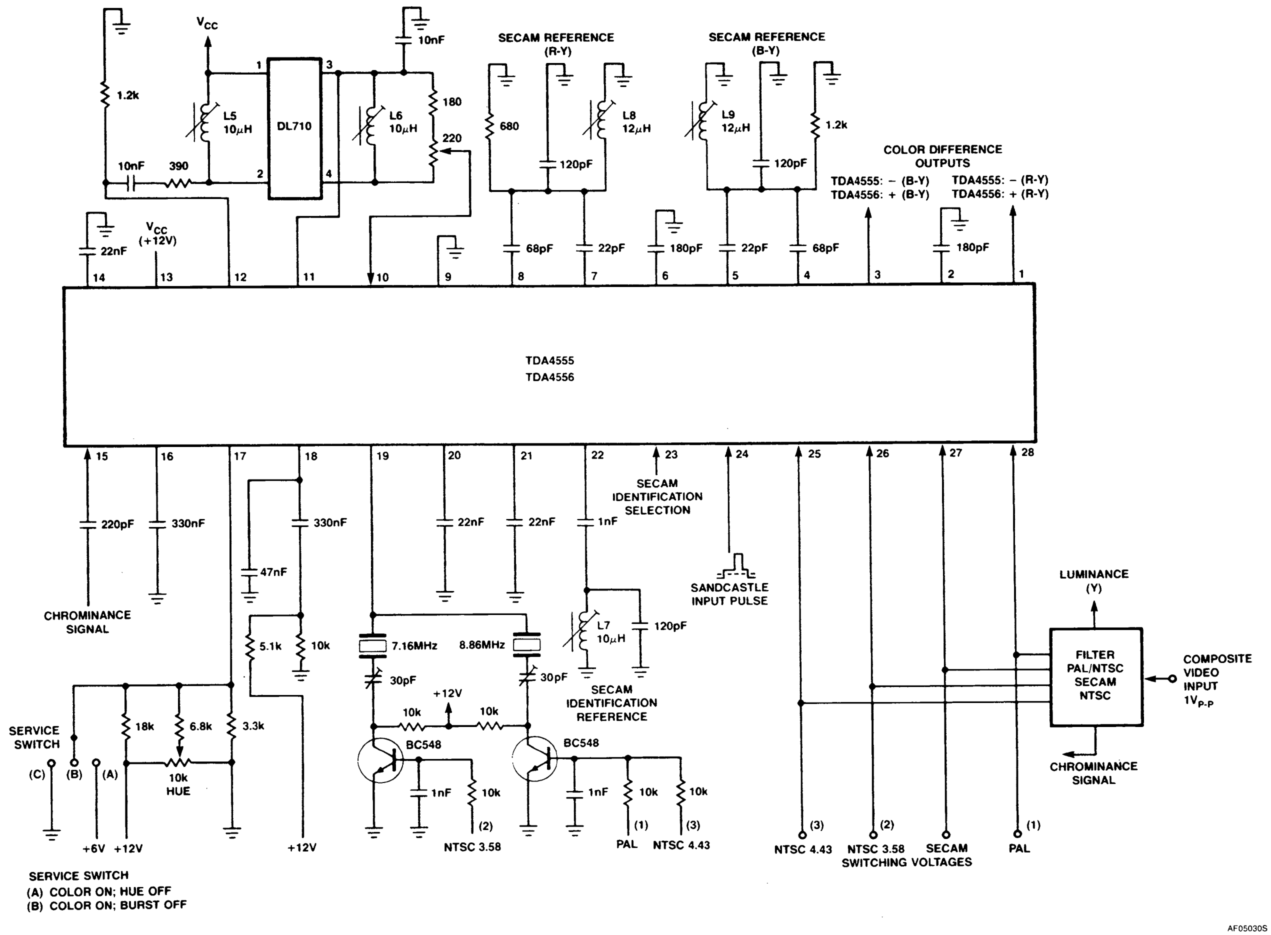


Figure 1. Application Diagram

Signetics

Linear Products

AN1551
Single-Chip Multistandard Color Decoder TDA4555/ TDA4556

Application Note

In areas where TV transmissions to more than one color standard can be received, color receivers are required which can handle multistandard transmissions without additional manual switching. This requirement will greatly increase with the introduction of satellite TV.

Such receivers have, in the past, incorporated a multistandard color decoder (MSD) using several integrated circuits to automatically select the standard of the received signal. However, the growing need for these MSDs makes it economically and technically desirable to incorporate all the active parts in one IC and to reduce, as far as possible, the external circuitry.

This publication describes two new single-chip MSDs using bipolar technology, the TDA4555 and TDA4556. The ICs are similar except for the polarity of the color difference signals at the output. The TDA4555 provides −(R-Y) and −(B-Y) signals; the TDA4556 provides +(R-Y) and +(B-Y) signals. Only the TDA4555 will be described.

Since all the active parts of the MSD are in a single IC, the design and layout of the printed circuit board is considerably simplified and assembly cost is reduced. The greater reliability of "wiring on silicon" increases the overall reliability of the decoder and reduction of external circuitry simplifies assembly.

The ICs are universally applicable and allow the design of a range of TV receivers having a common chassis. Automatic selection of the required standard has been made more reliable and the maximum time required for identification and switching is a little over half a second.

When reception is difficult because signals are weak, noisy, or badly distorted, the automatic standard recognition (ASR) can be switched off and the standard chosen manually.

Although the ICs are capable of processing multistandard signals, their performance is as high as that for single-standard decoders.

Figure 1 is a block diagram of a typical multistandard color decoder incorporating the TDA4555.

The composite video input signal (CVBS) is fed via switchable filters to the input of the MSD. The filters separate the chrominance and luminance signals according to the standard selected and are controlled by the ASR circuit within the TDA4555.

Chrominance signals from the filters are AC coupled to the input of the TDA4555, which produces the color difference outputs that are, in turn, AC coupled to the Color Transient Improvement (CTI) TDA4565. This IC also contains an adjustable luminance delay-line (Y) formed by gyrators, so a conventional wirewound delay line is not needed.

The signals are then fed to the Video Combination IC, TDA3505, which converts the color difference signals −(R-Y) and −(B-Y) and the luminance signal (Y) into the RGB signals. The TDA3505 also incorporates the saturation, contrast, and brightness control circuits and allows for the insertion of external RGB signals. Finally, the processed video signals are applied, via the RGB output stage, to the picture tube.

The new MSD can decode color TV signals transmitted according to the following standards:

1. NTSC standards with any color subcarrier frequency, for example:
   - NTSC-M (f_O = 3.579545MHz), referred to as NTSC-3.5.
   - Non-standard NTSC systems, for example with $f_O = f_{OPAL}$ = 4.43361MHz.
     This is a de facto standard used for VCR signals in some European communities and the Middle East, and is referred to as NTSC 4.43. As the color subcarrier frequency is the same as that of the normal PAL system, the same crystal can be used without switching in the reference oscillator for both systems.
2. PAL standard, characterized by phase reversal of the (R-Y) signal on alternate scan lines. The color subcarrier frequency for normal PAL is 4.43361875MHz.
3. SECAM, characterized by transmission of the color difference signals (R-Y) and (B-Y) on alternate scan lines and frequency mod-

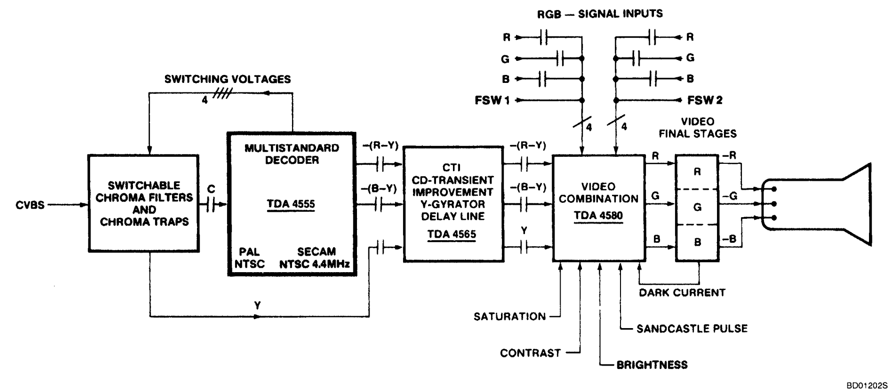


Figure 1. Block Diagram of a Color Decoder

ulation of the color subcarriers. The frequency of the color signals may vary between 3.900MHz and 4.756MHz. The frequencies of the color subcarriers are: f_{OB} = 4.250MHz for a "blue line" f_{OR} = 4.40625MHz for a "red line".

With these capabilities, the new decoders can handle most of the color TV transmissions used in the world.

DESIGN CONSIDERATIONS

To minimize the number of integrated components and reduce the required crystal area and power dissipation of the MSD, the same sections of the IC are used, where possible, for several standards. For example:

- the gain-controlled input stages
- the common switching pulse generators
- the PAL and NTSC quadrature demodulators and oscillators
- the PAL and SECAM delay line
- the common driver stage preceding the delay lines
- part of the stage following the delay line and the demodulator

The number of connections are kept to a minimum compatible with the required functions. With the new ICs, the reference oscillator, its filter, and the SECAM identification circuit, each require only a single pin. The sandcastle pulse is the only external pulse signal. These, and other measures, allow the TDA4555 chip to be housed in a 28-lead SO-117 encapsulation, despite the many functions it performs.

There are three alternative approaches to multi-standard color decoder design.

1. Separate parallel-connected decoders for each standard with the appropriate output selected by switching. This is the principle used in the three-standard decoder comprised of the TDA3510 for PAL, TDA3520 for SECAM, and TDA3570 for NTSC. The color ON/OFF switch voltages generated in each decoder are used for automatic switching of the standards, and each decoder has to be kept at least partially activated.
2. A single PAL decoder can be switched to handle NTSC signals. SECAM signals are converted into quasi-PAL signals by a SECAM-PAL transcoder. The PAL decoder derives the color-difference signals from this quasi-PAL signal. An example of this approach is the circuit using the single-chip PAL decoder TDA3562A with NTSC option and one of the SECAM circuits, TDA3590, TDA3590A, or TDA3591.
3. The methods described in 1 and 2 are not suited to a single-chip MSD because the multiple use of circuit blocks is limited. A much better usage can be obtained if the standards are scanned sequentially. In this approach, the decoder circuit, including the filters at the input, is switched to decode each standard in turn. The switching continues until the standard recognition circuit (SRC) indicates that the standard of the received signal corresponds to the standard of decoding selected at that moment. The scanning procedure is restarted if the standard of the input signal changes because of tuning to another transmitter or switching to an external signal source. The same thing applies if the signal temporarily becomes too weak or disappears. A major advantage of sequential standard switching is that it allows the complete decoder, including the external filters at its input, to be optimized for each standard. This is why the TDA4555 and TDA4556 are designed in this manner.

TDA4555 CIRCUIT DESCRIPTION

Figure 2 is the circuit of a multistandard color decoder using TDA4555/TDA4556.

Pulse Generation

The IC only requires a single sandcastle pulse at Pin 24 for the generation of all internal pulses (e.g., burst key, horizontal, and vertical blanking pulses). The sandcastle pulse levels are > 8V for the burst key, 4.5V for horizontal blanking, and 2.5V for vertical blanking.

Level detectors in the sandcastle pulse detector separate the three levels which are used to generate the required key pulse and clamp pulses.

Standard Control Circuit

A special System Control and Standard Scanning circuit (SCSS) provides the 4 switching voltages to set the MSD to the desired standard.

As long as no color standard is recognized, the SCSS circuit switches the decoder sequentially to the PAL, SECAM, NTSC-3.58 and NTSC-4.43 standards. If the standard of the received signal is not recognized after four field periods (80ms), the next decoding system is activated. This time interval, also called the standard scanning period, is a good compromise between fast switch-on of the color, and effective interference suppression with noisy signals. The maximum time between the start of scanning and switching on the color is 360ms, including the color switch-on delay of two field periods. However, in the TDA4555, a PAL priority circuit is incorporated to improve the reliability for SECAM, so the scanning can last for another two scanning periods (520ms maximum).

After recognition of a SECAM signal, the information is stored and the decoding is switched to PAL. A second SECAM recognition is only provided if no PAL recognition occurs. This gives reliable SECAM recognition when the SECAM-PAL transcoding at the source (e.g., in cable systems) is not perfect, or when PAL signals are distorted by reflections so that they simulate SECAM signals.

With b/w signals, the scanning is continuous and the color is kept switched off because there is no standard recognition.

The switching voltage corresponding to the recognized standard ramps from 2.5V to 6V during scanning while the remaining switching voltages are held at 0.5V maximum.

These 4 voltages are used to switch the filters at the inputs, the crystals of the reference oscillators, and the color subcarrier traps, and also to indicate the recognized standard (e.g., by LEDs).

To prevent unnecessary restarting of scanning because of momentary disturbances (e.g., short-term interruptions of the color signal), the TDA4555 incorporates a delay of two field periods (40ms) before scanning can start.

Finally, the IC allows the automatic standard recognition (ASR) to be switched off by forcing one of the decoding modes by applying at least 9V to Pin 28 for PAL; Pin 27 for SECAM; Pin 26 for NTSC-3.58; and Pin 25 for NTSC-4.43. These pins also serve as outputs for the internally-generated switch voltages which indicate the selected standard.

Color Signal Control

The MSD must provide color-difference output signals with an amplitude referred to a given test signal, despite amplitude variations (within limits) of the color input signal. This is required to maintain a fixed amplitude relationship between the luminance signal (Y) and the color-difference signals, independent of different IF filters or receiver detuning. The TDA4555/56 incorporates an Automatic Color Control circuit (ACC) for this purpose.

In the case of PAL and NTSC, the reference for the control is the burst amplitude. For SECAM, the complete color signal is used. The color signal is AC-coupled, via Pin 15, to a gain-controlled amplifier and the control voltage is obtained by in-phase synchronous demodulation of the burst or the color signal.

This approach has the advantage that the same demodulator, having only one external capacitor at Pin 16, can be used for all standards and also results in noise reduction with noisy signals. Unwanted increase of saturation with noisy signals (color bright-up effect) is prevented without an extra peak detector being required.

In-phase synchronous demodulation has the advantage that it is independent of synchronization and the state of the decoder, so the color gain can settle quickly and the color standard scanning period is therefore short. Special low-distortion symmetrical circuits were chosen for the gain-control stage and the following amplifier stage so that H/2 components in the color-difference channel are reduced as far as possible during SECAM reception. Biasing of the color gain-control stage is stabilized by a DC feedback loop decoupled by an external capacitor at Pin 14.

The nominal amplitude of the color input signal at Pin 15 is $100mV_{P-P}$ for a 75% colorbar signal. It may vary between $10mV_{P-P}$ and $200mV_{P-P}$. This range is chosen so that, for a normal $1V_{P-P}$ composite video signal at the input to the filters, transformation is not required.

For PAL and NTSC decoding, the amplitude-controlled color signal, including its burst, is then fed to the SRC, reference generation, and burst blanking stages. The output of the latter stage is applied to the color signal demodulators and the delay-line driver stage.

Standard Recognition Circuit

The SRC tells the SCSS whether the activated decoding mode is the same as that of the incoming signal. This task is performed using the signals occurring during the back porch of horizontal blanking.

For SECAM, it is necessary to distinguish between line (H) identification signals of carrier frequency at the back porch and field (V) identification (special lines carrying identification signals during the field blanking period).

The standard recognition comprises the following parts: a phase discriminator which compares the burst phase of PAL and NTSC signals with the internal reference signal, a frequency discriminator for generating an H/2 signal during SECAM reception, an H/2 demodulator for PAL and SECAM signals, and the logic circuits for the final recognition.

The two phase discriminators for PAL and NTSC signals are supplied with the color signal, and the amplitude-controlled burst. The phase detector for the PAL signals uses the (R-Y) reference signal for the phase comparison; the NTSC phase detector uses the (B-Y) reference signal. Both reference signals are generated by dividing the reference oscillator output. When the correct signals are received, the phase discriminators output the demodulated burst signal for standard recognition.

The discriminator for generating the H/2 signal comprises an internal phase discriminator and an external phase-shift circuit, known as the SECAM identification reference, connected to Pin 22.

The polarity of the PAL and SECAM phase discriminator output signals is reversed line-sequentially. With PAL, this is caused by a change of phase of the burst at line-frequency. With SECAM, it is the result of the color subcarrier frequency changing at line frequency.

Since the signal is changing polarity, it is of no use for the following circuitry. Therefore, the discriminator output signals are fed to the H/2 demodulator which line-sequentially reverses the signal polarity. The pulses are then integrated by external capacitors connected to Pin 21 (PAL and SECAM discriminator output) and to Pin 20 (NTSC phase discriminator output). The voltages on these capacitors are the identification signals which are used by the comparator and logic circuits to derive the control signals. They are dependent on the standard of the incoming signal and on the activated decoding standard and are composed of an internal biasing at half the supply voltage (6V) and a contribution from the identification signal. In the following explanation, only the latter part ΔV_{20} and ΔV_{21} is considered.

a. When the decoder is set to PAL, the frequency of the reference signal is about 4.43MHz. The NTSC discriminator is switched off and the voltage at C_{20} is only the bias voltage. The H/2 demodulator is therefore driven by the output of the PAL discriminator. The output of the SECAM discriminator is not used. With a PAL signal at the input, the H/2 demodulator delivers pulses with equal polarity so that capacitor C_{21} is charged to ΔV_{21} if the reference oscillator is correctly locked.

   With an NTSC-4.43 input signal, the H/2 modulator provides no pulses or, in case of phase faults, small pulses with a line-sequentially changing polarity. The latter is caused by the constant burst phase of NTSC signals which is line-sequentially reversed by the H/2 demodulator. The average charge current of C_{21} is, therefore, zero, and the capacitor voltage equals the biasing voltage.

   When a SECAM or NTSC-3.58 signal is received, the difference between the burst and f_O frequency is so large that the phase changes very rapidly and, as a result, the H/2 pulses are irregular. This causes the average charge current of C_{21} to be zero.

b. When the decoder is set to NTSC-4.43, the PAL and NTSC-4.4 phase discriminator is activated and the SECAM frequency discriminator is switched off. The PAL phase discriminator and the H/2 demodulator operate as previously described.

   With an NTSC-4.43 signal at the input, the output of the NTSC phase discriminator consists of pulses with the same polarity because the burst of the NTSC signal and the reference signal (B-Y) have the same phase.

   With a PAL input signal, the NTSC phase discriminator also outputs pulses with the same polarity, because the PAL burst comprises a component which is stable in the negative (B-Y) direction for each line. Capacitor C_{20} at the output of the NTSC phase discriminator is therefore charged by an NTSC-4.43, as well as a PAL, input signal, although the decoder is set to the NTSC-4.43 mode.

   With NTSC-3.58 and SECAM signals, the average output current of the NTSC phase discriminator is zero ($\Delta V_{20} = 0$) because the frequency of the burst of the carrier frequency does not match that of the reference.

c. When the decoder is set to NTSC-3.58, the oscillator circuit (including dividers) generates reference signals of about 3.58MHz and the SECAM frequency discriminator is switched off. The NTSC-3.58 phase discriminator provides demodulated burst pulses with constant polarity. At the H/2 demodulator output, no pulses, or, in case of phase faults, small pulses with alternating polarity, appear as in the NTSC-4.43 mode.

   For all other color input signals (PAL, SECAM, NTSC-4.43), the large difference between burst or carrier frequency and reference signal frequency prevents defined discriminator output pulses. As a result, the average charge currents of capacitor C_{20} and C_{21} are zero.

d. When decoding SECAM, the H/2 demodulator obtains its signals from the SECAM discriminator. The output of the PAL phase discriminator is not used and the NTSC phase discriminator is switched off so no output signal is available ($\Delta V_{20} = 0$).

   For SECAM decoding, a frequency discriminator in the recognition block is active. H/2 pulses with line-alternating polarity occur when the frequency of the applied signal is alternately higher and lower than the resonant frequency f_{RES} of the SECAM identification circuit.

   $$f_{RES} = (f_{OB} + f_{OR})/2 \cong 4.43MHz$$

   Therefore, the output of the H/2 demodulator is a train of equal polarity pulses charging the capacitor C_{21}. For PAL, NTSC-3.58 and NTSC-4.43 signals, the burst frequency is constant so the output of the frequen-

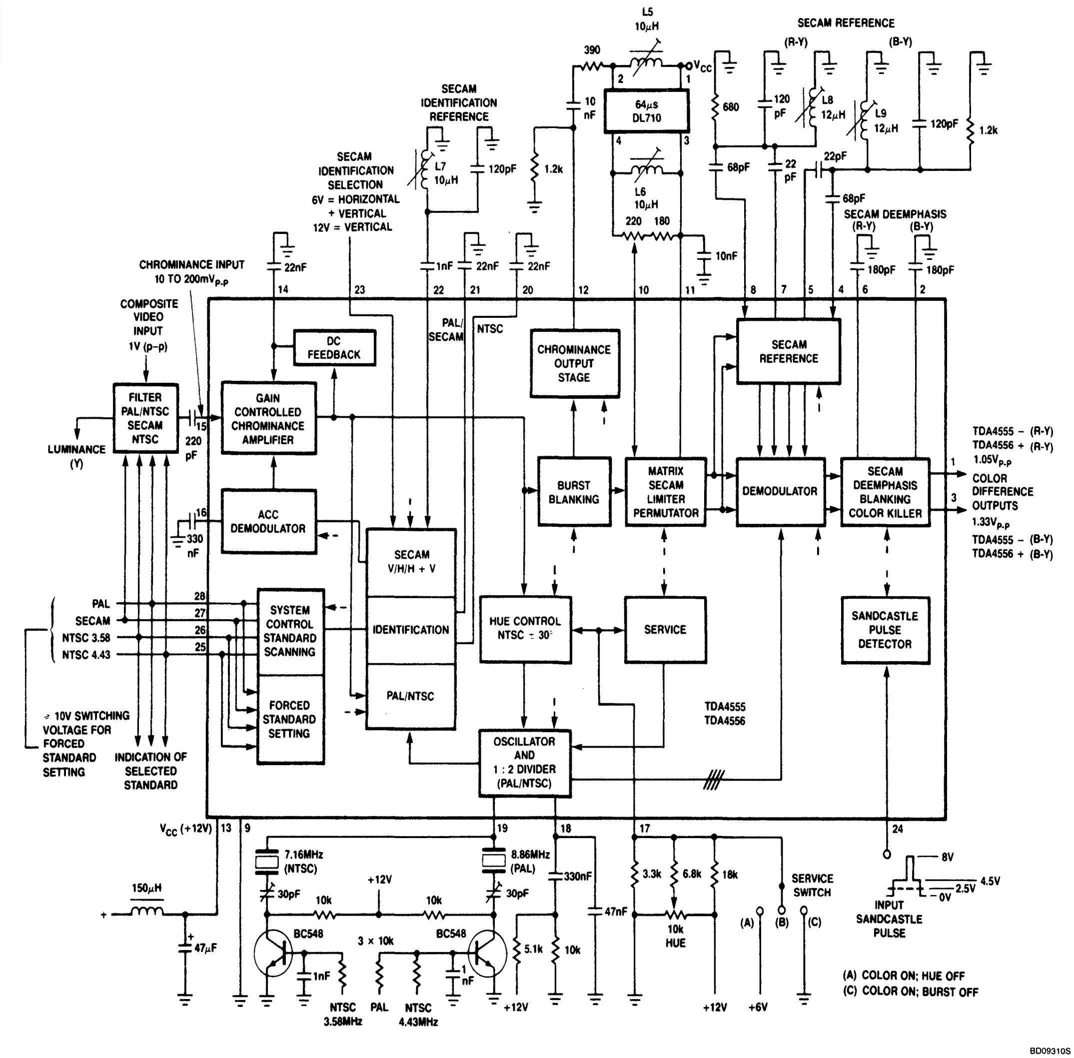


Figure 2. Block Diagram and Peripheral Circuitry

cy discriminator consists of unipolar pulses and the H/2 demodulator outputs alternating polarity pulses. The average charge current of capacitor C_{21} is therefore zero ($\Delta V_{21} = 0$).

The TDA4555 is designed so that identification of SECAM signals can be performed as required by using the special signals in each field blanking period (V-identification) or the burst signal at the back porch (H-identification), or both signals at the same time (H + V-ident). The required standard is selected by applying the appropriate voltage to Pin 23 as follows:

$V_{23} < 2V$ (e.g., ground), H-identification

$V_{23} > 10V$ (e.g., V_{SUPPLY}), V-identification

$V_{23} = 6V$ or floating, H + V-identification.

V-identification is more reliable than the H-identification because the identification signals are longer and have a greater frequency deviation ($\Delta f_{I,B} = 3.9MHz$; $\Delta f_{I,R} = 4.756MHz$). With H-identification, only the normal carrier signal at the end of the back porch is available for identification. When it is required to transmit other information during the field-blanking period, several transmitters (e.g.,in France) stop transmitting the V-identification signals. However, the TDA4555 can easily be adapted to such system changes.

Table 1 summarizes the foregoing. For b/w signals, the average charge current is zero, so no standard is recognized and the scanning is continuous.

Generation of PAL and NTSC Reference Signals

For demodulation and identification of the quadrature amplitude-modulated PAL and NTSC color signals, the reference signals Ref(R-Y) and Ref(B-Y) are needed. These signals are derived from the transmitted burst by a PLL which comprises a voltage-controlled oscillator (VCO), a 2:1 frequency divider, and a phase discriminator. The oscillator frequency is twice the subcarrier frequency ($2f_O$) and the circuit has the advantage that the two quadrature reference signals are available at the output of the divider.

With PAL and NTSC, the phase discriminator compares the (R-Y) reference signal and the burst. The burst and the color signal obtained from the ACC stage are applied to the discriminator directly for PAL and via the hue control for NTSC. In the hue control block, the phase of the burst signal can be shifted ±30° by an external voltage of between 2V and 4V at Pin 17. This voltage is derived from the supply by a simple resistor network. Pin 17 also receives the voltage from the "service" switch. If V_{17} is less than 1V (e.g., ground), the color is forced ON and the oscillator free runs because the burst is switched OFF. The oscillator frequency can be adjusted with the trimmers in series with the crystals. If V_{17} is greater than 6V (e.g., the supply voltage), the color is forced ON and the hue control is switched OFF.

The phase discriminator, which provides a VCO control voltage which depends on the phase difference between burst and reference signal, is activated by a burst key pulse. The control voltage is filtered by an external second-order, low-pass filter connected to Pin 18.

The two crystals for the reference oscillator are both connected between Pin 19 and ground via a switch circuit comprising two transistors driven by the external standard switch voltages. To prevent interference, the oscillator is switched off during SECAM decoding.

Color Signal Demodulators

Demodulation of the color signals is performed in the same way as in single standard predecessors.

In the PAL decoding mode, the burst signal is removed from the color signal derived from the gain-controlled chroma amplifier to prevent disturbances caused by reflections in the glass delay-line delayed by other than a single line period. The color signal is applied to an 18dB amplifier and driver stage (emitter-follower) which compensate for the "worst-case" loss in the external delay-line circuit. Color subcarrier signals CSC_{R-Y} and CSC_{B-Y} are separated by the delay line connected to Pin 12 and terminated at both input and output. Direct and delayed signals are matched by a potentiometer in the output termination. Phase matching can be obtained with coils L_5 and L_6, which compensate the delay-line capacitances.

The delayed signal is taken from the potentiometer slider and fed to the internal matrix via Pin 10, where the direct and delayed signal are added and subtracted to obtain the separated color subcarriers CSC_{R-Y} and CSC_{B-Y}. The matrixing is very simple because the demodulators have symmetrical differential inputs and the direct color signal is available in both polarities. Signals of one polarity are applied to one of the (B-Y) demodulator inputs, and signals of the other polarity to one of the (R-Y) demodulator inputs. The remaining input of both demodulators is supplied with the delayed signal. Unlike previous PAL decoders, the PAL switch is located just in front of the (R-Y) demodulator, i.e., in the CSC_{R-Y} signal path.

The actual color signal demodulators are conventional synchronous types comprising an analog multiplying differential stage with a current source in the emitter circuit and balanced, cross-coupled switching stages in the collector circuit. The latter are driven by reference signals Ref(R-Y) or Ref(B-Y) and one or both analog inputs receive the color signal $CSC_{(R-Y)}$ or $CSC_{(B-Y)}$. The color-difference signals CD, obtained after demodulation, are blanked during the line blanking interval to provide signals with clean levels.

For NTSC decoding, the color signal is demodulated in a similar manner except that only the direct (undelayed) signal is used. The PAL switch in the $CSC_{(R-Y)}$ path is not used.

For reception of the line sequential SECAM color signals, a parallel-crossover switch ("permutator") is required before the demodulators. This permutator alternately feeds both demodulators with a direct and (via the external delay line) a delayed color signal of the same subcarrier frequency.

After the permutator, both color channels incorporate a limiter stage to eliminate amplitude modulation. The color signals are demodulated by quadrature demodulators, each comprising an internal multiplier and an external single-tuned phase-shift circuit, known as the SECAM reference circuit. These reference circuits, connected to Pins 5.6 and 7.8, cause a phase shift of about 90° for the unmodulated subcarrier frequency. Thus, for unmodulated subcarrier signals, there is no output apart from the biasing voltage. The SECAM reference circuits are adjusted by L_8 and L_9 so that the reference levels appear at the CD outputs when the subcarrier is unmodulated or when the color is switched off.

In each color-difference channel, the demodulators are followed by internal low-pass deemphasis networks which remove the unwanted high-frequency components (harmonics of reference and color signals).

The color-difference signals pass, via the output emitter-followers with current sources

Table 1. Charge on Storage Capacitors C_{20} and C_{21} for Combinations of Input Signals and Decoding Mode

DECODING MODE	STANDARD OF THE COLOR INPUT SIGNAL									
	PAL		NTSC-4.433		NTSC-3.588		SECAM		B/W	
	C_{20}	C_{21}	C_{20}	C_{21}	C_{20}	C_{21}	C_{20}	C_{21}	C_{20}	C_{21}
PAL	0*	+	0*	0	0*	0	0*	0	0	0
NTSC-4.43	+	+	+	0	0	0	0	0	0	0
NTSC-3.58	0	0	0	0	+	0	0	0	0	0
SECAM	0*	0	0*	0	0*	0	0*	+	0	0

NOTES:
0 average charge current $I_{AV} = 0$, $\Delta V_C = 0$, $V_C = ½$ supply
+ average charge current $I_{AV} > 0$, $\Delta V_C > 0$ (assuming correct locking of the reference oscillator and proper switching of the H/2 demodulators)
* NTSC phase discriminators switched off

Single-Chip Multistandard Color Decoder TDA4555/TDA4556

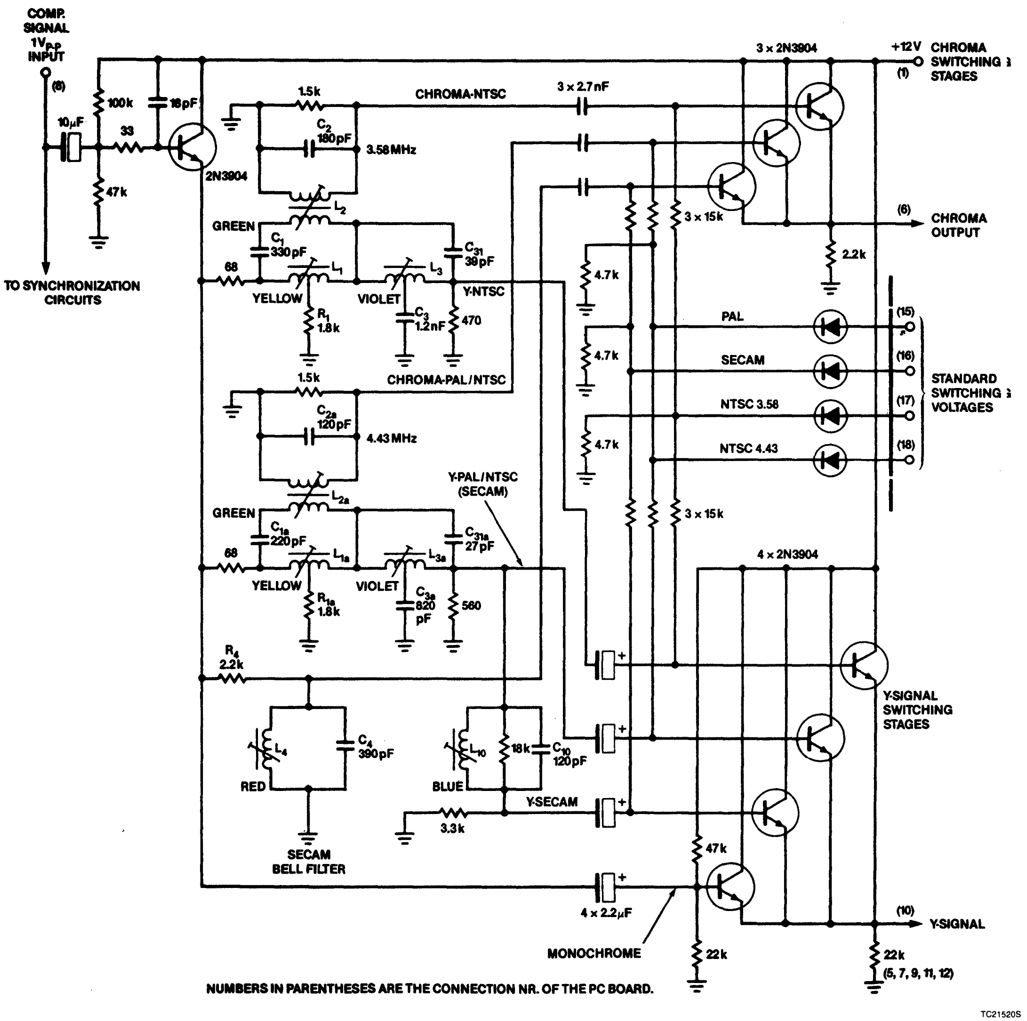


Figure 3. Input Filters and Standard Switching

in their emitter circuits, to Pins 1 and 3, no matter what decoding mode is selected. They have the following nominal amplitudes referred to a 75% saturated color bar:

$$V_{(R-Y)} = 1.05V_{P-P};\ V_{(B-Y)} = 1.33V_{P-P}.$$

For the TDA4555, the polarity of the signals is negative and therefore suitable for input to the video combination family TDA3500 (except TDA3506).

The TDA4556 is similar to the TDA4555 except for the positive polarity of the TDA4556 color difference output signals. Therefore, this TDA4556 can be used with the Video Combination TDA3506.

APPLICATION CONSIDERATIONS

Circuit Example

Figure 2 is a tested circuit of a multistandard decoder. A more detailed circuit of the input filters is shown in Figure 3. These filters separate the luminance signal (Y) from the color signals for the four decoding modes. The same filters can be used for PAL and NTSC-4.43 signals since they have a similar frequency spectrum. For SECAM signals, it is possible to use the 4.43MHz subcarrier trap of the PAL/NTSC-4.43 filter, but it is then necessary to add a trap tuned to about 4.05MHz in the Y channel. This filter suppresses the color signal components below about 4.2MHz, which mainly occur during the "blue SECAM line".

The filter circuits for PAL and NTSC signals are based on a separation filter which also equalizes phase delay. This means that, be-

Table 2. Coil Data for the Multistandard Decoder of Figure 2 and Figure 3

COIL NO	INDUCTANCE (μH)	Q	TOKO TYPE NO.[1]	NO. OF TURNS	COLOR	USE	FIGURE
L_1/L_{1a}	5.5	> 90 (4.43MHz)	119 LNS-A 4449 AH	8 + 8	Yellow	Separation filter	3
L_2/L_K L_{2a}/L_{Ka}	12.5	> 90 (4.43MHz)	119 LNS-A 4451 DY	24/1	Green	Color bandpass filter	3
L_3 L_{3a}	66	60 (2.52MHz)	KANS-K 4087 HU	19 + 46	Violet	Phase delay correction	3
L_4	3.8	60 (4.43MHz)	113 CNS-2 K 843 EG	17 (= 14 + 3)	Red	Bell filter	3
L_5, L_6, L_7	10	> 80 (4.43MHz)	119 LN-A 3753 GO	11 + 11	Blue	Decoder board and SECAM trap for f_{OB}	2
L_{10}	10	> 80 (4.43MHz)	119 LN-A 3753 GO	11 + 11	Blue	PAL/NTSC trap	3
L_8, L_9	12	> 80	119 LN-A 3753 GO	12 + 12	Blue	Decoder board	2

NOTE:
1. Toko America, Mt. Prospect, IL 312/297-0070

sides separating the luminance and color signals, the impulse response of the luminance channel is improved and has symmetrical overshoots, giving the impression of better resolution on the screen. This type of filter is only given as an example. Simpler filters can also be used. The SECAM circuit contains the obligatory "bell" filter. Coil data for the circuit shown in Figure 3 is given in Table 2.

Figure 4 shows oscillograms of the luminance and color filtering in the three signal paths. It can be seen that the color passband in the PAL and NTSC decoding mode has its minimum just below the color subcarrier frequency. This means that the lower sideband of the color signal is mainly used and, as a result, the filter may have a narrower bandwidth. Generally, the upper sideband of the color signal is already attenuated by the IF filter. The passband of the filter in the SECAM color signal path has the required "bell" shape as shown in Figure 4c.

From the low-pass characteristics of the luminance channels, it follows that the subcarriers (4.43MHz for PAL/NTSC-4.43 and 3.58MHz for NTSC-3.58) and the unmodulated carrier frequency ($f_{OB} \cong 4.41$MHz for SECAM) are strongly attenuated. Additionally, low-pass filter ($L_{10}C_{20}$) of the SECAM luminance channel resonates at about 4.05MHz which provides the required attenuation of frequencies below 4.2MHz for modulated carriers.

All three separation filters are fed with the CVBS input signal via an emitter-follower (transistor BC548B). Therefore, the complete decoder has a high input resistance and the filters are driven for a low impedance signal source.

Depending on the decoding mode, the luminance signal is fed from the appropriate filter, via the luminance delay line, to the video combination IC, and the color signal is fed via a small coupling capacitor (220pF) to input Pin 15 of the decoder IC.

Emitter-followers in the color signal path provide the required switching. There is one for each mode, PAL/NTSC-4.43, NTSC-3.58, and SECAM, feeding a common emitter-resistor. Three more emitter-followers in the luminance signal path are combined with a fourth which supplies the unfiltered video signal to the video combination IC during b/w reception, or while the standards are being scanned. The video signals are applied to the bases of the transistor switches via coupling capacitors, the switch voltages being supplied via resistor-diode networks. The fourth transistor switch in the luminance channel has fixed-base biasing of about 4.4V.

The resistors in parallel with the SECAM tuned circuits determine their Q and therefore the conversion efficiency (dV/df) of the demodulators in the SECAM mode and can be used to set the nominal output values of the CD signals (with a color bar signal). The switch transistors for the oscillator crystals at Pin 19 have their collectors connected, via 10kΩ resistors, to the supply line. Because they are either fully conducting or completely cutoff and the voltages are low (12V max.), the type of transistor is not critical.

The standard control voltage outputs (Pins 25 to 28) can deliver a current of 3mA which is insufficient to drive a LED to indicate the standard to which the circuit is set. An additional transistor amplifier such as that shown in Figure 5 is therefore required. Resistor R_{CS} determines the current through the LED, and R_{BS} limits the maximum base current.

If an indication is provided for each of the standard switch voltages, then it is easy to establish which standard, if any, is recognized. When all the diodes light up in sequence, the circuit is still scanning and no standard has been recognized.

Alignment of the Input Filter

The alignment of both the PAL/NTSC-4.43 and NTSC-3.58 separation filters consists of three procedures for each separation filter.

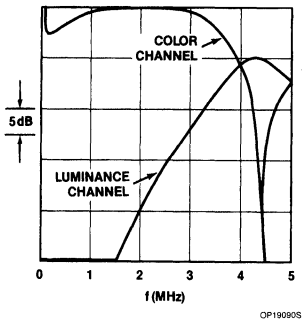
1. **Alignment of the Color Bandpass**
Apply a sweep signal [f = 3.5MHz (4MHz); $\Delta f \cong \pm 3$MHz (± 3MHz) to the filter input (PCB Pin 8). Connect an oscilloscope to PCB Pin 6 and make the filter output available at IC Pin 6 by applying an external switch voltage to the appropriate switch transistor. Adjust $L_2(L_{2a})$ for maximum output at 3.45MHz (4.2MHz).

2. **Alignment of the Compensation Circuit**
Apply a 3.58MHz (4.43MHz) subcarrier to the filter input (PCB Pin 8) and adjust $L_1(L_{1a})$ so that the voltage at the Y output of the filter is minimum. This Y output can be measured at the 470Ω (560Ω) terminating resistor, or at PCB Pin 10, if the proper switch transistor is activated by an external switch voltage.

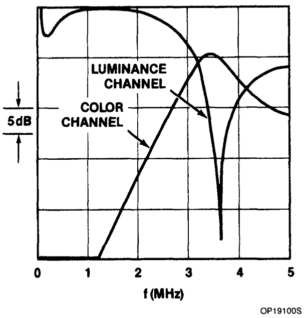
. **Alignment of the Phase Delay Equalizer**
Apply a 16 100kHz square wave to the filter input (PCB Pin 8) and connect an oscilloscope to the output of the luminance filter (470Ω or 560Ω terminating resistor).

Single-Chip Multistandard Color Decoder TDA4555/TDA4556

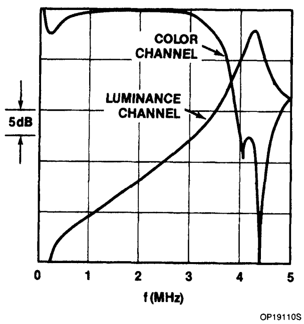
AN1551



a. **PAL/NTSC-4.43 (f_O = 4.433MHz)**



b. **NTSC-3.58 (f_O = 3.579MHz)**



c. **SECAM (f_{OB} = 4.250MHz, f_{OR} = 4.406MHz), "Bell" Filter (f_{RES} = 4.286MHz)**

Figure 4. Amplitude-frequency Characteristics of Input Filter

Alternatively, the oscilloscope can be connected to PCB Pin 10, if an external switch voltage is applied to the appropriate input. Adjust coil $L_3(L_{3a})$ to obtain a symmetrical overshoot at the leading and trailing edges of the pulse.

Because the impulse response of a receiver also depends on the IF filter, it is recommended that the filter be included in the test signal path when aligning L_3/L_{3a}. In practice, a square wave-modulated IF signal should be applied to the input of the IF circuit for this adjustment.

Filter $L_{10}C_{10}$ attenuates the SECAM color signal in the luminance channel below 4.2MHz. L_{10} is adjusted so that an applied 4.05MHz signal has minimum amplitude at the output of the SECAM Y-filter (terminating resistor 3.3kΩ, or PCB Pin 10, if an external switch voltage is applied to the appropriate input).

To align the SECAM "bell" filter, a SECAM color bar is applied to the filter input (PCB Pin 8) and an external switch voltage (e.g., the supply voltage) to PCB Pin 16 to force the SECAM decoding mode. L_4 is then adjusted for minimum amplitude-modulation of the filtered color signal (PCB Pin 6).

To locate the coils to be adjusted, it is useful to color code them as shown in Table 2 and Figure 3.

Decoder Alignment

PAL and NTSC-4.43 Signals

Force the PAL decoding mode by an external voltage exceeding 9V (e.g., the supply voltage) applied to Pin 28 of the IC (or PCB Pin 15) and apply a PAL color signal (e.g., color bar) to the filter input, PCB Pin 8. Connect IC Pin 17 to ground with the service switch. The color is forced ON and the oscillator is free-running because the PLL oscillator circuit does not receive the burst.

Adjust the trimmer in series with the 8.8MHz crystal for minimum color rolling. Alternatively, observe the color-difference signals at IC output Pins 1 and 3 and minimize the beat frequency with the trimmer. This 8.8MHz oscillator adjustment is also valid for the decoder in NTSC-4.43 mode.

To adjust the phase of the delay-line decoder, apply a PAL color bar signal to the input of the circuit (PCB Pin 8) with the service switch in its normal (middle) position. Adjust L_5 and L_6 to minimize amplitude differences of each color bar in the (B-Y) output signal (IC Pin 3 or PCB Pin 13).

Alternatively, minimize the PAL structure (pairing of the lines) observed on the screen. If the adjustment range of L_5 is too small, adjust L_6.

To adjust the amplitude of the delay-line decoder, apply an NTSC-4.43 color bar signal to the input of the circuit (PCB Pin 8) and connect IC Pin 17 to the supply line with the service switch. The color is forced ON and the hue control is switched off. Adjust the 220Ω potentiometer connected to Pin 4 of the DL711 delay line for minimum amplitude differences of each color bar in the (R-Y) output signal (IC Pin 1 or PCB Pin 14) using an oscilloscope, or, observing the picture-tube screen, minimize the PAL structure (pairing of the lines).

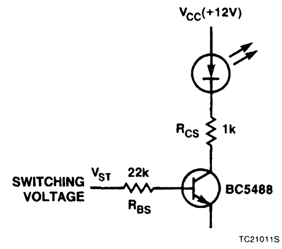


Figure 5. Example of Standard Indicator Circuit

Special test patterns can also be used for delay line adjustment.

Finally, remove the external switching voltage applied to Pin 28 and put the service switch in the mid (normal) position.

NTSC-3.58 Signals

In this case, only the 7.16MHz oscillator has to be adjusted. Force the circuit to the NTSC-3.58 decoding mode by connecting IC Pin 26 or PCB Pin 17 to the supply voltage. Apply an NTSC 3.58 color signal to the filter input (PCB Pin 8). Connect IC Pin 17 to ground with the service switch. The color is forced ON and the oscillator is free-running because the PLL oscillator does not receive burst signals.

Adjust the trimmer in series with the 7.16MHz crystal for minimum color rolling. Alternatively, observe the CD signals at the IC output Pins 1 and 3 and minimize the beat frequency.

Finally, remove the connection between PCB Pin 17 and the supply voltage and put the service switch back to its mid position.

Alignment for SECAM Signals

Force the circuit in the SECAM decoding mode by connecting the supply voltage to IC Pin 27 (or PCB Pin 16). Apply a SECAM color bar to the filter input (PCB Pin 8).

Connect IC Pin 23 (or PCB Pin 20) to the supply line to activate the H-identification. Connect a high-impedance (> 10MΩ) voltmeter between IC Pin 21 and ground. Adjust coil L_7 for the maximum voltage at IC Pin 21.

Observe the −(R-Y) output signal at IC Pin 1 (PCB Pin 14) with an oscilloscope. Adjust L_8 so that the levels of the black and white bars are in accordance with the level inserted during blanking.

Observe the −(B-Y) output signal at IC Pin 3 (PCB Pin 13) with an oscilloscope. Adjust L_9

so that the levels of the black and white bars are in accordance with the levels inserted during blanking.

Use of the PC Board for a PAL-Only Decoder With the TDA4510

To efficiently manufacture a family of receivers, based on the same main PC board, the TDA4555/TDA4556 can be used as a single standard decoder (e.g., a NTSC-only decoder), but the "pin-aligned" TDA4570 is a cheaper alternative. The connections of the TDA4570 and those of the TDA4555 are shown in Figure 6. Apart from the omission of many peripheral components, only small changes in the external circuitry are needed.

NOTE:
This application note, written by Klaus Juhnke and published as Technical Publication 169 by ELCOMA in 1985, has been revised and edited.

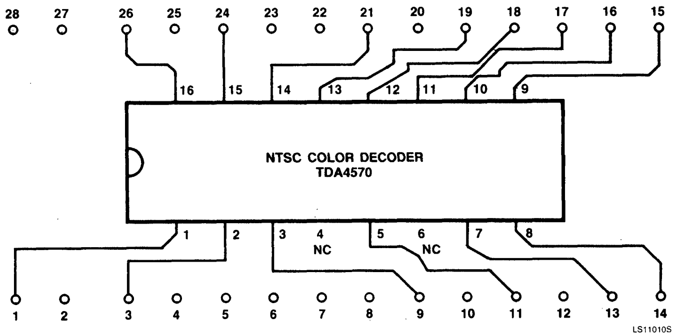


Figure 6

Signetics

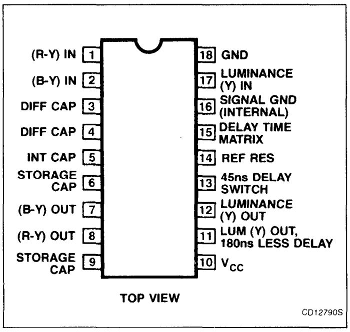
TDA4565
Color Transient Improvement Circuit

Product Specification

Linear Products

DESCRIPTION
The TDA4565 is a monolithic integrated circuit for color transient improvement (CTI) and luminance delay line in gyrator technique in color television receivers.

FEATURES
- **Color transient improvement for color difference signals (R-Y) and (B-Y) with transient detecting, storage, and switching stages resulting in high transients of color difference output signals**
- **A luminance signal path (Y) which substitutes the conventional Y-delay coil with an integrated Y-delay line**
- **Switchable delay time from 690ns to 1005ns in steps of 45ns**
- **Two Y output signals; one of 180ns less delay**

PIN CONFIGURATION


ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
18-Pin Plastic DIP (SOT-102CS)	0 to +70°C	TDA4565N

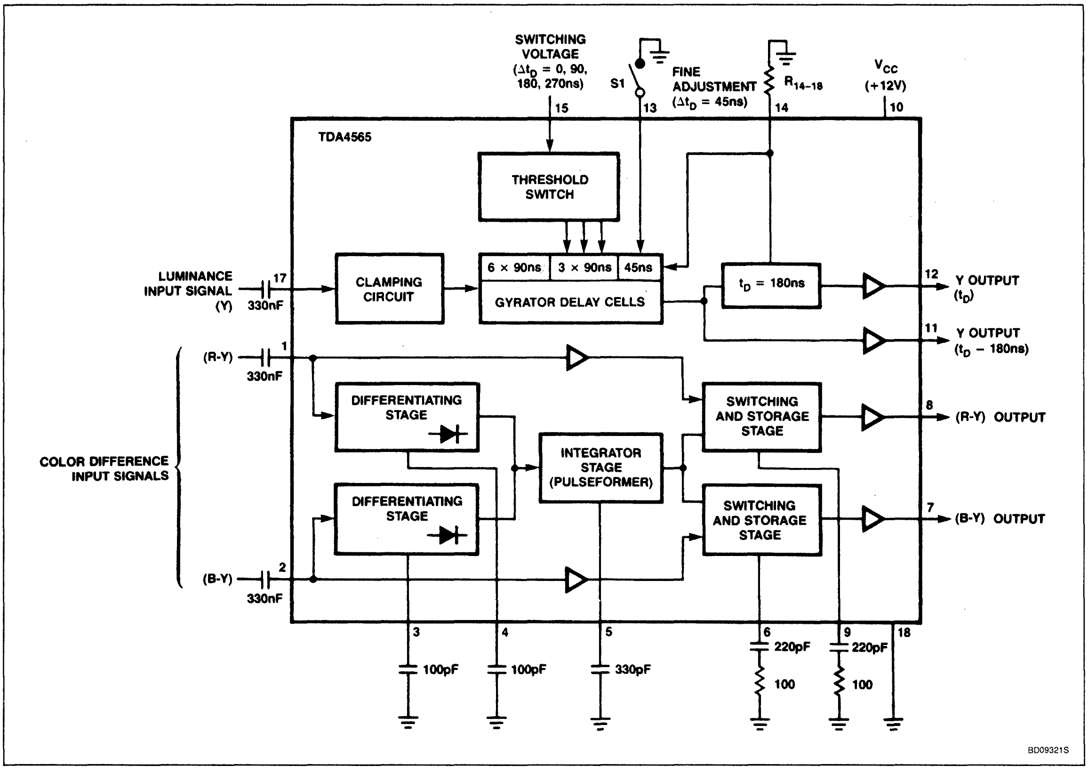
ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
$V_{CC} = V_{10-18}$	Supply voltage (Pin 10)	13.2	V
	Voltage ranges to Pin 18 (ground)		
V_{n-18}	at Pins 1, 2, 12, and 15	0 to V_{CC}	V
V_{11-18}	at Pin 11	0 to (V_{CC}–3V)	V
V_{17-18}	at Pin 17	0 to 7	V
	Voltage ranges		
V_{7-6}	at Pin 7 to Pin 6	0 to 5	V
V_{8-9}	at Pin 8 to Pin 9	0 to 5	V
	Currents		
$\pm I_{6,9}$ $I_{7,8,11,12}$	at Pins 6, 9 at Pins 7, 8, 11, and 12	15	mA
P_{TOT}	Total power dissipation	1.1	W
T_{STG}	Storage temperature range	-25 to +150	°C
T_A	Operating ambient temperature range	0 to +70	°C

NOTE:
DC potential not published for Pins 3, 4, 5, 6, 9, 13, and 14.

Color Transient Improvement Circuit TDA4565

BLOCK DIAGRAM



Color Transient Improvement Circuit TDA4565

DC ELECTRICAL CHARACTERISTICS $V_{CC} = V_{10-18} = 12V$; $T_A = 25°C$; measured in application circuit Figure 1, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS Min	Typ	Max	UNIT
Supply (Pin 10)					
$V_{CC} = V_{10-18}$	Supply voltage	10.8	12	13.2	V
$I_{CC} = I_{10}$	Supply current		35	50	mA
Color difference channels (Pins 1 and 2)					
V_{1-18}	(R-Y) input voltage (peak-to-peak value) 75% color bar signal		1.05		V
V_{2-18}	(B-Y) input voltage (peak-to-peak value) 75% color bar signal		1.33		V
$R_{1,\,2-18}$	Input resistance		12		kΩ
$V_{1,\,2-18}$	Internal bias (input)		4.3		V
α_{CD}	(B-Y), (R-Y) signal attenuation $\frac{V_8}{V_1}, \frac{V_7}{V_2}$		0		dB
$V_{7,\,8-18}$	Output voltage (DC)		4.3		V
$-I_{7,\,8}$	Output current (emitter-follower with constant-current source 0.6mA)		1.2		mA
t_{TR}	(R-Y) and (B-Y) output signal transient time		150		ns
Y-signal path (Pin 17)					
$V_{17-18(P-P)}$	Y-input voltage (composite signal) (peak-to-peak value)		1		V
V_{17-18}	Internal bias voltage (during clamping)		1.5		V
 I_{17} $-I_{17}$	Input current during picture content during synchronizing pulse		 8 100		 μA μA
α_Y	Y-signal attenuation $\frac{V_{11}}{V_{17}}$		6.5		dB
α_Y	Y-signal attenuation $\frac{V_{12}}{V_{17}}$		6.5		dB
V_{11-18}	Output voltage (DC)		2.3		V
V_{12-18}	Output voltage (DC)		10.3		V
$-I_{11,\,12}$	Output current (emitter-follower with constant-current source 0.6mA)		1.2		mA
$f_{11,\,12-17}$	Cut-off frequency[1, 3] $R_{14-18} = 1.2k\Omega$; $V_{15-18} = 12V$; S1 open		5		MHz
 t_D t_D t_D t_D	Adjustable delay[2, 3] (S1 open) at $V_{15-18} = 0$ to 2.5V; $R_{14-18} = 1.2k\Omega$ at $V_{15-18} = 3.5$ to 5.5V; $R_{14-18} = 1.2k\Omega$ at $V_{15-18} = 6.5$ to 8.5V; $R_{14-18} = 1.2k\Omega$ at $V_{15-18} = 9.5$ to 12V; $R_{14-18} = 1.2k\Omega$	 630 720 810 900	 690 780 870 960	 750 840 930 1020	 ns ns ns ns
Δt_D	Fine adjustment delay (S1 closed) at $V_{13-18} = 0V$		45		ns
t	Signal delay for velocity modulation (Pin 11)		$t_D - 180ns$		
θ_{JA}	Thermal resistance from junction to ambient (in free air)			70	°C/W

NOTES:
1. R_{14-18} influences the bandwidth.
2. Delay time is proportional to resistor R_{14-18}.
3. Devices with suffix "α" require the value of resistor R_{14-18} to be 1.1kΩ, but the cut-off frequency and delay times remain as stated in these characteristics.

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Color Transient Improvement Circuit TDA4565

FUNCTIONAL DESCRIPTION

The IC consists of two color difference channels (B-Y) and (R-Y) and a luminance signal path (Y) as shown in the Block Diagram.

Color Difference Channels

The (B-Y) and (R-Y) color difference channels consist of a buffer amplifier at the input, a switching stage, and an output amplifier. The switching stages, which are controlled by transient detecting stages (differentiators), switch to a value that has been stored at the beginning of the transients. The differentiating stages get their signal direct from the color difference detecting signal (Pins 1 and 2). Two parallel storage stages are incorporated in which the color difference signals are stored during the transient time of the signal. At the end of this transient time, they are switched immediately (transient time of 150 ns) to the outputs. The color difference channels are not attenuated.

Y-signal Path

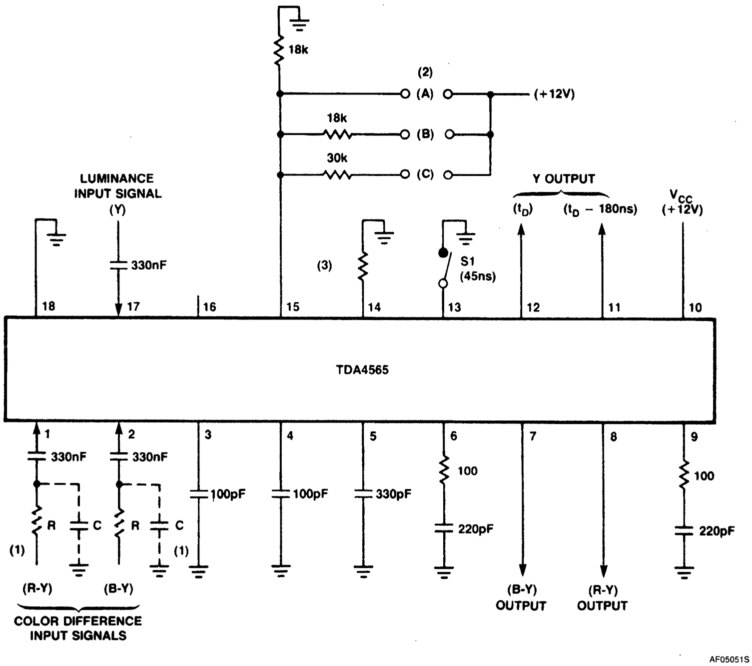
The Y-signal input (Pin 17) is capacitively coupled to an input clamping circuit. Gyrator delay cells provide a maximum delay of 1005ns, including an additional delay of 45ns via the fine adjustment switch (S1) at Pin 13. Three delay cells are switched with two interstage switches dependent on the voltage at Pin 15. Thus, three switchable delay times of 90ns, 180ns, or 270ns less than the maximum delay time are available. A tuning compensation circuit ensures accuracy of delay time despite process tolerances. The Y-signal path has a 6.5dB attenuation as a normal Y-delay coil and can replace this completely. The output is fed to Pin 12 via a buffer amplifier. An additional output stage provides a signal of 180ns less delay at Pin 11.

Table 1. Switching Sequence for Delay Times

CONNECTION			VOLTAGE AT PIN 15	DELAY TIME (ns)*
(A)	(B)	(C)		
O	O	O	0 to 2.5V	690
O	O	X	3.5 to 5.5V	780
O	X	X	6.5 to 8.5V	870
X	X	X	9.5 to 12V	960

Where:
X = connection closed; O = connection open.
*When switch (S1) is closed, the delay time is increased by 45ns.

APPLICATION INFORMATION



NOTES:
1. Residual carrier reduced to $20mV_{P-P}$ (R = 1kΩ, C = 100pF).
2. Switching sequence for delay times shown in Table 1.
3. R_{14-18} = 1.2kΩ for TDA4565.

Figure 1. Application Diagram and Test Circuit

Signetics

TDA4570
NTSC Color Difference Decoder

Product Specification

Linear Products

DESCRIPTION

The TDA4570 is a monolithic, integrated NTSC decoder for NTSC television receivers, which is decoder for NTSC television receivers, which is pin-sequence compatible with multistandard decoder TDA4555.

It can be used in applications with 3.58MHz subcarrier frequency as well as in applications with 4.43MHz subcarrier frequency.

FEATURES

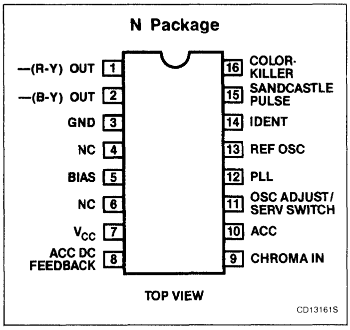
Chrominance part:

- **Gain-controlled amplifier with operating point control stage**
- **ACC (automatic chrominance control) with sampled rectifier during burst-key**
- **Blanking circuit for the color burst signal**
- **Voltage-controlled reference oscillator for double subcarrier frequency**
- **Divider stages which provide –(R-Y) and –(B-Y) reference signals with the correct 90° phase relation for the demodulators**
- **Phase comparator, which compares the –(R-Y) reference signal with the burst pulse and controls the frequency and phase of the reference oscillator**
- **Hue-control stage for phase-shifting via the combined service and hue-control input Pin 11**
- **Identification demodulator, which delivers a positive-going identification signal for NTSC signals at Pin 14; also used for the automatic color-killer**
- **Service switch with two functions. The first position ($V_{14-3} < 1V$) allows the adjustment of the reference oscillator; therefore, the color is switched on, the hue-control and the burst for the oscillator PLL is switched off. The second position ($V_{14-3} > 5V$) switches the color on, the hue-control is switched off, and the output signals can be observed**
- **Sandcastle pulse detector for burst gate, –line and +line vertical blanking pulse detection; the vertical part of the sandcastle pulse is needed for the internal color-on and color-off delay**
- **Pulse processing part which shall prevent a premature switching on of the color; the color-on delay, two or three field periods after identification of the NTSC signal, is achieved by a counter. The color is switched off immediately, or, at the latest, one field period after disappearance of the identification voltage**

Demodulator part:

- **Two synchronous demodulators for the (B-Y) and (R-Y) signals, which incorporate stages for blanking during line- and field-flyback**
- **Internal filtering of the residual carrier in the demodulated color difference signals**
- **Color switching stages controlled by the pulse processing part in front of the output stages**
- **–(B-Y) and –(R-Y) signal output stages; the output stages are low-resistance NPN emitter-followers**
- **Separate color switching output**

PIN CONFIGURATION



APPLICATIONS

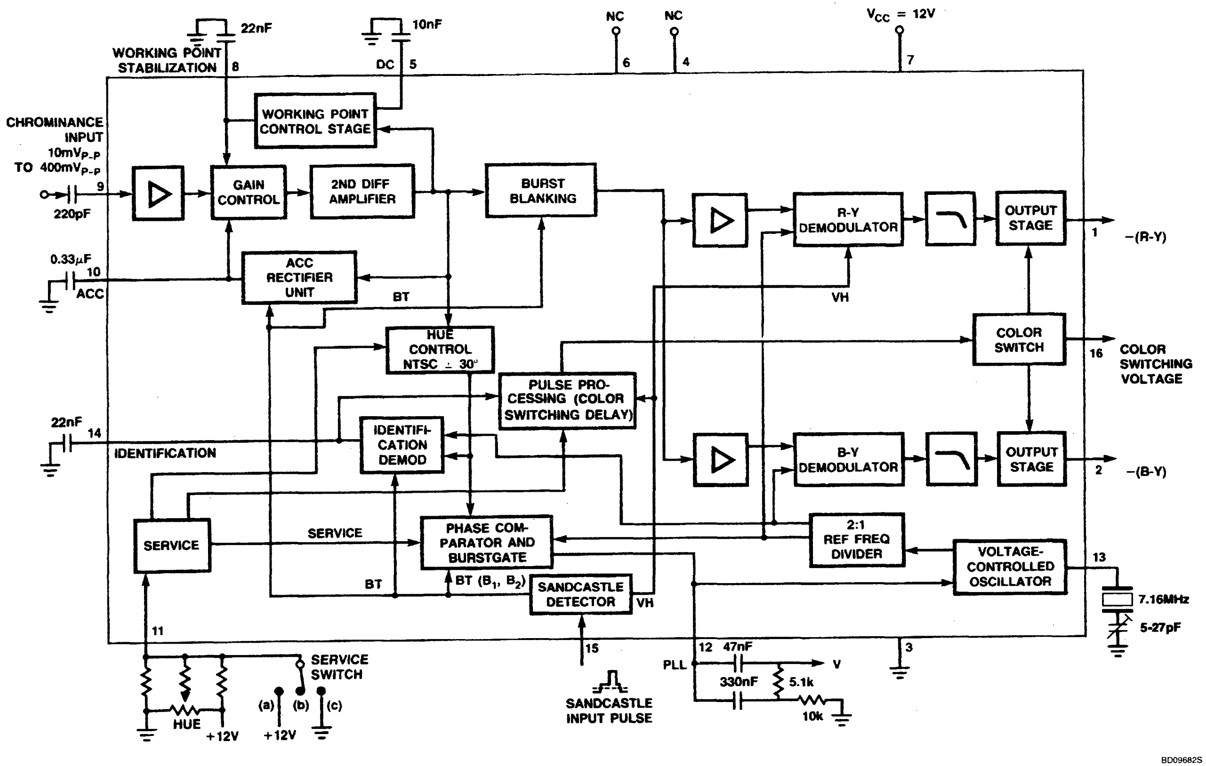
- Video processing
- TV receivers
- Graphic systems

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
16-Pin Plastic DIP (SOT-38)	0 to +70°C	TDA4570N

NTSC Color Difference Decoder TDA4570

BLOCK DIAGRAM



NOTES:
(A) Color ON: Hue OFF.
(B) Color ON: Hue OFF; f_O adjustment.

NTSC Color Difference Decoder — TDA4570

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
$V_{CC} = V_{7-3}$	Supply voltage range	10.8 to 13.2	V
	Currents		
$-I_{1,2}$	at Pins 1 and 2	5	mA
$-I_{16}$	at Pin 16	5	mA
θ_{JA}	Thermal resistance	80	°C/W
P_{TOT}	Total power dissipation	800	mW
T_{STG}	Storage temperature range	−65 to +150	°C
T_A	Operating ambient temperature range	0 to +70	°C

DC ELECTRICAL CHARACTERISTICS $V_{CC} = 12V$; $T_A = 25°C$; measured in Figure 1, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
I_7	Supply current		50		mA
Chrominance part					
$V_{9-3(P-P)}$	Input voltage range (peak-to-peak value)	10		400	mV
$V_{9-3(P-P)}$	Nominal input voltage (peak-to-peak values) with 75% color bar signal		100		mV
Z_{9-3}	Input impedance		3.3		kΩ
C_{9-3}	Input capacitance		4		pF
Oscillator and control voltage part					
f_O	Oscillator frequency for subcarrier frequency of 3.58MHz		7.16		MHz
R_{13-3}	Input resistance		350		Ω
Δf	Catching range (depending on RC network between Pins 12 and 3)	±300			Hz
	Control voltage				
V_{14-3}	without burst signal		6		V
V_{14-3}	color switching threshold		6.6		V
V_{14-3}	hysteresis of color switching		150		mV
t_D ON	Color-on delay			3	Field period
t_D OFF	Color-off delay			1	Field period
	Color-switching output (open NPN emitter)				
$-I_{16}$	output current			5	mA
V_{16-3}	color-on voltage		6		V
V_{16-3}	color-off voltage		0		V
Hue control and service switches					
ϕ	Phase shift of reference carrier relative to the input signal $V_{11-3} = 3V$	−5	0	5	Degree
	Phase shift of reference carrier relative to phase				
$-\phi$	at $V_{11-3} = 3V$ $V_{11-3} = 2V$	30			Degree
ϕ	$V_{11-3} = 4V$	30			Degree
	Internal source (open pin)		3		V
V_{11-3}	First service position (PLL is inactive for oscillator adjustment, color ON, hue OFF)	0		1	V
V_{11-3}	Second service position (color ON; hue OFF)	5		V_{CC}	V

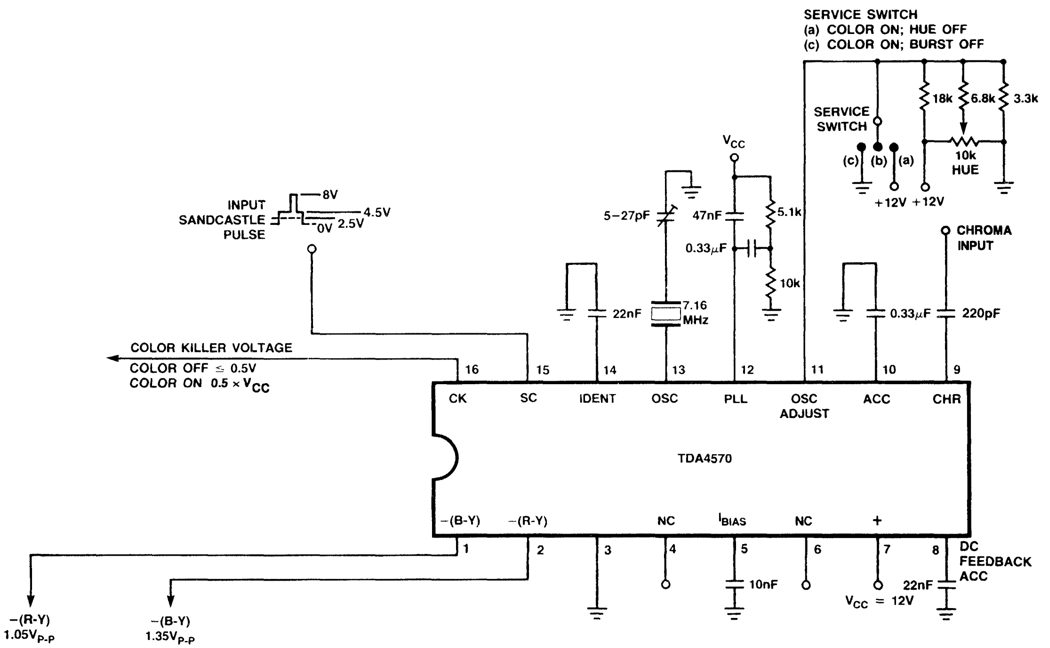
10

DC ELECTRICAL CHARACTERISTICS (Continued) V_{CC} = 12V; T_A = 25°C; measured in Figure 1, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Demodulator part					
	Color difference output signals (peak-to-peak value)				
$V_{1-3(P-P)}$	−(R-Y) signal	0.84	1.05	1.32	V
$V_{2-3(P-P)}$	−(B-Y) signal	1.06	1.33	1.67	V
$\frac{V_{1-3}}{V_{2-3}}$	Ratio of color difference output signals (R-Y)/(B-Y)	0.71	0.79	0.87	
$V_{1,\,2-3}$	DC voltage at color difference outputs		7.7		V
	Residual carrier at color difference outputs				
$V_{1,\,2-3(P-P)}$	(1 × subcarrier frequency)			20	mV
$V_{1,\,2-3(P-P)}$	(2 × subcarrier frequency)			30	mV
Sandcastle pulse detector					
The sandcastle pulse is compared to three internal threshold levels, which are proportional to the supply voltage.					
	Thresholds:				
V_{15-3}	Field- and line-pulse separation; pulse on	1.3	1.6	1.9	V
$V_{15-3(P-P)}$	Required pulse amplitude	2	2.5	3	V
V_{15-3}	Line-pulse separation; pulse on	3.3	3.6	3.9	V
$V_{15-3(P-P)}$	Required pulse amplitude	4.1	4.5	4.9	V
V_{15-3}	Burst-pulse separation; pulse on	6.6	7.1	7.6	V
$V_{15-3(P-P)}$	Required pulse amplitude	7.7			V
V_{15-3}	Input voltage during horizontal scanning			1.1	V
$-I_{15}$	Input current			100	μA

NTSC Color Difference Decoder

TDA4570



NOTE:
Crystal frequency = 7.16MHz; resonance resistance 60Ω; load capacitance 20pF, dynamic capacitance 22pF and static capacitance 5.5pF.

Figure 1

Signetics

Linear Products

TDA4580

Video Control Combination Circuit With Automatic Cut-Off Control

Product Specification

DESCRIPTION

The TDA4580 is a monolithic integrated circuit which performs video control functions in television receivers with a color difference interface. For example, it operates in conjunction with the multi-standard color decoder TDA4555. The required input signals are: luminance and negative color difference -(R-Y) and -(B-Y), and a 3-level sandcastle pulse for control purposes. Analog RGB signals can be inserted from two sources, one of which has full performance adjustment possibilities. RGB output signals are available for driving the video output stages. This circuit provides automatic cut-off control of the picture tube.

FEATURES

- **Capacitive coupling of the color difference, luminance, and RGB input signals with black level clamping**
- **Two sets of analog RGB inputs via fast switch 1 and fast switch 2**
- **First RGB inputs and fast switch 1 in accordance with peritelevision connector specification**
- **Saturation, contrast, and brightness control acting on first RGB inputs**
- **Brightness control acting on second RGB inputs**
- **Equal black levels for television and inserted signals**
- **Clamping, horizontal and vertical blanking, and timing of automatic cut-off, controlled by a 3-level sandcastle pulse**
- **Automatic cut-off control with compensation for leakage current of the picture tube**
- **Measuring pulses of cut-off control start immediately after end of vertical part of sandcastle pulse**
- **Three selectable blanking intervals for PAL, SECAM, and NTSC/PAL-M**
- **Two switch-on delays for run-in without discoloration**
- **Adjustable peak drive limiter**
- **Average beam current limiter**
- **G-Y and RGB matrix coefficients selectable for PAL/SECAM and NTSC (correction for FCC primaries)**
- **Bandwidth 10MHz (typ.)**
- **Emitter-follower outputs for driving the RGB output stages**

APPLICATIONS

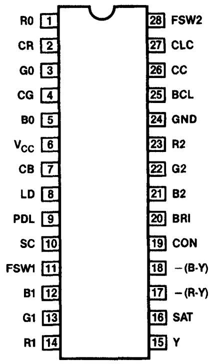
- **Video processing**
- **TV receivers**
- **Projection TV**

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
28-Pin Plastic DIP, (SOT-117)	0 to 70°C	TDA4580N

PIN CONFIGURATION

N Package

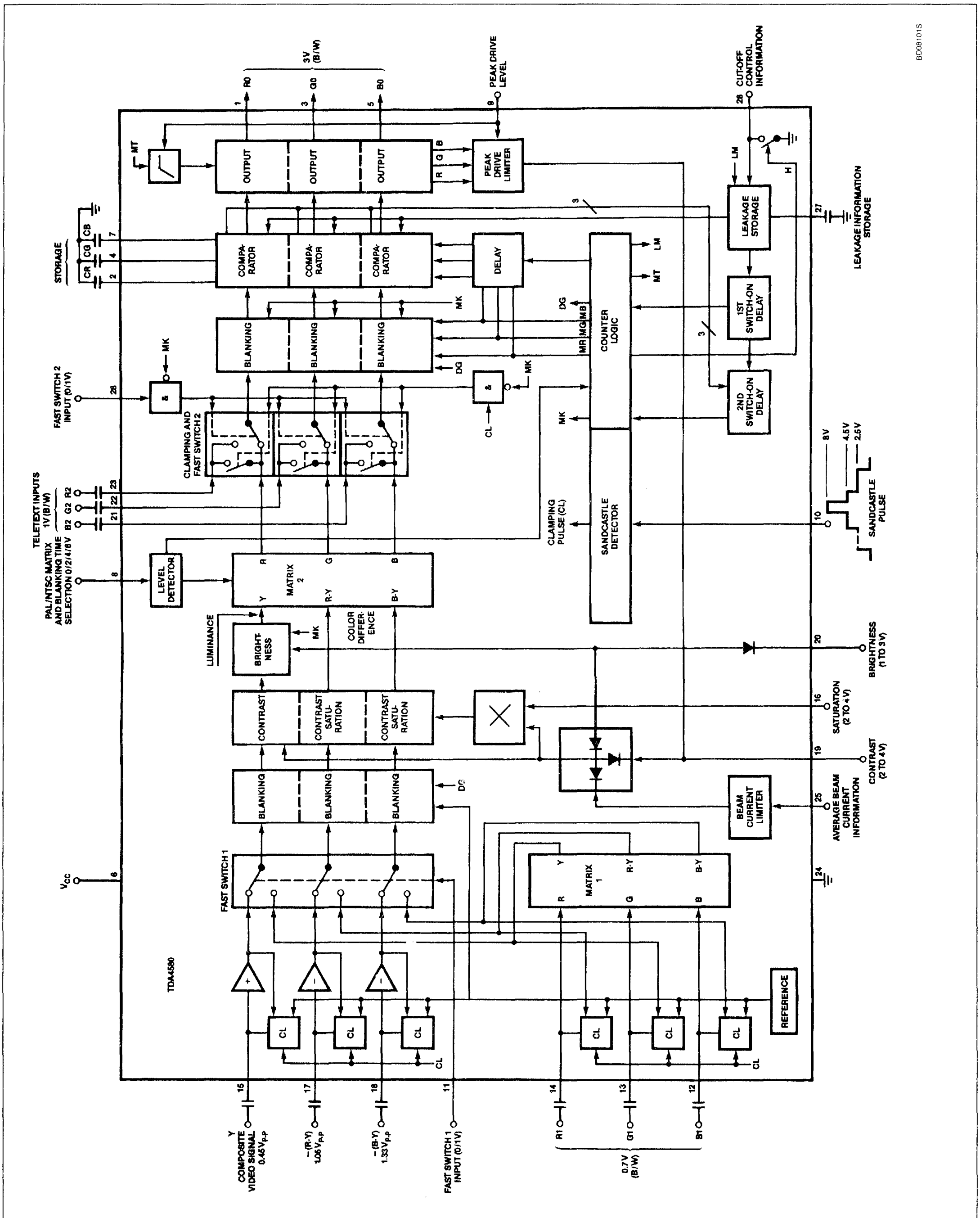


TOP VIEW

CD12030S

PIN NO.	SYMBOL	DESCRIPTION
1	R0	Red output
2	CR	Red storage capacitor for cut-off control
3	G0	Green output
4	CG	Green storage capacitor for cut-off control
5	B0	Blue output
6	V_{CC}	Positive supply voltage (+12V)
7	CB	Blue storage capacitor for cut-off control
8	LD	PAL/NTSC matrix and blanking time level detector input
9	PDL	Peak drive limiting input
10	SC	Sandcastle pulse input
11	FSW1	Fast switch 1 for Y, CD, and RGB inputs
12	B1	Blue input (external signal)
13	G1	Green input (external signal)
14	R1	Red input (external signal)
15	Y	Luminance input
16	SAT	Saturation control input
17	-(R - Y)	Color difference input -(R - Y)
18	-(B - Y)	Color difference input -(B - Y)
19	CON	Contrast control input
20	BRI	Brightness control input
21	B2	Teletext blue input
22	G2	Teletext green input
23	R2	Teletext red input
24	GND	Ground
25	BCL	Average beam current limiting input
26	CC	Automatic cut-off control input
27	CLC	Storage capacitor for leakage current
28	FSW2	Fast switch 2 for teletext inputs

BLOCK DIAGRAM



ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
$V_{CC} = V_{6-24}$	Supply voltage range (Pin 6)	0 to 13.2	V
V_{n-24}	Voltage range at Pins 2, 4, 7, 9, 12, 13, 14, 15, 16, 17, 18, 19, 20, 21, 22, 23, 25, 27 to Pin 24 (ground)	0 to V_{CC}	V
	Voltage ranges		
$V_{8, 11, 28-24}$	at Pins 8, 11, 28	-0.5 to V_{CC}	V
V_{10-24}	at Pin 10	0 to $V_{CC}+0.7$	V
V_{26-24}	at Pin 26	-0.7 to $V_{CC}+0.7$	V
	Currents		
$-I_{1, 3, 5(AV)}$	at Pins 1, 3, 5 (average)	3	mA
$-I_{1, 3, 5(M)}$	at Pins 1, 3, 5 (peak)	10	mA
$I_{19(AV)}$	at Pin 19 (average)	5	mA
I_{26}	at Pin 26	1	mA
P_{TOT}	Total power dissipation	2	W
T_{STG}	Storage temperature range	-65 to +150	°C
T_A	Operating ambient temperature range	0 to +70	°C
θ_{JA}	Thermal resistance from junction to ambient	37	°C /W

DC ELECTRICAL CHARACTERISTICS

$V_{CC} = 12V$; $T_A = 25°C$; measured in a circuit similar to Figure 2 at nominal settings (saturation, contrast, brightness), no beam current or peak drive limiting; all voltages with respect to Pin 24 (ground), unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supply (Pin 6)					
$V_{CC} = V_{6-24}$	Supply voltage range	10.8		13.2	V
$I_{CC} = I_6$	Supply current		110		mA
Color difference inputs (Pins 17 and 18)					
$V_{17-24(P-P)}$	−(R − Y) input signal at Pin 17 (peak-to-peak value)[1, 2]		1.05		V
$V_{18-24(P-P)}$	−(B − Y) input signal at Pin 18 (peak-to-peak value)[1, 2]		1.33		V
$\lvert I_{17, 18}\rvert$	Input current during scanning			0.3	μA
$R_{17, 18}$	Input resistance	5			MΩ
$V_{17, 18-24}$	Internal DC bias voltage during clamping time		7.5		V
Luminance input (Pin 15)[2]					
$V_{15-24(P-P)}$	Composite video input signal (VBS) (peak-to-peak value)		0.45		V
$\lvert I_{15}\rvert$	Input current during scanning			0.3	μA
R_{15}	Input resistance	5			MΩ
V_{15-24}	Internal DC bias voltage during clamping time		7.4		V
Signal switch 1 input (Pin 11)					
V_{11-24}	Input voltage level for insertion of Y and CD signals			0.4	V
V_{11-24}	RGB1 signals	0.9		3.0	V
R_{11}	Internal resistor to ground		10		kΩ

DC ELECTRICAL CHARACTERISTICS (Continued) V_{CC} = 12V; T_A = 25°C; measured in a circuit similar to Figure 2 at nominal settings (saturation, contrast, brightness), no beam current or peak drive limiting; all voltages with respect to Pin 24 (ground), unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
RGB1 inputs (R1 Pin 14, G1 Pin 13, B1 Pin 12) (signals controlled by saturation, contrast, and brightness)[2]					
$V_{12,\ 13,\ 14-24}$	Input signal (black to white value)		0.7		V
$\lvert I_{12,\ 13,\ 14}\rvert$	Input current during scanning			0.3	μA
$R_{12,\ 13,\ 14}$	Input resistance	5			MΩ
$V_{12,\ 13,\ 14-24}$	Internal DC bias voltage during clamping time		8.2		V
RGB/Y, (R - Y), (B - Y) — Matrix					
Matrixed according to the equations $V_{(R-Y)} = 0.7\ V_R - 0.59\ V_G - 0.11\ V_B$ $V_{(B-Y)} = -0.3\ V_R - 0.59\ V_G + 0.89\ V_B$ $V_{(Y)} = 0.3\ V_R + 0.59\ V_G + 0.11\ V_B$					
Contrast control input (Pin 19) (contrast control acts on Y and CD signals or RGB1 signals, respectively)[3]					
V_{19-24}	Maximum contrast		4		V
V_{19-24}	Nominal contrast (6dB below maximum)		3		V
	Attenuation of contrast at V_{19-24} = 2V (related to maximum)		22		dB
$-I_{19}$	Input current at V_{19-24} = 2 to 4V			3	μA
Peak drive limiting input (Pin 9)[4]					
V_{9-24}	Internal DC bias voltage		9		V
R_9	Input resistance at V_{9-24} > 9V		10		kΩ
I_{19}	Control current into contrast input (Pin 19) during peak drive $V_{1,\ 2,\ or\ 3-24} > V_{9-24}$		20		mA
Average beam current limiting input (Pin 25)[5]					
V_{25-24}	Start of contrast reduction at maximum contrast setting		8.5		V
ΔV_{25-24}	Input range for full contrast reduction		1.0		V
R_{25}	Input resistance at V_{25-24} < 6V		2.2		kΩ
Saturation control input (Pin 16) (saturation control acts on CD signals or RGB1 signals, respectively)					
V_{16-24}	Maximum saturation		4		V
V_{16-24}	Nominal saturation (6dB below maximum)		3		V
	Attenuation of saturation at V_{16-24} = 1.8V (related to maximum at 100kHz)	50			dB
I_{16}	Input current at V_{16-24} = 1.8 to 4V			10	μA
Brightness control input (Pin 20)[6, 7]					
V_{20-24}	Control voltage range	1		3	V
$-I_{20}$	Input current at V_{20-24} = 1 to 3V			10	μA
V_{20-24}	Control voltage for nominal brightness		2.2		V
	Change of black level in the control range related to the nominal output signal (black/white) for ΔV_{20-24} = 1V		33		%
V_{20-24}	Signal switched off and black level equal to cut-off level	11.5			V

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DC ELECTRICAL CHARACTERISTICS (Continued) V_{CC} = 12V; T_A = 25°C; measured in a circuit similar to Figure 2 at nominal settings (saturation, contrast, brightness), no beam current or peak drive limiting; all voltages with respect to Pin 24 (ground), unless otherwise specified.

SYMBOL	PARAMETER	LIMITS Min	Typ	Max	UNIT
Y, (R-Y), (B-Y)/RGB-Matrix[8]					
	PAL matrix (V_{8-24} = $\leqslant$4.5V)				
	Matrixed according to the equation $V_{(G-Y)} = -0.51V_{(R-Y)} - 0.19V_{(B-Y)}$				
	NTSC matrix (V_{8-24} = $\geqslant$5.5V)				
	(Adaption for NTSC-FCC primaries, nominal hue control set on −5°C)				
	Matrixed according to the equation $V_{(G-Y)}{}^{8} = -0.43V_{(R-Y)} - 0.11V_{(B-Y)}$ $V_{(R-Y)}{}^{8} = 1.57V_{(R-Y)} - 0.41V_{(B-Y)}$ $VI_{(B-Y)}{}^{8} = V_{(B-Y)}$				
RGB2 inputs (Teletext) (R2 Pin 23, G2 Pin 22, B2 Pin 21)[2]					
	(RGB signals controlled by brightness control)				
$V_{21,\ 22,\ 23-24}$	Input signal for 100% output signals (black to white value)		1		V
$I_{21,\ 22,\ 23}$	Input current during scanning			0.3	μA
$I_{21,\ 22,\ 23}$	Input resistance	5			MΩ
Signal switch 2 input (Pin 28)					
	Input voltage level for insertion of Y, CD signals or RGB1 signals, respectively				
V_{28-24} V_{28-24}	RGB signals from matrix[9] RGB2 signals[9]	 0.9		0.4 3.0	V V
R_{28-24}	Internal resistor to ground		10		kΩ
Automatic cut-off control input (Pin 26) (Leakage current measuring time and insertion of RGB cut-off measuring lines — see Figure 3; types of ultra-black level — see Figure 1.)[10]					
V_{26-24}	Allowed maximum external DC bias voltage	5.5			V
ΔV_{26-24}	Voltage difference between cut-off current measurement and leakage current measurement		0.5		V
$V_{1,\ 3,\ 5-24}$	Warm-up test pulse		$V_{9-24}{}^{8}$		V
V_{26-24}	Threshold for warm-up detector		8		V
Storage input for leakage current (Pin 27)					
R_{27}	Internal resistance during leakage current measuring time (current limiting at I_{27} = 0.2mA)		400		Ω
$\lvert I_{27} \rvert$	Input current except during cut-off control cycle			0.5	μA
Storage inputs for automatic cut-off control (Pins 2, 4, 7)					
$\lvert I_{2,\ 4,\ 7} \rvert$	Charge and discharge currents		0.3		mA
$\lvert I_{2,\ 4,\ 7} \rvert$	Input currents of storage inputs out of control time			0.1	μA

DC ELECTRICAL CHARACTERISTICS (Continued) V_{CC} = 12V; T_A = 25°C; measured in a circuit similar to Figure 2 at nominal settings (saturation, contrast, brightness), no beam current or peak drive limiting; all voltages with respect to Pin 24 (ground), unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Switch input for PAL/NTSC matrix and vertical blanking time (Pin 8)[11]					
  V_{8-24} V_{8-24} V_{8-24}	Switching voltage input for PAL matrix and vertical blanking period of 25 lines 22 lines 18 lines	 1.5 3.5	 0 2 4	 0.5 2.5 4.5	 V V V
V_{8-24}	NTSC matrix and vertical blanking period of 18 lines	5.5	6	12	V
I_8	Input current			50	μA
Sandcastle pulse detector (Pin 10)[12]					
  V_{10-24} V_{10-24} V_{10-24} t_D	The following amplitudes are required for separating the various pulses: horizontal and vertical blanking pulses horizontal pulses for counter logic clamping pulses delay of leading edge of clamping pulse	 2.0 4.0 7.5 	 2.5 4.5 1	 3.0 5.0 	 V V V μs
$-I_{10}$	Input current at V_{10-24} = 0V			100	μA
Outputs for positive RGB signals (R0 Pin 1, G0 Pin 3, B0 Pin 5)[13]					
$V_{1,\ 3,\ 5-24}$	Nominal signal amplitude (black/white)		3		V
	Spreads between channels			10	%
$V_{1,\ 3,\ 5-24}$	Maximum signal amplitude (black/white)	4			V
$I_{1,\ 3,\ 5}$	Internal current source		3		mA
$R_{1,\ 3,\ 5}$	Output resistance		160	220	Ω
$V_{1,\ 3,\ 5-24}$	Minimum output voltage		1		V
$V_{1,\ 3,\ 5-24}$	Maximum output voltage		10		V
	Horizontal and vertical blanking to ultra-black level 2, related to nominal signal black level in percentage of nominal signal amplitude	45	55		%
	Vertical blanking to ultra-black level 1, related to cut-off measuring level in percentage of nominal signal amplitude	25	35		%
	Recommendation: Range for cut-off measuring level 1.5 to 5.0V; nominal value at 3V[14]				
Gain data[15]					
d	Frequency response of Y path (0 to 8MHz) Pins 1, 3, and 5 to Pin 15			3	dB
d	Frequency response of CD path (0 to 8MHz) Pin 1 to Pin 17 = Pin 5 to Pin 18			3	dB
d	Frequency response of RGB1 path (0 to 8MHz) Pin 1 to Pin 14 = Pin 3 to Pin 13 = Pin 5 to Pin 12			3	dB
d	Frequency response of RGB2 path (0 to 10MHz) Pin 1 to Pin 23 = Pin 3 to Pin 22 = Pin 5 to Pin 21			3	dB

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NOTES:

1. The value of the color difference input signals, −(B − Y) and −(R − Y), is given for saturated color bar with 75% of maximum amplitude.
2. Capacitive coupled to a low ohmic source; recommended value 600Ω (maximum).
3. At Pin 19 for $V_{19-24} \leq 2.0V$, no further decrease of contrast is possible.
4. The peak drive limiting of output signals is achieved by contrast reduction. The limiting level of the output signals is equal to the voltage V_{9-24}, adjustable in the range 5 to 11V. After exceeding the adjusted limiting level at peak drive, limiter will not be active during the first line.
5. The average beam current limiting acts on contrast and at minimum contrast on brightness (the external contrast voltage at Pin 19 is not affected).
6. At nominal brightness the black level at the output is 0.3V (≈ −10% of nominal signal amplitude) below the measuring level.
7. The internal control voltage can never be more positive than 0.7V above the internal contrast voltage.
8. Matrix equation

$V_{(R-Y)}$, $V_{(B-Y)}$	: output of NTSC decoder of PAL type demodulating axis and amplitudes
$V_{(G-Y)}^*$, $V_{(R-Y)}^*$, $V_{(B-Y)}^*$	: for NTSC modified CD signals; equivalent to demodulation with the following axes and amplification factors:
(B − Y)* demodulator axis	0°
(R − Y)* demodulator axis	115° (PAL 90°)
(R − Y)* amplification factor	1.97 (PAL 1.14)
(B − Y)* amplification factor	2.03 (PAL 2.03)

$V_{(G-Y)}^* = -0.27V_{(R-Y)}^* - 0.22V_{(B-Y)}^*$.

9. During clamping time, in each channel the black level of the inserted signal is clamped on the black level of the internal signal behind the matrix (dependent on brightness control).
10. During warm-up time of the picture tube, the RGB outputs (Pins 1, 3, and 5) are blanked to minimum output voltage. An inserted white pulse during the vertical flyback is used for beam current detection. If the beam current exceeds the threshold of the warm-up detector at Pin 26, the cut-off current control starts operating, but the video signal is still blanked. After run-in of the cut-off current control loop, the video signal will be released. The first measuring pulse occurs in the first complete line after the end of the vertical part of the sandcastle pulse. The absolute minimum vertical part must contain 9 line-pulses. The cycle time of the counter is 63 lines. When the vertical pulse is longer than 61 lines, the IC is reset to the switch-on condition. In this event the video signal is blanked and the RGB outputs are blanked to minimum output voltage as during warm-up time. During leakage current measurement, all three channels are blanked to ultra-black level 1. With the measuring level only in the controlled channel, the other two channels are blanked to ultra-black level 1. The brightness control shifts both the signal black level and the ultra-black level 2. The brightness control is disabled from line 4 to the end of the last measuring line (see Figure 1).
    With the most adverse conditions (maximum brightness and minimum black level 2) the blanking level is located 30% of nominal signal amplitude below the cut-off measuring level.
11. The given blanking times are valid for the vertical part of the sandcastle pulse of 9 to 15 lines. If the vertical part is longer and the cut-off lines are outside the vertical blanking period of 18, 22, or 25 lines, respectively, the blanking of the signal ends with the end of the last of the three cut-off measuring pulses as shown in Figure 3.
12. The sandcastle pulse is compared with three internal thresholds (proportional to V_{CC}) to separate the various pulses. The internal pulses are generated when the input pulse at Pin 10 exceeds the thresholds. The thresholds are for:
    - Horizontal and vertical blanking $V_{10-24} = 1.5V$
    - Horizontal pulse $V_{10-24} = 3.5V$
    - Clamping pulse $V_{10-24} = 7.0V$
13. The outputs at Pins 1, 3, and 5 are emitter-followers with current sources and emitter protection resistors.
14. The value of the cut-off control range for the positive RGB output signals is given for a nominal output signal. If the signal amplitude is reduced, the cut-off range can be increased.
15. The gain data is given for a nominal setting of the contrast and saturation controls, measured without load at the RGB outputs (Pins 1, 3, and 5).

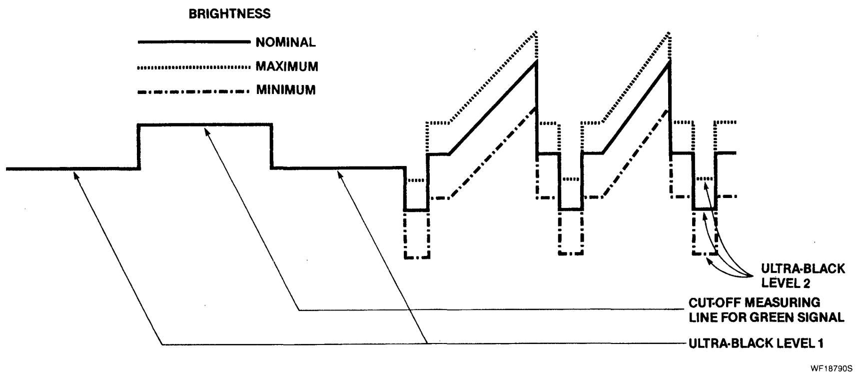
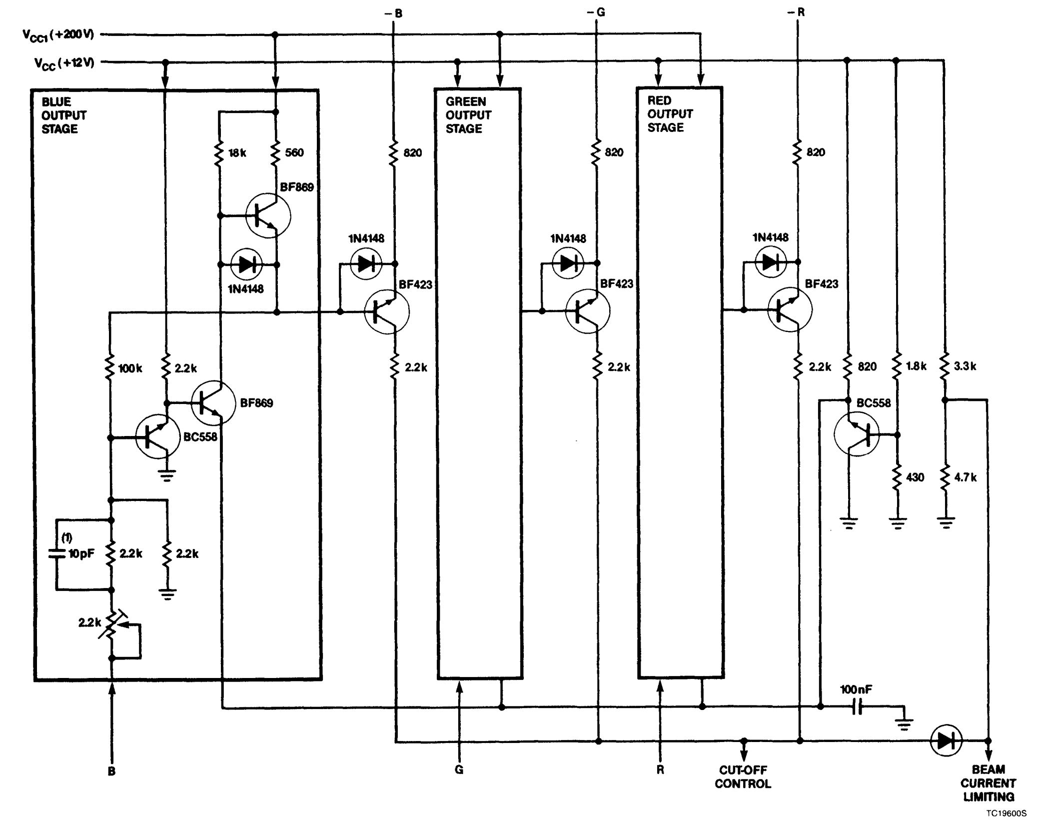


Figure 1. Types of Ultra-Black Levels

Video Control Combination Circuit With Automatic Cut-Off Control

TDA4580



NOTE:
1. Capacitor value depends on circuit layout.

Figure 2a. Part of Typical Application Circuit Diagram Using the TDA4580; Continued in Figure 2b

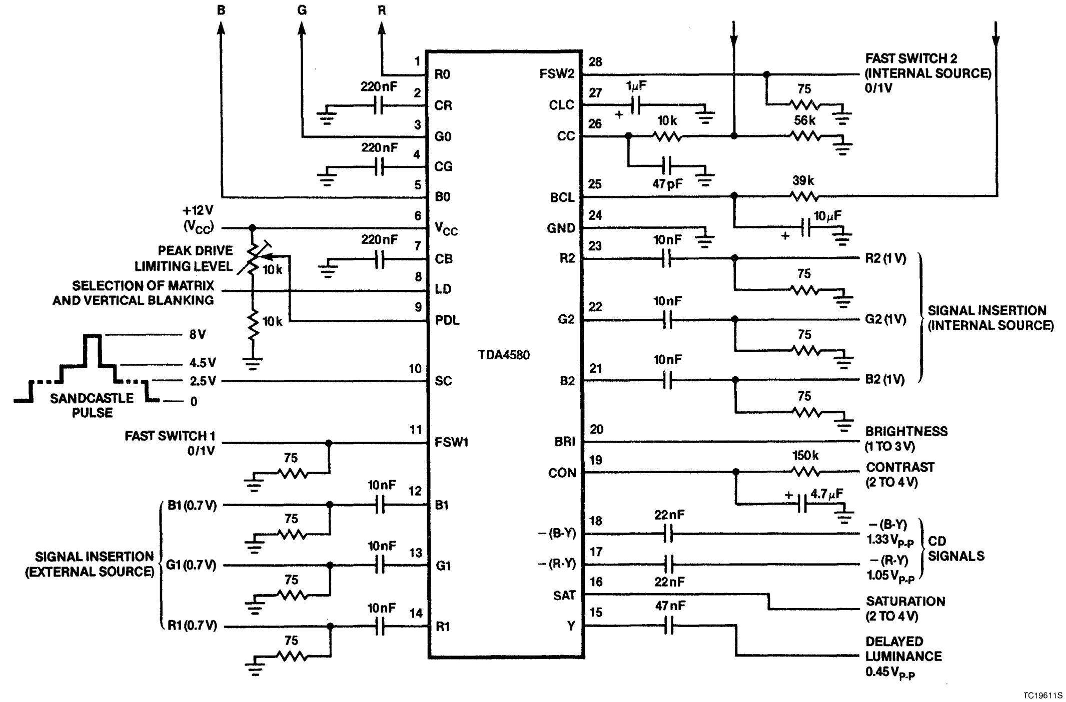
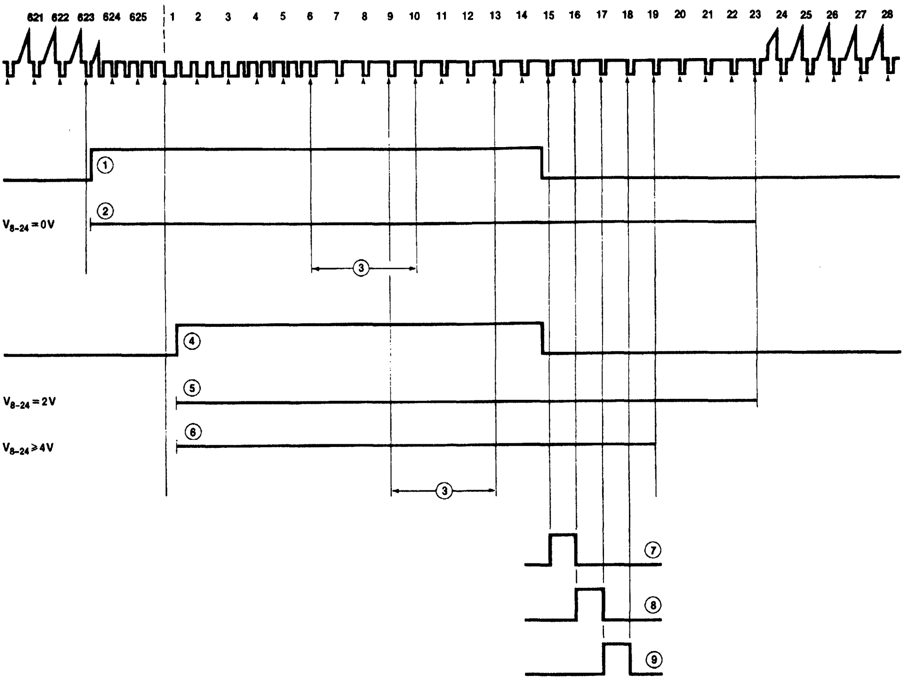


Figure 2b. Part of Typical Application Circuit Diagram Using the TDA4580; Continued from Figure 2a

Video Control Combination Circuit With Automatic Cut-Off Control

TDA4580



NOTES:
1. Vertical part of sandcastle pulse starts with equalizing pulses and ends with flyback.
2. Blanking period of 25 complete lines.
3. Leakage measuring period (LM).
4. Vertical part of sandcastle pulse starts and ends with flyback.
5. Blanking period of 22 complete lines.
6. Blanking period of 18 complete lines.
7. Cut-off measuring line for red signal (MR).
8. Cut-off measuring line for green signal (MG).
9. Cut-off measuring line for blue signal (MB).

Figure 3. Blanking and Measuring Lines

Signetics

TDA8442
Quad DAC With I^2C Interface

Product Specification

Linear Products

DESCRIPTION

The TDA8442 consists of four 6-bit D/A converters and 3 output ports. This IC was designed to provide I^2C control, by replacing the potentiometers, for the TDA3560-series single-chip color decoders. Control of the IC is performed via the two-line, bidirectional I^2C bus.

FEATURES

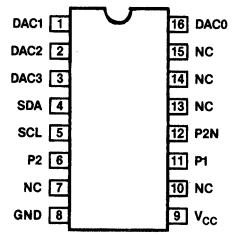
- **6-bit resolution**
- **3 output ports**
- **I^2C control**

APPLICATIONS

- **I^2C interface control**
- **System control**
- **Switching**

PIN CONFIGURATION

N Package



TOP VIEW

CD12850S

PIN NO.	SYMBOL	DESCRIPTION
1	DAC1	Analog output 1
2	DAC2	Analog output 2
3	DAC3	Analog output 3
4	SDA	Serial data line } I^2C bus
5	SCL	Serial clock line } I^2C bus
6	P2	Port 2 NPN collector output with internal pull-up resistor
7	NC	Not connected
8	GND	Supply return (ground)
9	V_{CC}	Positive supply voltage
10	NC	Not connected
11	P1	Port 1 open NPN emitter output
12	P2N	Inverted P2 output
13	NC	Not connected
14	NC	Not connected
15	NC	Not connected
16	DAC0	Analog output 0

ORDERING INFORMATION

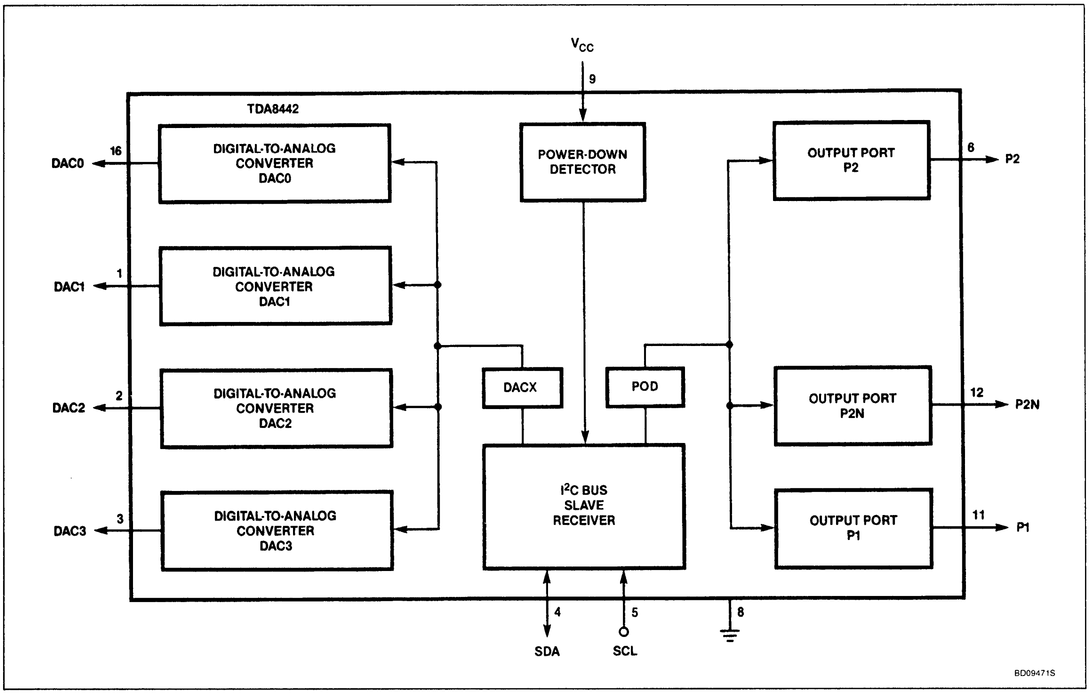
DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
16-Pin Plastic DIP (SOT-38)	−20°C to +70°C	TDA8442N

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{CC}	Supply voltage range (Pin 9)	−0.3 to +13.2	V
	Input/output voltage ranges		
V_{SDA}	(Pin 4)	−0.3 to +13.2	V
V_{SCL}	(Pin 5)	−0.3 to +13.2	V
V_{CC2}	(Pin 6)	−0.3 to V_{CC}[1]	V
V_{CC2N}	(Pin 12)	−0.3 to V_{CC}[1]	V
V_{CC1}	(Pin 11)	−0.3 to V_{CC}[1]	V
V_{DAX}	(Pins 1 to 3 and Pin 16)	−0.3 to V_{CC}[1]	V
P_{TOT}	Total power dissipation	1	W
T_A	Operating ambient temperature range	−20 to +70	°C
T_{STG}	Storage temperature range	−65 to +150	°C

NOTE:
1. Pin voltage may exceed V_{CC} if the current in that pin is limited to 10mA.

BLOCK DIAGRAM



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DC AND AC ELECTRICAL CHARACTERISTICS $T_A = +25°C$; $V_{CC} = 12V$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supplies					
V_{CC}	Supply voltage (Pin 9)	10.8	12	13.2	V
I_{CC}	Supply currents (no outputs loaded) (Pin 9)		12		mA
I^2C bus inputs SDA (Pin 4) and SCL (Pin 5)					
V_{IH}	Input voltage High[1]	3		$V_{CC} - 1$	V
V_{IL}	Input voltage Low	−0.3		1.5	V
I_{IH}	Input current High[1]			10	μA
I_{IL}	Input current Low[1]			10	μA
I^2C bus output SDA (Pin 4) (open-collector)					
V_{OL}	Output voltage Low at $I_{OL} = 3.0mA$			0.4	V
I_{OL}	Maximum output sink current		5		mA
Ports P2 and P2N (Pins 6 and 12) (NPN collector output with pull-up resistor to V_{CC})					
R_O	Internal pull-up resistor to V_{CC}	5	10	15	kΩ
V_{OL}	Output voltage Low at $I_{OL} = 2mA$			0.4	V
I_{OL}	Maximum output sink current	2	5		mA
Port P1 (Pin 11) (open NPN emitter output)					
I_{OH}	Output current High at $0 < V_O < V_{CC} - 1.5V$	14			mA
I_{OL}	Output leakage current at $0 < V_O < V_{CC}V$			100	μA
Digital-to-analog outputs Output DAC0 (Pin 16)					
V_{OMAX}	Maximum output voltage (unloaded)[2]	3			V
V_{OMIN}	Minimum output voltage (unloaded)[2]			1	V
V_{OLSB}	Positive value of smallest step[2] (1 LSB)	0		100	mV
	Deviation from linearity			150	mV
Z_O	Output impedance at $-2 < I_O < +2mA$			70	Ω
$-I_{OH}$	Maximum output source current	2		6	mA
I_{OL}	Maximum output sink current	2	8		mA
Output DAC1 (Pin 1)					
V_{OMAX}	Maximum output voltage (unloaded)[2]	4			V
V_{OMIN}	Minimum output voltage (unloaded)[2]			1.7	V
V_{OLSB}	Positive value of smallest step[2] (1 LSB)	0		120	mV
	Deviation from linearity			170	mV
Z_O	Output impedance at $-2 < I_O < +2mA$			70	Ω
$-I_{OH}$	Maximum output source current	2		6	mA
I_{OL}	Maximum output sink current	2	8		mA

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $T_A = +25°C$; $V_{CC} = 12V$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Output DAC2 (Pin 2)					
V_{OMAX}	Maximum output voltage (unloaded)[2]	4			V
V_{OMIN}	Minimum output voltage (unloaded)[2]			1.7	V
V_{OLSB}	Positive value of smallest step[2] (1 LSB)	0		120	mV
	Deviation from linearity			170	mV
Z_O	Output impedance at $-2 < I_O < +2mA$			70	Ω
$-I_{OH}$	Maximum output source current	2		6	mA
I_{OL}	Maximum output sink current	2	8		mA
Output DAC3 (Pin 3)					
V_{OMAX}	Maximum output voltage (unloaded)[2]	10			V
V_{OMIN}	Minimum output voltage (unloaded)[2]			1	V
V_{OLSB}	Positive value of smallest step[2] (1 LSB)	0		350	mV
	Deviation from linearity			0.50	V
Z_O	Output impedance at $-2 < I_O < +2mA$			70	Ω
$-I_{OH}$	Maximum output source current	2		6	mA
I_{OL}	Maximum output sink current	2	8		mA
Power-down reset					
V_{CCD}	Maximum value of V_{CC} at which power-down reset is active	6		10	V
t_R	Rise time of V_{CC} during power-on (V_{CC} rising from 0V to V_{CCD})	5			μs

NOTES:
1. If $V_{CC} < 1V$, the input current is limited to 10μA at input voltages up to 13.2V.
2. Values are proportional to V_{CC}.

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FUNCTIONAL DESCRIPTION

Control

Analog control is facilitated by four 6-bit digital-to-analog converters (DAC0 to DAC3). The values of the output voltages from the DACs are set via the I²C bus.

The high-current output port (P1) is suitable for switching between internal and external RGB signals. It is an open NPN emitter output capable of sourcing 14mA (minimum).

The two output ports (P2 and P2N) can be used for NTSC/PAL switching. These are NPN collector outputs with internal pull-up resistors of 10kΩ (typical). Both outputs are capable of sinking up to 2mA with a voltage drop of less than 400mV. If one output is programmed to be Low, the other output will be High, and vice versa.

Reset

The power-down reset mode occurs whenever the positive supply voltage falls below 8.5V (typical) and resets all registers to a defined state.

OPERATION

Write

The TDA8442 is controlled via the I²C bus. Programming of the TDA8442 is performed using the format shown in Figure 1.

Acknowledge (A) is generated by the TDA8442 only when a valid address is received and the device is not in the power-down reset mode ($V_{CC} > 8.5V$ (typ)).

Control

Control is implemented by the instruction bytes POD (port output data) and DACX (digital-to-analog converter control), and the corresponding data/control bytes (see Figure 2).

POD Bit P1 — If a '1' is programmed, the P1 output is forced High. If a '0' is programmed, or after a power-down reset, the P1 output is Low (high-impedance state).

POD Bit P2/P2N — If a '1' is programmed, the P2 output goes High and the P2N output goes Low. If a '0' is programmed, and after a power-down reset, the P2 output is Low and the P2N output is High.

DAX Bits AX5 to AX0 — The digital-to-analog converter selected corresponds to the decimal equivalent of the two bits X1 and X0. The output voltage of the selected DAC is programmed using Bits AX5 to AX0, the lowest value being all AX5 to AX0 data at '0', or when power-down reset has been activated.

	MODULE ADDRESS									INSTRUCTION BYTE									DATA/CONTROL BYTE									
S	1	0	0	0	1	0	0	0	A	I_7	I_6	I_5	I_4	I_3	I_2	I_1	I_0	A	D_7	D_6	D_5	D_4	D_3	D_2	D_1	D_0	A	P
	MSB							R/W̄		MSB									MSB									

AF04662S

Figure 1. TDA8442 Programming Format

	INSTRUCTION BYTE								DATA/CONTROL BYTE							
	I_7	I_6	I_5	I_4	I_3	I_2	I_1	I_0	D_7	D_6	D_5	D_4	D_3	D_2	D_1	D_0
POD	X	X	X	X	1	0	0	0				P2				
DACX	X	X	X	X	0	0	X1	X0			AX5	AX4	AX3	AX2	AX1	AX0

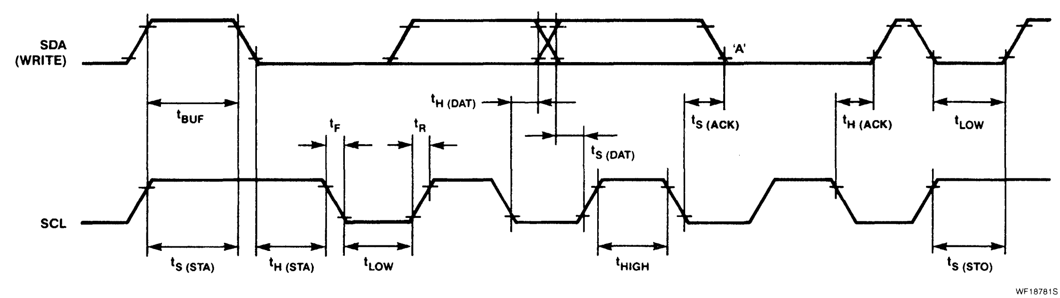
AF04682S

Figure 2. Control Programming

I²C BUS TIMING

Bus loading conditions: 4kΩ pull-up resistor to +5V; 200pF capacitor to GND.
All values are referred to V_{IH} = 3V and V_{IL} = 1.5V.

SYMBOL	PARAMETER	LIMITS Min	Typ	Max	UNIT
t_{BUF}	Bus free before start	4			μs
t_{SU}, t_{STA}	Start condition setup time	4			μs
t_{HD}, t_{STA}	Start condition hold time	4			μs
t_{LOW}	Low period SCL, SDA	4			μs
t_{HIGH}	High period SCL	4			μs
t_R	Rise time SCL, SDA			1	μs
t_F	Fall time SCL, SDA			0.30	μs
t_{SU}, t_{DAT}	Data setup time (write)	0.25			μs
t_{HD}, t_{DAT}	Data hold time (write)	0			μs
t_{SU}, t_{ACK}	Acknowledge (from TDA8442) setup time			2	μs
t_{HD}, t_{ACK}	Acknowledge (from TDA8442) hold time	0			μs
t_{SU}, t_{STO}	Stop condition setup time	4			μs



NOTE:
Reference levels are 10 and 90%.

Figure 3. I²C Bus Timing, TDA8442

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Signetics

TDA8443, TDA8443A
RGB/YUV Switch

Preliminary Specification

Linear Products

DESCRIPTION

The TDA8443/8443A is intended to be used in color TV sets which have more than one base-band video source. The IC has two sets of inputs. The first (Inputs 1) is intended for the internal video signals (R-Y), Y, (B-Y), and the associated synchronization pulse coming from the color decoder; the second (Inputs 2) is intended for external video signals R, G, B, and the associated synchronization pulse coming from the accessory inputs. The latter ones (Inputs 2) can also consist of the video signals (R-Y), Y, (B-Y), and the associated synchronization pulse. The RGB signals at Inputs 2 can also be matrixed internally into the luminance signal Y and the color-difference signals (R-Y) and (B-Y) before they become available at the outputs. By means of I^{2}C bus mode or manual control (control by DC voltages), one of these inputs can be selected and will be available at the outputs. The IC contains three pins for programming the sub-address; this means that within one TV set the system can be expanded up to seven ICs. The TDA8443 is designed to be used with the CCTV levels, while the TDA8443A is designed to be used for the standard decoder signal levels.

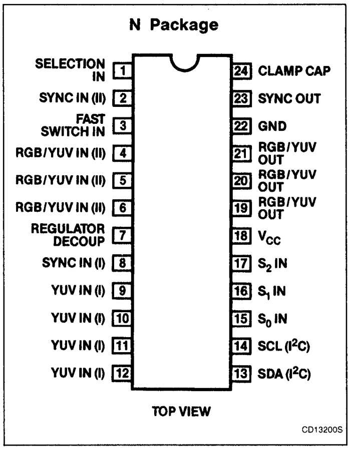
FEATURES

- **Two RGB/YUV selectable clamped inputs with associated sync**
- **An RGB/YUV matrix**
- **3-State switching with an OFF state**
- **Four amplifiers with selectable gain**
- **Fast switching to allow for mixed mode**
- **I^{2}C or non-I^{2}C mode (control by DC voltages)**
- **Slave receiver in the I^{2}C mode**
- **External OFF command**
- **System expansion possible up to 7 devices**

APPLICATIONS

- **TV receivers**
- **Video switching**

PIN CONFIGURATION



ORDERING INFORMATION

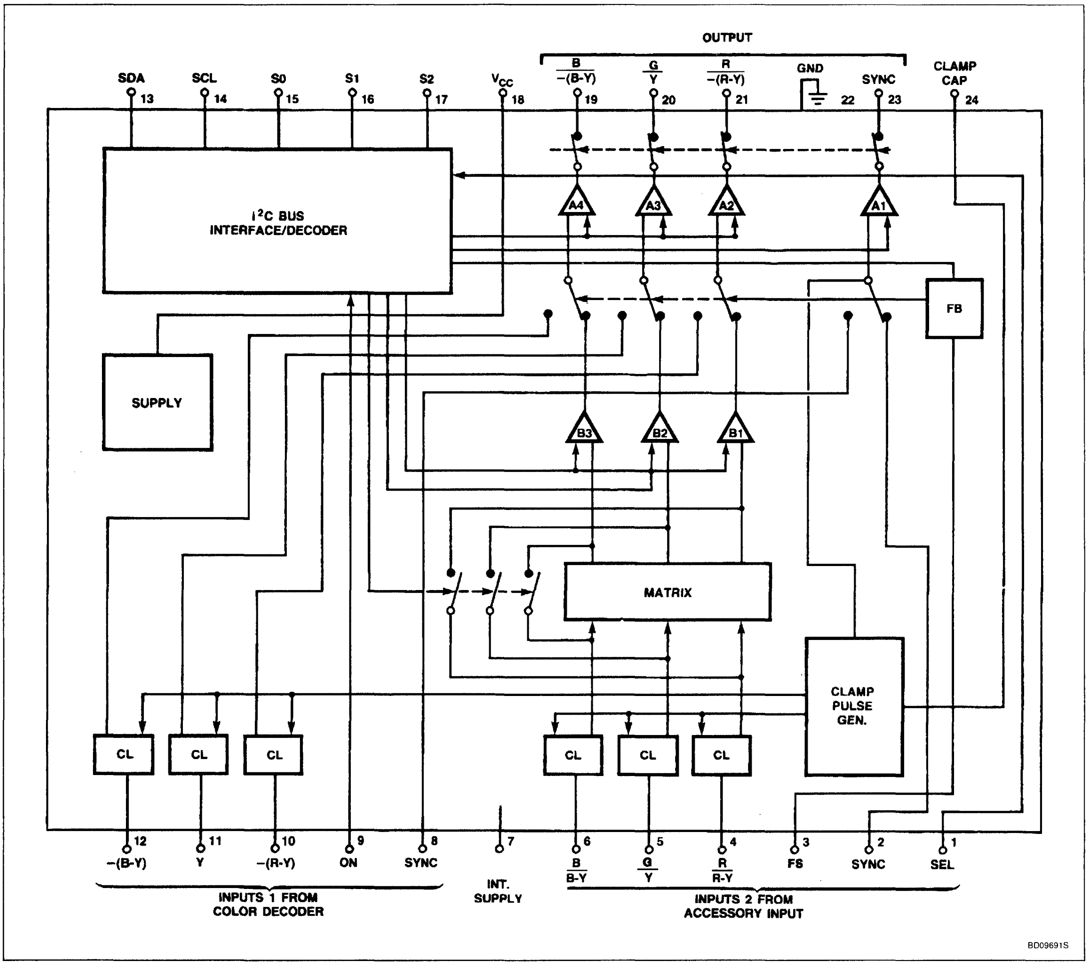
DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
24-Pin Plastic DIP (SOT-101)	0 to +70°C	TDA8443N
24-Pin Plastic DIP (SOT-101)	0 to +70°C	TDA8443AN

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
T_{STG}	Storage temperature range	-65 to +150	°C
T_A	Operating ambient temperature range	0 to +70	°C
V_{18-7}	Supply voltage	14	V
P_D	Total power dissipation		W
T_{JMAX}	Maximum junction temperature	125	°C
V_{SDA} V_{SCL}	Input voltage range Pin 13 14 other pins	-0.3 to 14 -0.3 to 14 -0.3 to V_{CC}+0.3	V V V
I_{OMAX}	Maximum output current	TBD	mA

RGB/YUV Switch | TDA8443, TDA8443A

BLOCK DIAGRAM



DC ELECTRICAL CHARACTERISTICS $T_A = 25°C$ and $V_{CC} = 12V$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
V_{18-7}	Supply voltage	10		13.2	V
I_{18}	Supply current		TBF	TBF	mA
RGB/YUV channels					
	Absolute gain difference with respect to programmed value		0	10	%
	Relative gain difference between any 2 channels of one input		0	5	%
I_{IN}	Input current		TBF	0.3	μA
Z_{OUT}	Output impedance		TBF	30	Ω
	3dB bandwidth (mode 0 or 2)		10		MHz
	3dB bandwidth mode 1		10		MHz
	Mutual time difference at output if all inputs of one source are connected together		TBF	25	ns
	Maximum output amplitude of YUV signals	2.8			V_{P-P}
	Crosstalk between inputs of same source, at 5MHz[1]			−30	dB
	Crosstalk between different sources			−50	dB
	Isolation (OFF state) at 10MHz	50			dB
	Differential gain at nominal output signals: R-Y = $1.05V_{P-P}$ B-Y = $1.33V_{P-P}$ Y = $0.34V_{P-P}$			10	%
S/N	Signal-to-noise ratio at nominal input	50			dB
BW	Bandwidth = 5MHz[2]				
	Supply voltage rejection[3]	50			dB
	DC level of outputs during clamp		5.3		V
Sync channels					
	Gain difference with respect to programmed value			10	%
BW	3dB bandwidth		TBF		MHz
	Input amplitude of sync pulse for proper operation of clamp pulse generator	0.2		2.5	V_{P-P}
Z_{OUT}	Output impedance		TBF	30	Ω
	Maximum output amplitude (undistorted)	2.5			V_{P-P}
	DC level on top of sync pulse at output	TBF	1.8	TBF	V
I^2C bus inputs/outputs					
	SDA input (Pin 13)				
	SCL input (Pin 14)				
V_{IH}	Input voltage High	3		V_{CC}	V
V_{IL}	Input voltage Low	−0.3		1.5	V
I_{IH}	Input current High			10	μA
I_{IL}	Input current Low			10	μA
	SDA output (open-collector)				
V_{OL}	Output voltage Low at IO-L = 3mA			0.4	V
I_{OL}	Maximum output sink current		5		mA

DC ELECTRICAL CHARACTERISTICS (Continued) $T_A = 25°C$ and $V_{CC} = 12V$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Sub-address inputs S0 (Pin 15), S1 (Pin 16), S2 (Pin 17)					
V_{IH}	Input voltage High	3		V_{CC}	V
V_{IL}	Input voltage Low	-0.3		0.4	V
I_{IH}	Input current High			TBF	µA
I_{IL}	Input current Low			TBF	µA
Fast switching pin					
V_{3-7}	Input voltage High	1		3	V
V_{3-7}	Input voltage Low	-0.3		0.4	V
I_3	Input current High			TBF	µA
I_3	Input current Low			TBF	µA
	Switching delay[4]			TBF	
	Switching time[4]			TBF	
SEL pin					
V_{1-7}	Input voltage High	3		V_{CC}	V
V_{1-7}	Input voltage Low	-0.3		0.4	V
I_1	Input current High			TBF	µA
I_1	Input current Low			TBF	µA
ON pin					
V_{9-7}	Input voltage High	3		V_{CC}	V
V_{9-7}	Input voltage Low	-0.3		1.5	V
I_9	Input current High			TBF	µA
I_9	Input current Low			TBF	µA

NOTES:

1. Crosstalk is defined as the ratio between the output signal originating from another input and the nominal output signal on the same output.
2. $S/N = 20\log \frac{V_{OP\text{-}P}}{V_O \text{ noise RMS } B = 5MHz}$.
3. Supply voltage rejection $= 20\log \frac{V_R \text{ supply}}{V_R \text{ on output}}$.
4. Fast switching input signal
   Output signal: YUV
   Input : 0V input 1, mode 2
   0.75V RGB input 2, mode 1

RGB/YUV Switch TDA8443, TDA8443A

FUNCTIONAL DESCRIPTION

The circuit contains two sets of inputs: input 1 from the color decoder (color difference signals), and input 2 from the accessory input, RGB, or possibly YUV, both with associated synchronization inputs.

In the RGB mode, the signals are matrixed internally to color difference signals for further processing in a control circuit (e.g., TDA8461).

The inputs are clamped, thus the clamp pulse is internally derived from the sync signals. The outputs can be made high-ohmic (OFF) in order to be able to put several circuits in parallel.

Control

The circuit can be controlled by an I²C bus or directly by DC voltages. The fast switching input can be operated by Pin 16 of the accessory input.

I²C BUS MODE

The protocol for the TDA8443 for I²C bus mode is:

STA	A6	A5	A4	A3	A2	A1	A0	R/W	AC	D7	D6	D5	D4	D3	D2	D1	D0	AC	STO

STA : Start condition
A6 : 1 } fixed address bits
A5 : 1 }
A4 : 0 }
A3 : 1 }
A2 : Sub-address bit set by S2
A1 : Sub-address bit set by S1
A0 : Sub-address bit set by S0
R/W : Read/Write bit (= 0 only write mode allowed)

AC : Acknowledge, generated by the TDA8443
D7 : MOD1 } mode control bits, see Table 2
D6 : MOD0 }
D5 : G2 } gain control bits, see Table 4
D4 : G1 }
D3 : G0 }
D2 : PRIOR, priority bit
D1 : ON/OFF bit
D0 : ON/OFF active bit

Table 1. Sub-Addressing

SLAVE ADDRESS BITS			ADDRESS SELECT PINS		
A2	A1	A0	S2	S1	S0
0	0	0	GND	GND	GND
0	0	1	GND	GND	V_{CC}
0	1	0	GND	V_{CC}	GND
0	1	1	GND	V_{CC}	V_{CC}
1	0	0	V_{CC}	GND	GND
1	0	1	V_{CC}	GND	V_{CC}
1	1	0	V_{CC}	V_{CC}	GND
1	1	1	V_{CC}	V_{CC}	V_{CC}

NOTE:
Non-I²C bus operation, see Table 5.

Table 2. Mode Control

MOD1	MOD0	MODE	FUNCTION
0	0	0	Inputs 2 are selected directly
0	1	1	Inputs 2 are selected via RGB/YUV matrix
1	0	2	Inputs 1 are selected directly
1	1	3	Reserved; not to be used

Table 3. Priority Fast Switching Action

PRIOR	FS	MODE SELECTED
0	X	As set by mode control (Table 2)
1	0.4V	Mode 2
1	1 – 3V	Mode 1 if mode 1 is selected Mode 0 if mode 0 or 2 is selected

Table 4. Gain Settings (see Block Diagram)

G2	G1	G0	TDA8443/C3			TDA8443A/C3	
			A1	A2, A3, A4	B1, B3	B1, B3	B2
0	0	0	1	1	−0.6	−1	0.45
0	0	1	1	1	1	1	1
0	1	0	Reserved; not to be used				
0	1	1	1	1	−0.6	−1	0.45
1	0	0	2	2	−0.6	−1	0.45
1	0	1	2	1	1	1	1
1	1	0	2	2	1	1	1
1	1	1	2	1	−0.6	−1	0.45

NOTES:

Matrix equations: relations between output and input signals of the matrix

$Y = 0.3R + 0.59V + 0.11B$

$R\text{-}Y = 0.7R - 0.59V - 0.11B$

$B\text{-}Y = -0.3R - 0.59V + 0.89B$

ON BIT

ON	FUNCTION
0	OFF, no output signal, outputs high-ohmic
1	ON, normal functioning

OFFACT-ON (Pin 9) Function

OFFACT	ON	FUNCTIONING
0	L	OFF
0	H	In accordance with last defined D7 – D1 (may be entered while OFF = L)
1	X	In accordance with last defined D7 – D1

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POWER-ON RESET

When the circuit is switched on in the I^2C mode, bits D0 – D7 are set to zero.

Table 5. Non-I^2C Bus Mode (S2 = S1 = S0 = 0)

CONTROL			MODE SWITCHED BY FS	GAIN SETTINGS				
				TDA8443			TDA8443A	
SDA	SCL	SEL		A1	A4, A3, A2	B1, B3	B1, B3	B2
L	L	L	2/0	1	1	1	1	1
L	L	H	2/0	1	2	1	1	1
L	H	L	2/1	1	1	−0.6	−1	0.45
L	H	H	2/0	1	1	−0.6	−1	0.45
H	L	L	2/0	2	1	1	1	1
H	L	H	2/0	2	2	1	1	1
H	H	L	2/1	2	1	−0.6	−1	0.45
H	H	H	2/0	2	1	−0.6	−1	0.45

Fast Switching Input

F S	MODE SELECTED
⩽0.4V	Mode 2
1 – 3V	Mode 0 or 1 as set by control

ON Input

ON	FUNCTION
L	OFF, no output signal, outputs high-ohmic
H	Functioning as determined in Table 5

I^2C BUS LOAD CONDITIONS

4kΩ pull-up resistor to +5V; 200pF capacitor to GND.

All values are referred to V_{IH} = 3V and V_{IL} = 1.5V.

SYMBOL	PARAMETER	RATING			UNIT
		Min	Typ	Max	
t_{BUF}	Bus free before start	4			μs
t_{SU}, t_{STA}	Start condition setup time	4			μs
t_{HD}, t_{STA}	Start condition hold time	4			μs
t_{LOW}	SCL, SDA Low period	4			μs
t_{HIGH}	SCL High period	4			μs
t_R	SCL, SDA rise time			1	μs
t_F	SCL, SDA fall time			0.3	μs
t_{SU}, t_{DAT}	Data setup time (write)	1			μs
t_{HD}, t_{DAT}	Data hold time (write)	1			μs
t_{SU}, t_{CAC}	Acknowledge (from TDA8443) setup time			2	μs
t_{HD}, t_{CAC}	Acknowledge (from TDA8443) hold time	0			μs

NOTE:
Timings t_{SU}, t_{DAT} and t_{HD}, t_{DAT} deviate from the I^2C bus specification.
After reset has been activated, transmission may only be started after 50μs delay.

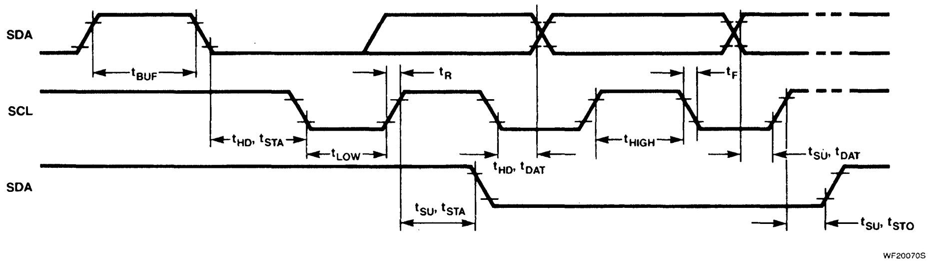

Figure 1. I^2C Timing Diagram

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Table 6. Application Information

INPUT 1	INPUT 2	OUTPUT	MODE	G2	G1	G0
YUV/S 0.34/−1.33/−1.05/0.3		YUV/S 0.34/−1.33/−1.05/0.6	2	1	1	1
	RGB/S 0.75/0.75/0.75/0.3		1	1	1	1
YUV/S 0.34/−1.33/−1.05/0.3			2	1	0	0
	RGB/S 0.75/0.75/0.75/0.3	YUV/S 0.68/−2.66/−2.1/0.6	1	1	0	0
YUV/S 0.34/−1.33/−1.05/0.3		YUV/S 0.34/−1.33/−1.05/0.6	2	1	0	1
	YUV/S 0.34/−1.43/−1.05/0.3		0	1	0	1
YUV/S 0.34/−1.33/−1.05/0.3		YUV/S 0.68/−2.66/−2.1/0.6	2	1	1	0
	YUV/S 0.34/−1.33/−1.05/0.3		0	1	1	0

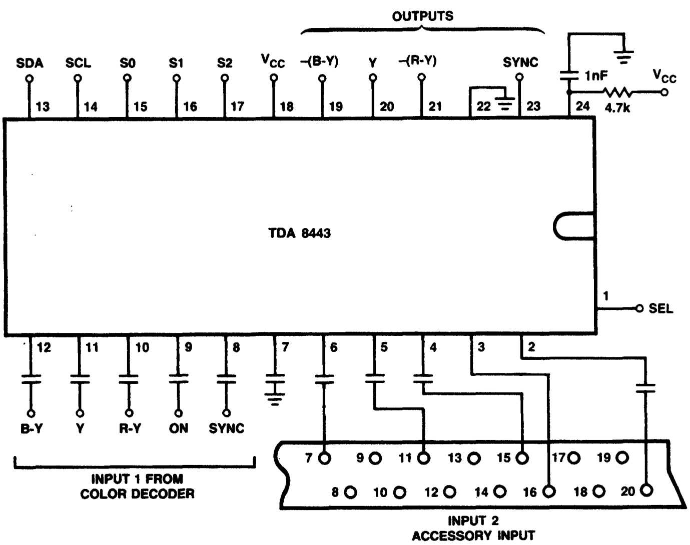


Figure 2. Application of the TDA8443

Signetics

Section 11 Special-Purpose Video Processing

Linear Products

INDEX

11

Signetics

NE568
150MHz Phase-Locked Loop

Preliminary Specification

Linear Products

DESCRIPTION

The NE568 is a monolithic phase-locked loop (PLL) which operates from 1Hz to frequencies in excess of 150MHz. The integrated circuit consists of a limiting amplifier, a current-controlled oscillator (ICO), a phase detector, a level shift circuit, V/I and I/V converters, an output buffer, and bias circuitry with temperature and frequency compensating characteristics. The design of the NE568 is particularly well-suited for demodulation of FM signals with extremely large deviation in systems which require a highly linear output. In satellite receiver applications with a 70MHz IF, the NE568 will demodulate ±20% deviations with less than 1.0% typical non-linearity. In addition to high linearity, the circuit has a loop filter which can be configured with series or shunt elements to optimize loop dynamic performance. The NE568 is available in 20-pin dual in-line and 20-pin SO (surface-mounted) plastic packages.

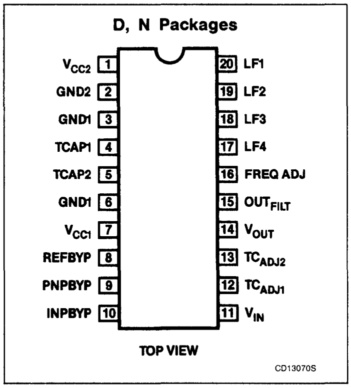
FEATURES

- **Operation to 150MHz**
- **High linearity buffered output**
- **Series or shunt loop filter component capability**
- **Temperature compensated**

APPLICATIONS

- **Satellite receivers**
- **Fiber-optic video links**
- **VHF FSK demodulators**
- **Clock recovery**

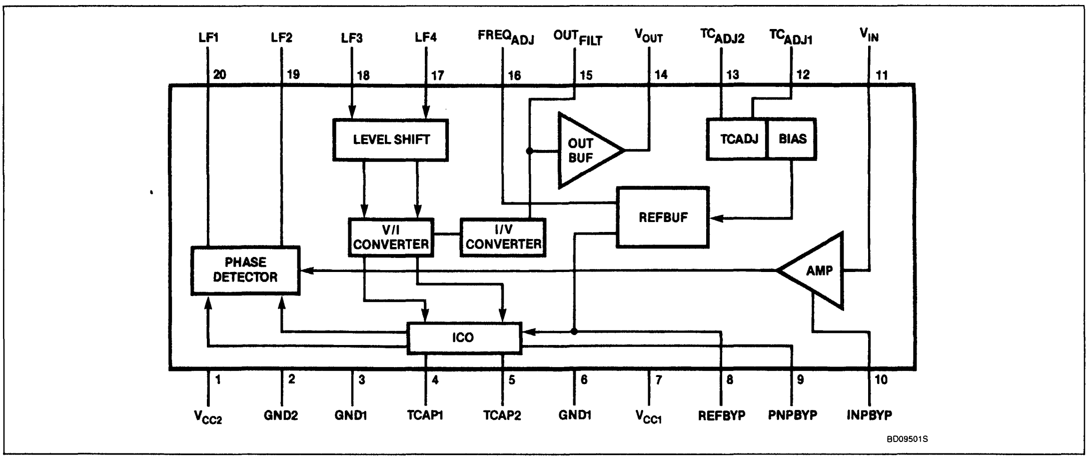
PIN CONFIGURATION



ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
20-Pin Plastic SOL Package	0 to +70°C	NE568D
20-Pin Plastic DIP	0 to +70°C	NE568N

BLOCK DIAGRAM



150MHz Phase-Locked Loop — NE568

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{CC}	Supply voltage	6	V
T_A	Operating free-air ambient temperature range	0 to +70	°C
T_J	Junction temperature	+150	°C
T_{STG}	Storage temperature range	-65 to +150	°C
P_{DMAX}	Maximum power dissipation	500	mW

ELECTRICAL CHARACTERISTICS

The electrical characteristics listed below are actual tests (unless otherwise stated) performed on each device with an automatic IC tester prior to shipment. Performance of the device in automated test setup is not necessarily optimum. The NE568 is layout-sensitive. Evaluation of performance for correlation to the data sheet should be done with the circuit and layout of Figures 1 – 3 with the evaluation unit soldered in place. (Do not use a socket!)

DC ELECTRICAL CHARACTERISTICS

$T_A = 25°C$, $V_{CC} = 5V$, $f_O = 70MHz$, Test Circuit Figure 1, $f_{IN} = -20dBm$, $R_4 = 0\Omega$ (ground), unless otherwise specified.

SYMBOL	PARAMETER	TEST CONDITIONS	LIMITS			UNIT
			Min	Typ	Max	
V_{CC}	Supply voltage		4.75	5	5.25	V
I_{CC}	Supply current			60	75	mA

AC ELECTRICAL CHARACTERISTICS

SYMBOL	PARAMETER	TEST CONDITIONS	LIMITS			UNIT
			Min	Typ	Max	
f_{OSC}	Maximum oscillator operating frequency[3]		150			MHz
	Input signal level		50 −20[1]		2000 +10	mV_{P-P} dBm
BW	Demodulated bandwidth			$f_O/7$		MHz
	Non-linearity[5]	Dev = ±20%, Input = −20dBm		1.0	4.0	%
	Lock range[2]	Input = −20dBm	±25	±35		% of f_O
	Capture range[2]	Input = −20dBm	±20	±30		% of f_O
	TC of f_O	Figure 1		100		ppm/°C
R_{IN}	Input resistance[4]		1			kΩ
	Output impedance			6		Ω
	Demodulated V_{OUT}	Dev = ±20% of f_O measured at Pin 14	0.40	0.52		V_{P-P}
	AM rejection	$V_{IN} = -20dBm$ (30% AM) referred to ± 20% deviation		50		dB
f_O	Distribution[6]	Centered at 70MHz, $R_2 = 1.2k\Omega$, $C_2 = 17pF$, $R_4 = 0\Omega$ ($C_2 + C_{STRAY} = 20pF$)	−15	0	+15	%
f_O	Drift with supply	4.75V to 5.25V		1		%/V

NOTES:
1. Signal level to assure all published parameters. Device will continue to function at lower levels with varying performance.
2. Limits are set symmetrical to f_O. Actual characteristics may have asymmetry beyond the specified limits.
3. Not 100% tested, but guaranteed by design.
4. Input impedance depends on package and layout capacitance. See Figures 4 and 5.
5. Linearity is tested with incremental changes in input frequency and measurement of the DC output voltage at Pin 14 (V_{OUT}). Nonlinearity is then calculated from a straight line over the deviation range specified.
6. Free-running frequency is measured as feedthrough to Pin 14 (V_{OUT}) with no input signal applied.

150MHz Phase-Locked Loop | NE568

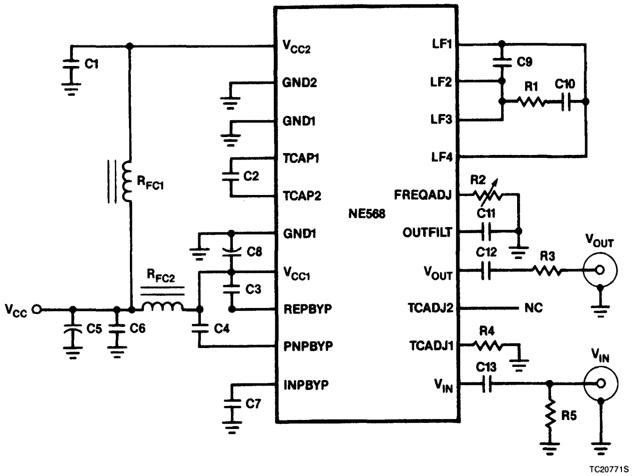


Figure 1. Test and Application Circuit

FUNCTIONAL DESCRIPTION

The NE568 is a high-performance phase-locked loop (PLL). The circuit consists of conventional PLL elements, with special circuitry for linearized demodulated output, and high-frequency performance. The process used has NPN transistors with $f_T > 6GHz$. The high gain and bandwidth of these transistors make careful attention to layout and bypass critical for optimum performance. The performance of the PLL cannot be evaluated independent of the layout. The use of the application layout in this data sheet and surface-mount capacitors are highly recommended as a starting point.

The input to the PLL is through a limiting amplifier with a gain of 200. The input of this amplifier is differential (Pins 10 and 11). For single-ended applications, the input must be coupled through a DC-blocking capacitor with low impedance at the frequency of interest. The single-ended input is normally applied to Pin 11 with Pin 10 AC-bypassed with a low-impedance capacitor. The input impedance is characteristically slightly above 500Ω. Impedance match is not necessary, but loading the signal source should be avoided. When the source is 50 or 75Ω, a DC-blocking capacitor is usually all that is needed.

Input amplification is low enough to assure reasonable response time in the case of large signals, but high enough for good AM rejection. After amplification, the input signal drives one port of a multiplier-cell phase detector. The other port is driven by the current-controlled oscillator (ICO). The output of the phase comparator is a voltage proportional to the phase difference of the input and ICO signals. The error signal is filtered with a low-pass filter to provide a DC-correction voltage, and this voltage is converted to a current which is applied to the ICO, shifting the frequency in the direction which causes the input and ICO to have a 90° phase relationship.

The oscillator is a current-controlled multivibrator. The current control affects the charge/discharge rate of the timing capacitor. It is common for this type of oscillator to be referred to as a voltage-controlled oscillator (VCO), because the output of the phase comparator and the loop filter is a voltage. To control the frequency of an integrated ICO multivibrator, the control signal must be conditioned by a voltage-to-current converter. In the NE568, special circuitry predistorts the control signal to make the change in frequency a linear function over a large control-voltage range.

The free-running frequency of the oscillator depends on the value of the timing capacitor connected between Pins 4 and 5. The value of the timing capacitor depends on internal resistive components and current sources. When $R_2 = 1.2k\Omega$ and $R_4 = 0\Omega$, a very close approximation of the correct capacitor value is:

$$C^* = \frac{0.0014}{f_O}\ F$$

where

$$C^* = C_2 + C_{STRAY}.$$

The temperature-compensation resistor, R_4, affects the actual value of capacitance. This equation is normalized to 70MHz. See Figure 6 for correction factors.

The loop filter determines the dynamic characteristics of the loop. In most PLLs, the phase detector outputs are internally connected to the ICO inputs. The NE568 was designed with filter output to input connections from Pins 20 (ϕ DET) to 17 (ICO), and Pins 19 (ϕ DET) to 18 (ICO) external. This allows the use of both series and shunt loop-filter elements. The loop constants are:

$K_D = 0.127V/Radian$ (Phase Detector Constant)

$K_O = 4.2 \times 10^9 \frac{\text{Radians}}{\text{V-sec}}$ (ICO Constant)

The loop filter determines the general characteristics of the loop. Capacitors C_9, C_{10}, and resistor R_1, control the transient output of the phase detector. Capacitor C_9 suppresses 70MHz feedthrough by interaction with 100Ω load resistors internal to the phase detector.

$$C_9 = \frac{1}{2\pi\ (50)(f_O)}\ F$$

At 70MHz, the calculated value is 45pF. Empirical results with the test and application board were improved when a 56pF capacitor was used.

The natural frequency for the loop filter is set by C_{10} and R_1. If the center frequency of the loop is 70MHz and the full demodulated bandwidth is desired, i.e., $f_{BW} = f_O/7 = 10MHz$, and a value for R_1 is chosen, the value of C_{10} can be calculated.

$$C_{10} = \frac{1}{2\pi\ R_1 f_{BW}}\ F$$

PARTS LIST AND LAYOUT 70MHz APPLICATION NE568D

C_1	100nF	±10%	Ceramic chip	1206
C_2[1]	18pF	±2%	Ceramic chip	0805
C_2[2]	34pF	±2%	Ceramic OR chip	
C_3	100nF	±10%	Ceramic chip	1206
C_4	100nF	±10%	Ceramic chip	1206
C_5	6.8μF	±10%	Tantalum	35V
C_6	100nF	±10%	Ceramic chip	1206
C_7	100nF	±10%	Ceramic chip	1206
C_8	100nF	±10%	Ceramic chip	1206
C_9	56pF	±2%	Ceramic chip	0805 or 1206
C_{10}	560pF	±2%	Ceramic chip	0805 or 1206
C_{11}	47pF	±2%	Ceramic chip	0805 or 1206
C_{12}	100nF	±10%	Ceramic chip	1206
C_{13}	100nF	±10%	Ceramic chip	1206
R_1	27Ω	±10%	Chip	⅛W
R_2	1.2kΩ		Trim pot	⅛W
R_3[3]	43Ω	±10%	Chip	⅛W
R_4[4]	4.7kΩ	±10%	Chip	⅛W
R_5[3]	50Ω	±10%	Chip	⅛W
RFC_1[5]	10μH	±10%	Surface mount	
RFC_2[5]	10μH	±10%	Surface mount	

NOTES:
1. $C_2 + C_{STRAY} = 20pF$.
2. $C_2 + C_{STRAY} = 36pF$ for temperature-compensated configuration with $R_4 = 4.7k\Omega$.
3. For 50Ω setup. $R_1 = 62\Omega$, $R_3 = 62\Omega$, $R_5 = 75\Omega$ for 75Ω application.
4. For test configuration $R_4 = 0\Omega$ (GND) and $C_2 = 18pF$.
5. 0Ω chip resistors (jumpers) may be substituted with minor degradation of performance.

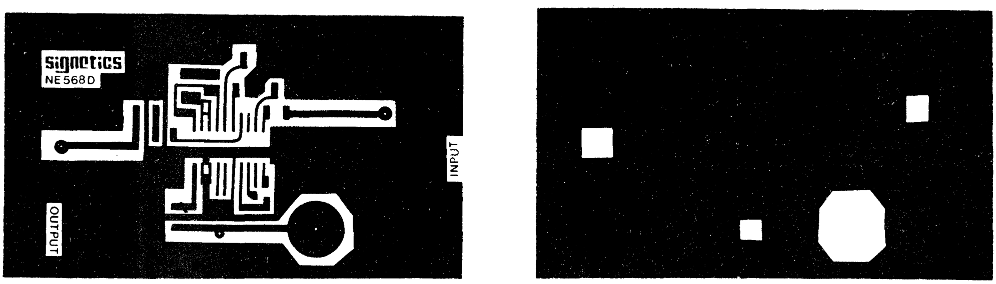
For the test circuit, R_1 was chosen to be 27Ω. The calculated value of C_{10} is 590pF; 560pF was chosen as a production value. (In actual satellite receiver applications, improved video with low carrier/noise has been observed with a wider loop-filter bandwidth.)

A typical application of the NE568 is demodulation of FM signals. In this mode of operation, a second single-pole filter is available at Pin 15 to minimize high frequency feedthrough to the output. The roll-off frequency is set by an internal resistor of 350Ω ±20%, and an external capacitor from Pin 15 to ground. The value of the capacitor is:

$$C11 = \frac{1}{2\pi\ (350) f_{BW}}\ F$$

Two final components complete the active part of the circuitry. A resistor from Pin 12 to ground sets the temperature stability of the circuit, and a potentiometer from Pin 16 to ground permits fine tuning of the free-running oscillator frequency. The Pin 16 potentiometer is normally 1.2kΩ. Adjusting this resistance controls current sources which affect the charge and discharge rates of the timing capacitor and, thus, the frequency. The value of the temperature stability resistor is chosen from the graph in Figure 6; the respective timing capacitor needs to be changed.

The final consideration is bypass capacitors for the supply lines. The capacitors should be ceramic chips, preferably surface-mount types. They must be kept very close to the device. The capacitors from Pins 8 and 9 return to V_{CC1} before being bypassed with a separate capacitor to ground. This assures that no differential loops are created which might cause instability. The layouts for the test circuits are recommended.



DF07740S DF07730S

a. Component Side Top of Board b. Back of Board

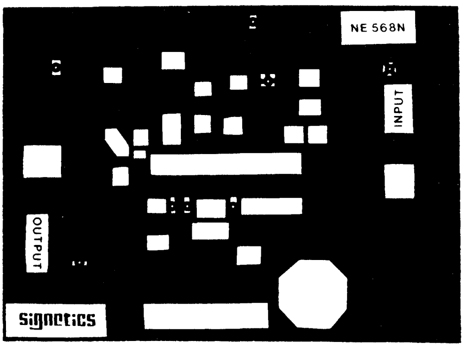
NOTES:
1. Board is laid out for King BNC Connector P/N KC-79-243-M06 or equivalent. Mount on bottom (back) of board. Add stand-off in each corner.
2. Back and top side ground must be connected at 8 point minimum.

Figure 2

PARTS LIST AND LAYOUT 70MHz APPLICATION NE568N

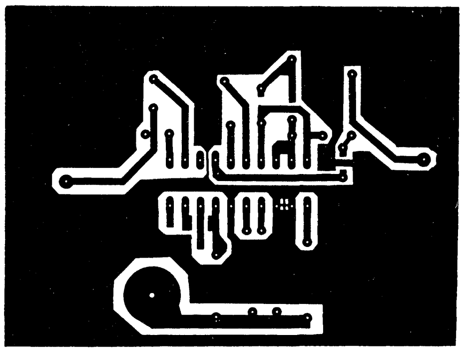
C_1	100nF	±10%	Ceramic chip	50V
C_2[1]	17pF	±2%	Ceramic OR chip	50V
C_2[2]	34pF	±2%	Ceramic chip	0805
C_3	100nF	±10%	Ceramic chip	50V
C_4	100nF	±10%	Ceramic chip	50V
C_5	6.8μF	±10%	Tantalum	35V
C_6	100nF	±10%	Ceramic OR chip	50V
C_7	100nF	±10%	Ceramic chip	50V
C_8	100nF	±10%	Ceramic chip	50V
C_9	56pF	±2%	Ceramic chip	50V
C_{10}	560pF	±2%	Ceramic chip	50V
C_{11}	47pF	±2%	Ceramic OR chip	50V
C_{12}	100nF	±10%	Ceramic OR chip	50V
C_{13}	100nF	±10%	Ceramic OR chip	50V
R_1	27Ω	±10%	Carbon	¼W
R_2	1.2kΩ		Trim pot	
R_3[3]	43Ω	±10%	Carbon	¼W
R_4[4]	4.7kΩ	±10%	Carbon	¼W
R_5[3]	50Ω	±10%	Carbon	¼W
RFC_1	10μH	±10%		
RFC_2	10μH	±10%		

NOTES:
1. $C_2 + C_{STRAY} = 20pF$ for test configuration with $R_4 = 0\Omega$.
2. $C_2 = 34pF$ for temperature-compensated configuration with $R_4 = 4.7k\Omega$.
3. For 50Ω setup. $R_1 = 62\Omega$; $R_3 = 75\Omega$ for 75Ω applications.
4. For test configuration $R_4 = 0\Omega$ (GND) and $C_2 = 17pF$.



DF07750S

a. **Component Side for Leaded Components**



DF07760S

b. **Solder Side of Board and Chip Capacitors**

NOTES:
1. Board is laid out for King BNC Connector P/N KC-79-243-M06 or equivalent mounted on the component side of the board.
2. Component side and solder side ground planes must be connected at 8 points minimum.

Figure 3

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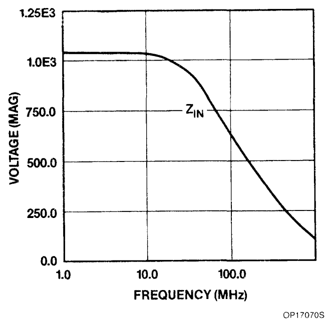


Figure 4. NE568 Input Impedance with CP = 0.5pF 20-Pin SO Package

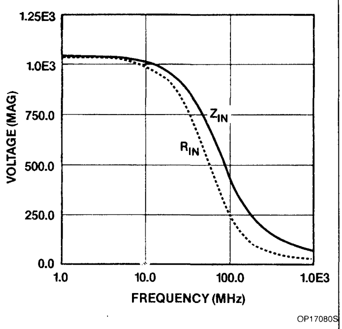


Figure 5. NE568 Input Impedance with CP = 1.49pF 20-Pin Dual In-Line Plastic Package

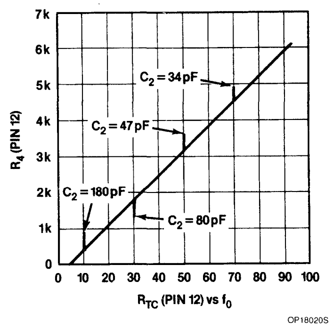


Figure 6

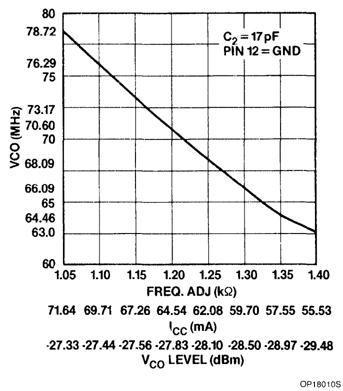


Figure 7. Typical VCO Frequency vs R_2 Adjustment

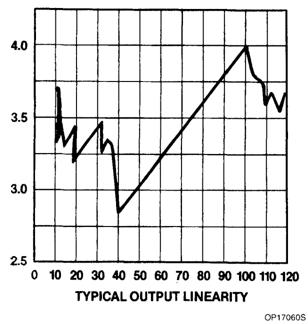


Figure 8. Typical Output Linearity

Signetics

PNA7509
7-Bit Analog-to-Digital Converter

Preliminary Specification

Linear Products

DESCRIPTION

The PNA7509 is a monolithic NMOS 7-bit analog-to-digital converter designed for video applications. The device converts the analog input signal into 7-bit binary coded digital words at a sampling rate of 22MHz.

The circuit comprises 129 comparators, a reference resistor chain, combining logic, transcoder stages, and TTL output buffers which are positive edge-triggered and can be switched into 3-State mode. The digital output is selectable in two's complement or binary coding.

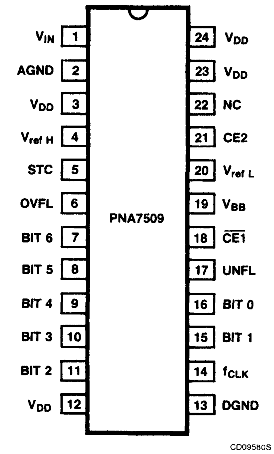
The use of separate outputs for overflow and underflow detection facilitates full-scale driving.

FEATURES

- **7-bit resolution**
- **22MHz clock frequency**
- **No external sample and hold required**
- **High input impedance**
- **Binary or two's complement 3-State TTL outputs**
- **Overflow and underflow 3-State TTL outputs**
- **Low reference current (250μA typ.)**
- **Positive supply voltages (+5V, +10V)**
- **Low power consumption (400mW typ.)**
- **Available in SO package**

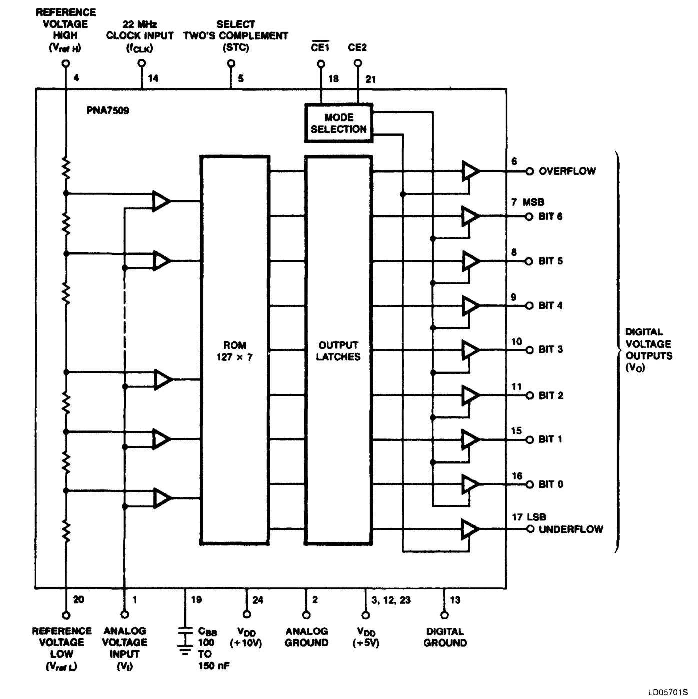
PIN CONFIGURATION

D, N Packages



PIN NO.	SYMBOL	DESCRIPTION
1	V_{IN}	Analog voltage input
2	AGND	Analog ground
3	V_{DD}	Positive supply voltage (+5V)
4	V_{REFH}	Reference voltage High
5	STC	Select two's complement
6	OVFL	Overflow
7	Bit 6	Most significant bit (MSB)
8	Bit 5	
9	Bit 4	
10	Bit 3	
11	Bit 2	
12	V_{DD}	Positive supply voltage (+5V)
13	DGND	Digital ground
14	f_{CLK}	22MHz clock input
15	Bit 1	
16	Bit 0	Least significant bit (LSB)
17	UNFL	Underflow
18	$\overline{CE1}$	Chip enable input 1
19	V_{BB}	Back bias output
20	V_{REFL}	Reference voltage Low
21	CE2	Chip enable input 2
22	NC	Not connected
23	V_{DD}	Positive supply voltage (+5V)
24	V_{DD}	Positive supply voltage (+10V)

BLOCK DIAGRAM



NOTE:
Pins 3, 12, and 23 must be connected to positive supply voltage +5V.

APPLICATIONS

- **High-speed A/D conversion**
- **Video signal digitizing**
- **Radar pulse analysis**
- **High energy physics research**
- **Transient signal analysis**

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7-Bit Analog-to-Digital Converter PNA7509

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
24-Pin Plastic DIP (SOT-101A)	0 to +70°C	PNA7509N
24-Pin Plastic SO (SOT-101)	0 to +70°C	PNA7509D

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{DD}	Supply voltage range (Pins 3, 12, 23)	7	V
V_{DD}	Supply voltage range (Pin 24)	12	V
V_{IN}	Input voltage range	7	V
I_{OUT}	Output current	5	mA
P_D	Power dissipation	1	W
T_{STG}	Storage temperature range	−65 to +150	°C
T_A	Operating ambient temperature range	0 to +70	°C

7-Bit Analog-to-Digital Converter — PNA7509

DC ELECTRICAL CHARACTERISTICS

$V_{DD} = V_{3,\ 12,\ 23-13} = 4.5$ to 5.5V; $V_{DD} = V_{24-2} = 9.5$ to 10.5V; $C_{BB} = 100nF$; $T_A = 0$ to +70°C, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS Min	Typ	Max	UNIT
Supply					
V_{DD}	Supply voltage (Pins 3, 12, 23)	4.5		5.5	V
V_{DD}	Supply voltage (Pin 24)	9.5		10.5	V
I_{DD}	Supply current (Pins 3, 12, 23)		51	85	mA
I_{DD}	Supply current (Pin 24)		11	18	mA
Reference voltages					
V_{REFL}	Reference voltage Low (Pin 20)	2.4	2.5	2.6	V
V_{REFH}	Reference voltage High (Pin 4)	5.0	5.1	5.2	V
I_{REF}	Reference current	150		450	μA
Inputs					
	Clock input (Pin 14)				
V_{IL}	Input voltage Low	−0.3		0.8	V
V_{IH}	Input voltage High	3.0		5.5	V
	Digital input levels (Pins 5, 18, 21)*				
V_{IL}	Input voltage Low	0		0.8	V
V_{IH}	Input voltage High	2.0		5.5	V
	Input current				
$-I_5$	at $V_5 = 0V$; $V_{13} = GND$	15		70	μA
I_{18}	at $V_{18} = 5V$; $V_{13} = GND$	15		70	μA
	Input leakage current				
$-I_{21}$	at $V_{21} = 0V$; $V_{13} = GND$	25		120	μA
I_{LI}	Input leakage current (except Pins 5, 18, 21)			10	μA
	Analog Input levels (Pin 1)				
	at $V_{REFL} = 2.5V$; $V_{REFH} = 5.1V$				
$V_{IN\ P\text{-}P}$	Input voltage amplitude (peak-to-peak value)		2.6		V
V_{IN}	Input voltage (underflow)		2.5		V
V_{IN}	Input voltage (overflow)		5.1		V
$V_I - V_{REFL}$	Offset input voltage (underflow)		10		mV
$V_I - V_{REFH}$	Offset input voltage (overflow)		−10		mV
$C_{1,\ 2}$	Input capacitance			60	pF
Outputs					
	Digital voltage outputs (Pins 6 to 11 and 15 to 17)				
	Output voltage Low				
V_{OL}	at $I_O = 2mA$	0		−0.4	V
	Output voltage High				
V_{OH}	at $-I_O = 0.5mA$	2.4		5.5	V

*When Pin 5 is Low, binary coding is selected.
When Pin 5 is High, two's complement is selected.
If Pins 5, 18 and 21 are open-circuit, Pins 5, 21 are High and Pin 18 is Low.
For output coding, see Table 1; for mode selection, see Table 2.

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AC ELECTRICAL CHARACTERISTICS

$V_{DD} = V_{3,\ 12,\ 23-13} = 4.5$ to 5.5V; $V_{DD} = V_{24-2} = 9.5$ to 10.5V; $V_{REFL} = 2.5V$; $V_{REFH} = 5.1V$; $f_{CLK} = 22MHz$; $C_{BB} = 100nF$; $T_A = 0$ to $+70°C$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Timing (see also Figure 1)					
	Clock input (Pin 14)				
f_{CLK}	clock frequency	1	25	22	MHz
t_{LOW}	clock cycle time Low	20			ns
t_{HIGH}	clock cycle time High	20			ns
	Input rise and fall times[1]				
t_R	rise time			3	ns
t_F	fall time			3	ns
	Analog input[1]				
BW	Bandwidth (−3 dB)	11	20		MHz
dG	Differential gain				
	at $f_I = \leqslant 4.5MHz$[2]		±3	±5	%
dp	Differential phase				
	at $f_I = \leqslant 4.5MHz$[2]		±1	±2.5	deg
P_E	Phase error				
	at $f_I = \leqslant 4.5MHz$[3]		10	±12	deg
S/N	Signal-to-noise ratio (non-harmonic noise)		−40	−36	dB
	Peak error			3	LSB
	Harmonics (full-scale)				
f_0	Fundamental			0	dB
$f_{2,\ 3}$	2nd and 3rd harmonics		−31	−28	dB
f_{4-7}	4th +5th +6th +7th harmonics		−39	−35	dB
	Digital outputs[2, 4]				
t_{HOLD}	Output hold time	6			ns
t_D	Output delay time $C_L = 15pF$			38	ns
	Output delay time $C_L = 50pF$			48	ns
t_{CY}	Internal delay		3		clocks
t_{DT}	3-State delay time (see Figure 2)			25	ns
C_{OL}	Capacitive output load	0		15	pF
	Transfer function				
	Non-linearity at $f_I = 1.1kHz$				
INL	integral		±¼	±½	LSB
DNL	differential		±⅓	±½	LSB

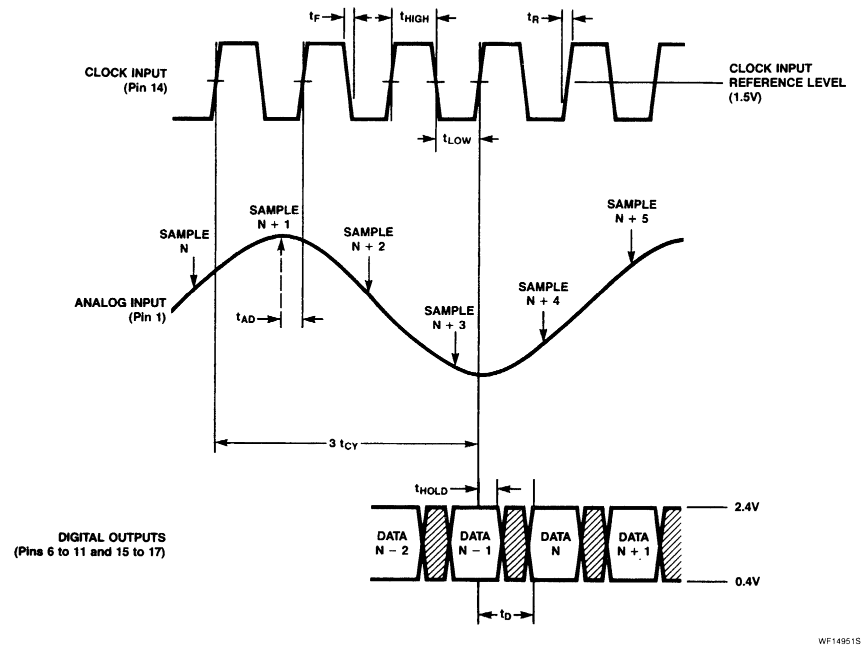
NOTES:
1. Clock input rise and fall times are at the maximum clock frequency (10% and 90% levels).
2. Low frequency sine wave (peak-to-peak value of the analog input voltage at $V_{IN} = 1.8V$) amplitude modulated with a sine wave voltage ($V_{IN} = 0.7V$) at $f_I = 5MHz$.
3. Sine wave voltage with increasing amplitude at $f_I = 5MHz$ (minimum amplitude $V_{IN} = 0.25V$; maximum amplitude $V_{IN} = 2.5V$).
4. The timing values of the digital output Pins 6 to 11 and 15 to 17 are measured with the clock input reference level at 1.5V.

Table 1. Output Coding (V_{REFL} = 2.5V; V_{REFH} = 5.1V)

STEP	$V_{1,2}$ (Typ)	UNFL	OVFL	BINARY Bit 6 – Bit 0	TWO's COMPLEMENT Bit 6 – Bit 0
Underflow	< 2.51	1	0	0 0 0 0 0 0 0	1 0 0 0 0 0 0
0	2.51	0	0	0 0 0 0 0 0 0	1 0 0 0 0 0 0
1	2.53	0	0	0 0 0 0 0 0 1	1 0 0 0 0 0 1
•	•	•	•	• • • • • • •	• • • • • • •
•	•	•	•	• • • • • • •	• • • • • • •
•	•	•	•	• • • • • • •	• • • • • • •
126	5.03	0	0	1 1 1 1 1 1 0	0 1 1 1 1 1 0
127	5.05	0	0	1 1 1 1 1 1 1	0 1 1 1 1 1 1
Overflow	≥ 5.07	0	1	1 1 1 1 1 1 1	0 1 1 1 1 1 1

Table 2. Mode Selection

CE1	CE2	BIT 0 to BIT 6	UNFL, OVFL
X	0	High-impedance	High-impedance
0	1	Active	Active
1	1	High-impedance	Active



NOTE:
There is a delay of 3 clock cycles between sampling of an analog input signal and the corresponding digital output.

Figure 1. Timing Diagram

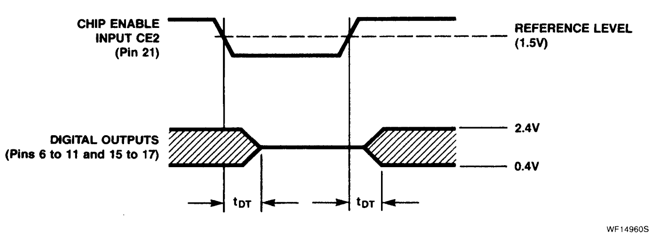


Figure 2. Timing Diagram for 3-State Delay

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APPLICATION INFORMATION

The minimum and maximum values provided in the data sheet are guaranteed over the whole voltage and temperature range. This note gives additional information to the data sheet where the typical values indicate the behavior under nominal conditions; V_{DD5} = 5V, V_{DD10} = 10V, T_A = 22°C.

SYMBOL	PARAMETER	TYP	UNIT
I_{DD5}	Supply current (Pins 3, 12, 23)	51	mA
I_{DD10}	Supply current (Pin 24)	11	mA
f_{CLK}	Maximum clock frequency	25	MHz
B	Bandwidth (−3dB)	20	MHz
P_D	Total power dissipation	365	mW
	Peak error (non-harmonic noise)	1.5	LSB
	Suppression of harmonics sum of:		
$f_{2,3}$	$f_{2nd} + f_{3rd}$	31	dB
f_{4-7}	$f_{4th} + f_{5th} + f_{6th} + f_{7th}$	39	dB
	Non-linearity		
INL	integral	± 1/4	LSB
DNL	differential	± 1/3	LSB
D_G	Differential gain	±3	%
D_P	Differential phase	±1	%
P_e	Large-signal phase error	10	deg
S/N	Signal-to-noise ratio (non-harmonic noise)	−40	dB

NOTE:
1. Typical values are measured on sample base.

Application Recommendation

- Spikes at the 10V supply input have to be avoided (e.g., overshoots during switching). Even a spike duration of less than 1μs can destroy the device.

Test Philosophy

Figure 3 is a block diagram showing analog-to-digital testing with a phase-locked signal source. The signal generator provides a 5MHz sine wave for the device under test (except for the linearity test). The 22MHz clock input is provided by the clock generator.

The phase relationship between signal and clock generator is shifted by 100ps each signal period to provide an effective clock rate of 10GHz for analysis.

Most calculations are carried out in the spectral domain using Fast Fourier Transformation (FFT) and the inverse FFT to return to time domain.

The successive processing completes the specific measurement (Figures 4, 5, 6, and 7).

The non-linearities of the converter, integral (INL) and differential (DNL), are measured using a low frequency ramp signal. Within a general uncertain range of conversion between two steps, the output signal of the converter randomly switches.

After low-pass filtering, the different step width is used for calculating the line of least squares to obtain integral non-linearity.

To calculate differential non-linearity, a counter is used to count the frequency of each step. A histogram is calculated from the counter result to provide the basis for further computation (Figure 6).

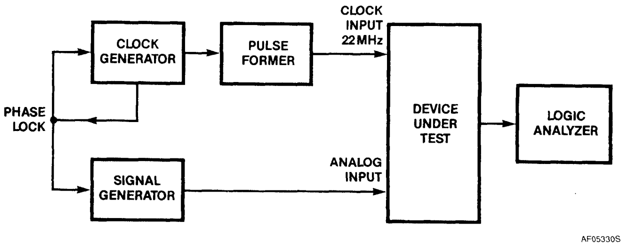
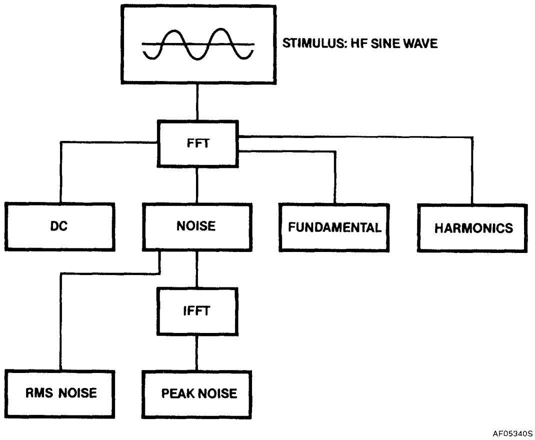


Figure 3. Analog-to-Digital Converter Testing with Phase-Locked Signal Source



Where: FFT = Fast Fourier Transformation.
IFFT = Inverse Fast Fourier Transformation.

Figure 4. Sine Wave Test; Non-Harmonic Noise and Peak Error

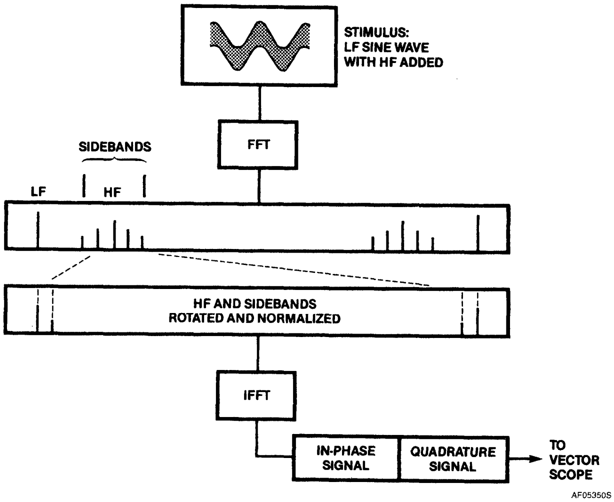
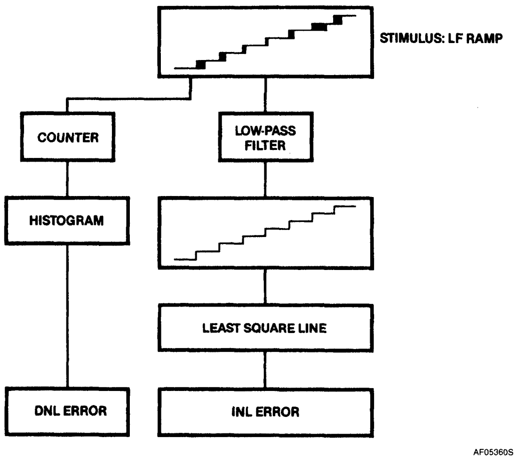


Figure 5. Differential Gain and Phase



Where: INL = Integral Non-Linearity.
DNL= Differential Non-Linearity.

Figure 6. Low Frequency Ramp Test; Linearity

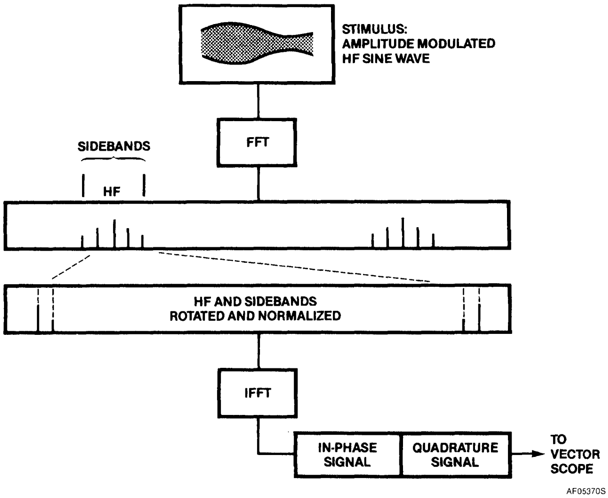


Figure 7. Large-Signal Phase Error

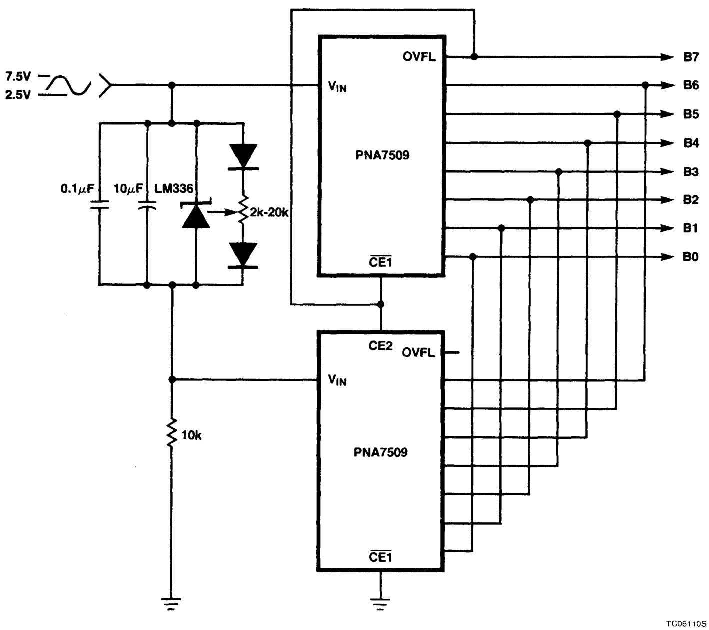


Figure 8. 8-Bit Resolution

Signetics

AN108
An Amplifying, Level-Shifting Interface for the PNA7509 Video A/D Converter A/D Converter

Linear Products

Application Note

The NE5539 is well-suited for use as a level-shifting amplifier at the input of the PNA7509 video speed analog-to-digital converter. Designing this circuit is straightforward and relatively simple.

The first step is to determine the gain that is required. Since the PNA7509 requires a maximum input of 5.0 V_{DC} and a minimum input of 2.5V_{DC} the required amplifier gain is

$$A_V = \frac{5.0 - 2.5}{V_{MAX} - V_{MIN}} = \frac{2.5}{V_{MAX} - V_{MIN}}.$$

where V_{MAX} is the maximum level of the amplifier input signal, and V_{MIN} is the minimum level of the amplifier input signal.

This gain must be greater than unity as the gain of a non-inverting amplifier such as this is

$$A_V = 1 + (R_F/R_I).$$

The ratio of R_F to R_I is then

$$R_F/R_I = A_V - 1.$$

The task is now to select R_F and R_I. These resistors should be low enough to swamp out the effects of any stray capacitance. If R_I is arbitrarily chosen, R_F is found to be

$$R_F = \frac{1.5\ R_I}{V_{MAX} - V_{MIN}}.$$

The required offset voltage, V_O, is then found to be

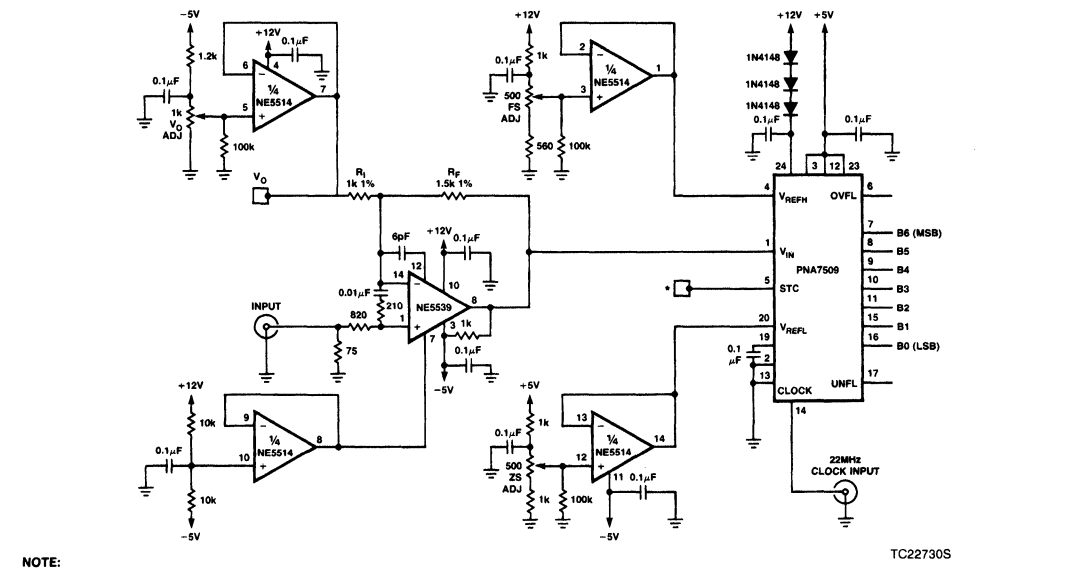
$$V_O = V_{MAX} - [(5 - V_{MAX})\ (R_I/R_F)].$$

Because the NE5539 input cannot be driven closer to its negative supply than about 4.7V, that negative supply must be −4.7V or more negative in order to accommodate an input signal whose minimum potential is 0V. The NE5539 output must never come any closer to the supply rail than about 5.5V, and the maximum output required to drive the PNA7509 is 5V, so the positive supply must be at least 5 + 5.5V, or 10.5V. If we use standard power supply potentials of +12V and −5V, this would satisfy these requirements, except we must insure that the negative supply is at least as negative as −4.7V. Tests have been conducted that indicate satisfactory operation with the positive supply between 10.5V and 13.5V, and the negative supply between −4.7V and −5.7V. Furthermore, because the NE5539 is sensitive to unbalance in the supplies, it is necessary to insure that its Pin 7 potential is close to halfway between the positive and the negative supply. Two resistors and an op amp driving Pin 7 nicely provide this balance. Another op amp is used to set the offset voltage.

The three diodes are used to drop the 12V supply to 10V for the PNA7509. If available and desired, a separate 10V supply could be used without the diodes.

Other components are shown for the convenience of the user. The potentiometer at Pin 5 of the NE5514 is used to adjust V_O. The potentiometer at Pin 12 of the NE5514 sets the voltage at the low end of the PNA7509 reference ladder, so is a zero-scale adjustment. The potentiometer at Pin 3 of the NE5514 sets the high end voltage on the PNA7509 reference ladder and is, effectively, a full-scale adjustment. It is also possible to use a signal divider at the NE5539 input for full-scale adjustment. R_F can also be made variable to provide full-scale adjustment. Care should be exercised, however, when introducing potentiometers into feedback loops or into high-frequency signal paths.

The NE5514 was chosen for its low input offset voltage temperature coefficient.



NOTE:
*Pin 5 should be grounded for binary output, or tied to a logic high for two's complement output.

Signetics

Linear Products

NE5150/5151/5152
Triple 4-Bit RGB D/A Converter With and Without Memory

Product Specification

DESCRIPTION

The NE5150/5151/5152 are triple 4-bit DACs intended for use in graphic display systems. They are a high performance — yet cost effective — means of interfacing digital memory and a CRT. The NE5150/5152 are single integrated circuit chips containing special input buffers, an ECL static RAM, high-speed latches, and three 4-bit DACs. The input buffers are user-selectable as either ECL or TTL compatible for the NE5150. The NE5152 is similar to the NE5150, but is TTL compatible only, and operates off of a single +5V supply. The RAM is organized as 16 × 12, so that 16 "color words" can be down-loaded from the pixel memory into the chip memory. Each 12-bit word represents 4 bits of red, 4 bits of green and 4 bits of blue information. This system gives 4096 possible colors. The RAM is fast enough to completely reload during the horizontal retrace time. The latches resynchronize the digital data to the DACs to prevent glitches. The DACs include all the composite video functions to make the output waveforms meet RS-170 and RS-343 standards, and produce $1V_{P-P}$ into 75Ω. The composite functions (reference white, bright, blank, and sync) are latched to prevent screen-edge distortions generally found on "video DACs." External components are kept to an absolute minimum (bypass capacitors only as needed) by including all reference generation circuitry and termination resistors on-chip, by building in high-frequency PSRR (eliminating separate V_{EE}s and costly power supplies and filtering), and by using a single-ended clock. The guaranteed maximum operating frequency for the NE5150/5152 is 110MHz over the commercial termperature range. The devices are housed in a standard 24-pin package and consume less than 1W of power.

The NE5151 is a simplified version of the NE5150, including all functions except the memory. Maximum operating frequency is 150MHz.

FEATURES

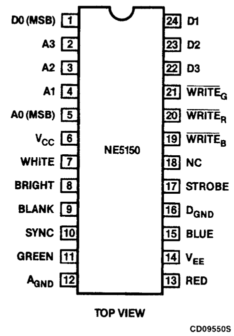
- **Single-chip**
- **On-board ECL static RAM**
- **4096 colors**
- **ECL and TTL compatible**
- **110MHz update rate (NE5150, 5152)**
- **150MHz update rate (NE5151)**
- **Low power and cost**
- **Drives 75Ω cable directly**
- **Internal reference**
- **40dB PSRR**
- **No external components necessary**

APPLICATIONS

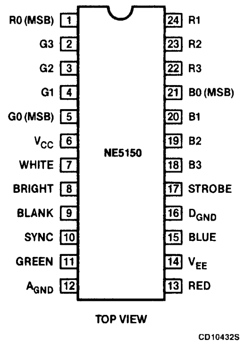
- **Bit-mapped graphics**
- **Super high-speed DAC**
- **Home computers**
- **Raster-scan displays**

PIN CONFIGURATIONS

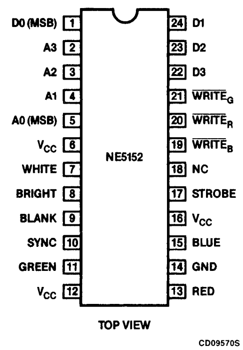
NE5150 F Package



NE5151 F Package



NE5152 D[1] and F Packages



NOTE:
1. Available in large SO pakage only.

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
24-Pin Ceramic DIP	0°C to +70°C	NE5150F
24-Pin Ceramic DIP	0°C to +70°C	NE5151F
24-Pin Ceramic DIP	0°C to +70°C	NE5152F
24-Pin Plastic SOL	0°C to +70°C	NE5152D

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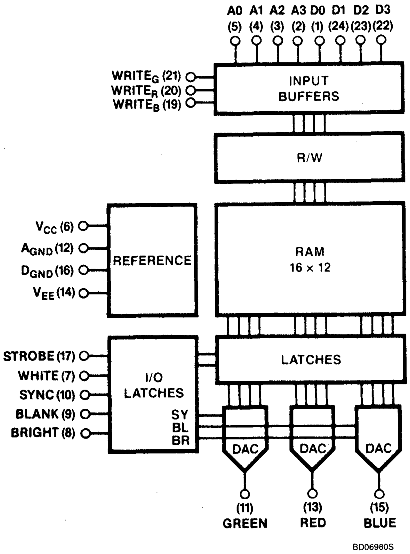
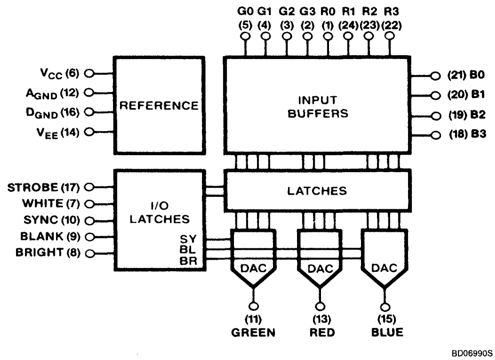
BLOCK DIAGRAMS

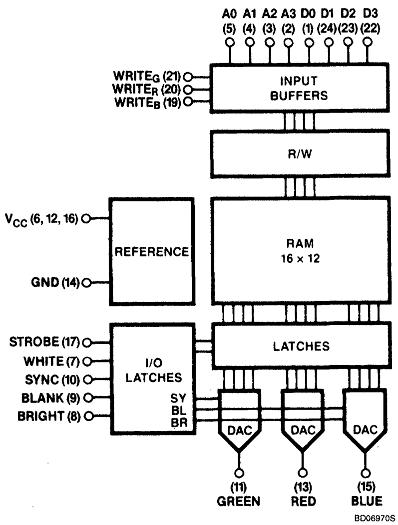
NE5150



NE5151



NE5152



Triple 4-Bit RGB D/A Converter With and Without Memory

NE5150/5151/5152

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
	Temperature range		
T_A	Operating	0 to +70	°C
T_{STG}	Storage	-65 to +150	°C
V_{CC}	Power supply	7.0	V
V_{EE}		-7.0	V
	Logic levels		
	TTL-high	5.5	V
	TTL-low	-0.5	V
	ECL-high	0.0	V
	ECL-low	0 to V_{EE}	V

DC ELECTRICAL CHARACTERISTICS

V_{CC} = +5V (TTL), 0V (ECL), V_{EE} = -5V, 0°C < T_A < +70°C, for NE5150/5151; V_{CC} = +5V (TTL), GND = 0V for NE5152, unless otherwise noted.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
	Resolution	4			bits
	Monotonicity	4			bits
NL	Non-linearity		± 1/16	± 1/2	LSB
DNL	Differential non-linearity		± 1/8	± 1	LSB
	Offset error (25°C) [1111] (BRT = 1)		- 1/5	± 1	LSB
	Gain error (25°C) [0000] (BRT = 1)		± 1/2	± 1	LSB
V_{CC}	Positive power supply (TTL mode) (NE5150)	4.5	5.0	5.5	V
	(TTL mode) (NE5151)	4.75	5.0	5.5	V
	(ECL mode)	-0.1	0.0	0.1	V
V_{EE}	Negative power supply (TTL or ECL mode) (NE5150/5151)	-4.75	-5.0	-5.5	V
I_{CC}	Positive supply current (NE5150/5151)		15	25	mA
	(NE5152)		175	210	mA
I_{EE}	Negative supply current (NE5150)		175	210	mA
	(NE5151)		145	175	mA
	Analog voltage range (ZS to FS)		603		mV
	Gain tracking (any two channels)			± 1/4	LSB
LSB	Least significant bit		40.2		mV
EWH	Enhanced white level (25°C)[2]		0		mV
BS	Bright shift (25°C)(0 to 1)		71.4		mV
EBL	Enhanced blanking level (25°C)[2]		-674		mV
ESY	Enhanced sync level (25°C)[2]		-960		mV
R_O	Output resistance (25°C)	67.5	75.0	82.5	Ω
V_{IH}	TTL logic input high	2.0			V
V_{IL}	TTL logic input low			0.8	V
I_{IH}	TTL logic high input current (V_{IN} = 2.4V)			20	μA
I_{IL}	TTL logic low input current (V_{IN} = 0.4V)			-1.6	mA
V_{IH}	ECL logic input high	-1.045			V
V_{IL}	ECL logic input low			-1.48	V
I_{IH}	ECL logic high input current (V_{IN} = -0.8V)			-1.0	mA
I_{IL}	ECL logic low input current (V_{IN} = -1.8V)			-1.0	mA

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Triple 4-Bit RGB D/A Converter With and Without Memory

NE5150/5151/5152

TEMPERATURE CHARACTERISTICS V_{CC} = +5V (TTL), 0V (ECL), V_{EE} = −5V, 0°C < T_A < +70°C, for NE5150/5151; V_{CC} = +5V (TTL), GND = 0V for NE5152, unless otherwise noted.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
	Offset TC[1]		±50	±100	ppm/°C
	Gain TC[1]		±70	±200	ppm/°C
	Gain Tracking TC (any two channels)		±20	±50	ppm/°C
	Enhanced white level TC[1]		±50	±100	ppm/°C
	Bright shift TC		±70	±200	ppm/°C
	Enhanced blanking level TC		±100	±300	ppm/°C
	Enhanced sync level TC		±100	±300	ppm/°C
	Output resistance TC		+1000	+2000	ppm/°C

NOTES:
1. Normalized to full-scale (603mV).
2. With respect to [1111] (BRT = 1).

AC ELECTRICAL CHARACTERISTICS V_{CC} = +5V (TTL), 0V (ECL), V_{EE} = −5V, 0°C < T_A < +70°C, for NE5150/5151; V_{CC} = +5V (TTL), GND = 0V for NE5152, unless otherwise noted.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
f_{MAX}	Maximum operating frequency (NE5150/5152)	110			MHz
t_{WAS}	Write address setup (NE5150/5152)	0			ns
t_{WAH}	Write address hold (NE5150/5152)	0			ns
t_{WDS}	Write data setup (NE5150/5152)	4			ns
t_{WDH}	Write data hold (NE5150/5152)	2			ns
t_{WEW}	Write enable pulse width (NE5150/5152)	3			ns
t_{RCS}	Read composite[1] setup (NE5150/5152)	3			ns
t_{RCH}	Read composite[1] hold (NE5150/5152)	2			ns
t_{RAS}	Read address setup (NE5150/5152)	3			ns
t_{RAH}	Read address hold (NE5150/5152)	2			ns
t_{RSW}	Read strobe pulse width (NE5150/5152)	3			ns
t_{RDD}	Read DAC delay (NE5150/5152)		8		ns
f_{MAX}	Maximum operating frequency (NE5151)	150			MHz
t_{CS}	Composite[1] setup (NE5151)	3			ns
t_{CH}	Composite[1] hold (NE5151)	2			ns
t_{DS}	Data-bits setup (NE5151)	1			ns
t_{DH}	Data-bits hold (NE5151)	5			ns
t_{SW}	Strobe pulse width (NE5151)	3			ns
t_{DD}	DAC delay (NE5151)		8		ns
t_R	DAC rise time (10 – 90%)		3		ns
t_S	DAC full-scale settling time[2]		10		ns
C_{OUT}	Output capacitance (each DAC)		10		pF
SR	Slew rate		200		V/µs

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
GE	Glitch energy			30	pV-s
PSRR[3]	Power supply rejection ratio (to red, green or blue outputs)				
	V_{EE} at 1kHz		43		dB
	V_{EE} at 10MHz		28		dB
	V_{EE} at 50MHz		14		dB
	V_{CC} at 1kHz		80		dB
	V_{CC} at 10MHz		50		dB
	V_{CC} at 50MHz		36		dB

NOTES:
1. Composite implies any of the WHITE, BRIGHT, BLANK or SYNC signals.
2. Setting to ± ½ LSB, measured from STROBE 50% point (rising edge). This time includes the delay throught the strobe input buffer and latch.
3. Listed PSRR is for the NE5150/51. The NE5152 PSRR specs are identical to the V_{EE} numbers in the table.

NE5150 PIN DESCRIPTION

Write enable inputs use negative-true logic while all other inputs are positive-true. All inputs operate synchronously with the positive edge-triggered strobe input. When V_{CC} is taken high (5V), all inputs are TTL compatible. When V_{CC} is grounded, all inputs are ECL compatible. All DACs are complementary, so that all ones is the highest absolute voltage and all zeroes is the lowest. All ones is called zero-scale (ZS) and all zeroes is called full-scale (FS). The analog output voltage is approximately 0V (ZS) to −1V (SYNC).

Pins 1, 24, 23, 22: **DATA** bits D0 (MSB) through D3, used to input digital information to the memory during the write phase. During this phase, the data bits are presented to the internal latches (noninverted) and the DACs will output the analog equivalent of the stored word, unless overridden by WHITE, BLANK or SYNC.

Pins 5, 4, 3, 2: **ADDRESS** lines A0 (MSB) through A3, used for selecting a memory address to write to or read from.

Pin 7: **WHITE** command. Presets the latches to all ones [1111] and outputs 0V absolute on all DACs. Can be modified to −71mV absolute when BRIGHT is taken low. Will be overridden by either a BLANK or SYNC command.

Pin 8: **BRIGHT** command. A low input here turns on an additional −71mV (10 IRE unit) switch, shifting all other levels downward. Not overridden by any other input.

Pin 9: **BLANK** command. Presets the latches to all zeroes [0000] and turns on an additional −71mV (10 IRE unit) switch. Absolute output is −671mV. Can be modified another −71mV to −742mV absolute when BRIGHT is taken low. Will override WHITE, and will be overridden by SYNC.

Pin 10: **SYNC** command. Presets the latches to all zeroes [0000] and turns on the BLANK switch. Additionally turns on a −286mV (40 IRE unit) switch in the green channel only. Absolute output is −671mV for the red and blue channels, and −957mV for the green channel. All levels can be shifted −71mV by taking BRIGHT low. Overrides WHITE and BLANK.

Pins 11, 13, 15: **GREEN, RED, BLUE.** Analog outputs with 75Ω internal termination resistors. Can directly drive 75Ω cable and should be terminated at the display end of the line with 75Ω. Output voltage range is approximately 0V to −1V, independent of whether the digital inputs are ECL or TTL compatible. All outputs are simultaneously affected by the WHITE, BLANK or BRIGHT commands. Only the GREEN channel carries SYNC information.

NOTE:
There are 100 IRE units from WHITE to BLANK. One IRE unit is approximately 7.1mV. Full-scale is 90 IRE units and 10 IRE units is ⅑ of full-scale (e.g., BRIGHT function).

Pins 19, 20, 21: $\overline{\text{WRITE}}_B$, $\overline{\text{WRITE}}_R$, $\overline{\text{WRITE}}_G$. Write enable commands for each of the three 16 × 4 memories. When all write commands are high, then the READ operation is selected. This is the normal display mode. To write data into memory, the write enable pin is taken low. Data D0 - D3 will be written into address A0 - A3 of each memory when its corresponding write enable pin goes low.

Pin 17: **STROBE.** The strobe signal is the main system clock and is used for resynchronizing digital signals to the DACs. Preventing data skew eliminates glitches which would otherwise become visible color distortions on a CRT display. The strobe command has no special drive requirements and is TTL or ECL compatible.

Pins 12, 16: $\mathbf{A_{GND}}$, $\mathbf{D_{GND}}$. Both **A**nalog and **D**igital ground carry a maximum of approximately 100mA of DC current. For proper operation, the difference voltage between A_{GND} and D_{GND} should be no greater than 50mV, preferably less.

Pin 14: $\mathbf{V_{EE}}$. The negative power supply is the main chip power source. V_{CC} is only used for TTL input buffers. As is usual, good bypassing techniques should be used. The chip itself has a good deal of power supply rejection — well up into the VHF frequency range — so no elaborate power supply filtering is necessary.

Pin 18: **NC.** This unused pin should be tied high or low.

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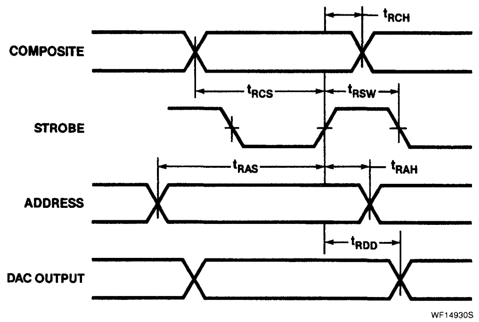
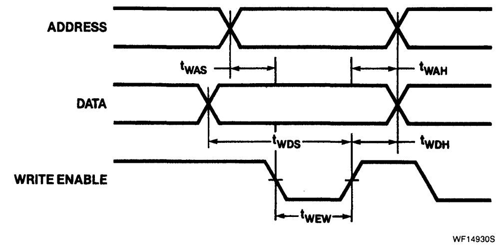
NE5150/5152 TIMING DIAGRAMS



Read Cycle



Write Cycle

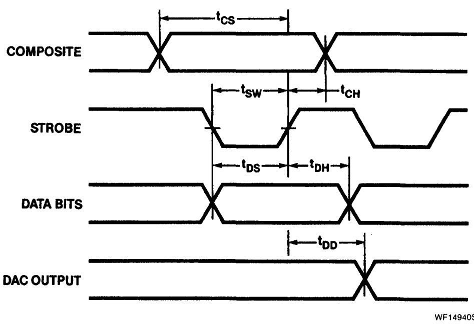
NE5151 PIN DESCRIPTION AND TIMING DIAGRAM

The eleven digital inputs D0 – D3, A0 – A3, $\overline{\text{WRITE}}_{G/R/B}$, and the unused Pin 18 of the NE5150 are replaced in the NE5151 with the three 4-bit DAC digital inputs G0 – G3, R0 – R3, and B0 – B3. All other pin functions (e.g., composite functions, power supplies, strobe, etc.) are identical to the NE5150.

NE5152 PIN DESCRIPTION

The NE5152 is a TTL-compatible-only version of the NE5150, operating off of a single +5V supply. V_{CC} Pins 6, 12 and 16 should be connected to +5V and Pin 14 to 0V. DAC output is referenced to V_{CC}.

NE5151 TIMING DIAGRAM

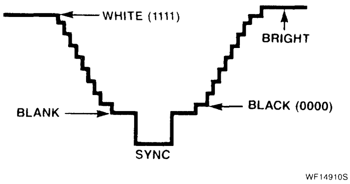


NE5150/NE5151/NE5152 LOGIC TABLE

SYNC	BLANK	WHITE	BRIGHT	DATA	ADDRESS	OUTPUT[3]	CONDITION
1	X	X	0	X	X	-1031mV	SYNC[1]
1	X	X	1	X	X	-960mV	Enhanced SYNC[1]
0	1	X	0	X	X	-746mV	BLANK
0	1	X	1	X	X	-674mV	Enhanced BLANK
0	0	1	0	X	X	-71mV	WHITE
0	0	1	1	X	X	0mV	Enhanced WHITE
0	0	0	0	[0000]	Note 2	-674mV	BLACK (FS)
0	0	0	1	[0000]	Note 2	-603mV	Enhanced BLACK (EFS)
0	0	0	0	[1111]	Note 2	-71mV	WHITE (ZS)
0	0	0	1	[1111]	Note 2	0mV	Enhanced WHITE (EZS)

NOTES:
1. Green channel output only. RED and BLUE will output BLANK or Enhanced BLANK under these conditions.
2. For the NE5150/5152 the DATA column represents the memory data accessed by the specific address. For the NE5151, the DATA is the direct digital inputs.
3. Note output voltages in Logic Table are referenced to V_{CC} for the NE5152 only.

COMPOSITE VIDEO WAVEFORM



Signetics

AN1081
NE5150/51/52 Family of Video Digital-to-Analog Converters

Application Note

Linear Products

Author: Michael J. Sedayao

INTRODUCTION

Raster-scan systems and bit-mapped graphics are here to stay. For a computer to be of use, it needs an interactive means of communicating with the user. So for every computer, whether it is a 10MFLOP (millions of floating-point operations per second) supercomputer or a home computer for playing video games, some type of terminal or graphics display device is needed. Not long ago, inputs to the computer were made using stacks of Hollerith cards pushed into a hopper and then read into the computer. Results would then come from a printer. The hardcopy results were exactly what they looked like: final judgment from the computer. In order to respond, it was back to the punch-card machine. Needless to say, debugging programs became quite laborious. This problem led to the interactive display, allowing the user to enter information and see the results immediately. A new age in computing had arrived.

The areas of word processing, on-screen circuit simulation, and computer graphics developed with great rapidity. As technology improved, so did the ability to make larger displays having more colors and better resolution. As software developed, so did techniques such as windowing, the use of icons, and the ability to use graphic input devices such as mouses, light pens, and joysticks. Three-dimensional images and photographic quality reproduction soon followed.

Of the different technologies, how did raster scanning predominate over other forms? What differentiates bit-mapped graphics systems from character or vector-map systems? In the following sections it will become clear how technology and economics drove the market and, consequently, product development.

Displays: Raster, Vector Refresh, Storage Tube

A *raster* is technically a display of horizontal lines. How the display is created is what makes it unique. An electron beam generated by a CRT (Cathode Ray Tube) and containing video information, starts at the top left of the screen and traces a path to the right part of the screen (see Figure 1). It makes a slight angle as it travels across. The gun is then turned off as the beam rapidly returns to the left. It then repeats this zig-zag path until it reaches the bottom of the screen. The gun is again turned off as the beam travels back to the top of the screen. This entire process is repeated from 30 to 60 times per second so flicker is decreased (motion pictures or film typically display 24 images per second). What the electron beam has done is *scanned* its information onto the screen. This process is called *raster scanning.*

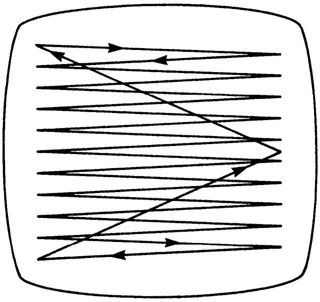


Figure 1. Raster-Scanned Display

All television sets display information in this manner. For television sets in the United States, the screen is redrawn 30 times per second. Additionally, the screen is *interlaced*, meaning that every other line is scanned and then the lines in between are scanned. This gives the illusion that the image is continuous. Since the television sets have 525 lines, 262.5 lines are scanned first (the odd field) and then the other 262.5 (the even field) are scanned. To visualize this, consider a 21-line system (see Figure 2). Scanning occurs at the above-mentioned 30Hz rate which is also known as the *frame rate*. Two fields (odd and even) equal one frame. Scanning 525 lines 30 times a second equals 15,750 horizontal lines scanned in a second. This is called the *horizontal scan frequency*. These are standard in the U.S., coming under the standard known as NTSC (National Television Standards Committee). In Europe, television has 625 lines and has a frame rate of 25Hz, or half the power line frequency, 50Hz.

Vector refresh displays, or stroke-writers, work on the principle that one line is the base unit of information. Each line then corresponds to a vector. Instead of scanning continuously, information is drawn line-by-line, hence the name stroke-writer. These systems off-load the refreshing tasks to special hardware, making the system slightly more cost-effective. Still, during the 1960's making them proved too expensive for everyday applications.

In 1971, Tektronix introduced the Direct View Storage Tube (DVST) for displaying and interfacing graphic data. It was based on oscilloscope techniques, storing information in a special, long-persistence phosphor which coats the inside of the screen. The display resolution is limited only by the phosphor grain size and the quality of the deflection circuitry. Although inexpensive, these devices were fine for oscilloscopes in the lab, but too cumbersome for fully interactive work. When the screen would redraw itself after the entry of new information, the sudden disappearance and reappearance was almost like looking at the light of a camera flashbulb. Another problem with the storage refresh screen was that when new information entered, it would write directly over the existing information. Only upon refreshing the screen would the new information be clear and readable. In many cases, the annoyance did not justify the low cost.

Bit-Mapped Graphics

In a bit-mapped graphics system, the screen is divided into individual elements called pixels, short for picture elements. When they say "bit-mapped", each pixel corresponds to a bit, or, in most cases, an address or memory location. This is what differentiates television from bit-mapped computer displays. Although both systems use raster scanning techniques, the information transmitted on television is continuous — a stream of analog information between horizontal sync pulses (the pulses used to denote the beginning and end of a horizontal line) — whereas in bit-mapped systems, each line is divided into discrete elements (the aforementioned pixels). The approximation of analog images would then be determined by the pixel density or screen resolution. As an example, Figure 3 shows a line approximated by a finite number of pixels.

The lines seem to staircase rather than flow because of the enlargement of the pixels. The effect is known in some computer graphics circles as "jaggies", short for jagged edges.

So, with more pixels, better resolution is possible. This is not without a price, though. Since each pixel corresponds to a memory

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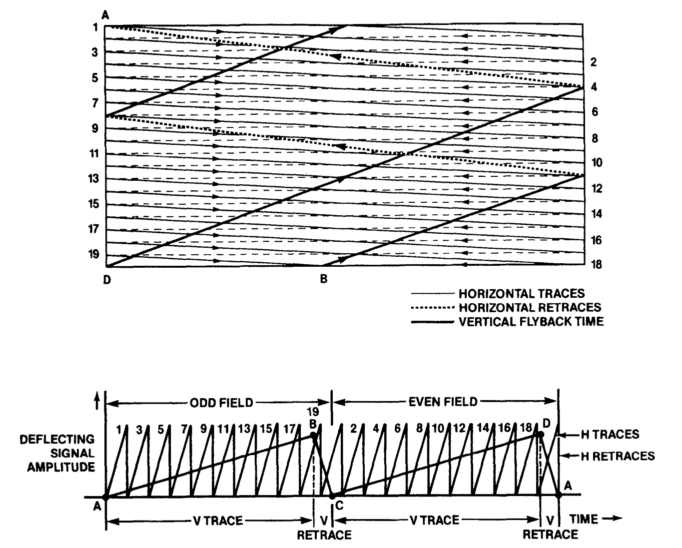

NOTES:
A sample scanning pattern for 21 interlaced lines per frame and 10½ lines per field. The corresponding H and V sawtooth deflection waveforms are shown below pattern. Starting at point A, the scanning motion continues through B, C, and D, and back to A again.

Figure 2. Interlaced Raster for 21-Line System

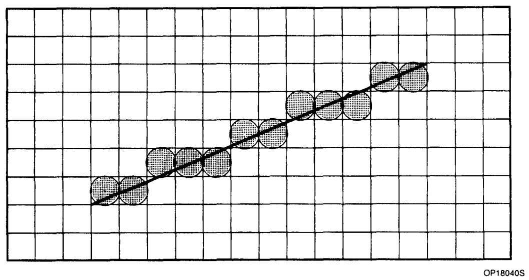

Figure 3. Ideal Line and Its Discrete Pixel Representation

location, memory cost rises dramatically as pixel resolution increases. Drawing speed must also increase since more pixels have to be drawn to maintain the $\geqslant$30Hz frame rate needed to avoid flicker. Clearly then, the increase in bit-mapped graphics systems can be tied to the continuing price reductions in memory, specifically, the Dynamic Random Access Memory (DRAM). Fortunately, as the price has dropped, the memory size has not stood still. The last 14 years have seen size increases from 4k to 16k, 16k to 64k, 64k to 256k, and now, 256k to 1M bits of memory. One might expect to see DRAMs on the order of 4Mb within 2 to 3 years. Additionally, the continuing development of video RAMs cannot be ignored.

A bit-mapped system might be described in one of three ways. First, assume the display is monochrome and that each pixel can be represented by a certain number, for instance, 4 bits of information. This means that there are $2^4 = 16$ possible values of shading. Each bit of information can be represented by a "plane" of information. The plane would correspond to the area that was mapped by the pixels, namely the drawing area or display. Imagine an 8 $\times$ 8 pixel display. This means that there are 4 bit-planes and each pixel would have to pierce all four planes to give the proper information (see Figure 4). This is a fairly quick way to draw the screen since the data goes directly from the bit-map to the DAC (Digital/Analog Converter; DAC is singular here since the display is monochrome).

A direct conversion system for color is the second step. This is just an upgrade of the first case. Instead of 4 bit-planes, there are 12: three sets of the 4 planes for the three primary colors red, green, and blue. The advantage here is that there are now $2^{12} = 4096$ different colors, but the corresponding disadvantage is that the memory requirement has tripled. For more bit resolution per pixel, the associated memory demands increase by 3 times the pixel size times n, where n is the additional bit of resolution per pixel.

The third type of bit-map system uses a color look-up table (CLUT) as the driver for the display. The operation is straightforward. As the controller scans the bit-map each time it comes upon a pixel, it retrieves the bits which are then decoded into an address. This address is a pointer to the look-up table where sixteen 12-bit words (colors) are stored (see Figure 5). Once selected, that word is then sent to the color DACs and, from there, to the screen. The idea is similar to that of having cache memory in a computer, a fast memory used when the information in the memory is frequently accessed. Note that the bit-planes grow as n for 2^n additional colors while memory grows for 3n in the direct conversion case, a definite savings in memory.

The limitation in this case is that only 16 colors can be displayed at a time. In some systems, however, the CLUT is fast enough to be reloaded during the horizontal retrace time (CLUT size is sometimes referred to as the maximum number of colors that can be displayed on one horizontal line). This is especially important if the image is to simulate a smooth motion such as the rotation of a merry-go-round or the movement of an object with mirrored surfaces. In most cases, 16 colors is sufficient for any single display. 64 colors (6 bit-planes) is extremely good. 256 colors (8 bit-planes) is definitely a luxury.

It's clear that the memory speed and memory density, which are direct functions of the color and screen resolution, play a large part in the feasibility of a bit-mapped system. For that reason, the enormous gains and technological advancements in the field of memory design have made bit-mapped raster-scan graphic systems the best choice for both cost and performance.

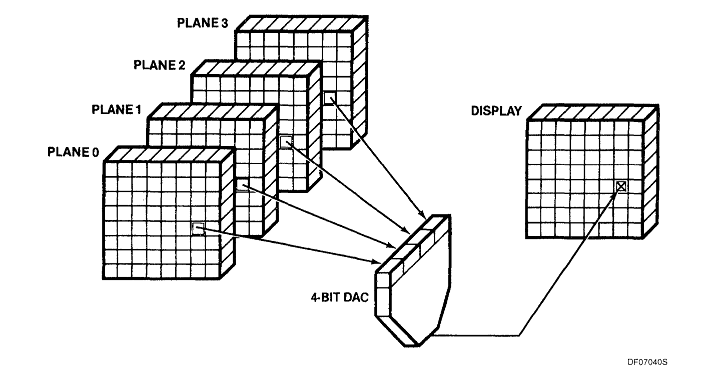

Figure 4. Monochrome Bit-Map With Direct Conversion to Display

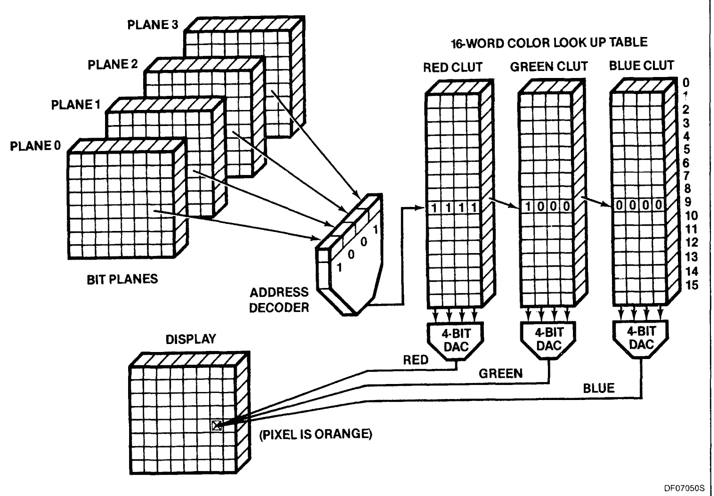

Figure 5. Color Bit-Map With 16-Word Color Look-Up Table

ISSUES FOR GRAPHIC DISPLAY SYSTEMS

Making the DAC Fit the Application

When designing graphic display systems, there are many decisions to be made in specifying the hardware and software needed for a system. What kind of speed is necessary in a given application? What kind of resolution will the users of the system require? Is color needed or will monochrome be adequate? If color, how many colors? Will images be viewed in two or three dimensions? How much memory is needed? How should the microprocessor/CRT controller/video DAC/frame buffer be matched with the rest of the system? What's the best type of software for a particular application? and on and on...

These questions could form the subject of an entire book and so will not be discussed in detail. This section will, however, discuss the few issues needed in the selection of the proper video DAC for a system.

Display Resolution vs Bit Resolution

When the quality of a display terminal is being evaluated, one primary consideration is the kind of resolution it has. There are two different types of resolution: display resolution, which is determined by the monitor and cannot be changed by the design; and bit resolution, which is dependent on the design of the video DAC used.

Display resolution determines how many pixels can be projected onto the monitor at any one time. (Actually, only one pixel is displayed on the screen at a time, in rapid succession). Table 1 shows commonly-used screen resolutions corresponding to various applications.

However, since each pixel must correspond to a memory element, the more pixels per screen the faster the DAC and video RAM must be in order to write the information to the screen fast enough to avoid flicker. This imposes speed requirements that have to be satisfied.

The other type of resolution, bit resolution, depends on the type of DAC used. The number of bits converted also determines the size of the color palette which is the number of possible colors that can be displayed. This should not be confused with the number of colors displayed at once (see Section on Color Look-Up Tables). Assuming that the monitor is an RGB-type, the bit resolution, n, must be multiplied by 3 to get the total bit resolution, 3n. Taking this number as 2^{3n} gives the size of our color palette. Table 2 shows common bit sizes for video DACs with their corresponding palettes.

It should be clear that, if imaging is the goal, a higher bit resolution gives access to the assorted tones and mixtures of colors that make color graphics as realistic as possible. The major problems associated with higher-resolution DACs are that they are larger and more complex than lower-resolution DACs and tend to take longer for their signals to settle. This has a direct effect on selection of the proper DAC for a particular system because of the DAC's bandwidth and because of the need to weigh advantages and disadvantages of higher and lower bit resolutions.

For a low-end personal computer graphics screen on which the pixels can actually be seen at arm's length, it makes little sense to have a bit resolution that shows flesh tones because the benefit of the large palette is defeated by a screen that shows jagged edges. On the other hand, having a high screen resolution with a limited amount of colors does not defeat the purpose in the same way — if many colors aren't needed.

Integrated circuit layout, for instance, may not require thousands of colors — only enough to distinguish 12 – 15 masks; but sharply defined edges and zooming ability are needed to examine the circuit. The need for this user could be a bit resolution of 2 (64 colors) and a display resolution of 1024 $\times$ 1280.

For all this talk of colors and bit resolution, monochrome should not be totally ignored. After all, people got along fine with black and white TV for years before color came along. For applications such as word processing or

Table 1. Display Resolutions With Applications

DISPLAY RESOLUTION	APPLICATION
250 × 500	Low-end personal computers (home computers)
640 × 480	High-end personal computers
600 × 800	Next-generation personal computers
768 × 576	Next-generation personal computers
1024 × 800	Workstations
1024 × 1024	High-end workstations
1024 × 1280	High-end graphics terminals (CAE/CAD)
1024 × 1500	High-end graphics terminals (3-D Imaging)
1500 × 1500	High-end graphics terminals
2048 × 2048	High-end graphics terminals (photo quality)

Table 2. Bit Resolution With Palette Size

BITS/DAC	RGB	PALETTE SIZE	APPLICATION
1	3	8	Digital RGB, "rainbow colors"
2	6	64	Some home and personal computers
4	12	4096	Color workstations, CAD/CAE
6	18	262,144	High-end CAD/CAE, medical imaging
8	24	16,777,216	Photographic quality reproduction

Table 3. Display Resolution With Minimum DAC Speed

DISPLAY RESOLUTION	# PIXELS	MINIMUM DAC SPEED
250 × 500	125,000	10MHz
640 × 480	308,000	25MHz
600 × 800	480,000	38MHz
768 × 576	443,000	35MHz
1024 × 800	820,000	65MHz
1024 × 1024	1,049,000	85MHz
1024 × 1280	1,311,000	105MHz
1024 × 1500	1,536,000	125MHz
1500 × 1500	2,250,000	180MHz
2048 × 2048	4,195,000	330MHz

circuit design, monochrome is fine. To achieve different shades of black and white, no chrominance operation is necessary. All of the bit resolution can be done with one DAC to operate on the luminance, or brightness signal. In this case, the brightness resolution can be said to be 2^n. Remember, the decision to go with color or monochrome does not rest upon the designers of the graphics board. A monitor is either color or monochrome to begin with. Adding a color video DAC won't change that.

DAC Speed

The DAC's update rate or bandwidth is a crucial consideration in choosing a DAC if the type of monitor has already been specified.

For raster-scan systems, a few calculations can be made to determine the minimum speed required for the DAC.

First of all, assume that the screen needs to be refreshed at 60Hz to avoid flicker. To account for the electron beam going back to the top to start the next frame, assume that the retrace time is 30% of the drawing time. Multiply the frame rate by 1.3 to account for the retrace. Thus, the minimum bandwidth for the DAC would be determined by the following formula:

Speed (Hz) = 1.3 (retrace factor) × # pixels × 60Hz (frame rate)

For the screen resolutions noted earlier, a new table can be generated for the minimum DAC speed required (see Figure 8).

For the 60Hz frame rate, the screen is probably not interlaced. Interlacing the screen at 30Hz would give the same effect because interlacing gives the *illusion* that the screen is being refreshed at a faster rate. The DAC would only have to operate at a quarter of the speed of the 60Hz non-interlaced rate because only half of the lines are being drawn at a speed that's half the 60Hz frame rate. This is how scanning operates under the NTSC television standard. The FCC says that televisions can't refresh the screen faster than 30Hz, so interlacing was developed to get around it. There are no such restrictions in graphics monitors. In fact, there are monitors that have horizontal scan rates as much as 4 times faster (65kHz) than that for television (15.75kHz).

Color Look-Up Tables: Yes or No?

As mentioned in the Bit-Mapped Graphics section, graphic systems may have direct conversion from a bit-map or they can use color look-up tables (CLUTs). It should be pointed out that one is not necessarily faster than the other. Speed depends primarily on the system. A fast CLUT is of no use if the external frame buffer can't load a new set of colors into the CLUT during the retrace time (horizontal or vertical). A video DAC without the CLUT may be faster since it can bypass the memory accesses needed for the CLUT, but, as seen in the Bit-Mapped Graphics section, the extra cost of the bit-planes (1 million additional bits for a 1024 × 1024 display) may be excessive, and accessing the additional planes may produce some design problems.

If a CLUT is needed, the size of the CLUT should also be a major consideration. Each bit-plane added requires 2^n more memory cells. Constraints on die-size and power requirements become apparent. Also, one must ask whether one needs 16, 32, 64, 128, or 256 colors on every line. This depends on the color resolution desired *for the entire screen*. An easy way to determine the system needs is to picture the most common scene that would be displayed. The general rule is that the more complex and three-dimensional the images that are required, the more variations and shading are needed to truly represent them. Conversely, if the image is simple and two-dimensional, fewer colors would be needed. An example of the former would be geological formations. For the latter, consider the colors of flags of the world's nations. Almost all of them can be displayed with a CLUT of 16 colors. Remember, this refers to the number of colors needed at any one time.

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No flag has more than 16 colors. The range of colors available for display after CLUT refresh depends on the color resolution or the number of data bits for each pixel.

Gamma Correction

A problem encountered in both television systems and in display monitors in general is the gamma effect. This is due to the nonlinear relationship between light output and the signal voltage applied to a cathode-ray tube. Although it would be desirable to have the luminous output of the phosphors on the display to vary directly with the changes in the signal applied to it, they usually do not. Each monitor has its own characteristic, but the international convention is to assume that the fractional value of the luminous output can be approximated by raising the percentage of display signal input to the 2.2 power. For example, a 60% of full-scale input signal will result in 33% of the full-scale luminous output ($0.6^{2.2} = 0.33$).

In Figure 6, the monitor does not respond linearly for a linear input signal. Adding a gamma correction circuit can take care of this problem.

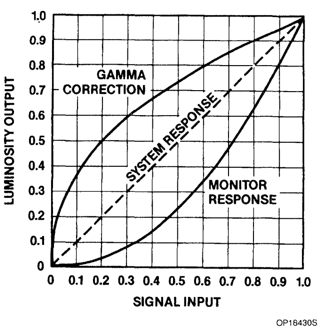

Figure 6. Monitor and System Response With Gamma Correction

In the television industry, correction for this non-linearity takes place at the camera as the image is recorded. The camera takes the 2.2 root of its full-scale fractional value. This cancels the gamma effect and produces a linear system response.

In graphics systems for which the image is generated from digital information, DACs convert the digital information into a voltage that drives the guns of the CRT. Basically, the systems designer has three choices:

1. Correct for gamma in the software. This can be done by using the 2.2 power/root compensation to pixel values before they are stored into the frame buffer. This could be an expensive addition to the software and might slow the overall system because of the added computation time.
2. Apply analog gamma correction in the hardware. The correction factor could be done with additional circuitry to the output of the DAC before it drives the monitor. As mentioned before, this presents an additional hardware overhead. This is not done, however, without some risks. Since every monitor has individual characteristics, the resulting correction would not look the same on every monitor.
3. Ignore the whole subject and accept the non-linearity of the luminous output as a characteristic of the system. Since most graphics applications are for the generation of images for specific problems and not for the lifelike reproduction of scenes (although it would be desirable), a gamma correction mechanism is unnecessary.

This last approach seems to be the most prevalent solution since few, if any, DACs contain gamma correction circuitry. When graphics software designers select their colors, they do so for the best visual performance. This fine-tuning for colors and shading is really software gamma correction because they can select the digital information needed for colors and intensity and see the results from the other side of the monitor.

CIRCUIT FEATURES AND OPERATION

This section covers the basic features and operation of the NE5150/51/52. The first two sections briefly discuss RS-170 and RS-343A, the standards for color and monochrome video systems. The next section covers the composite video signal (CVS) that is specified in the two previous standards.

RS-343A and RS-170

RS-170, the Electrical Performance Standards for Monochrome Television Studio Facilities, and RS-343A, the Electrical Performance Standards for High Resolution Monochrome Closed Circuit Television Cameras, were issued in November 1957 and September 1969, respectively, by the EIA (Electronic Industries Association). The specifications outlined in RS-343A determine the voltage levels required for the part.

Composite Video Signal

Shown in Figure 7 is a section of a composite video signal. With the exception of the BRIGHT function, the levels and tolerances are specified by RS-343A.

Sync, Blank, and Setup

The sync signal is situated 286mV (40 IRE) below the blanking level which lies 714mV (100 IRE) below the reference white level (next section). The sync signal synchronizes the monitor horizontal and vertical scanning. This, and the rest of the composite video signal, is not to be confused with the composite sync signal which is often used for a combined horizontal *and* vertical sync signal.

The blank level lies just below the reference black level, separated by an amount known as the setup. The difference between reference white and the blanking level is defined as 100 IRE. Applying the blanking level voltage to the monitor input will reduce the CRT electron beam current so that there will be no visible trace of the electron gun on the phosphor.

For television, the setup is defined as the *ratio* between the reference white and the reference black level measured from the blanking level. It is usually expressed as a percentage. Basically, it's the difference between the reference black level and the composite blanking level. RS-343A has set the limits of the setup as 7.5 ±5 IRE. Any value between 2.5 to 12.5% of the blanked picture signal can be designated as the setup (2.5 – 12.5 IRE or 17.85 – 89.25mV). Since the full-scale range of the video signal represents 100 IRE, a percentage of the signal is synonymous with its IRE value. For the NE5150, the setup is 71mV or 10 IRE.

Reference Black and White

Reference black and white correspond to the signal levels for a maximum limit of black and white peaks. White corresponds to having all color guns on and black to having all guns off. The gray scale, which refers to the rest of the color values and contains a majority of the signal information, is defined by the amplitude between reference white and reference black. Since the reference white to blanking level is fixed at 100 IRE, the reference black level is determined by the setup. Since the setup can be between 2.5 and 12.5 IRE, the gray scale range must reflect those tolerances and so has a range of 92.5 ±5 IRE (660.5mV ±35.7mV).

To allow for a BRIGHT function, the NE5150/51/52 family of video DACs were designed for a full-scale range (blank to reference white) of 675mV (about 94 IRE) and a gray-scale range of 643mV (about 90 IRE). Using the BRIGHT function adds 71mV (10 IRE) to the reference white value.

For instance, in a 12-bit system like the NE5150/51/52, using 4 bits/DAC would enable us to resolve the gray scale range into 16 parts. For the NE5150, that would be about 40.1mV (5.6 IRE) = 1 LSB. For 6 bits, 64 parts could be resolved, and for 8 bits, 256 parts.

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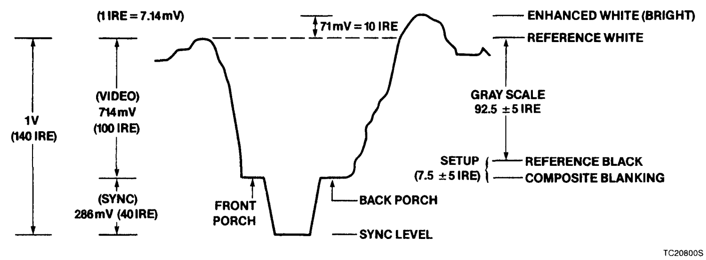

Figure 7. RS-343A Video and Sync Levels

NE5150/NE5151/NE5152 LOGIC TABLE

SYNC	BLANK	WHITE	BRIGHT	DATA	ADDRESS	OUTPUT[3]	CONDITION
1	X	X	0	X	X	−1031mV	SYNC[1]
1	X	X	1	X	X	−960mV	Enhanced SYNC[1]
0	1	X	0	X	X	−746mV	BLANK
0	1	X	1	X	X	−674mV	Enhanced BLANK
0	0	1	0	X	X	−71mV	WHITE
0	0	1	1	X	X	0mV	Enhanced WHITE
0	0	0	0	[0000]	Note 2	−674mV	BLACK (FS)
0	0	0	1	[0000]	Note 2	−603mV	Enhanced BLACK (EFS)
0	0	0	0	[1111]	Note 2	−71mV	WHITE (ZS)
0	0	0	1	[1111]	Note 2	0mV	Enhanced WHITE (EZS)

NOTES:
1. Green channel output only. RED and BLUE will output BLANK or ENHANCED BLANK (BRIGHT ON) under these conditions.
2. For the NE5150/5152, the DATA column represents the memory data accessed by the specific address. For the NE5151, the DATA is the direct digital inputs.
3. Note output voltages in Logic Table are referenced to V_{CC} for the NE5152 only.

Device Description and Operation

Corresponding to the RS-343A requirements outlined in the previous section, the logic table indicates the output voltages given the digital inputs shown. Although the output voltages for the DACs are shown, the user should also know what is happening to the circuit and how the priority given to each function influences the output. [All ones (1111) is called zero-scale (ZS) and all zeroes (0000) is called full-scale (FS).]

The BLANK command presets all the latches to all zeroes (0000) and sends the output to its blanking level of 100 ± 5 IRE below reference white (−71mV) or about −746mV. When BRIGHT is on (a '1'), the output is raised 10 IRE (71mV or 1/9th of full-scale) to −674mV. BLANK overrides WHITE and is overridden by SYNC.

The WHITE command presets the latches to all ones (1111) and outputs −71mV to all DACs. When the BRIGHT command is on, this value is raised to 0V. WHITE will be overridden by both SYNC and BLANK.

The SYNC command presets all of the latches to zeroes and turns on the BLANK switch. In addition, it turns on a 40 IRE switch (drops voltage 286mV) in the GREEN channel only. So the GREEN channel sits at 140 IRE down and the RED and BLUE channels will be 100 IRE below ground.

The BRIGHT command turns off one current switch within the circuit and adds 10 IRE (71mV) to the output levels of all three guns. This comes in handy if using a cursor (optional blinking) to brighten other parts of the screen. This switch cannot be overridden by any other switch.

Referring to the pinouts of both the NE5150/52 and the NE5151 (see Figure 8), there are additional considerations.

The $\overline{WRITE_G}$, $\overline{WRITE_R}$, and $\overline{WRITE_B}$ pins are the write enable pins for each of the 16 × 4 memories in the CLUT. When these pins are pulled High, the memory is then in the READ mode. This is the normal mode of operation. To write to the memory, **one** of the pins must be pulled Low. The data on D0 – D3 will then be written to the memory location A0 – A3 of the corresponding WRITE pin.

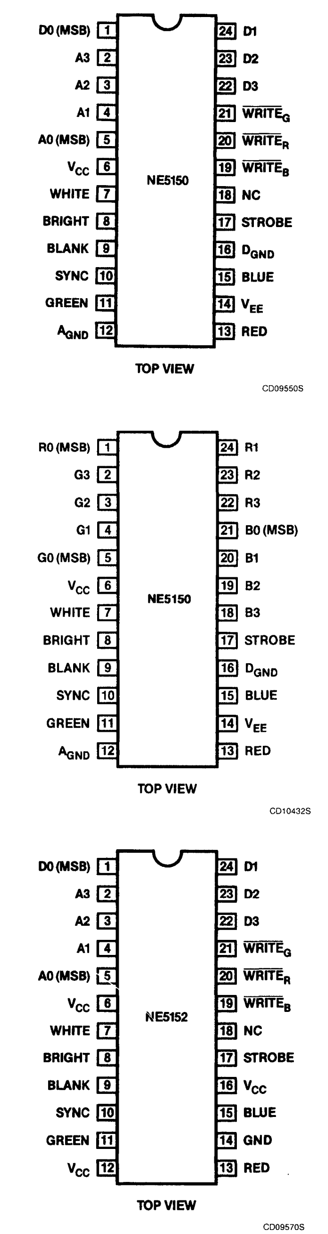

Figure 8. Pinouts of NE5150/52 and NE5151

STROBE is the main system clock and synchronizes all digital operations on the DAC.

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The strobe is ECL and TTL compatible and demands no special drive requirements. The positive edge of STROBE clocks the latches.

The GREEN, RED, and BLUE pins are the analog outputs of the DACs. The DACs are voltage output and need no external components (75Ω resistors are on-chip). The output voltage range is approximately 0 to −1V and is independent of the input logic (either TTL or ECL).

The DATA and ADDRESS bits are designated so that D0 and A0 represent the most significant data and address bits (MSB), respectively. Similarly, D3 and A3 correspond to the least significant data and address bits (LSB). Since the NE5151 has no CLUT, there is no need for the address pins (4) or the write enable pins (3). Adding the NC (no connection) pin (1) gives the eight additional input pins for two 4-bit DACs. The original data bus now carries the logic for the RED gun.

Analog and digital ground (A_{GND} and D_{GND}) should always be connected together in any configuration and should not have more than 50mV of potential between them to insure proper operation of the device. The next section will cover connection of V_{CC} and V_{EE}, in addition to A_{GND} and D_{GND}, on different system configurations.

Using Different Logic and Supply Voltages

Different users have different needs. Some have access to dual supplies, other only to single-ended supplies. Signal logic may be TTL or ECL. In any case or configuration, the NE5150/51/52 family can be used. The following configurations cover most cases.

Explanation of the configurations are as follows:

A. Case A shows a basic ECL configuration for the NE5150 and NE5151. The signal voltage is basic ECL with a −1.3V threshold and is powered from ground and −5V (or −5.2V). Since the TTL buffers are no longer needed, V_{CC} is tied to analog and digital ground (A_{GND} and D_{GND}), excluding the buffers from the circuit.

B. In some cases, people use ECL logic but run it off a single supply, +5V and ground. In this case, operation is the same except that the supplies are shifted up 5V. In this new ECL mode, the threshold −1.3V is moved up by 5V to +3.7V. ECL operation is not available for the NE5152.

C. For TTL operation in the NE5150 and NE5151, dual supplies are normally needed. If available, standard TTL-level signals with a +1.4V threshold (between a logic '1' Low of 2.0V and a logic '0' High of 0.8V) can be connected directly.

D. In some situations, a dual supply is not available. Single-supply TTL operation is made possible by making similar connections and by pulling up the inputs of each pin with a 10kΩ resistor connected to V_{CC} = +5V. This is necessary because the threshold is now 3.7V.

E. Case (D) necessitated the construction of the NE5152, which has only one mode using a single 5V supply and accepts TTL inputs. A_{GND} and D_{GND} become V_{CCA} and V_{CCD} and are tied to V_{CC}.

In some cases, a single supply is used and the internal ECL mode has been shifted up to the positive supply; the output voltage will be swinging from 0V to −1V, but, referenced from V_{CC} = +5V, it will swing from 5V to 4V. If the monitor accepts only positive sync pulses or video information, DC-offsetting the outputs or AC-coupling them with 1μF capacitors would make the signal acceptable to the monitor.

Since the outputs have internal 75Ω resistors, the monitor should have a 75Ω resistor to ground in order to doubly-terminate the cable and to prevent reflections.

Unused Inputs

For ECL mode (NE5150), any unused inputs, regardless of desired permanent stage, should be tied to a fixed-level output of an unused gate.

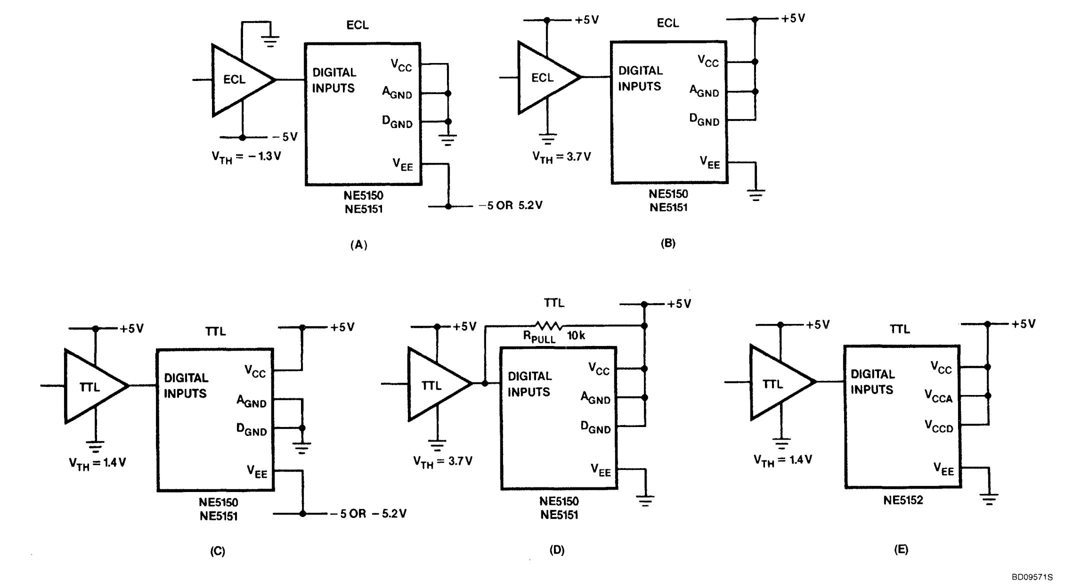

Figure 9. Video DAC Modes of Operation

BLOCK DIAGRAMS

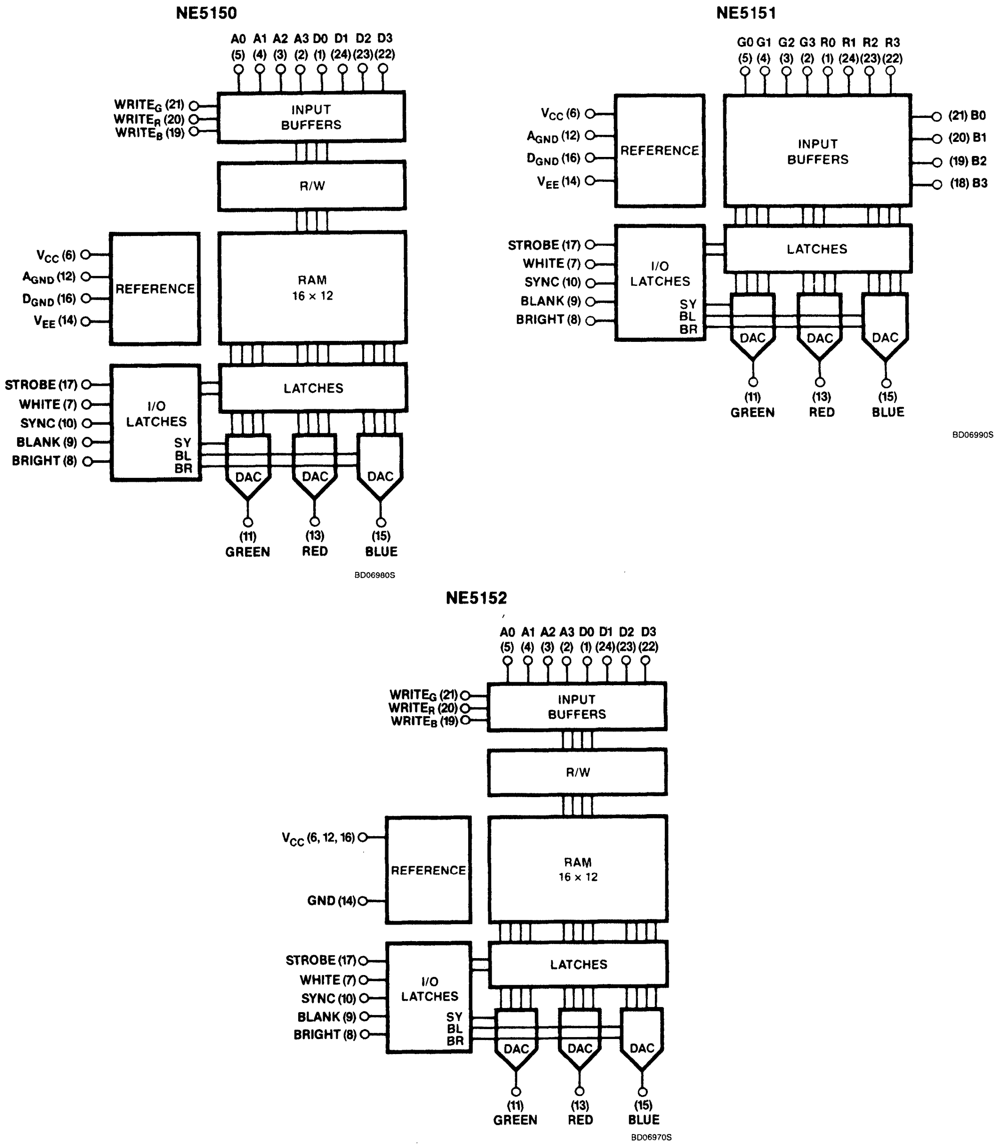

Figure 10. NE5150/51/52 Block Diagrams

NE5150/51/52 Family of Video Digital-to-Analog Converters AN1081

Circuit Description

As can be seen from the block diagrams in Figure 13, the only difference between the NE5150/52 and the NE5151 is the lack of a color look-up table on the NE5151. Bypassing the CLUT with its assorted address decoding, sense amplifiers, and read/write logic enables it to not only use 200mW less power, but also to increase its update rate to 150MHz.

The NE5151 is basically the same die as the NE5150/52, with the exception of a metal mask option that permits it to bypass all of the circuitry associated with the CLUT. It is also bonded differently to enable all 12 bits to be loaded into the DAC at any one time instead of being multiplexed 4 bits at a time to the NE5150/52 CLUT.

DAC Reference

The need for separate references for the DACs resulted from the problems associated with glitching and crosstalk between the DACs. When one DAC maintains a constant value through pixel updates, while another undergoes major transitions such as the 1111 to 0000 on/off switching of currents through the DAC, feedthrough can be expected if all 3 DACs derive their reference voltage from the same source. Having separate references solves this problem. It also isolates the DACs from each other and the other parts of the circuit.

The reasons for choosing the DAC shown in Figure 12 are its simplicity, the bandgap's insensitivity to temperature variations, and its excellent supply rejection (PSRR) through high frequencies. It consists of a PTAT current source supplying a bandgap reference. The output of the bandgap is approximately −1.2V.

To provide the bias for the different current sources on each of the DAC stages, the circuit uses a control amplifier that provides negative feedback to maintain its stability. BIT and its complement drive the differential pair that (along with QS2) makes up one part of the DAC. The bandgap drives the current sources through the control amplifier. If the bias line voltage should rise or fall, the negative feedback in the QS1 and QS3 current path would correct for it.

The control amplifier consists of a transconductance stage driving an emitter-follower. The output of the emitter-follower provides a low-output impedance line that drives QS4. The inclusion of QS4 prevents switching transients from degrading settling time. The control amplifier has a 60MHz unity-gain bandwidth, providing power supply rejection up into the VHF range.

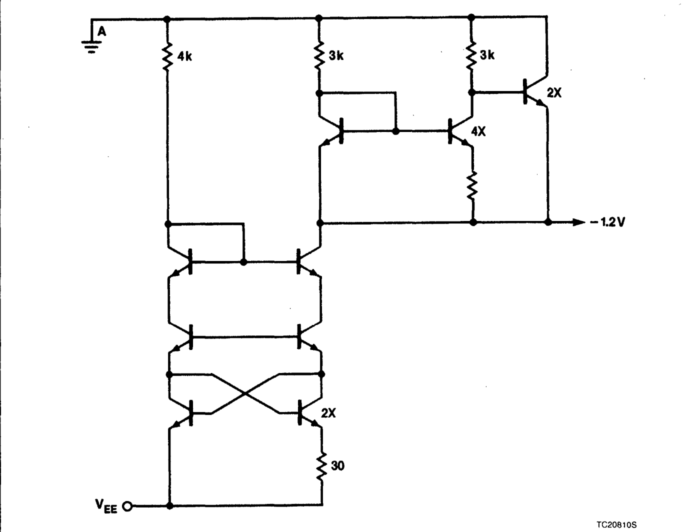

Figure 11. Bandgap Reference for DAC (1 of 3)

Digital-to-Analog Converters

The three DACs consist of differential pairs that are switched on or off depending on the value of the bits. Each of the transistors switches a different amount of current depending on the significance of each bit (see Figure 13). Although only one transistor is shown for each bit, the circuit actually has several transistors in parallel to get the required current. In this case, B3 is the least significant bit since it switches the least amount of current and would produce the smallest voltage drop across the 75Ω load resistor. The reverse is true for B0, the most significant bit, since it draws the most current.

So for all bits low, 0000, all of the current would go through the load resistor, bringing the output voltage to its lowest point. If all three DACs are low, this would correspond to reference BLACK. All bits high, or 1111, shunt current away from the load and leave the output voltage at reference WHITE. Different combinations of bits give 16 values between WHITE and BLACK. One additional 2mA switch is turned on by the input value of BRIGHT, which level-shifts the output by 1/9th the full-scale value, or about 10%. The BLANK and SYNC pins work in a similar manner. Refer to the Logic Table beside Figure 8 for the output voltages for each of these functions.

Some of the problems associated with DACs can be attributed to switching glitches, usually measured in terms of glitch energy. Glitching occurs when digital switching of the transistors causes spikes onto the collectors of the current sources to each of the differential pairs. These current spikes charge the collector-base capacitance, C_{JC}, of the collector transistor, and result in a slower settling time. The asymmetrical turn-on/off behavior of bipolar transistors and mismatched load bit-wiring capacitances also contribute to glitches. This can also be seen as an overshoot of the waveform, a "glitch" on the rising or falling edge of what should look like a square wave. Signals that overshoot the desired analog output level consequently take longer to settle to their final value. The measure of this overshoot is the glitch energy, usually given in pV-sec. The units do not actually work out as units of energy or Joules, which is C-V (Coulomb-Volts), but result from measuring the area of the glitch [Area = Height (V) × Width (psec)].

The NE5150/51/52 resolves this problem by putting the current sources in series with another set of transistors (see Figure 14). The stage below the differential pair is then biased by a low-impedance line which reduces the effect of the current spiking. The biasing for the lower transistor comes from the control amplifier mentioned in the DAC Reference Section.

Video DAC Timing

For the NE5150 and NE5152, the presence of the memory dictates both a READ and a WRITE cycle, whereas the NE5151 needs only one diagram. The explanation of each of the waveforms can be found in the timing glossary. For the guaranteed specifications, the user is referred to the data sheet.

NE5150/52 (With CLUT)

In the NE5150/52 READ cycle, the COMPOSITE signal refers to either the WHITE, BRIGHT, BLANK, or SYNC signals. The read composite hold time, t_{RCH}, is defined from the rising edge of the strobe to the end of the composite pulse. This is the required time the composite signal must remain on the bus for latching. The time between the end of the composite pulse to the next rising edge of the strobe defines the read composite setup time, t_{RCS}. This is the same as the read address setup time, t_{RAS}. The read DAC delay time, t_{RDD}, is the propagation time of the signal through the device clocked from the strobe to the 50% change of the DAC output.

In the WRITE cycle, t_{WAS}, the write address setup time is defined by the start of address to the falling edge of the write enable strobe. At the end of this time, data can be written to the CLUT. Both ADDRESS and DATA must remain latched until they reach the rising edge of the WRITE ENABLE. This defines the WRITE ENABLE pulse width, t_{WEW}. The data should also be latched at the same time as the address. The start of the data (and address) to the end of the write enable pulse is defined as t_{WDS}, or the write data setup time. After the write pulse finishes, an address and data hold time is also specified.

NE5151 (No CLUT)

Since the NE5151 has no memory for the signal to propagate through, it typically has a faster conversion time. As can be seen from the pinouts, the three 4-bit words enter the DAC simultaneously as opposed to the sequential 4-bit loading scheme used in the NE5150/52. With no memory, there's no need for READ or WRITE cycles and so there is only one standard timing diagram. (See Figure 16).

This timing diagram is similar to the READ cycle of the NE5150/52 with the exception that addresses are not clocked to the CLUT; instead, data bits are sent directly to the DACs. In this case, t_{DH} is analogous to the address hold time in the NE5150/52. All other definitions are analogous to the earlier READ case.

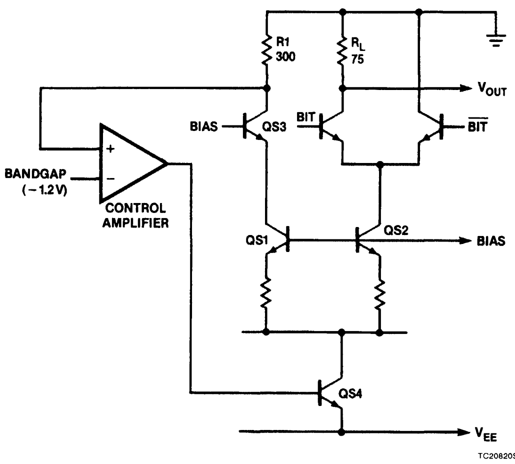

Figure 12. Negative Feedback Referenced to Bandgap

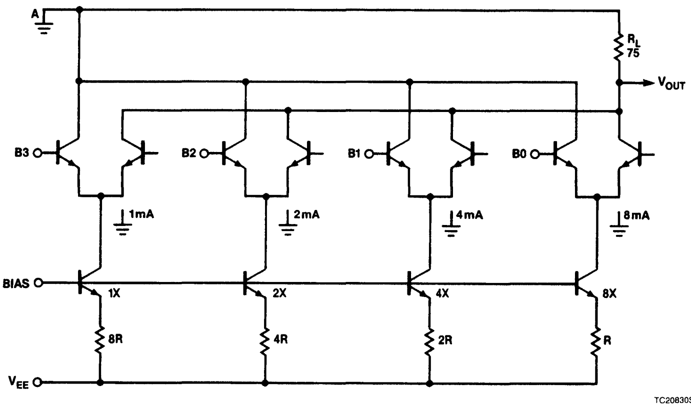

Figure 13. Simplified Schematic of DAC (1 of 3)

WORKSTATION APPLICATION

Introduction

This section describes the design of a color graphics interface for the Modula, Inc. Lilith Workstation. The workstation initially loads 16 colors (it only requires 16) into the NE5150's color look-up table. After the colors are loaded, the workstation then generates addresses to the look-up table. The entire color range (4096) is not required in this application.

The LILITH Workstation

The Lilith Workstation is a 16-bit workstation manufactured by Modula, Inc. It was originally designed by Niklaus Wirth and his students at the Swiss Federal Institute of Technology (ETH). The Lilith is a Modula-2 computing engine. In its original package, the Lilith includes 256kB of memory, a 15MB Winchester disk drive, a floppy disk, a mouse, and an 832 × 640 monochrome graphics tube.

The Signetics Logic Design Group in Orem, Utah, has modified the Lilith by adding 2MB of memory and a high-resolution 1024 × 1024 color monitor. The changes made to the Lilith graphics section comprise the bulk of this application description. Benchmarks of the modified workstation have shown that its performance on applications ranging from matrix multiplications to complete circuit analysis is approximately half as fast as a VAX 11/780 minicomputer. In addition to the circuit simulator used, the Signetics-modified Lilith also supports a layout editor, SLED, that uses about 10,000 lines of Modula-2 source code. More detailed information on the Lilith can be obtained from the manufacturer and from the articles listed in the reference section.

For the purposes of this application, it is sufficient to know that the Lilith contains a 16-bit data bus for interaction with the SCC63484 Advanced CRT Controller and a

14-bit bus that is used to initialize the color look-up table in the NE5150 video DAC. Read/write, I/O lines, CLOCK, data acknowledge, and chip select signals are also sent to the SCC63484 for data and control purposes.

Software Aspects (Pascal and Modula-2)
Modula-2 is a superset of Pascal. Anyone with a working knowledge of Pascal should have no trouble programming a Lilith workstation or in understanding the initialization program outlined in this section. Some noteworthy features about Modula-2 and its influence on the architecture of the Lilith (the M-machine) is the fact that the Lilith instruction set (M-code) has only 256 carefully chosen instructions. This limits any instruction to a 1B length and increases the speed of operation. The Modula-2 language constructs map neatly to M-code. There are no excess instructions to add extra baggage. For additional details, the reader is referred to the August 1984 issue of BYTE magazine that contains several good articles on Modula-2.

Considering each '1' as ON and each '0' as OFF, the binary values for each color can be specified for each of the respective guns. Starting from the top, all guns OFF = BLACK. Similarly, all guns ON corresponds to word 7, WHITE. In the software definition module used to load the values, two constants were declared: black = 0 and white = 15. These correspond to the addresses shown in the table and were predefined because of their frequent use. Single guns completely ON give 1, 2, and 4 — the primary colors RED, GREEN, and BLUE, respectively.

System Hardware

The basic system configuration for the color graphics interface is shown below. The Lilith workstation sends data to the SCC63484 and the NE5150. The information sent to the NE5150 is the data for the CLUT initialization. Control signals are sent to the ACRTC. The ACRTC in turn controls the video DAC. The frame buffer sends and receives data (via an address/data buffer stage) to and from the ACRTC for video DAC addressing. The ACRTC also provides horizontal and vertical sync to the CRT while the video DAC supplies the video information. One stage not shown is the address and data buffering for the frame buffer and the pixel stage. This stage, in addition to assorted logic and timing chips, merely facilitates functionality between the major blocks shown in Figure 21.

The host microprocessor, system memory, and DMA control are local to the workstation and will not be described. The horizontal and vertical deflection sections are local to the CRT and will also be omitted. The rest of this section supplies an overall parts list and then describes each of the graphics blocks in somewhat greater detail. Although the actual pin numbers have been omitted, the functionality of each pin is shown for understanding. For actual pinouts and more detailed information, refer to the appropriate data sheet.

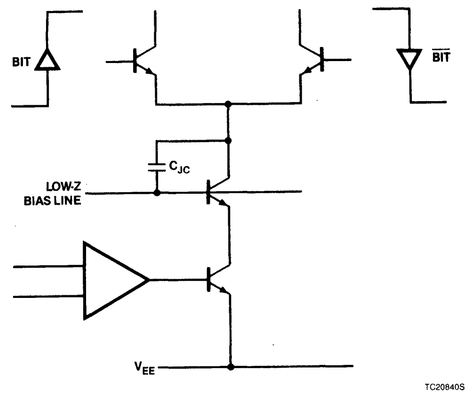

Figure 14. Low-Z Bias Line to Improve Settling Time of DAC

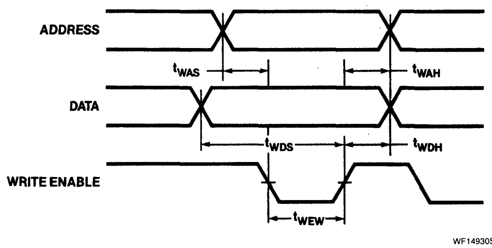

Figure 15. NE5150/52 READ and WRITE Cycle Timing Diagrams

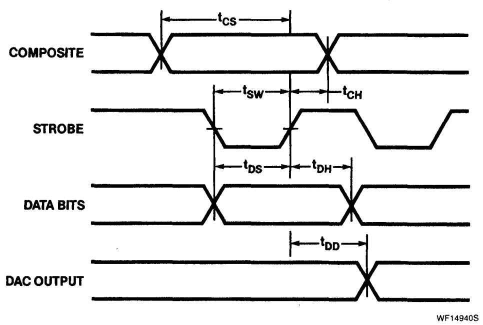

Figure 16. NE5151 Timing Diagram

Parts List

The following parts were used in the design of the color graphics interface (the actual quantity of each part is not listed). The "F" designation stands for Signetics FAST-type logic.

- NE5150 Video DAC
- SCC63484 Advanced CRT Controller
- MB85103-10 64k × 8 Dynamic RAM modules (Fujitsu)
- 7404 Hex Inverter
- 7432 2-Input NAND Gate

- 7474 Dual D-Type Flip-Flop
- 74123 Dual Retriggerable Monostable Multivibrator
- 74138 1-of-8 Decoder/Demultiplexer
- 74F139 Dual 1-of-4 Decoder/Multiplexer
- 74F157 Quad 2-Input Data Selector/ Multiplexer (Non-Inverted)
- 74F161 4-Bit Binary Counter
- 74F164 8-Bit Serial-In/Parallel-Out Shift Register
- 74F166 8-Bit Serial/Parallel-In, Serial-Out Shift Register
- 74F245 Octal Transceiver (3-State)
- 74F373 Octal Transparent Latch
- 7905 5V Voltage Regulator
- M1001 40MHz Crystal (MF Electronics)

PC Board Layout Considerations

Whenever dealing with high-frequency systems, analog or digital, care must be taken with PC board layout in order to insure good, reliable operation. Video DACs are hybrid devices in the sense that they are both analog and digital. They are also run at frequencies well into the RF range. This makes them especially susceptible to RF interference and different types of radiation. Signal traces should be kept as short as possible and 90° turns should be avoided. Power supplies should have adequate decoupling.

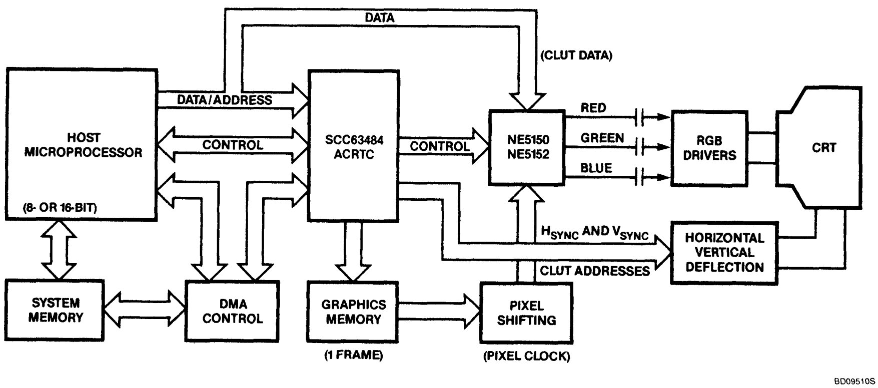

Figure 17. Block Diagram of Color Graphics Interface

More details are provided in the reference section under Reference Number 4, "Getting the Best Performance From Your Video Digital-to-Analog Converter".

Functional Description

The interface is designed to drive a Mitsubishi C-6919 or 6920 19-inch monitor. The monitor has 1024 × 1024 display resolution. Of these, 1024 × 768 pixels are actually drawn, giving us about 790,000 pixels, and, according to our earlier formulas, requiring a DAC with a conversion frequency of about 62MHz. That, however, assumes a non-interlaced display with a frame rate of 60Hz. This application uses a 30Hz interlaced display and so it needs only one-fourth that speed since it is drawing half as many lines at half of the frame rate. The pixel clock is derived from a 40MHz crystal. Other timing signals are also derived from the same crystal.

Table 4. Colors with Corresponding Bit Values

WORD #	COLOR	BLUE	GREEN	RED
0	BLACK	0000	0000	0000
1	RED	0000	0000	1111
2	GREEN	0000	1111	0000
3	YELLOW	0000	1111	1111
4	BLUE	1111	0000	0000
5	VIOLET	1111	0000	1111
6	TURQUOISE	1111	1111	0000
7	WHITE	1111	1111	1111
8	GREY	1010	1010	1010
9	ORANGE	0000	1000	1111
10	AVOCADO	0000	1010	1000
11	LIME	0101	1111	1111
12	NAVY	1111	1000	1000
13	ROUGE	1000	0000	1111
14	LAVENDER	1111	1111	1000
15	PEA	1000	1111	1000

NOTE:
The colors listed are for an application example only. The colors were randomly ordered and their gun and bit values in no way represent the de facto standard values or colors.

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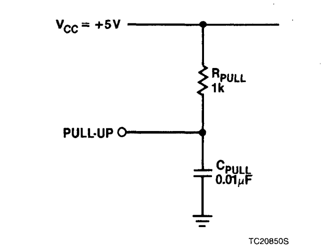

Figure 18. Circuit for Pull-Up to V_{CC}

The interface uses a 512kByte frame buffer that is organized as 64k by 64-bit words. Within each 16-bit block of memory (1 of 4 per word), there are 4 pixels of 4 bits each. Each bit supplies an address to the Color Look-Up Table in the Video DAC. The interface shifts out 64-bits or 16 pixels of information during each display cycle.

In each of the following schematics certain pins have been pulled up to V_{CC}, indicated by an arrow. For each arrow pointing to PULL-UP, the connection goes into the pull-up circuit shown below.

C_{PULL} is used for decoupling any power line ripple. Each point has a similar circuit.

ADVANCED CRT CONTROLLER

The Signetics SCC63484 is a state-of-the-art device ideal for controlling raster-scan-type CRTs. It is a CMOS VLSI system that can control both text and graphics. One of the advantages of this part is its ability to do on-board graphic processing in its Drawing and Display Processor, relieving some of the computational overhead from the Lilith.

Another attractive feature of the part is its flexibility. It has three different operating modes: character only, graphic only, and multiplexed character/graphic mode. In addition, it offers three scanning modes: non-interlace, interlace sync (this application), and interlace sync and video modes. With 2MB of graphic memory and a maximum drawing speed of 2 million pixels/second, it can supply the information to almost any type of high-resolution display (4096 × 4096 pixels maximum).

For additional information on the command set and a full listing of features, please refer to the data sheet and user's manual. This application note will concentrate on only the interconnections relevant to this application.

In this configuration (Figure 19), the SCC63484 Graphics Controller provides the horizontal and vertical sync pulses to the CRT and important timing pulses to the address and data buffers. It supplies timing to the frame buffer, the pixel-shifting stage, and to the frame buffer through direct and logical modifications made to the following system outputs:

1. MRD — Memory Read or the Bus Direction Control Line. This determines the bus direction for the Frame Buffer Data Bus.
2. $\overline{\text{DRAW}}$ — the Drawing/Refresh Cycle pin. This differentiates between drawing cycles and CRT display refresh cycles.
3. $\overline{\text{AS}}$ — Address Strobe. This provides the address strobe for demultiplexing the frame buffer/data bus (MAD0/MAD15).
4. MCYC — Memory Clock. Provides the frame buffer memory access timing. Equal to one-half the frequency of 2CLK signal.
5. $\overline{\text{DISP1}}$ — Display Enable Timing. This is a programmable display enable timing signal used to selectively enable, disable, and blank logical screens.
6. MAD0 - MAD15 — Address and Data Bus. Multiplexed frame buffer address/data bus.
7. MA16, MA17 — Address Bits/Raster Address Outputs. Gives the higher-order address bits for graphic screens and the raster address outputs for character screens. (lower 2 bits of MA16 - MA19).

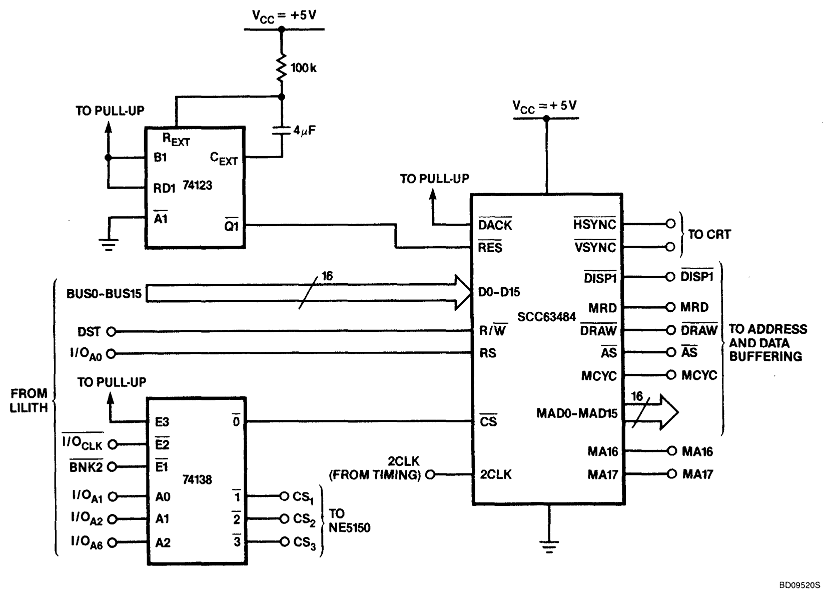

Figure 19. SCC63484 Advanced CRT Controller

The 2CLK signal provides the main clock input to the SCC63484 and is derived from the pixel clock (see System Timing).

The ACRTC also provides horizontal and vertical sync pulses directly to the CRT via the $\overline{HSYNC}$ and $\overline{VSYNC}$ outputs.

In Figure 19, the 16-bit bus of the Lilith is connected directly to the data inputs. The Lilith also provides a write signal (DST) to the R/$\overline{W}$ input. The first I/O line (I/OA0) is connected to the RS (Register Select) input. In addition, there is a high-order I/O bank select, three lower-order address lines, and a negative true I/O clock that, used with the 74138 Decoder, selects one of 4 devices: the ACRTC or 3 areas in the NE5150's color look-up table.

On the ACRTC, a 74123 one-shot produces a reset pulse ($\overline{RES}$) on power-up. The Data Acknowledge pin is not used and is pulled up to V_{CC}.

ADDRESS AND DATA BUFFERING

The address and data buffer stage provides an interface between the SCC63484 and the rest of the circuit. This stage takes the address/data lines MAD0 – MAD15 and separates them into two blocks. The 74F373 latches the upper bank for the addresses; this is the first bank. The second bank consists of 74F245 transceivers in the lower bank for the data.

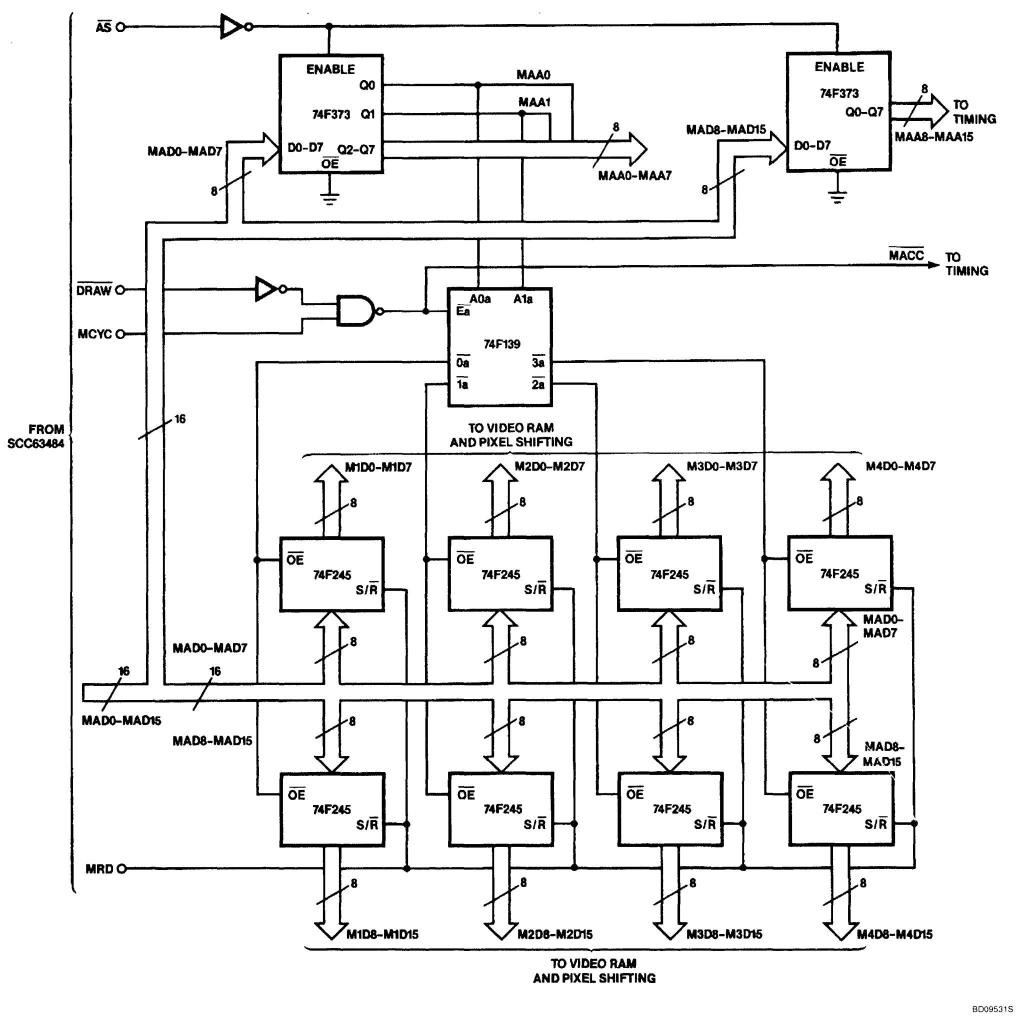

Figure 20. Address and Data Buffering

The 74F373s are used to latch the addresses at the beginning of every memory cycle. The latches are enabled by the $\overline{AS}$ signal coming from the ACRTC. Since the ACRTC is configured to increment its display addresses by four between display cycles, 4 words or 64 bits are shifted out every cycle. For modifying memory cycles, the two lower address lines are used to enable one of four sets of 74F245 transceivers (2 per set). Enabling is performed by the 74F139 Decoder. The signal that clocks the decoder is a combination of MCYC (Memory Cycle) and $\overline{DRAW}$, that results in a new signal, $\overline{MACC}$. This signal is also used in the timing block.

The transceiver outputs are now written into the frame buffer. From there, they will be sent to the pixel-shifting stage and then to the DAC. Each set of four 4-bit pixels in a serial string of displayed pixels is contained in a different block of memory. This is the reason the two lower-order address signals are used to select one of the four banks in the Video RAM (frame buffer).

SYSTEM TIMING

In a system as complicated as a graphics display board, the timing of the various elements grows exceedingly complicated as the number of components grows. It becomes even more apparent when the components are individual systems with their own set of timing considerations. In our case, this means the Lilith, the ACRTC, and the frame buffer.

Figure 21 shows the many elements it takes to generate the timing signals for the system. In the middle of the diagram, there are two 74F164 8-bit serial-in/parallel-out shift registers that count the timing states for the rest of the interface. The Address Strobe ($\overline{AS}$) signal, coming from the ACRTC, starts and ends this timing train. Because of the pulse width of $\overline{AS}$, many states at the end of the train are unusable. The video RAM $\overline{RAS}$ signal (Row Address Strobe) starts at the beginning of State 1, and terminates as $\overline{AS}$ goes Low, activating the register's MR (Master Reset). The precharge requirement of $\overline{RAS}$ is met by the $\overline{AS}$ pulse width.

The 74F157 Multiplexers are connected in such a way that the lower-order addresses are used for the video RAM row addresses (the 157 on top). At the beginning of State 3, the higher-order addresses are presented at the Video RAM address inputs as the column address. At State 5 the $\overline{CAS}$ signal becomes valid. Because of changes in the data hold (WRITE cycle) and data setup (READ cycle) of the ACRTC, the timing edge of $\overline{CAS}$ might have to be changed to insure proper operation.

MRD (Memory Read) along with a combination of MCYC and $\overline{DRAW}$ from the Address and Data Buffer called $\overline{MACC}$, are used with the two lowest-order address lines from the 74F373s (MAA0 and MAA1) to write-select one of the four memory planes (this memory plane runs orthogonal to the bit-planes discussed earlier). Because this signal comes well before the $\overline{CAS}$ signal, this qualifies as an early WRITE cycle, allowing the use of DRAMs with Data-In and Data Out signals connected together.

Using two flip-flops, the output of the lower shift register generates the PE (Parallel Enable) signal for the pixel-shifting stage. Because it is clocked from the fifth point in the shifter, this pulse occurs between States 10 and 11.

The upper left-hand corner of Figure 21 shows the creation of the 2CLK signal derived from the 40MHz pixel clock by using a 74F161 Counter that performs a divide-by-eight operation.

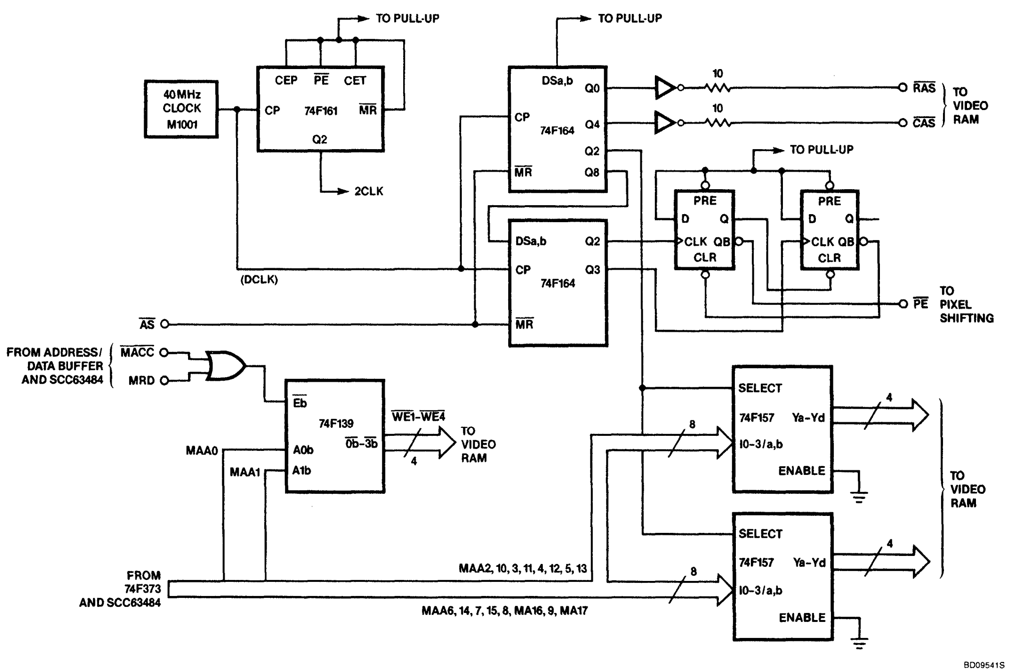

Figure 21. Components for System Timing

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PIXEL SHIFTING

The pixel-shifting stage consists of 8 very fast 74F166 Shift Registers divided into 4 banks, one for each address bit. These shift registers have maximum operating frequencies of 120MHz.

The data comes from the address and data buffering and the video RAM. The $\overline{PE}$ (Parallel Enable Input) signal from the system timing block activates the register, while the pixel clock, DCLK, strobes each of the registers. All chips are permanently enabled by grounding their chip enable ($\overline{CE}$) pins. The master reset ($\overline{MR}$) is permanently disabled by tying it to a pull-up.

The connection between the registers and the memory is such that all the bits of each pixel are shifted out simultaneously before going to the 74F157 multiplexer. From there, they address the colors of the CLUT on the Video DAC.

VIDEO RAM

The phrase "Video RAM" refers to a set of dynamic RAMs used as the memory section in this application. It is not meant to be confused with the Video RAM which is a dedicated device for video applications.

The Video RAM or frame buffer section consists of 8 Fujitsu MB85103-10 modules. The 10 suffix signals a 100ns row access time. The cycle time is about 200ns, or about 5MHz. This is fine because only the pixel clock has to travel at the high screen draw speeds. These modules are SIPs (single in-line packages) and were used because of space considerations. Each module consists of eight 64k × 1-bit DRAMs, giving eight modules of 64k × 8 or a 64k × 64 buffer. This buffer is divided into four sections (64k × 16) that represent the four bits of address that are shifted out to the NE5150's CLUT.

One can see how the frame buffer is set up to shift out data to the pixel shifter. The memory is divided into 4 banks that are write-selected by the $\overline{WE1}$ – $\overline{WE4}$ pulses. Two modules (64k × 16 bits) make up one bank. This makes up the four 16-bit words that are shifted out. But where is the information for each pixel? Taking the 1st bank as an example, it can be divided into 4 quadrants:

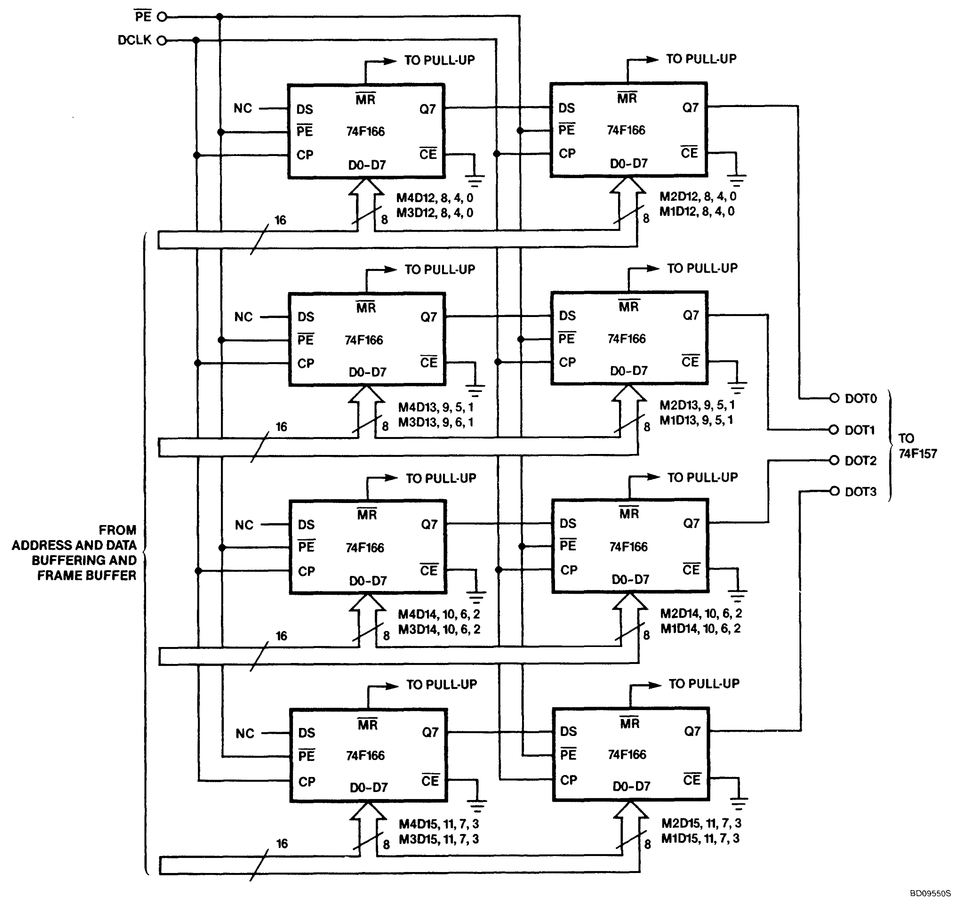


Figure 22. Shift Registers for Pixel-Shifting

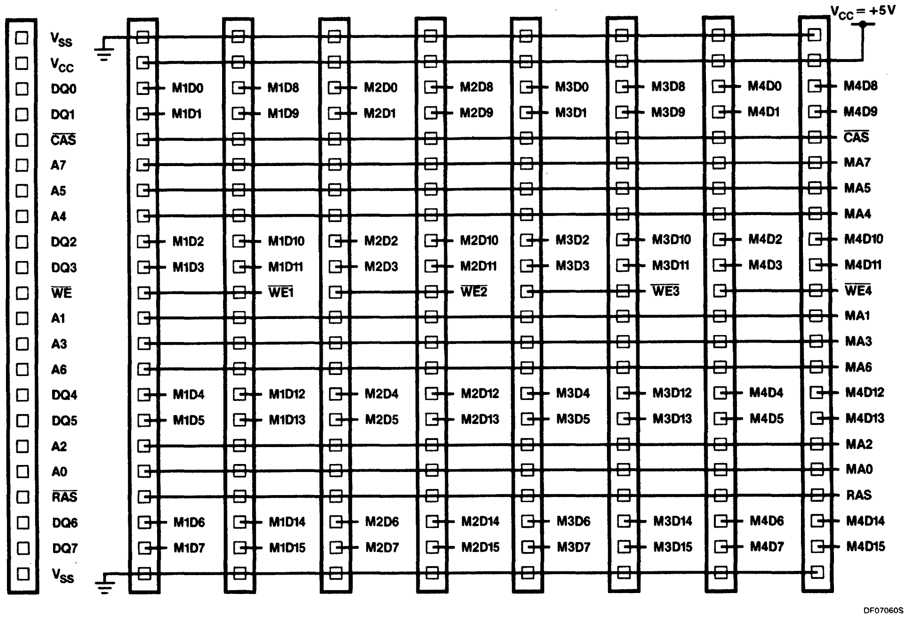


Figure 23. Memory Configuration to Store Pixels

M1D0 – M1D3, M1D4 – M1D7, M1D8 – M1D11, and M1D12 – M1D15. Each of these quadrants represents a dot. By tracking each dot in parallel back to the shift register in the pixel-shifting stage, they turn out to be each of the four quadrants in parallel. Comparing diagrams reveals the same to be true for each of the quadrants in each of the four banks of memory. Each quadrant, then, corresponds to one pixel, and all of the pixels for one bank are written out to the shift register during a write cycle.

VIDEO DAC INTERFACE

The interface to the NE5150 is shown in Figure 24. The 8-bit data bus comes from the lower 8 bits of the Lilith. The low 4 bits are connected directly to the Video DAC data inputs. Bits 4 – 7 are tied to the 74F157 Multiplexer. This provides the address to the CLUT when it is initialized.

The other set of inputs to the multiplexer comes from the pixel-shifting stage. After the first CLUT initialization, all of the addresses come from the pixel-shifter. The inverters, NAND gates, and OR gates are used to delay the write pulses $\overline{WRR}$, $\overline{WRG}$, and $\overline{WRB}$ so that they fit into the address setup window. The chip select pulses come from the 74F138 which are selected by the Lilith. $\overline{I/OCLK}$ clocks the 74138 and the OR gates for the chip select.

DCLK drives the STROBE of the DAC and clocks the two D-type flip-flops which provide the BLANKing signal. Both of these signals come from the ACRTC and the system timing section. The WHITE, BRIGHT, and SYNC inputs are not utilized and are connected to ground. V_{EE} is run off a 7905 voltage regulator powered by a –12V power supply.

The capacitors to the monitor and voltage regulator are polarized with the positive end to the monitor for the RGB outputs and to ground for the regulator. The regulator uses Tantalum capacitors.

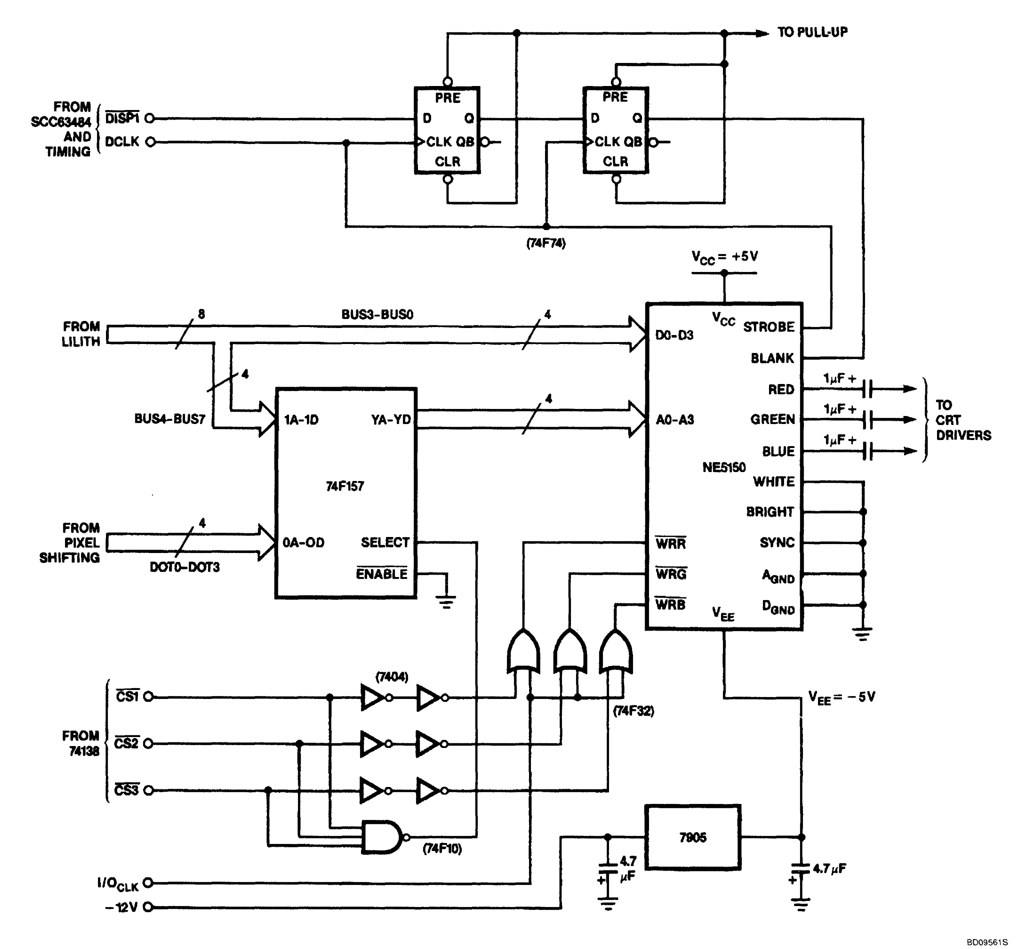


Figure 24. NE5150 Video DAC Interface

NE5150/51/52 Family of Video Digital-to-Analog Converters AN1081

GLOSSARY

This glossary consists of three parts: a section for graphics terminology, one for the timing of the NE5150 used in the Lilith workstation application, and a list of references. For the glossary section, many analogies are made with television to clarify some terminology.

GRAPHICS TERMINOLOGY

ACRTC — Short for Advanced CRT Controller. A device that helps to interface a microprocessor or microcomputer with a monitor. Advanced refers to the Signetics ACRTC, the SCC63484, called advanced because of its ability to do most of its graphics computations on-board, thus relieving some of the workload from the microprocessor and increasing its overall efficiency.

Bit-Map, Bit-Plane — A memory representation in which one or more bits correspond to a pixel. For each bit used in the representation of a pixel, there is a plane on which it can be mapped. To represent each pixel by 4 bits, 4 bit planes are needed. This is the case whether the bits store the actual data for the pixel or hold the address of the memory location containing the data.

Blanking — The process of turning off an electron gun so that it leaves no trace on the screen as it returns to the left or top of the screen in a raster-scan system. Applies to both television sets and monitors. The period for the blanking is defined as the horizontal blanking and the vertical blanking interval for their respective cases.

CRT — Short for Cathode Ray Tube, a type of electron tube that produces an electron beam that strikes the phosphor-coated screen, causing that screen to emit light.

Chrominance — The color information supplied in a signal. While this information has to be extracted by color decoders in television (via phase differencing with a fixed-frequency subcarrier), in computer monitors and bit-mapped systems it is supplied digitally and then converted to analog to directly drive color guns.

Color Look-up Table — Sometimes referred to as the CLUT, it is associated with a Video DAC and speeds system access of often-used colors. The time savings results because a color can be generated by sending a CLUT address to the DAC instead of loading a word from external memory. Current CLUTs range in size from 16 to 256 words. Word length depends on the bit resolution of the DAC.

DAC — Short for Digital-to-Analog Converter. Most DACs have a single output. Some have as many as eight. RGB Video DACs have three — one for each of the primary colors. Video DACs typically operate at very high speeds since they have to supply a new piece of information for each pixel on the screen at rates of 30 to 80 times per second.

ECL — Short for Emitter-Coupled-Logic. A fast, non-saturating form of bipolar logic that usually operates from 0 to −5.2V. It has a threshold of −1.3V.

Frame Buffer — Sometimes used interchangeably with video RAM. A frame buffer is a large, fast-access store of memory that contains the digital information necessary to display part or all of a display. It is used in conjunction with bit-mapped graphic systems. It actually "stores" the bit-plane.

Glitch Energy — The area displaced by an analog signal as it overshoots or undershoots its ideal value. This is a problem usually found in DACs. Units are usually given in pV-s. When glitch energy is high, settling times tend to be longer and may result in visual color aberrations on the screen.

Hue — The actual color(s) on a monitor. The hue depends on the frequency of the light striking the human eye. For television transmission, it is determined by the video signal's phase difference with a color subcarrier reference frequency. For computer graphics systems, it is determined by the combination of binary values applied to the DAC. The resolution of hue/colors is determined by the bit length of each word of information.

Lilith — The brand name of the workstation manufactured by Modulo, Inc. of Provo, UT.

Luminance — The brightness information in a video signal. A black and white (monochrome) monitor displays only variations in brightness. Only a luminance signal is being manipulated. The same holds true for television. Although chrominance information is also present in a television signal, B/W TV sets do not have the necessary decoders.

Modula-2 — A language that is the superset of Pascal. This was also invented by Niklaus Wirth of the Swiss Technological Institute.

NTSC — Short for the National Television Standards Committee, the ruling body for television standards in the United States. Other countries also use this standard as is, or with a different frequency for the color subcarrier.

Orthogonal — Defined as being mutually perpendicular. The product of two orthogonal vectors is zero. In bit-mapped systems, the bit length of a word lies orthogonal to the plane itself. Hence, each plane supplies only one bit of information for each pixel.

Pixel — Short for "picture element". The smallest resolvable element on a graphics display. Each pixel usually corresponds to at least one bit. The entire display is made up of a map of pixels. The term bit-map comes from the bit association. There is no equivalent in television. What is seen is the true analog representation of what is being recorded by a camera and then retraced on horizontal lines.

Raster-Scan — The form of visual display transmission used in all television sets and in most monitors. It consists of an electron beam tracing a path from left-to-right while going top-to-bottom.

Saturation — The "deepness" of a color. Usually depends on the amplitude of the color signal in television systems. Red and pink are the same hue, but red is actually more saturated than pink. In graphics systems, there is no true equivalent. Changing bit-values changes the color itself. The closest analogy would be to raise or lower the voltages on all three color guns simultaneously (the BRIGHT function on the NE5150/51/52). This would, however, depending on the amplitude change, give the impression of brightening or dimming the color (changing luminance) rather than saturating it.

Sync — The voltage level specified in RS-343A as being 140 IRE (1V) below the enhanced white level (ground). It is also 40 IRE (286mV) below the blanking level. Generically it is also used to refer to vertical and horizontal sync pulses that synchronize the timing and movement of the electron beam on a CRT. It should not be confused with "composite sync".

Teletext — A form of data transmission via television signals. In many cases, digital information is sent during the vertical blanking interval (VBI). In some cases, it is sent during every retrace. This is known as full-field teletext.

TTL — Short for Transistor-Transistor Logic. It has a threshold voltage of approximately 1.4V and is the most widely-used form of logic in the world today.

DEFINITIONS FOR NE5150/51/52 TIMING DIAGRAMS

This section contains explanations for the NE5150/51/52 Video DAC's timing diagram specifications. For the typical, minimum, and maximum values, please refer to Signetics' data sheet.

t_{WAS} — Write Address Setup (NE5150/52)

t_{WAH} — Write Address Hold (NE5150/52)

t_{WDS} — Write Data Setup (NE5150/52)

t_{WDH} — Write Data Hold (NE5150/52)

t_{WEW} — Write Enable Pulse Width (NE5150/52)

t_{RCS} — Read Composite Setup (NE5150/52)

t_{RCH} — Read Composite Hold (NE5150/52)

t_{RAS} — Read Address Setup (NE5150/52)

t_{RAH} — Read Address Hold (NE5150/52)

t_{RSW} — Read Strobe Pulse Width (NE5150/52)

t_{RDD} — Read DAC Delay (NE5150/52)

t_{CS} — Composite Setup (NE5151)

t_{CH} — Composite Hold (NE5151)

t_{DS} — Data bits Setup (NE5151)

t_{DH} — Data bits Hold (NE5151)

t_{SW} — Strobe Pulse Width (NE5151)

t_{DD} — DAC Delay (NE5151)

t_R — DAC Rise Time (NE5151)

t_S — DAC Full-Scale Settling Time (NE5151)

REFERENCES

The following books, articles, notes, and correspondences were used in the preparation of this application note.

1. *Raster Graphics Handbook*, 2nd edition, by the Conrac Corporation
2. "Trends in Graphics Hardware", paper by Randall R. Bird, Genisco Computers Corporation; presented at WESCON '85
3. *Basic Television and Video Systems*, 5th edition, by Bernard Grob, McGraw-Hill
4. *Getting the Best Performance from Video Digital-to-Analog Converters*, (AN-1) by Dennis Packard, Brooktree Corporation, San Diego
5. "A Cost-Effective Custom CAD System", paper by R.C. Burton, D.G. Brewer, R.E. Penman, and R. Schilimoeller, Computer Science Department, Brigham Young University and Signetics Corporation
6. "Lilith and Modula-2", by Richard Ohran, *Byte Magazine*, pgs. 181 – 192; August 1984
7. "Monolithic Color Palette Fills in the Picture for High-Speed Graphics", by Steven Sidman and John C. Kuklewicz, *Electronic Design*; November 29, 1984
8. *EIA Standard RS-343A: Electrical Performance Standards for High-Resolution Monochrome Closed-Circuit Television Camera*, by the Video Engineering Department of the Electronic Industries Association; September, 1969
9. "A Single-Chip RGB Digital-to-Analog Converter with High-Speed Color-Map Memory", by W. Mack and M. Horowitz, *Digest of the International Conference on Consumer Electronics*, p. 90; 1985

Signetics

TDA8440
Video and Audio Switch IC

Product Specification

Linear Products

DESCRIPTION

The TDA8440 is a versatile video/audio switch, intended to be used in applications equipped with video/audio inputs.

It provides two 3-State switches for audio channels and one 3-State switch for the video channel and a video amplifier with selectable gain (times 1 or times 2).

The integrated circuit can be controlled via a bidirectional I^2C bus or it can be controlled directly by DC switching signals. Sufficient sub-addressing is provided for the I^2C bus mode.

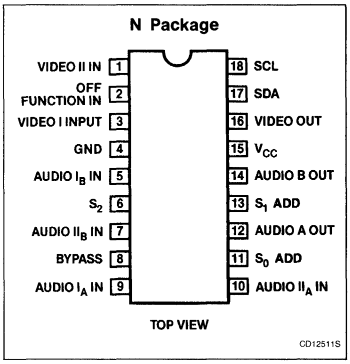
FEATURES

- **Combined analog and digital circuitry gives maximum flexibility in channel switching**
- **3-State switches for all channels**
- **Selectable gain for the video channels**
- **Sub-addressing facility**
- **I^2C bus or non-I^2C bus mode (controlled by DC voltages)**
- **Slave receiver in the I^2C bus mode**
- **External OFF command**
- **System expansion possible up to 7 devices (14 sources)**
- **Static short-circuit proof outputs**

APPLICATIONS

- **TVRO**
- **Video and audio switching**
- **Television**
- **CATV**

PIN CONFIGURATION



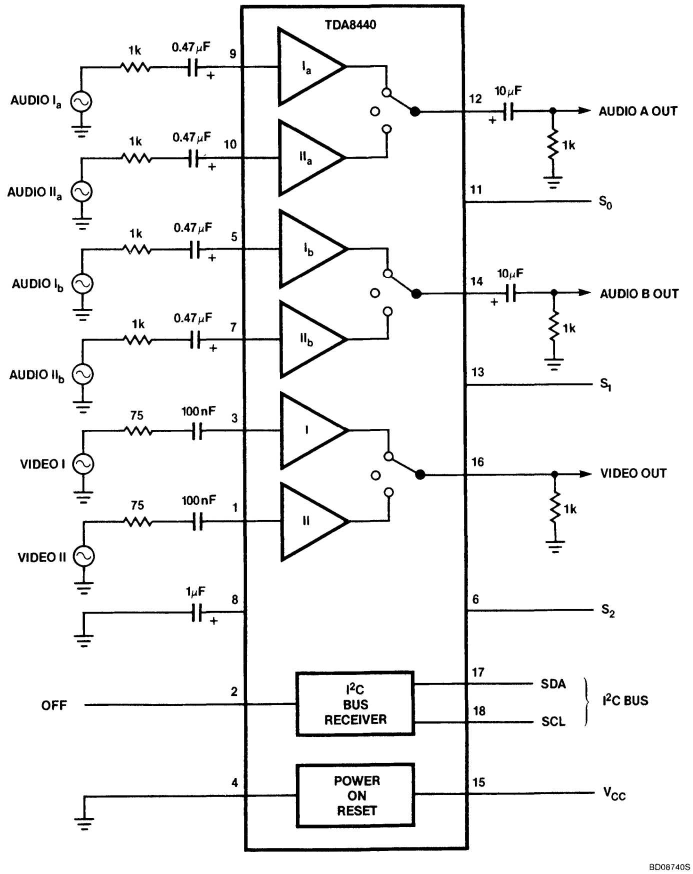
ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
18-Pin Plastic DIP (SOT-102)	0 to 70°C	TDA8440N

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{CC}	Supply voltage Pin 15	14	V
	Input voltage		
V_{SDA}	Pin 17	-0.3 to V_{CC} $+0.3$	V
V_{SCL}	Pin 18	-0.3 to V_{CC} $+0.3$	V
V_{OFF}	Pin 2	-0.3 to V_{CC} $+0.3$	V
V_{S0}	Pin 11	-0.3 to V_{CC} $+0.3$	V
V_{S1}	Pin 13	-0.3 to V_{CC} $+0.3$	V
V_{S2}	Pin 6	-0.3 to V_{CC} $+0.3$	V
$-I_{16}$	Video output current Pin 16	50	mA
T_{STG}	Storage temperature range	-65 to $+150$	°C
T_A	Operating ambient temperature range	0 to $+70$	°C
T_J	Junction temperature	$+150$	°C
θ_{JA}	Thermal resistance from junction to ambient in free-air	50	°C/W

BLOCK DIAGRAM AND TEST CIRCUIT



NOTE:
S0, S1, S2, and OFF (Pins 11, 13, 6, and 2) connected to V_{CC} or GND. If more than 1 device is used, the outputs and Pin 8 (bias decoupling of the audio inputs) may be connected in parallel.

DC ELECTRICAL CHARACTERISTICS $T_A = 25°C$; $V_{CC} = 12V$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supply					
V_{15-4}	Supply voltage	10		13.2	V
I_{15}	Supply current (without load)		37	50	mA
Video switch					
C_1C_3	Input coupling capacitor	100			nF
A_{3-16} A_{3-16}	Voltage gain (times 1; SCL = L) (times 2; SCL = H)	−1 +5	0 +6	+1 +7	dB dB
A_{1-16} A_{1-16}	Voltage gain (times 1; SCL = L) (times 2; SCL = H)	−1 +5	0 +6	+1 +7	dB dB
V_{3-4}	Input video signal amplitude (gain times 1)			4.5	V
V_{1-4}	Input video signal amplitude (gain times 1)			4.5	V
Z_{16-4}	Output impedance		7		Ω
Z_{16-4}	Output impedance in 'OFF' state	100			kΩ
	Isolation (off-state) ($f_O = 5MHz$)	60			dB
S/S + N	Signal-to-noise ratio[2]	60			dB
V_{16-4}	Output top-sync level	2.4	2.8	3.2	V
G	Differential gain			3	%
V_{16-4}	Minimum crosstalk attenuation[1]	60			dB
RR	Supply voltage rejection[3]	36			dB
BW	Bandwidth (1dB)	10			MHz
∝	Crosstalk attenuation for interference caused by bus signals (source impedance 75Ω)	60			db
Audio switch "A" and "B"					
V_{9-4} (RMS) V_{10-4} (RMS) V_{5-4} (RMS) V_{7-4} (RMS)	Input signal level			2 2 2 2	V V V V
Z_{9-4} Z_{10-4} Z_{5-4} Z_{7-4}	Input impedance	50 50 50 50	100 100 100 100		kΩ kΩ kΩ kΩ
Z_{12-4} Z_{14-4}	Output impedance			10 10	Ω Ω
Z_{14-4}	Output impedance (off-state)	100			kΩ
V_{9-12} V_{10-12} V_{5-14} V_{7-14}	Voltage gain	−1 −1 −1 −1	0 0 0 0	+1 +1 +1 +1	dB dB dB dB
	Isolation (off-state) (f = 20kHz)	90			dB
S/S + N	Signal-to-noise ratio[4]	90			dB
THD	Total harmonic distortion[6]			0.1	%

DC ELECTRICAL CHARACTERISTICS (Continued) $T_A = 25°C$; $V_{CC} = 12V$, unless otherwise specified.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
	Crosstalk attenuation for interferences caused by video signals[5]				
α	Weighted	80			dB
α	Unweighted	80			dB
α	Crosstalk attenuation for interferences caused by sinusoidal sound signals[5]	80			dB
	Crosstalk attenuation for interferences caused by the bus signal (weighted) (source impedance = 1kΩ)	80			dB
RR	Supply voltage rejection	50			dB
BW	Bandwidth (−1dB)	50			kHz
I^2C bus inputs/outputs SDA (Pin 17) and SCL (Pin 18)					
V_{IH}	Input voltage HIGH	3		V_{CC}	V
V_{IL}	Input voltage LOW	−0.3		+1.5	V
I_{IH}	Input current HIGH[7]			10	μA
I_{IL}	Input current LOW[7]			10	μA
V_{OL}	Output voltage LOW at $I_{OL} = 3mA$			0.4	V
I_{OL}	Maximum output sink current		5		mA
C_I	Capacitance of SDA and SCL inputs, Pins 17 and 18			10	pF
Sub-address inputs S_0 (Pin 11), S_1 (Pin 13), S_2 (Pin 6)					
V_{IH}	Input voltage HIGH	3		V_{CC}	V
V_{IL}	Input voltage LOW	−0.3		+0.4	V
I_{IH}	Input current HIGH			10	μA
I_{IL}	Input current LOW	−50		0	μA
OFF input (Pin 2)					
V_{IH}	Input voltage HIGH	+3		V_{CC}	V
V_{IL}	Input voltage LOW	−0.3		+0.4	V
I_{IH}	Input current HIGH			20	μA
I_{IL}	Input current LOW	−10		2	μA

NOTES:

1. Caused by drive on any other input at maximum level, measured in B = 5MHz, source impedance for the used input 75Ω, crosstalk $= 20\log \frac{V_{OUT}}{V_{IN}\ max}$.
2. S/N $= 20\log \frac{V_O\ \text{video noise}_{(P-P)}\ (2V)}{V_O\ \text{noise}_{RMS}\ B = 5MHz}$.
3. Supply voltage ripple rejection $= 20\log \frac{V_R\ \text{supply}}{V_R\ \textit{on output}}$ at f = max. 100kHz.
4. S/N $= 20\log \frac{V_O\ \text{nominal}\ (0.5V)}{V_O\ \text{noise}\ B = 20kHz}$.
5. Caused by drive of any other input at maximum level, measured in B = 20kHz, source impedance of the used input = 1kΩ, crosstalk $= 20\log \frac{V_{OUT}}{V_{IN}\ max}$ according to DIN 45405 (CCIR 468).
6. f = 20Hz to 20kHz.
7. Also if the supply is switched off.

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AC ELECTRICAL CHARACTERISTICS

I^2C bus load conditions are as follows: 4kΩ pull-up resistor to +5V; 200pF to GND. All values are referred to V_{IH} = 3V and V_{IL} = 1.5V.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
t_{BUF}	Bus free before start	4			μs
$t_{S\ (STA)}$	Start condition setup time	4			μs
$t_{H\ (STA)}$	Start condition hold time	4			μs
t_{LOW}	SCL, SDA LOW period	4			μs
t_{HIGH}	SCL, HIGH period	4			μs
t_R	SCL, SDA rise time			1	μs
t_F	SCL, SDA fall time			0.3	μs
$t_{S\ (DAT)}$	Data setup time (write)	1			μs
$t_{H\ (DAT)}$	Data hold time (write)	1			μs
$t_{S\ (CAC)}$	Acknowledge (from TDA8440) setup time			2	μs
$t_{H\ (CAC)}$	Acknowledge (from TDA8440) hold time	0			μs
$t_{S\ (STO)}$	Stop condition setup time	4			μs

Table 1. Sub-Addressing

S_2	S_1	S_0	SUB-ADDRESS		
			A_2	A_1	A_0
L	L	L	0	0	0
L	L	H	0	0	1
L	H	L	0	1	0
L	H	H	0	1	1
H	L	L	1	0	0
H	L	H	1	0	1
H	H	L	1	1	0
H	H	H	non I^2C addressable		

FUNCTIONAL DESCRIPTION

The TDA8440 is a monolithic system of switches and can be used in CTV receivers equipped with an auxiliary video/audio plug. The IC incorporates 3-State switches which comprise:

a) An electronic video switch with selectable gain (times 1 or times 2) for switching between an internal video signal (from the IF amplifier) with an auxiliary input signal.

b) Two electronic audio switches, for two sound channels (stereo or dual language), for switching between internal audio sources and signals from the auxiliary video/audio plug.

A selection can be made between two input signals and an OFF-state. The OFF-state is necessary if more than one TDA8440 device is used.

The SDA and SCL pins can be connected to the I^2C bus or to DC switching voltages. Inputs S_0 (Pin 11), S_1 (Pin 13), and S_2 (Pin 6) are used for selection of sub-addresses or switching to the non-I^2C mode. Inputs S_0, S_1, and S_2 can be connected to the supply voltage (H) or to ground (L). In this way, no peripheral components are required for selection.

NON-I^2C BUS CONTROL

If the TDA8440 switching device has to be operated via the auxiliary video/audio plug, inputs S_2, S_1, and S_0 must be connected to the supply line (12V).

The sources (internal and external) and the gain of the video amplifier can be selected via the SDA and SCL pins with the switching voltage from the auxiliary video/audio plug:

- Sources I are selected if SDA = 12V (external source)
- Sources II are selected if SDA = 0V (TV mode)
- Video amplifier gain is 2 × if SCL = 12V (external source)
- Video amplifier gain is 1 × if SCL = 0V (TV mode)

If more than one TDA8440 device is used in the non-I^2C bus system, the OFF pin can be used to switch off the desired devices. This can be done via the 12V switching voltage on the plug.

- All switches are in the OFF position if OFF = H (12V)
- All switches are in the selected position via SDA and SCL pins if OFF = L (0V)

I^2C BUS CONTROL

Detailed information on the I^2C bus is available on request.

Table 2. TDA8440 I^2C Bus Protocol

STA	A_6	A_5	A_4	A_3	A_2	A_1	A_0	R/W	AC	D_7	D_6	D_5	D_4	D_3	D_2	D_1	D_0	AC	STO

STA = start condition
A_6 = 1
A_5 = 0
A_4 = 0
A_3 = 1
(Fixed address bits: A_6–A_3)
A_2 = sub-address bit, fixed via S_2 input
A_1 = sub-address bit, fixed via S_1 input
A_0 = sub-address bit, fixed via S_0 input
R/W = read/write bit (has to be 0, only write mode allowed)
AC = acknowledge bit (= 0) generated by the TDA8440
D_7 = 1 audio I_a is selected to audio output a
D_7 = 0 audio I_a is not selected
D_6 = 1 audio II_a is selected to audio output a
D_6 = 0 audio II_a is not selected
D_5 = 1 audio I_b is selected to audio output b
D_5 = 0 audio I_b output is not selected
D_4 = 1 audio II_b is selected to audio output b
D_4 = 0 audio II_b is not selected
D_3 = 1 video I is selected to video output
D_3 = 0 video I is not selected
D_2 = 1 video II is selected to video output
D_2 = 0 video II is not selected
D_1 = 1 video amplifier gain is times 2
D_1 = 0 video amplifier gain is times 1
D_0 = 1 OFF-input inactive
D_0 = 0 OFF-input active
STO = stop condition

D_0/OFF Gating

D_0	OFF input	Outputs
0 (off input active)	H	OFF
0	L	In accordance with last defined $D_7 - D_1$ (may be entered while OFF = HIGH)
1 (off input inactive)	H	In accordance with $D_7 - D_1$
1	L	In accordance with $D_7 - D_1$

OFF FUNCTION

With the OFF input all outputs can be switched off (high-ohmic mode), depending on the value of D_0.

Power-on Reset

The circuit is provided with a power-on reset function.

When the power supply is switched on, an internal pulse will be generated that will reset the internal memory S_0. In the initial state all the switches will be in the off position and the OFF input is active ($D_7 - D_0 = 0$), (I^2C mode). In the non-I^2C mode, positions are defined via SDA and SCL input voltages.

When the power supply decreases below 5V, a pulse will be generated and the internal memory will be reset. The behavior of the switches will be the same as described above.

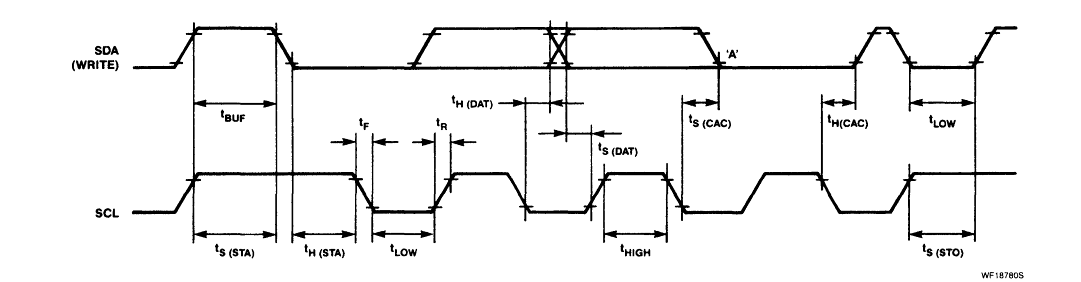


Figure 1. I^2C Bus Timing Diagram

Signetics

NE/SA5204
Wide-band High-Frequency Amplifier

Product Specification

Linear Products

DESCRIPTION

The NE/SA5204 is a high-frequency amplifier with a fixed insertion gain of 20dB. The gain is flat to ± 0.5dB from DC to 200MHz. The -3dB bandwidth is greater than 350MHz. This performance makes the amplifier ideal for cable TV applications. The NE/SA5204 operates with a single supply of 6V, and only draws 25mA of supply current, which is much less than comparable hybrid parts. The noise figure is 4.8dB in a 75Ω system and 6dB in a 50Ω system.

The NE/SA5204 is a relaxed version of the NE5205. Minimum guaranteed bandwidth is relaxed to 350MHz and the "S" parameter Min/Max limits are specified as typicals only.

Until now, most RF or high-frequency designers had to settle for discrete or hybrid solutions to their amplification problems. Most of these solutions required trade-offs that the designer had to accept in order to use high-frequency gain stages. These include high power consumption, large component count, transformers, large packages with heat sinks, and high part cost. The NE/SA5204 solves these problems by incorporating a wideband amplifier on a single monolithic chip.

The part is well matched to 50 or 75Ω input and output impedances. The standing wave ratios in 50 and 75Ω systems do not exceed 1.5 on either the input or output over the entire DC to 350MHz operating range.

Since the part is a small, monolithic IC die, problems such as stray capacitance are minimized. The die size is small enough to fit into a very cost-effective 8-pin small-outline (SO) package to further reduce parasitic effects.

No external components are needed other than AC-coupling capacitors because the NE/SA5204 is internally compensated and matched to 50 and 75Ω. The amplifier has very good distortion specifications, with second and third-order intermodulation intercepts of +24dBm and +17dBm, respectively, at 100MHz.

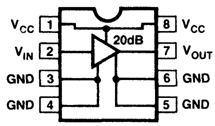
The part is well matched for 50Ω test equipment such as signal generators, oscilloscopes, frequency counters, and all kinds of signal analyzers. Other applications at 50Ω include mobile radio, CB radio, and data/video transmission in fiber optics, as well as broadband LANs and telecom systems. A gain greater than 20dB can be achieved by cascading additional NE/SA5204s in series as required, without any degradation in amplifier stability.

FEATURES

- **Bandwidth (min.)**
  200 MHz, ±0.5dB
  350 MHz, -3dB
- **20dB insertion gain**
- **4.8dB (6dB) noise figure**
  $Z_O = 75\Omega$ ($Z_O = 50\Omega$)
- **No external components required**
- **Input and output impedances matched to 50/75Ω systems**
- **Surface-mount package available**
- **Cascadable**

PIN CONFIGURATION

N, D Packages



TOP VIEW

CD12760S

APPLICATIONS

- **Antenna amplifiers**
- **Amplified splitters**
- **Signal generators**
- **Frequency counters**
- **Oscilloscopes**
- **Signal analyzers**
- **Broadband LANs**
- **Networks**
- **Modems**
- **Mobile radio**
- **Security systems**
- **Telecommunications**

ORDERING INFORMATION

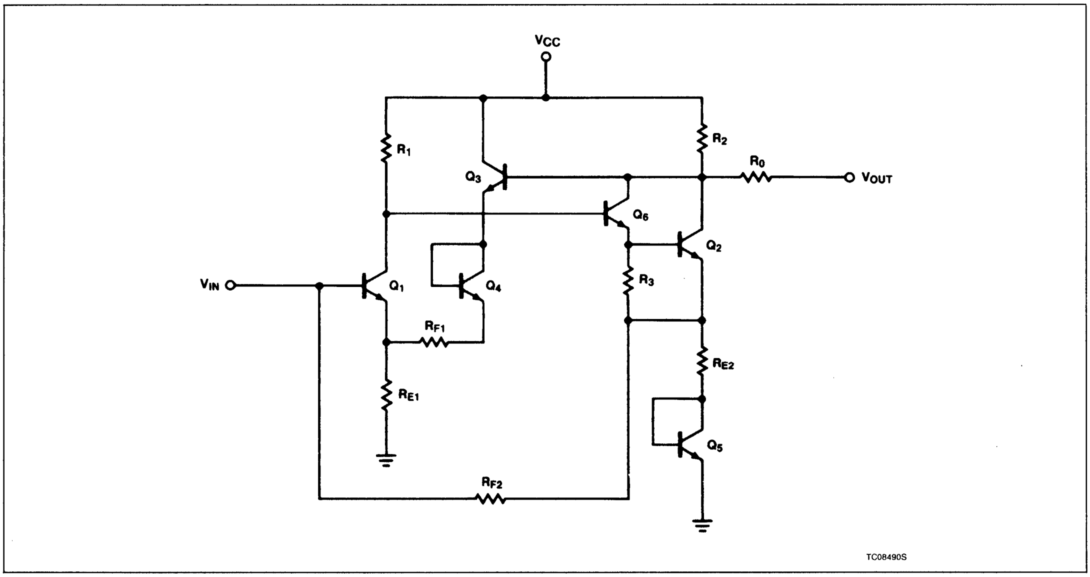
DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
8-Pin Plastic DIP	0 to +70°C	NE5204N
	-40 to +85°C	SA5204N
8-Pin Plastic SO package	0 to +70°C	NE5204D
	-40 to +85°C	SA5204D

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{CC}	Supply voltage	9	V
V_{IN}	AC input voltage	5	V_{P-P}
T_A	Operating ambient temperature range NE grade SA grade	 0 to +70 −40 to +85	 °C °C
P_{DMAX}	Maximum power dissipation[1, 2] T_A = 25°C (still-air) N package D package	 1160 780	 mW mW
T_J	Junction temperature	150	°C
T_{STG}	Storage temperature range	−55 to +150	°C
T_{SOLD}	Lead temperature (soldering 60s)	300	°C

NOTES:
1. Derate above 25°C, at the following rates
   N package at 9.3mW/°C
   D package at 6.2mW/°C.
2. See "Power Dissipation Considerations" section.

EQUIVALENT SCHEMATIC



DC ELECTRICAL CHARACTERISTICS at V_{CC} = 6V, $Z_S = Z_L = Z_O = 50\Omega$ and T_A = 25°C, in all packages, unless otherwise specified.

SYMBOL	PARAMETER	TEST CONDITIONS	LIMITS			UNIT
			Min	Typ	Max	
V_{CC}	Operating supply voltage range	Over temperature	5		8	V
I_{CC}	Supply current	Over temperature	19	24	31	mA
S21	Insertion gain	f = 100MHz, over temperature	16	19	22	dB
S11	Input return loss	f = 100MHz		25		dB
		DC –550MHz		12		dB
S22	Output return loss	f = 100MHz		27		dB
		DC –550MHz		12		dB
S12	Isolation	f = 100MHz		–25		dB
		DC –550MHz		–18		dB
BW	Bandwidth	±0.5dB	200	350		MHz
BW	Bandwidth	–3dB	350	550		MHz
	Noise figure (75Ω)	f = 100MHz		4.8		dB
	Noise figure (50Ω)	f = 100MHz		6.0		dB
	Saturated output power	f = 100MHz		+7.0		dBm
	1dB gain compression	f = 100MHz		+4.0		dBm
	Third-order intermodulation intercept (output)	f = 100MHz		+17		dBm
	Second-order intermodulation intercept (output)	f = 100MHz		+24		dBm
t_R	Rise time			5		ps
	Propagation delay			5		ps

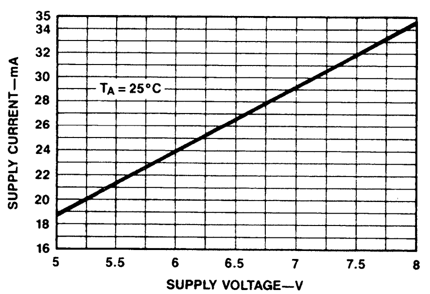


Figure 1. Supply Current vs Supply Voltage

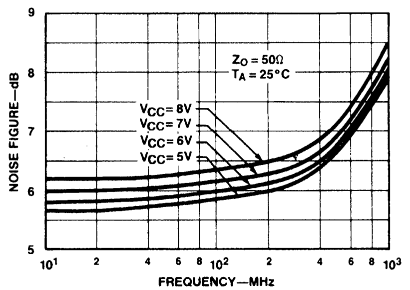


Figure 2. Noise Figure vs Frequency

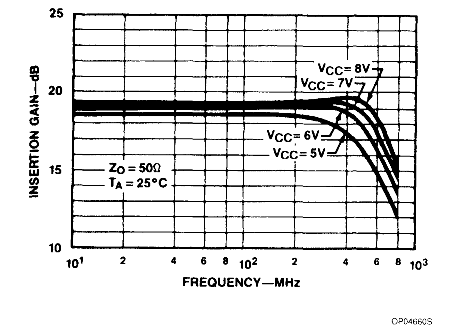


Figure 3. Insertion Gain vs Frequency (S_{21})

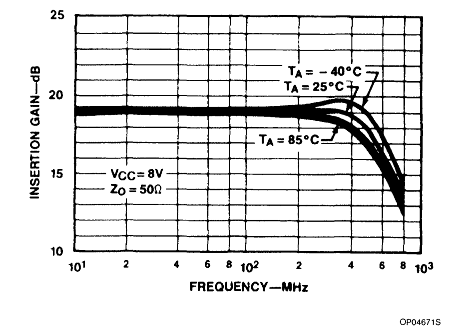


Figure 4. Insertion Gain vs Frequency (S_{21})

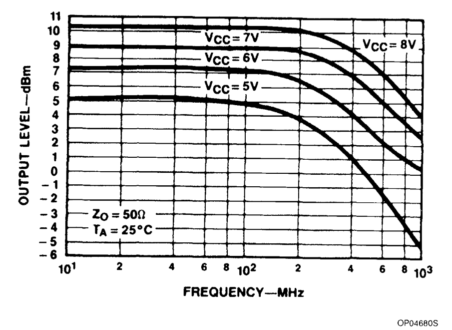


Figure 5. Saturated Output Power vs Frequency

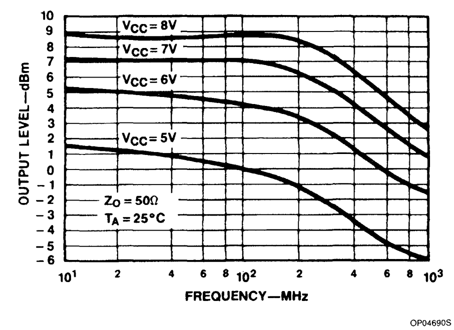


Figure 6. 1dB Gain Compression vs Frequency

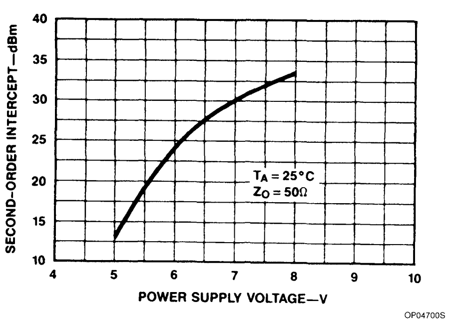


Figure 7. Second-Order Output Intercept vs Supply Voltage

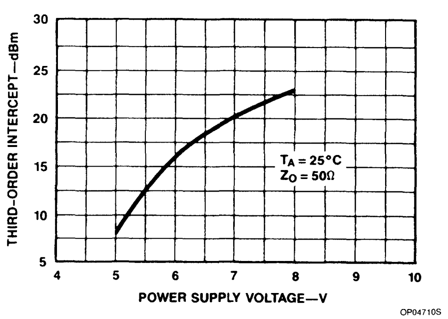


Figure 8. Third-Order Intercept vs Supply Voltage

11

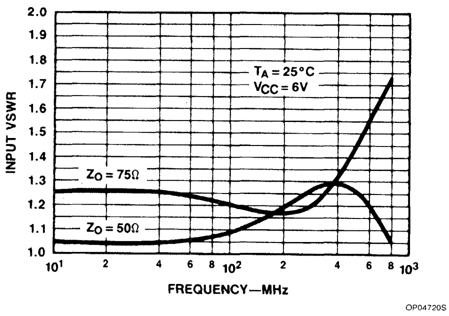


Figure 9. Input VSWR vs Frequency

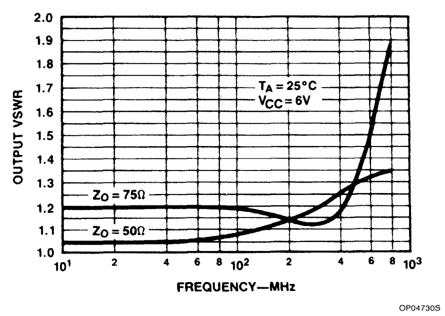


Figure 10. Output VSWR vs Frequency

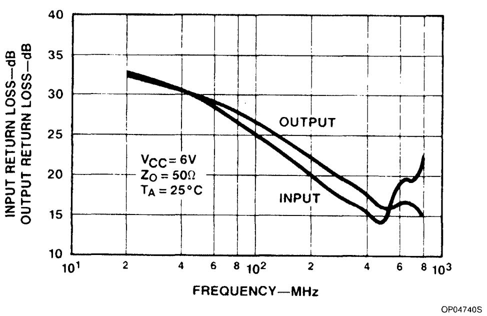


Figure 11. Input (S_{11}) and Output (S_{22}) Return Loss vs Frequency

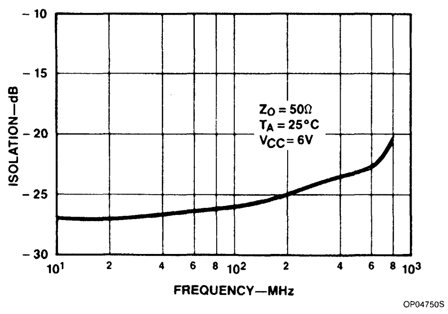


Figure 12. Isolation vs Frequency (S_{12})

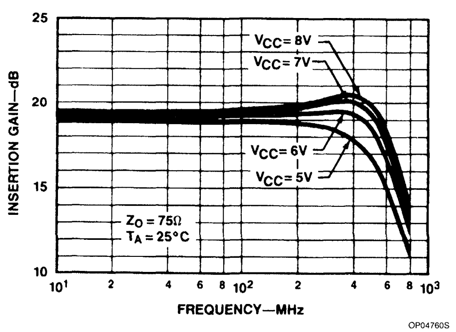


Figure 13. Insertion Gain vs Frequency (S_{21})

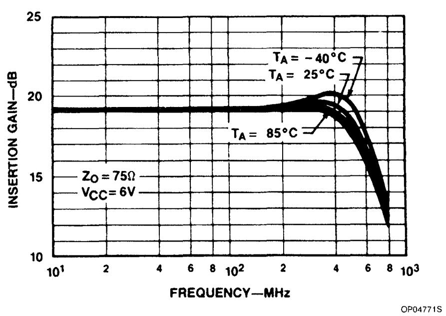


Figure 14. Insertion Gain vs Frequency (S_{21})

THEORY OF OPERATION

The design is based on the use of multiple feedback loops to provide wide-band gain together with good noise figure and terminal impedance matches. Referring to the circuit schematic in Figure 15, the gain is set primarily by the equation:

$$\frac{V_{OUT}}{V_{IN}} = (R_{F1} + R_{E1})/R_{E1} \qquad (1)$$

which is series-shunt feedback. There is also shunt-series feedback due to R_{F2} and R_{E2} which aids in producing wide-band terminal impedances without the need for low value input shunting resistors that would degrade the noise figure. For optimum noise performance, R_{E1} and the base resistance of Q_1 are kept as low as possible, while R_{F2} is maximized.

The noise figure is given by the following equation:

$$NF = 10\text{Log}\left\{1 + \frac{\left[r_b + R_{E1} + \frac{KT}{2qI_{C1}}\right]}{R_O}\right\}\text{dB} \qquad (2)$$

where $I_{C1} = 5.5\text{mA}$, $R_{E1} = 12\Omega$, $r_b = 130\Omega$, $KT/q = 26\text{mV}$ at 25°C and $R_O = 50$ for a 50Ω system and 75 for a 75Ω system.

The DC input voltage level V_{IN} can be determined by the equation:

$$V_{IN} = V_{BE1} + (I_{C1} + I_{C3})\, R_{E1} \qquad (3)$$

where $R_{E1} = 12\Omega$, $V_{BE} = 0.8\text{V}$, $I_{C1} = 5\text{mA}$ and $I_{C3} = 7\text{mA}$ (currents rated at $V_{CC} = 6\text{V}$).

Under the above conditions, V_{IN} is approximately equal to 1V.

Level shifting is achieved by emitter-follower Q_3 and diode Q_4, which provide shunt feedback to the emitter of Q_1 via R_{F1}. The use of an emitter-follower buffer in this feedback loop essentially eliminates problems of shunt-feedback loading on the output. The value of $R_{F1} = 140\Omega$ is chosen to give the desired nominal gain. The DC output voltage V_{OUT} can be determined by:

$$V_{OUT} = V_{CC} - (I_{C2} + I_{C6})R2, \qquad (4)$$

where $V_{CC} = 6\text{V}$, $R_2 = 225\Omega$, $I_{C2} = 7\text{mA}$ and $I_{C6} = 5\text{mA}$.

From here, it can be seen that the output voltage is approximately 3.3V to give relatively equal positive and negative output swings. Diode Q_5 is included for bias purposes to allow direct coupling of R_{F2} to the base of Q_1. The dual feedback loops stabilize the DC operating point of the amplifier.

The output stage is a Darlington pair (Q_6 and Q_2) which increases the DC bias voltage on the input stage (Q_1) to a more desirable value, and also increases the feedback loop gain. Resistor R_O optimizes the output VSWR (Voltage Standing Wave Ratio). Inductors L_1 and L_2 are bondwire and lead inductances which are roughly 3nH. These improve the high-frequency impedance matches at input and output by partially resonating with 0.5pF of pad and package capacitance.

POWER DISSIPATION CONSIDERATIONS

When using the part at elevated temperature, the engineer should consider the power dissipation capabilities of each package.

At the nominal supply voltage of 6V, the typical supply current is 25mA (30mA max). For operation at supply voltages other than 6V, see Figure 1 for I_{CC} versus V_{CC} curves. The supply current is inversely proportional to temperature and varies no more than 1mA between 25°C and either temperature extreme. The change is 0.1% per °C over the range.

The recommended operating temperature ranges are air-mount specifications. Better heat-sinking benefits can be realized by mounting the SO and N package bodies against the PC board plane.

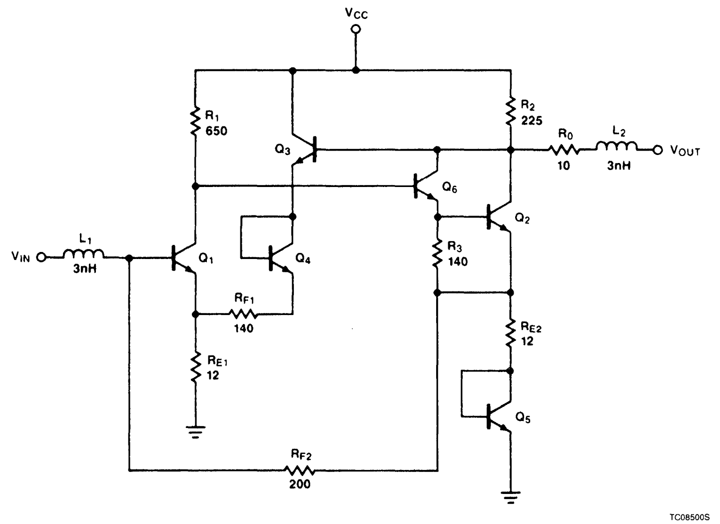


Figure 15. Schematic Diagram

Wide-band High-Frequency Amplifier

NE/SA5204

PC BOARD MOUNTING

In order to realize satisfactory mounting of the NE5204 to a PC board, certain techniques need to be utilized. The board must be double-sided with copper and all pins must be soldered to their respective areas (i.e., all GND and V_{CC} pins on the package). The power supply should be decoupled with a capacitor as close to the V_{CC} pins as possible, and an RF choke should be inserted between the supply and the device. Caution should be exercised in the connection of input and output pins. Standard microstrip should be observed wherever possible. There should be no solder bumps or burrs or any obstructions in the signal path to cause launching problems. The path should be as straight as possible and lead lengths as short as possible from the part to the cable connection. Another important consideration is that the input and output should be AC-coupled. This is because at $V_{CC} = 6V$, the input is approximately at 1V while the output is at 3.3V. The output must be decoupled into a low-impedance system, or the DC bias on the output of the amplifier will be loaded down, causing loss of output power. The easiest way to decouple the entire amplifier is by soldering a high-frequency chip capacitor directly to the input and output pins of the device. This circuit is shown in Figure 16. Follow these recommendations to get the best frequency response and noise immunity. The board design is as important as the integrated circuit design itself.

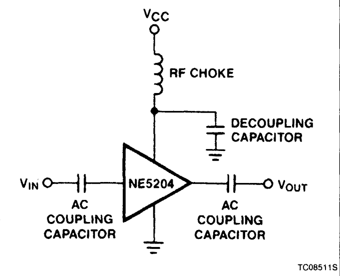
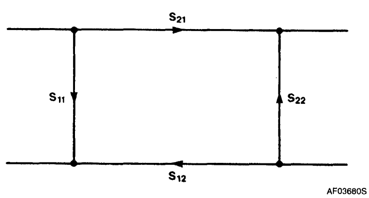


Figure 16. Circuit Schematic for Coupling and Power Supply Decoupling

SCATTERING PARAMETERS

The primary specifications for the NE5204 are listed as S-parameters. S-parameters are measurements of incident and reflected currents and voltages between the source, amplifier, and load as well as transmission losses. The parameters for a two-port network are defined in Figure 17.



a. Two-Port Network Defined

S_{11} — INPUT RETURN LOSS

$$S_{11} \equiv \sqrt{\frac{\text{POWER REFLECTED FROM INPUT PORT}}{\text{POWER AVAILABLE FROM GENERATOR AT INPUT PORT}}}$$

S_{12} — REVERSE TRANSMISSION LOSS OR ISOLATION

$$S_{12} \equiv \sqrt{\text{REVERSE TRANSDUCER POWER GAIN}}$$

S_{21} — FORWARD TRANSMISSION LOSS OR INSERTION GAIN

$$S_{21} \equiv \sqrt{\text{TRANSDUCER POWER GAIN}}$$

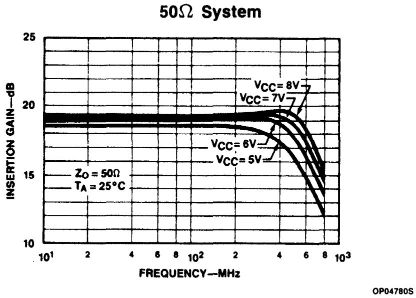
S_{22} — OUTPUT RETURN LOSS

$$S_{22} \equiv \sqrt{\frac{\text{POWER REFLECTED FROM OUTPUT PORT}}{\text{POWER AVAILABLE FROM GENERATOR AT OUTPUT PORT}}}$$

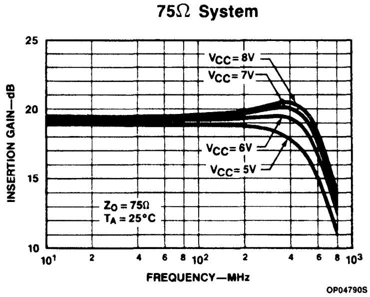
AF03690S

b.

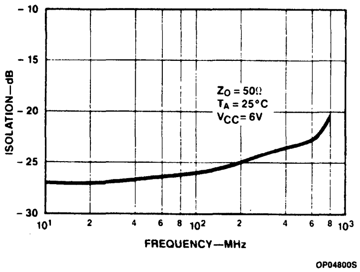
Figure 17



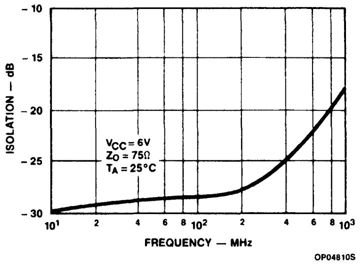
a. Insertion Gain vs Frequency (S_{21})



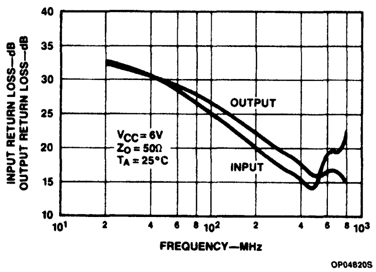
b. Insertion Gain vs Frequency (S_{21})



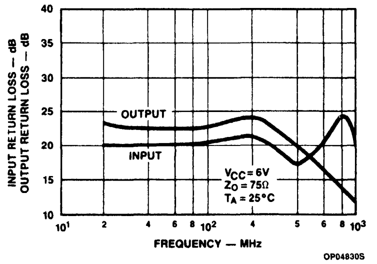
c. Isolation vs Frequency (S_{12})



d. S_{12} Isolation vs Frequency



e. Input (S_{11}) and Output (S_{22}) Return Loss vs Frequency



f. Input (S_{11}) and Output (S_{22}) Return Loss vs Frequency

Figure 18

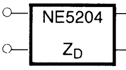
11

Actual S-parameter measurements, using an HP network analyzer (model 8505A) and an HP S-parameter tester (models 8503A/B), are shown in Figure 18.

Values for Figure 20 are measured and specified in the data sheet to ease adaptation and comparison of the NE5204 to other high-frequency amplifiers. The most important parameter is S_{21}. It is defined as the square root of the power gain, and, in decibels, is equal to voltage gain as shown below:

$Z_D = Z_{IN} = Z_{OUT}$ for the NE5204

$P_{IN} = \frac{V_{IN}^2}{Z_D}$



$P_{OUT} = \frac{V_{OUT}^2}{Z_D}$

$$\therefore \frac{P_{OUT}}{P_{IN}} = \frac{\frac{V_{OUT}^2}{Z_D}}{\frac{V_{IN}^2}{Z_D}} = \frac{V_{OUT}^2}{V_{IN}^2} = P_I$$

$P_I = V_I^2$

P_I = Insertion Power Gain
V_I = Insertion Voltage Gain

Measured value for the NE5204 = $|S_{21}|^2 = 100$

$$\therefore P_I = \frac{P_{OUT}}{P_{IN}} = |S_{21}|^2 = 100$$

and $V_I = \frac{V_{OUT}}{V_{IN}} = \sqrt{P_I} = S_{21} = 10$

In decibels:

$P_{I(dB)} = 10\text{Log } |S_{21}|^2 = 20\text{dB}$

$V_{I(dB)} = 20\text{Log } S_{21} = 20\text{dB}$

$\therefore P_{I(dB)} = V_{I(dB)} = S_{21(dB)} = 20\text{dB}$

Also measured on the same system are the respective voltage standing-wave ratios. These are shown in Figure 19. The VSWR can be seen to be below 1.5 across the entire operational frequency range.

Relationships exist between the input and output return losses and the voltage standing wave ratios. These relationships are as follows:

INPUT RETURN LOSS = S_{11}dB
$S_{11}\text{dB} = 20\text{Log } |S_{11}|$

OUTPUT RETURN LOSS = S_{22}dB
$S_{22}\text{dB} = 20\text{Log } |S_{22}|$

$$\text{INPUT VSWR} = \frac{|1 + S_{11}|}{|1 - S_{11}|} \leqslant 1.5$$

$$\text{OUTPUT VSWR} = \frac{|1 + S_{22}|}{|1 - S_{22}|} \leqslant 1.5$$

1dB GAIN COMPRESSION AND SATURATED OUTPUT POWER

The 1dB gain compression is a measurement of the output power level where the small-signal insertion gain magnitude decreases 1dB from its low power value. The decrease is due to non-linearities in the amplifier, an indication of the point of transition between small-signal operation and the large-signal mode.

The saturated output power is a measure of the amplifier's ability to deliver power into an external load. It is the value of the amplifier's output power when the input is heavily overdriven. This includes the sum of the power in all harmonics.

INTERMODULATION INTERCEPT TESTS

The intermodulation intercept is an expression of the low level linearity of the amplifier. The intermodulation ratio is the difference in dB between the fundamental output signal level and the generated distortion product level. The relationship between intercept and intermodulation ratio is illustrated in Figure 20, which shows product output levels plotted versus the level of the fundamental output for two equal strength output signals at different frequencies. The upper line shows the fundamental output plotted against itself with a 1dB to 1dB slope. The second and third order products lie below the fundamentals and exhibit a 2:1 and 3:1 slope, respectively.

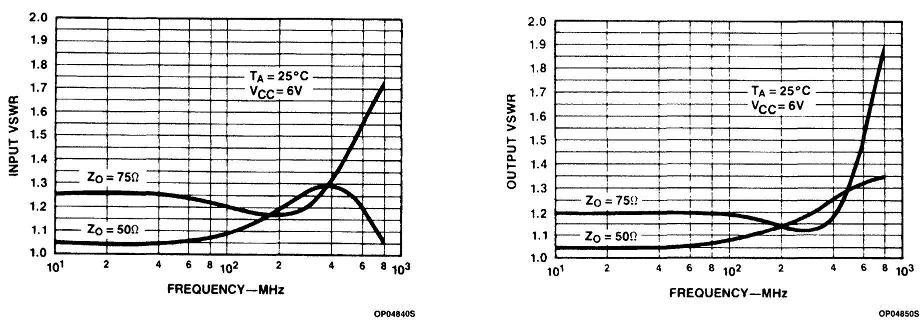
The intercept point for either product is the intersection of the extensions of the product curve with the fundamental output.

The intercept point is determined by measuring the intermodulation ratio at a single output level and projecting along the appropriate product slope to the point of intersection with the fundamental. When the intercept point is known, the intermodulation ratio can be determined by the reverse process. The second-order IMR is equal to the difference between the second-order intercept and the fundamental output level. The third-order IMR is equal to twice the difference between the third-order intercept and the fundamental output level. These are expressed as:

$IP_2 = P_{OUT} + IMR_2$

$IP_3 = P_{OUT} + IMR_3/2$

where P_{OUT} is the power level in dBm of each of a pair of equal level fundamental output signals, IP_2 and IP_3 are the second- and third-order output intercepts in dBm, and IMR_2 and IMR_3 are the second- and third- order intermodulation ratios in dB. The intermodulation intercept is an indicator of intermodulation performance only in the small-signal operating range of the amplifier. Above some output level which is below the 1dB compression point, the active device moves into large-signal operation. At this point, the intermodulation products no longer follow the straight-line output slopes, and the intercept description is no longer valid. It is therefore important to measure IP_2 and IP_3 at output levels well below 1dB compression. One must be care-



a. Input VSWR vs Frequency

b. Output VSWR vs Frequency

Figure 19. Input/Output VSWR vs Frequency

ful, however, not to select levels which are too low, because the test equipment may not be able to recover the signal from the noise. For the NE5204, an output level of −10.5dBm was chosen with fundamental frequencies of 100.000 and 100.01MHz, respectively.

ADDITIONAL READING ON SCATTERING PARAMETERS

For more information regarding S-parameters, please refer to *High-Frequency Amplifiers*; by Ralph S. Carson of the University of Missouri, Rolla, Copyright 1985, published by John Wiley & Sons, Inc.

S-Parameter Techniques for Faster, More Accurate Network Design, HP App Note 95-1, Richard W. Anderson, 1967, HP Journal.

S-Parameter Design, HP App Note 154, 1972.

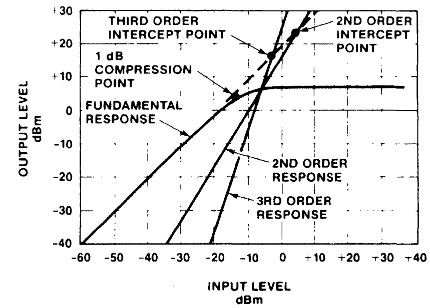


Figure 20

Signetics

NE/SA/SE5205
Wide-band High-Frequency Amplifier

Product Specification

Linear Products

DESCRIPTION

The NE/SA/SE5205 is a high-frequency amplifier with a fixed insertion gain of 20dB. The gain is flat to ± 0.5dB from DC to 450MHz, and the -3dB bandwidth is greater than 600MHz in the EC package. This performance makes the amplifier ideal for cable TV applications. For lower frequency applications, the part is also available in industrial standard dual in-line and small outline packages. The NE/SA/SE5205 operates with a single supply of 6V, and only draws 24mA of supply current, which is much less than comparable hybrid parts. The noise figure is 4.8dB in a 75Ω system and 6dB in a 50Ω system.

Until now, most RF or high-frequency designers had to settle for discrete or hybrid solutions to their amplification problems. Most of these solutions required trade-offs that the designer had to accept in order to use high-frequency gain stages. These include high-power consumption, large component count, transformers, large packages with heat sinks, and high part cost. The NE/SA/SE5205 solves these problems by incorporating a wide-band amplifier on a single monolithic chip.

The part is well matched to 50 or 75Ω input and output impedances. The Standing Wave Ratios in 50 and 75Ω systems do not exceed 1.5 on either the input or output from DC to the -3dB bandwidth limit.

Since the part is a small monolithic IC die, problems such as stray capacitance are minimized. The die size is small enough to fit into a very cost-effective 8-pin small-outline (SO) package to further reduce parasitic effects. A TO-46 metal can is also available that has a case connection for RF grounding which increases the -3dB frequency to 600MHz. The Cerdip package is hermetically sealed, and can operate over the full -55°C to +125°C range.

No external components are needed other than AC coupling capacitors because the NE/SA/SE5205 is internally compensated and matched to 50 and 75Ω. The amplifier has very good distortion specifications, with second and third-order intermodulation intercepts of +24dBm and +17dBm respectively at 100MHz.

The device is ideally suited for 75Ω cable television applications such as decoder boxes, satellite receiver/decoders, and front-end amplifiers for TV receivers. It is also useful for amplified splitters and antenna amplifiers.

The part is matched well for 50Ω test equipment such as signal generators, oscilloscopes, frequency counters and all kinds of signal analyzers. Other applications at 50Ω include mobile radio, CB radio and data/video transmission in fiber optics, as well as broad-band LANs and telecom systems. A gain greater than 20dB can be achieved by cascading additional NE/SA/SE5205s in series as required, without any degradation in amplifier stability.

FEATURES

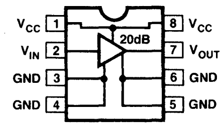
- **600MHz bandwidth**
- **20dB insertion gain**
- **4.8dB (6dB) noise figure $Z_O = 75\Omega$ ($Z_O = 50\Omega$)**
- **No external components required**
- **Input and output impedances matched to 50/75Ω systems**
- **Surface mount package available**
- **MIL-STD processing available**

APPLICATIONS

- **75Ω cable TV decoder boxes**
- **Antenna amplifiers**
- **Amplified splitters**
- **Signal generators**
- **Frequency counters**
- **Oscilloscopes**
- **Signal analyzers**
- **Broad-band LANs**
- **Fiber-optics**
- **Modems**
- **Mobile radio**
- **Security systems**
- **Telecommunications**

PIN CONFIGURATIONS

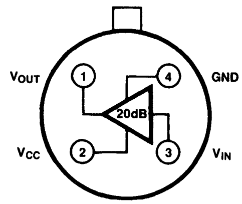
N, FE, D Packages



TOP VIEW

CD12760S

EC Package



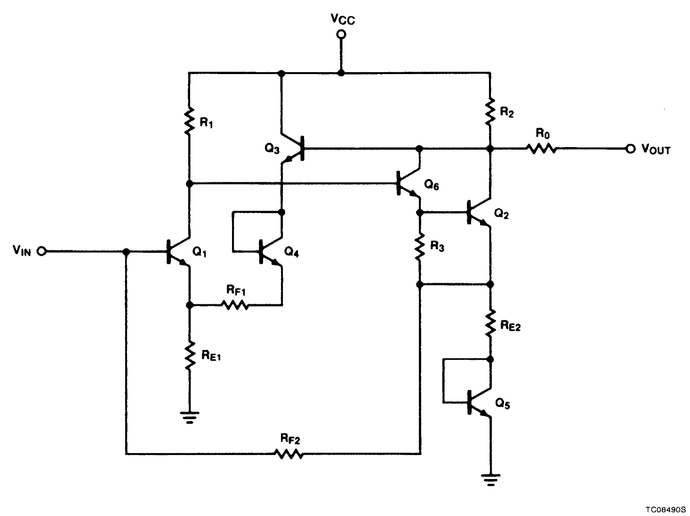
CD09850S

NOTE:
Tab denotes Pin 1.

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
8-Pin Plastic SO	0 to +70°C	NE5205D
4-Pin Metal can	0 to +70°C	NE5205EC
8-Pin Cerdip	0 to +70°C	NE5205FE
8-Pin Plastic DIP	0 to +70°C	NE5205N
8-Pin Plastic SO	−40°C to +85°C	SA5205D
8-Pin Plastic DIP	−40°C to +85°C	SA5205N
8-Pin Cerdip	−40°C to +85°C	SA5205FE
8-Pin Cerdip	−55°C to +125°C	SE5205FE
8-Pin Plastic DIP	−55°C to +125°C	SE5205N

EQUIVALENT SCHEMATIC



ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{CC}	Supply voltage	9	V
V_{AC}	AC input voltage	5	V_{P-P}
T_A	Operating ambient temperature range NE grade SA grade SE grade	 0 to +70 -40 to +85 -55 to +125	 °C °C °C
P_{DMAX}	Maximum power dissipation, T_A = 25°C (still-air)[1, 2] FE package N package D package EC package	 780 1160 780 1250	 mW mW mW mW

NOTES:
1. Derate above 25°C, at the following rates:
   FE package at 6.2mW/°C
   N package at 9.3mW/°C
   D package at 6.2mW/°C
   EC package at 10.0mW/°C
2. See "Power Dissipation Considerations" section.

DC ELECTRICAL CHARACTERISTICS at V_{CC} = 6V, $Z_S = Z_L = Z_O = 50\Omega$ and T_A = 25°C, in all packages, unless otherwise specified.

SYMBOL	PARAMETER	TEST CONDITIONS	SE5205			NE/SA5205			UNIT
			Min	Typ	Max	Min	Typ	Max	
	Operating supply voltage range	Over temperature	5 5		6.5 6.5	5 5		8 8	V V
I_{CC}	Supply current	Over temperature	20 19	24	30 31	20 19	24	30 31	mA mA
S21	Insertion gain	f = 100MHz Over temperature	17 16.5	19	21 21.5	17 16.5	19	21 21.5	dB
S11	Input return loss	f = 100MHz D, N, FE		25			25		dB
		DC – f_{MAX} D, N, FE	12			12			dB
S11	Input return loss	f = 100MHz EC package					24		dB
		DC – f_{MAX} EC				10			dB
S22	Output return loss	f = 100MHz D, N, FE		27			27		dB
		DC – f_{MAX}	12			12			dB
S22	Output return loss	f = 100MHz EC package					26		dB
		DC – F_{MAX}				10			dB
S12	Isolation	f = 100MHz		-25			-25		dB
		DC – f_{MAX}	-18			-18			dB
t_R	Rise time			5			5		ps
	Propagation delay			5			5		ps

DC ELECTRICAL CHARACTERISTICS at $V_{CC} = 6V$, $Z_S = Z_L = Z_O = 50\Omega$ and $T_A = 25°C$, in all packages, unless otherwise specified.

SYMBOL	PARAMETER	TEST CONDITIONS	SE5205			NE/SA5205			UNIT
			Min	Typ	Max	Min	Typ	Max	
BW	Bandwidth	±0.5dB D, N					450		MHz
f_{MAX}	Bandwidth	±0.5dB EC					500		MHz
f_{MAX}	Bandwidth	±0.5dB FE		300			300		MHz
f_{MAX}	Bandwidth	−3dB D, N				550			MHz
f_{MAX}	Bandwidth	−3dB EC				600			MHz
f_{MAX}	Bandwidth	−3dB FE	400			400			MHz
	Noise figure (75Ω)	f = 100MHz		4.8			4.8		dB
	Noise figure (50Ω)	f = 100MHz		6.0			6.0		dB
	Saturated output power	f = 100MHz		+7.0			+7.0		dBm
	1dB gain compression	f = 100MHz		+4.0			+4.0		dBm
	Third-order intermodulation intercept (output)	f = 100MHz		+17			+17		dBm
	Second-order intermodulation intercept (output)	f = 100MHz		+24			+24		dBm

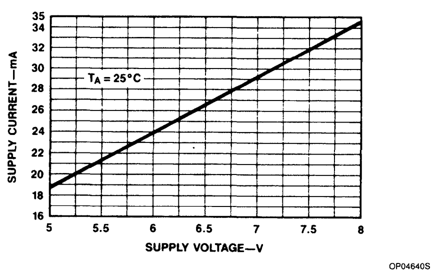


Figure 1. Supply Current vs Supply Voltage

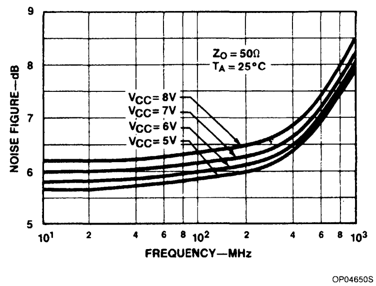


Figure 2. Noise Figure vs Frequency

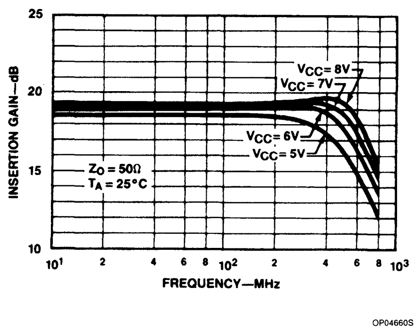


Figure 3. Insertion Gain vs Frequency (S_{21})

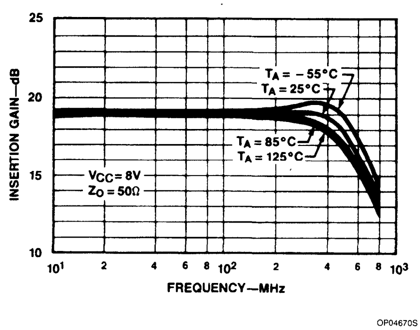


Figure 4. Insertion Gain vs Frequency (S_{21})

Wide-band High-Frequency Amplifier

NE/SA/SE5205

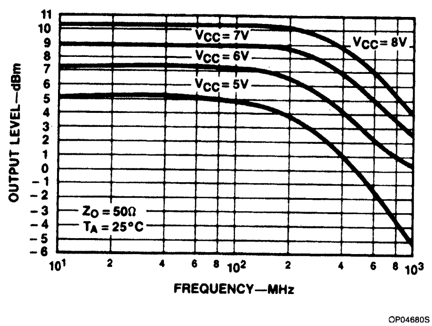


Figure 5. Saturated Output Power vs Frequency

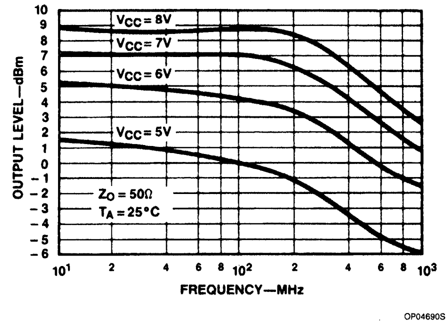


Figure 6. 1dB Gain Compression vs Frequency

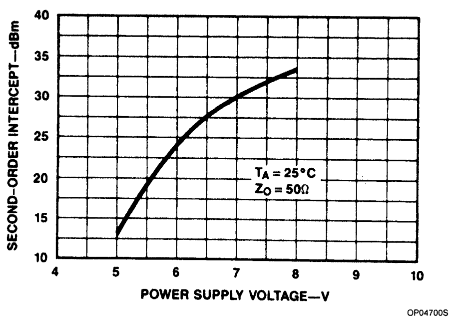


Figure 7. Second-Order Output Intercept vs Supply Voltage

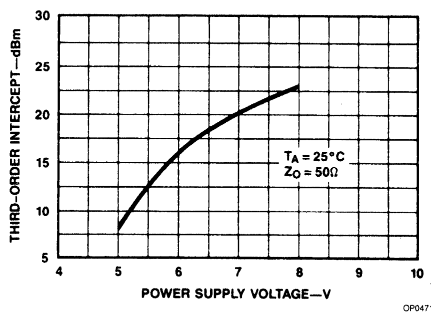


Figure 8. Third-Order Intercept vs Supply Voltage

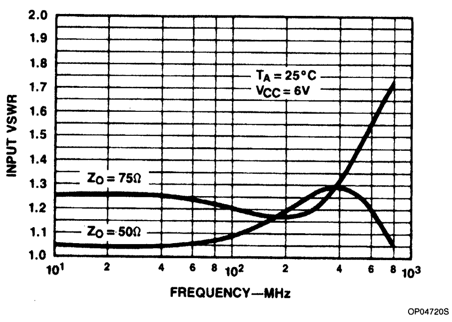


Figure 9. Input VSWR vs Frequency

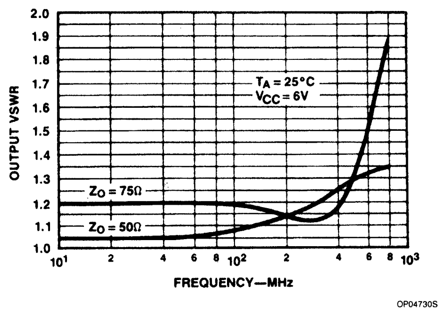


Figure 10. Output VSWR vs Frequency

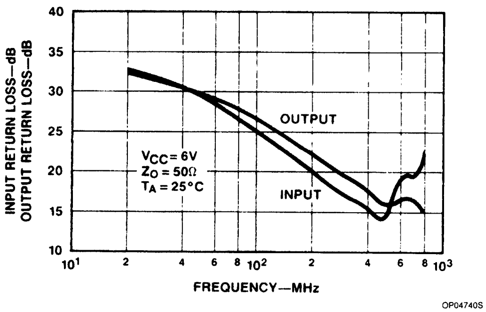


Figure 11. Input (S_{11}) and Output (S_{22}) Return Loss vs Frequency

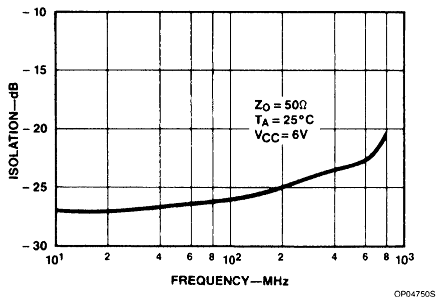


Figure 12. Isolation vs Frequency (S_{12})

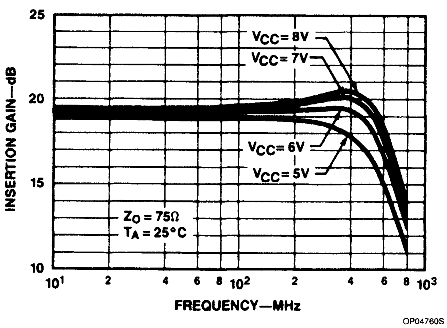


Figure 13. Insertion Gain vs Frequency (S_{21})

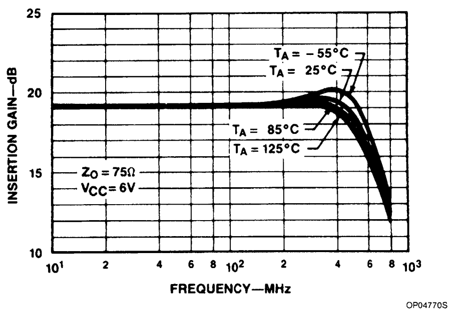


Figure 14. Insertion Gain vs Frequency (S_{21})

11

THEORY OF OPERATION

The design is based on the use of multiple feedback loops to provide wide-band gain together with good noise figure and terminal impedance matches. Referring to the circuit schematic in Figure 15, the gain is set primarily by the equation:

$$\frac{V_{OUT}}{V_{IN}} = (R_{F1} + R_{E1})/R_{E1} \qquad (1)$$

which is series-shunt feedback. There is also shunt-series feedback due to R_{F2} and R_{E2} which aids in producing wideband terminal impedances without the need for low value input shunting resistors that would degrade the noise figure. For optimum noise performance, R_{E1} and the base resistance of Q_1 are kept as low as possible while R_{F2} is maximized.

The noise figure is given by the following equation:

$$NF = 10 \text{ Log} \left\{ 1 + \frac{\left[r_b + R_{E1} + \frac{KT}{2qI_{C1}} \right]}{R_O} \right\} \text{dB} \qquad (2)$$

where $I_{C1} = 5.5mA$, $R_{E1} = 12\Omega$, $r_b = 130\Omega$, $KT/q = 26mV$ at 25°C and $R_0 = 50$ for a 50Ω system and 75 for a 75Ω system.

The DC input voltage level V_{IN} can be determined by the equation:

$$V_{IN} = V_{BE1} + (I_{C1} + I_{C3})\ R_{E1}$$

where $R_{E1} = 12\Omega$, $V_{BE} = 0.8V$, $I_{C1} = 5mA$ and $I_{C3} = 7mA$ (currents rated at $V_{CC} = 6V$).

Under the above conditions, V_{IN} is approximately equal to 1V.

Level shifting is achieved by emitter-follower Q_3 and diode Q_4 which provide shunt feedback to the emitter of Q_1 via R_{F1}. The use of an emitter-follower buffer in this feedback loop essentially eliminates problems of shunt feedback loading on the output. The value of $R_{F1} = 140\Omega$ is chosen to give the desired nominal gain. The DC output voltage V_{OUT} can be determined by:

$$V_{OUT} = V_{CC} - (I_{C2} + I_{C6})R2, \qquad (4)$$

where $V_{CC} = 6V$, $R_2 = 225\Omega$, $I_{C2} = 7mA$ and $I_{C6} = 5mA$.

From here it can be seen that the output voltage is approximately 3.3V to give relatively equal positive and negative output swings. Diode Q_5 is included for bias purposes to allow direct coupling of R_{F2} to the base of Q_1. The dual feedback loops stabilize the DC operating point of the amplifier.

The output stage is a Darlington pair (Q_6 and Q_2) which increases the DC bias voltage on the input stage (Q_1) to a more desirable value, and also increases the feedback loop gain. Resistor R_0 optimizes the output VSWR (Voltage Standing Wave Ratio). Inductors L_1 and L_2 are bondwire and lead inductances which are roughly 3nH. These improve the high-frequency impedance matches at input and output by partially resonating with 0.5pF of pad and package capacitance.

POWER DISSIPATION CONSIDERATIONS

When using the part at elevated temperature, the engineer should consider the power dissipation capabilities of each package.

At the nominal supply voltage of 6V, the typical supply current is 25mA (30mA Max). For operation at supply voltages other than 6V, see Figure 1 for I_{CC} versus V_{CC} curves. The supply current is inversely proportional to temperature and varies no more than 1mA between 25°C and either temperature extreme. The change is 0.1% per °C over the range.

The recommended operating temperature ranges are air-mount specifications. Better heat sinking benefits can be realized by mounting the D and EC package body against the PC board plane.

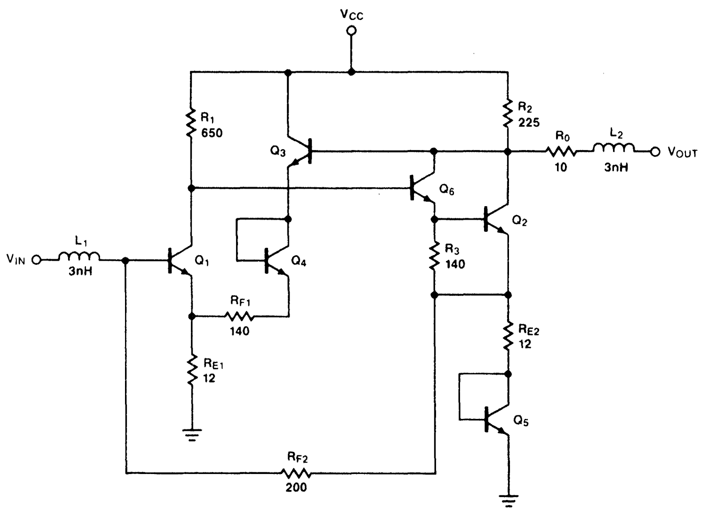


Figure 15. Schematic Diagram

Wide-band High-Frequency Amplifier NE/SA/SE5205

PC BOARD MOUNTING

In order to realize satisfactory mounting of the NE5205 to a PC board, certain techniques need to be utilized. The board must be double-sided with copper and all pins must be soldered to their respective areas (i.e., all GND and V_{CC} pins on the SO package). In addition, if the EC package is used, the case should be soldered to the ground plane. The power supply should be decoupled with a capacitor as close to the V_{CC} pins as possible and an RF choke should be inserted between the supply and the device. Caution should be exercised in the connection of input and output pins. Standard microstrip should be observed wherever possible. There should be no solder bumps or burrs or any obstructions in the signal path to cause launching problems. The path should be as straight as possible and lead lengths as short as possible from the part to the cable connection. Another important consideration is that the input and output should be AC coupled. This is because at V_{CC} = 6V, the input is approximately at 1V while the output is at 3.3V. The output must be decoupled into a low impedance system or the DC bias on the output of the amplifier will be loaded down causing loss of output power. The easiest way to decouple the entire amplifier is by soldering a high frequency chip capacitor directly to the input and output pins of the device. This circuit is shown in Figure 16. Follow these recommendations to get the best frequency response and noise immunity. The board design is as important as the integrated circuit design itself.

SCATTERING PARAMETERS

The primary specifications for the NE/SA/SE5205 are listed as S-parameters. S-parameters are measurements of incident and reflected currents and voltages between the source, amplifier and load as well as transmission losses. The parameters for a two-port network are defined in Figure 17.

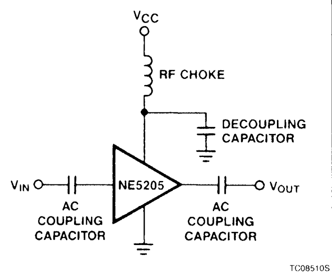


Figure 16. Circuit Schematic for Coupling and Power Supply Decoupling

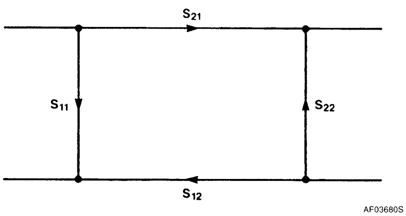


Figure 17a. Two-Port Network Defined

S_{11} — INPUT RETURN LOSS

$$S_{11} \equiv \sqrt{\frac{\text{POWER REFLECTED FROM INPUT PORT}}{\text{POWER AVAILABLE FROM GENERATOR AT INPUT PORT}}}$$

S_{12} — REVERSE TRANSMISSION LOSS OR ISOLATION

$$S_{12} \equiv \sqrt{\text{REVERSE TRANSDUCER POWER GAIN}}$$

S_{21} — FORWARD TRANSMISSION LOSS OR INSERTION GAIN

$$S_{21} \equiv \sqrt{\text{TRANSDUCER POWER GAIN}}$$

S_{22} — OUTPUT RETURN LOSS

$$S_{22} \equiv \sqrt{\frac{\text{POWER REFLECTED FROM OUTPUT PORT}}{\text{POWER AVAILABLE FROM GENERATOR AT OUTPUT PORT}}}$$

AF03690S

Figure 17b

Actual S-parameter measurements using an HP network analyzer (model 8505A) and an HP S-parameter tester (models 8503A/B) are shown in Figure 18.

Values for the figures below are measured and specified in the data sheet to ease adaptation and comparison of the NE/SA/SE5205 to other high-frequency amplifiers.

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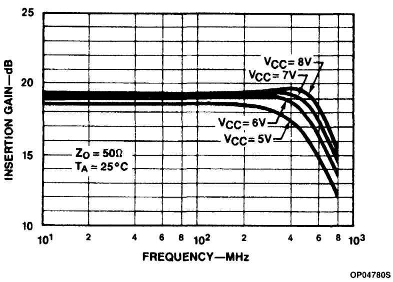
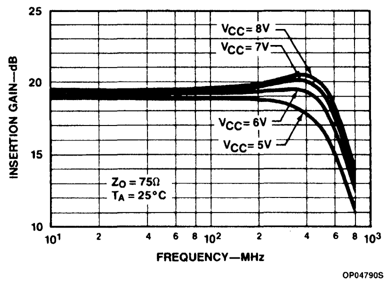
Wide-band High-Frequency Amplifier NE/SA/SE5205

50Ω System

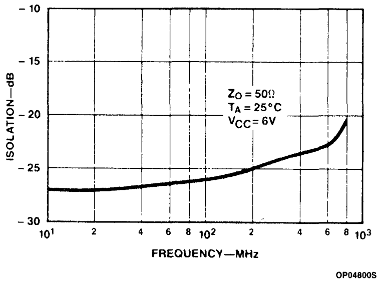


a. Insertion Gain vs Frequency (S_{21})

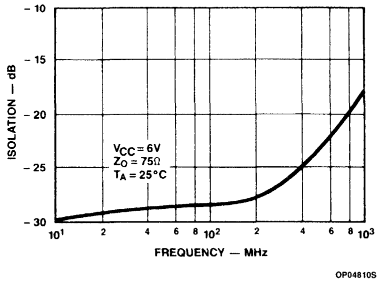
75Ω System



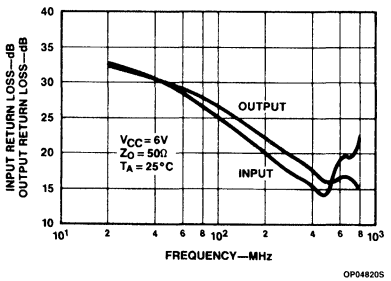
b. Insertion Gain vs Frequency (S_{21})



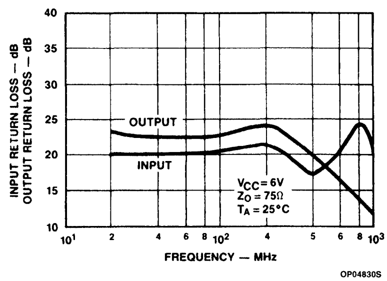
c. Isolation vs Frequency (S_{12})



d. S_{12} Isolation vs Frequency



e. Input (S_{11}) and Output (S_{22}) Return Loss vs Frequency



f. Input (S_{11}) and Output (S_{22}) Return Loss vs Frequency

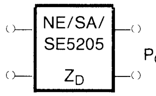
Figure 18

Wide-band High-Frequency Amplifier NE/SA/SE5205

The most important parameter is S_{21}. It is defined as the square root of the power gain, and, in decibels, is equal to voltage gain as shown below:

$Z_D = Z_{IN} = Z_{OUT}$ for the NE/SA/SE5205

$$P_{IN} = \frac{V_{IN}{}^2}{Z_D} \qquad P_{OUT} = \frac{V_{OUT}{}^2}{Z_D}$$



$$\therefore \frac{P_{OUT}}{P_{IN}} = \frac{\frac{V_{OUT}{}^2}{Z_D}}{\frac{V_{IN}{}^2}{Z_D}} = \frac{V_{OUT}{}^2}{V_{IN}{}^2} = P_I$$

$P_I = V_I{}^2$

P_I = Insertion Power Gain

V_I = Insertion Voltage Gain

Measured value for the NE/SA/SE5205 = $|S_{21}|^2 = 100$

$$\therefore P_I = \frac{P_{OUT}}{P_{IN}} = |S_{21}|^2 = 100$$

$$\text{and } V_I = \frac{V_{OUT}}{V_{IN}} = \sqrt{P_I} = S_{21} = 10$$

In decibels:

$P_{I(dB)} = 10 \text{ Log } |S_{21}|^2 = 20\text{dB}$

$V_{I(dB)} = 20 \text{ Log } S_{21} = 20\text{dB}$

$\therefore P_{I(dB)} = V_{I(dB)} = S_{21(dB)} = 20\text{dB}$

Also measured on the same system are the respective voltage standing wave ratios. These are shown in Figure 19. The VSWR can be seen to be below 1.5 across the entire operational frequency range.

Relationships exist between the input and output return losses and the voltage standing wave ratios. These relationships are as follows:

INPUT RETURN LOSS = S_{11}dB
$S_{11}\text{dB} = 20 \text{ Log } |S_{11}|$

OUTPUT RETURN LOSS = S_{22}dB
$S_{22}\text{dB} = 20 \text{ Log } |S_{22}|$

$$\text{INPUT VSWR} = \frac{|1 + S_{11}|}{|1 - S_{11}|} \leqslant 1.5$$

$$\text{OUTPUT VSWR} = \frac{|1 + S_{22}|}{|1 - S_{22}|} \leqslant 1.5$$

1dB GAIN COMPRESSION AND SATURATED OUTPUT POWER

The 1dB gain compression is a measurement of the output power level where the small-signal insertion gain magnitude decreases 1dB from its low power value. The decrease is due to nonlinearities in the amplifier, an indication of the point of transition between small-signal operation and the large signal mode.

The saturated output power is a measure of the amplifier's ability to deliver power into an external load. It is the value of the amplifier's output power when the input is heavily overdriven. This includes the sum of the power in all harmonics.

INTERMODULATION INTERCEPT TESTS

The intermodulation intercept is an expression of the low level linearity of the amplifier. The intermodulation ratio is the difference in dB between the fundamental output signal level and the generated distortion product level. The relationship between intercept and intermodulation ratio is illustrated in Figure 20, which shows product output levels plotted versus the level of the fundamental output for two equal strength output signals at different frequencies. The upper line shows the fundamental output plotted against itself with a 1dB to 1dB slope. The second and third order products lie below the fundamentals and exhibit a 2:1 and 3:1 slope, respectively.

The intercept point for either product is the intersection of the extensions of the product curve with the fundamental output.

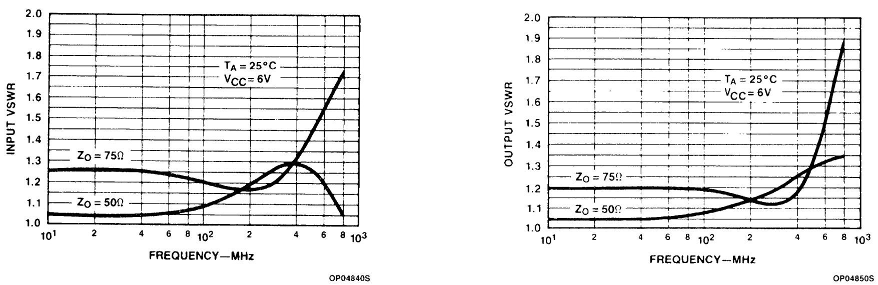
The intercept point is determined by measuring the intermodulation ratio at a single output level and projecting along the appropriate product slope to the point of intersection with the fundamental. When the intercept point is known, the intermodulation ratio can be determined by the reverse process. The second order IMR is equal to the difference between the second order intercept and the fundamental output level. The third order IMR is equal to twice the difference between the third order intercept and the fundamental output level. These are expressed as:

$IP_2 = P_{OUT} + IMR_2$

$IP_3 = P_{OUT} + IMR_3/2$

where P_{OUT} is the power level in dBm of each of a pair of equal level fundamental output signals, IP_2 and IP_3 are the second and third order output intercepts in dBm, and IMR_2 and IMR_3 are the second and third order intermodulation ratios in dB. The intermodulation intercept is an indicator of intermodulation performance only in the small signal operating range of the amplifier. Above some output level which is below the 1dB compression point, the active device moves into large-signal operation. At this point the intermodulation products no longer follow the straight line output slopes, and the intercept description is no longer valid. It is therefore important to measure IP_2 and IP_3 at output levels well below 1dB compression. One must be careful, however, not to select too low levels because the test equipment may not be able to recover the signal from the noise. For the NE/SA/SE5205 we have chosen an output level of −10.5dBm with fundamental frequencies of 100.000 and 100.01MHz, respectively.

11



a. Input VSWR vs Frequency **b. Output VSWR vs Frequency**

Figure 19. Input/Output VSWR vs Frequency

ADDITIONAL READING ON SCATTERING PARAMETERS

For more information regarding S-parameters, please refer to *High-Frequency Amplifiers* by Ralph S. Carson of the University of Missouri, Rolla, Copyright 1985; published by John Wiley & Sons, Inc.

"S-Parameter Techniques for Faster, More Accurate Network Design", HP App Note 95-1, Richard W. Anderson, 1967, HP Journal.

"S-Parameter Design", HP App Note 154, 1972.

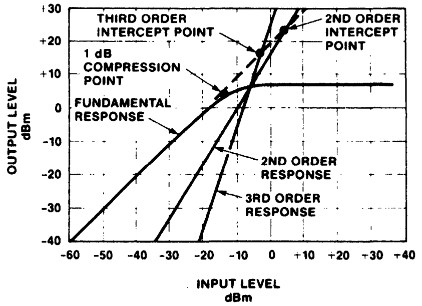


Figure 20

Signetics

NE/SE5539
High Frequency Operational Amplifier

Product Specification

Linear Products

DESCRIPTION

The NE/SE5539 is a very wide bandwidth, high slew rate, monolithic operational amplifier for use in video amplifiers, RF amplifiers, and extremely high slew rate amplifiers.

Emitter-follower inputs provide a true differential high input impedance device. Proper external compensation will allow design operation over a wide range of closed-loop gains, both inverting and non-inverting, to meet specific design requirements.

FEATURES

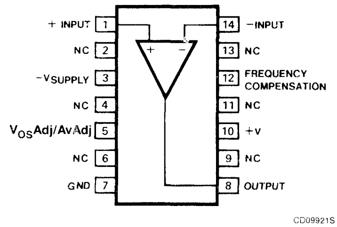
- **Bandwidth**
  - **Unity gain - 350MHz**
  - **Full power - 48MHz**
  - **GBW - 1.2 GHz at 17dB**
- **Slew rate: 600/Vμs**
- **A_{VOL}: 52dB typical**
- **Low noise - 4nV/$\sqrt{Hz}$ typical**
- **MIL-STD processing available**

APPLICATIONS

- **High speed datacomm**
- **Video monitors & TV**
- **Satellite communications**
- **Image processing**
- **RF instrumentation & oscillators**
- **Magnetic storage**
- **Military communications**

PIN CONFIGURATION

D, F, N Packages



TOP VIEW

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
14-Pin Plastic DIP	0 to +70°C	NE5539N
14-Pin Plastic SO	0 to +70°C	NE5539D
14-Pin Cerdip	0 to +70°C	NE5539F
14-Pin Plastic DIP	-55°C to +125°C	SE5539N
14-Pin Cerdip	-55°C to +125°C	SE5539F

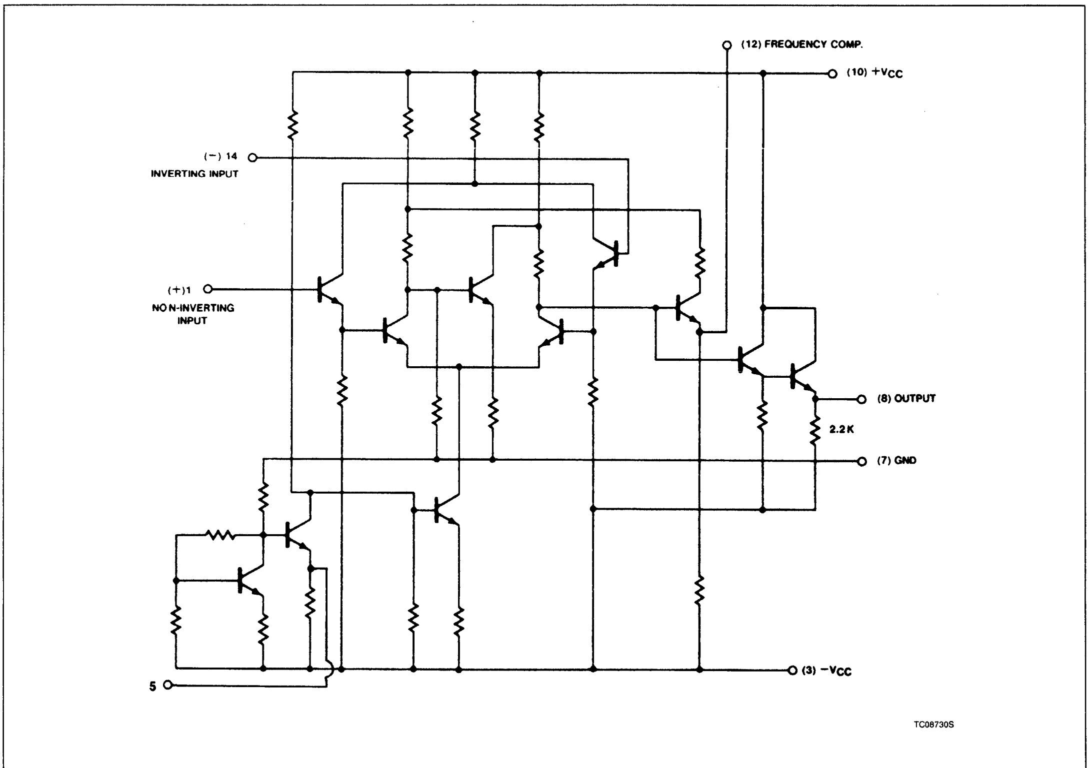
ABSOLUTE MAXIMUM RATINGS[1]

SYMBOL	PARAMETER	RATING	UNIT
V_{CC}	Supply voltage	±12	V
P_{DMAX}	Maximum power dissipation, T_A = 25°C (still-air)[2] F package N package D package	 1.17 1.45 0.99	 W W W
T_{STG}	Storage temperature range	-65 to +150	°C
T_J	Max junction temperature	150	°C
T_A	Operating temperature range NE SE	 0 to 70 -55 to +125	 °C °C
T_{SOLD}	Lead temperature (10sec max)	300	°C

NOTES:

1. Differential input voltage should not exceed 0.25V to prevent excessive input bias current and common-mode voltage 2.5V. These voltage limits may be exceeded if current is limited to less than 10mA.
2. Derate above 25°C, at the following rates:
   F package at 9.3 mW/°C
   N package at 11.6 mW/°C
   D package at 7.9 mW/°C

EQUIVALENT CIRCUIT



DC ELECTRICAL CHARACTERISTICS $V_{CC} = \pm 8V$, $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	TEST CONDITIONS		SE5539			NE5539			UNIT
				Min	Typ	Max	Min	Typ	Max	
V_{OS}	Input offset voltage	$V_O = 0V$, $R_S = 100\Omega$	Over temp		2	5				mV
			$T_A = 25°C$		2	3		2.5	5	
	$\Delta V_{OS}/\Delta T$				5			5		µV/°C
I_{OS}	Input offset current		Over temp		0.1	3				µA
			$T_A = 25°C$		0.1	1			2	
	$\Delta I_{OS}/\Delta T$				0.5			0.5		nA/°C
I_B	Input bias current		Over temp		6	25				µA
			$T_A = 25°C$		5	13		5	20	
	$\Delta I_B/\Delta T$				10			10		nA/°C
CMRR	Common-mode rejection ratio	F = 1kHz, $R_S = 100\Omega$, V_{CM} ±1.7V		70	80		70	80		dB
			Over temp	70	80					dB
R_{IN}	Input impedance				100			100		kΩ
R_{OUT}	Output impedance				10			10		Ω

DC ELECTRICAL CHARACTERISTICS (Continued) $V_{CC} = \pm 8V$, $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	TEST CONDITIONS			SE5539			NE5539			UNIT
					Min	Typ	Max	Min	Typ	Max	
V_{OUT}	Output voltage swing	$R_L = 150\Omega$ to GND and 470Ω to $-V_{CC}$		+Swing				+2.3	+2.7		V
				−Swing				−1.7	−2.2		
V_{OUT}	Output voltage swing	$R_L = 2k\Omega$ to GND	Over temp	+Swing	+2.3	+3.0					V
				−Swing	−1.5	−2.1					
			$T_A = 25°C$	+Swing	+2.5	+3.1					V
				−Swing	−2.0	−2.7					
$I_{CC}+$	Positive supply current	$V_O = 0$, $R_1 = \infty$		Over temp		14	18				mA
				$T_A = 25°C$		14	17		14	18	
$I_{CC}-$	Negative supply current	$V_O = 0$, $R_1 = \infty$		Over temp		11	15				mA
				$T_A = 25°C$		11	14		11	15	
PSRR	Power supply rejection ratio	$\Delta V_{CC} = \pm 1V$		Over temp		300	1000				$\mu V/V$
				$T_A = 25°C$					200	1000	
A_{VOL}	Large signal voltage gain	$V_O = +2.3V, -1.7V$ $R_L = 150\Omega$ to GND, 470Ω to $-V_{CC}$						47	52	57	dB
A_{VOL}	Large signal voltage gain	$V_O = +2.3V, -1.7V$ $R_L = 2\Omega$ to GND									dB
				$T_A = 25°C$				47	52	57	
A_{VOL}	Large signal voltage gain	$V_O = +2.5V, -2.0V$ $R_L = 2k\Omega$ to GND		Over temp	46		60				dB
				$T_A = 25°C$	48	53	58				

DC ELECTRICAL CHARACTERISTICS $V_{CC} = \pm 6V$, $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	TEST CONDITIONS			SE5539			UNIT
					Min	Typ	Max	
V_{OS}	Input offset voltage			Over temp		2	5	mV
				$T_A = 25°C$		2	3	
I_{OS}	Input offset current			Over temp		0.1	3	μA
				$T_A = 25°C$		0.1	1	
I_B	Input bias current			Over temp		5	20	μA
				$T_A = 25°C$		4	10	
CMRR	Common-mode rejection ratio	$V_{CM} = \pm 1.3V$, $R_S = 100\Omega$			70	85		dB
$I_{CC}+$	Positive supply current			Over temp		11	14	mA
				$T_A = 25°C$		11	13	
$I_{CC}-$	Negative supply current			Over temp		8	11	mA
				$T_A = 25°C$		8	10	
PSRR	Power supply rejection ratio	$\Delta V_{CC} = \pm 1V$		Over temp		300	1000	$\mu V/V$
				$T_A = 25°C$				
V_{OUT}	Output voltage swing	$R_L = 150\Omega$ to GND and 390Ω to $-V_{CC}$	Over temp	+Swing	+1.4	+2.0		V
				−Swing	−1.1	−1.7		
			$T_A = 25°C$	+Swing	+1.5	+2.0		
				−Swing	−1.4	−1.8		

AC ELECTRICAL CHARACTERISTICS $V_{CC} = \pm 8V$, $R_L = 150\Omega$ to GND & 470Ω to $-V_{CC}$, unless otherwise specified.

SYMBOL	PARAMETER	TEST CONDITIONS	SE5539			NE5539			UNIT
			Min	Typ	Max	Min	Typ	Max	
BW	Gain bandwidth product	$A_{CL} = 7$, $V_O = 0.1\ V_{P-P}$		1200			1200		MHz
	Small-signal bandwidth	$A_{CL} = 2$, $R_L = 150\Omega^1$		110			110		MHz
t_S	Settling time	$A_{CL} = 2$, $R_L = 150\Omega^1$		15			15		ns
SR	Slew rate	$A_{CL} = 2$, $R_L = 150\Omega^1$		600			600		V/µs
t_{PD}	Propagation delay	$A_{CL} = 2$, $R_L = 150\Omega^1$		7			7		ns
	Full power response	$A_{CL} = 2$, $R_L = 150\Omega^1$		48			48		MHz
	Full power response	$A_V = 7$, $R_L = 150\Omega^1$		20			20		MHz
	Input noise voltage	$R_S = 50\Omega$, 1MHz		4			4		$nV/\sqrt{Hz}$
	Input noise current	1MHz		6			6		$pA/\sqrt{Hz}$

NOTE:
1. External compensation.

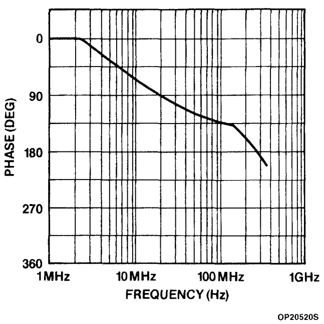
AC ELECTRICAL CHARACTERISTICS $V_{CC} = \pm 6V$, $R_L = 150\Omega$ to GND and 390Ω to $-V_{CC}$, unless otherwise specified.

SYMBOL	PARAMETER	TEST CONDITIONS	SE5539			UNIT
			Min	Typ	Max	
BW	Gain bandwidth product	$A_{CL} = 7$		700		MHz
	Small-signal bandwidth	$A_{CL} = 2^1$		120		MHz
t_S	Settling time	$A_{CL} = 2^1$		23		ns
SR	Slew rate	$A_{CL} = 2^1$		330		V/µs
t_{PD}	Propagation delay	$A_{CL} = 2^1$		4.5		ns
	Full power response	$A_{CL} = 2^1$		20		MHz

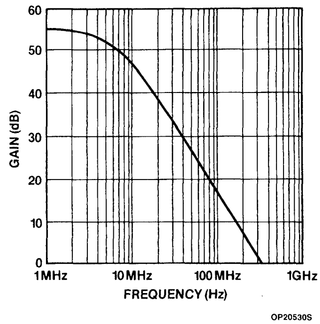
NOTE:
1. External compensation.

TYPICAL PERFORMANCE CURVES

NE5539 Open-Loop Phase

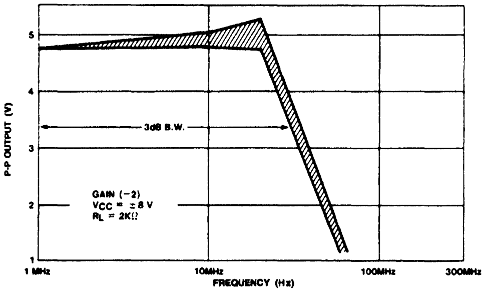


NE5539 Open-Loop Gain



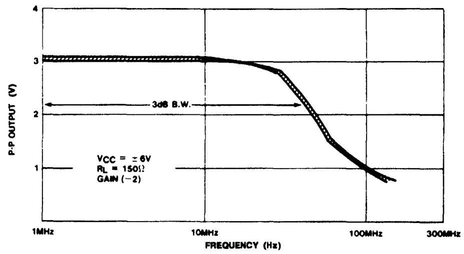
TYPICAL PERFORMANCE CURVES (Continued)

Power Bandwidth (SE)



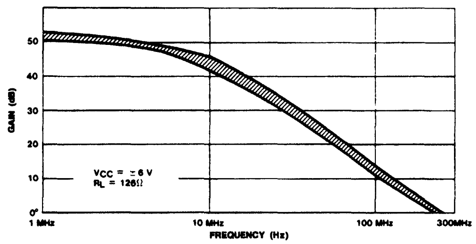
OP05202S

Power Bandwidth (NE)



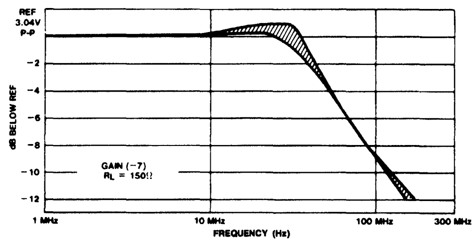
OP05212S

SE5539 Open-Loop Gain vs Frequency



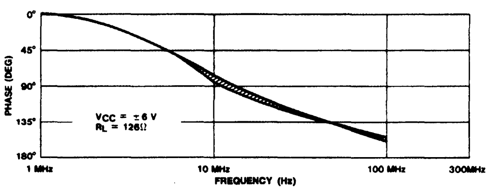
OP05222S

Power Bandwidth

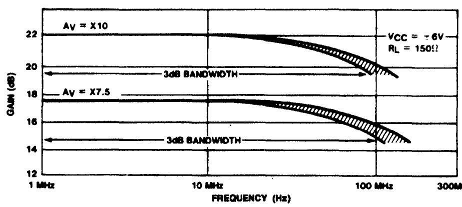


OP05231S

SE5539 Open-Loop Phase vs Frequency



Gain Bandwidth Product vs Frequency



OP05251S

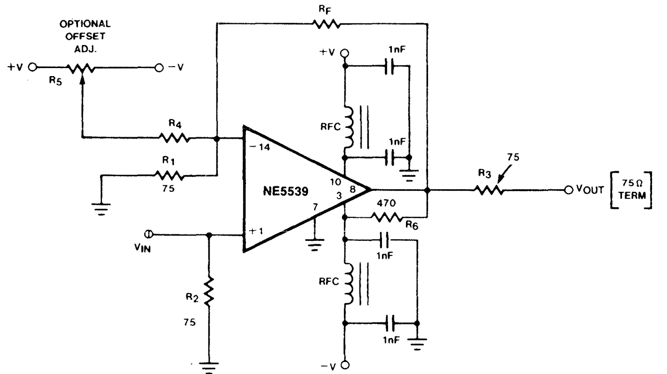
NOTE


Indicates typical distribution $-55°C \leq T_A \leq 125°C$

OP05241S

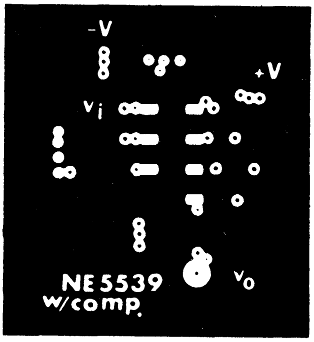
CIRCUIT LAYOUT CONSIDERATIONS

As may be expected for an ultra-high frequency, wide-gain bandwidth amplifier, the physical circuit layout is extremely critical. Breadboarding is not recommended. A double-sided copper-clad printed circuit board will result in more favorable system operation. An example utilizing a 28dB non-inverting amp is shown in Figure 1.



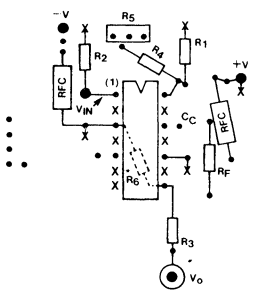
NOTES:
$R_1 = 75\Omega$ 5% CARBON
$R_2 = 75\Omega$ 5% CARBON
$R_3 = 75\Omega$ 5% CARBON
$R_4 = 36k$ 5% CARBON
$R_5 = 20k$ TRIMPOT (CERMET)
$R_F = 1.5k$ (28dB GAIN)
$R_6 = 470\Omega$ 5% CARBON
RFC 3T # 26 BUSS WIRE ON FERROXCUBE VK 200 09/3B CORE
BYPASS CAPACITORS 1nF CERAMIC (MEPCO OR EQUIV.)

Top Plane Copper[1] (Component Side)



DF05910S

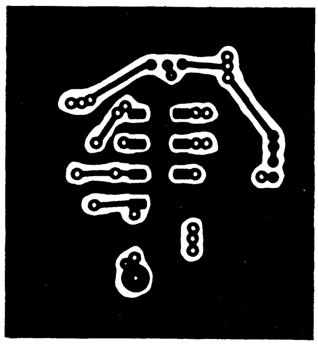
Component Side (Component Layout)



DF05920S

NOTES:
(X) indicates ground connection to top plane.
*R6 is on bottom side.

Bottom Plane Copper[1]



DF05930S

NOTE:
1. Bond edges of top and bottom ground plane copper.

Figure 1. 28dB Non-Inverting Amp Sample PC Layout

NE5539 COLOR VIDEO AMPLIFIER

The NE5539 wideband operational amplifier is easily adapted for use as a color video amplifier. A typical circuit is shown in Figure 2 along with vector-scope[1] photographs showing the amplifier differential gain and phase response to a standard five-step modulated staircase linearity signal (Figures 3, 4 and 5). As can be seen in Figure 4, the gain varies less than 0.5% from the bottom to the top of the staircase. The maximum differential phase shown in Figure 5 is approximately +0.1°.

The amplifier circuit was optimized for a 75Ω input and output termination impedance with a gain of approximately 10 (20dB).

NOTE:

1. The input signal was 200mV and the output 2V. V_{CC} was ±8V.

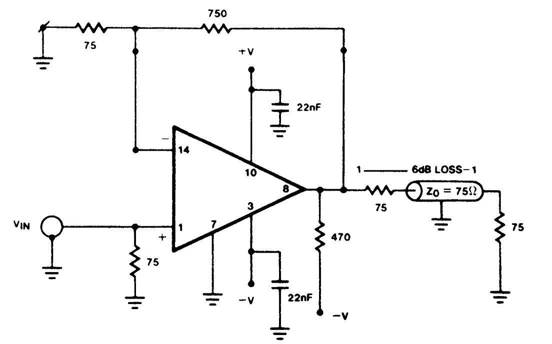
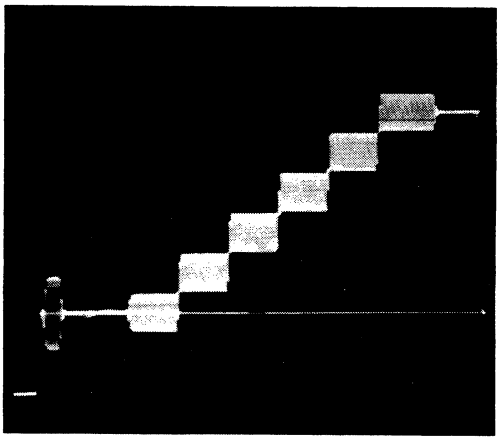
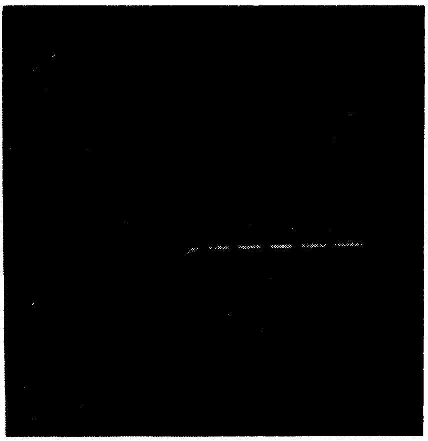


Figure 2. NE5539 Video Amplifier



DF05940S

Figure 3. Input Signal



DF05950S

Figure 4. Differential Gain < 0.5%

NOTE:
Instruments used for these measurements were Tektronix 146 NTSC test signal generator, 520A NTSC vectorscope, and 1480 waveform monitor.

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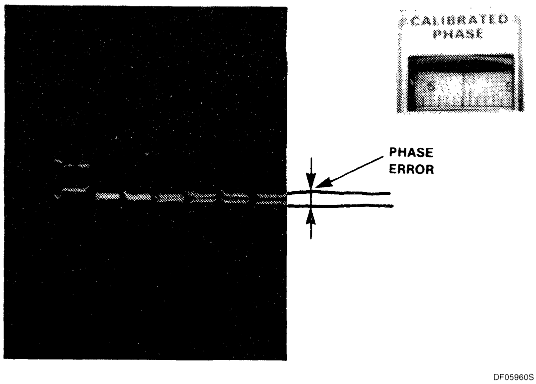


Figure 5. Differential Phase +0.1°

APPLICATIONS

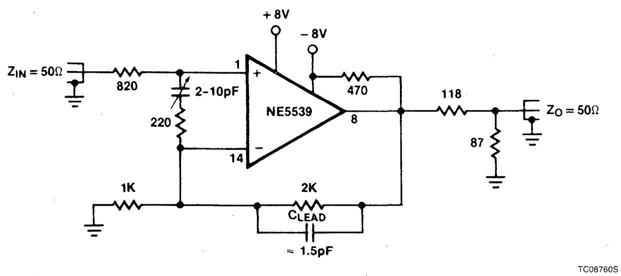


Figure 6. Non-Inverting Follower

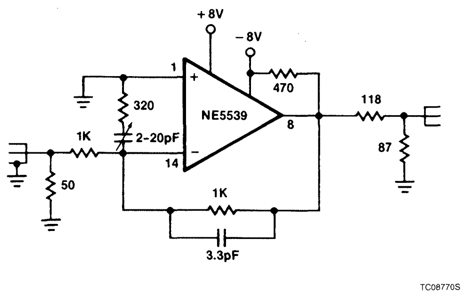


Figure 7. Inverting Follower

Signetics

AN140
Compensation Techniques for Use with the NE/SE5539

Application Note

Linear Products

NE5539 DESCRIPTION

The Signetics NE/SE5539 ultra-high frequency operational amplifier is one of the fastest monolithic amplifiers made today. With a unity gain bandwidth of 350MHz and a slew rate of 600V/μs, it is second to none. Therefore, it is understandable that to attain this speed, standard internal compensation would have to be left out of its design. As a consequence, the op amp is not unconditionally stable for all closed-loop gains and must be externally compensated for gains below 17dB. Properly done, compensation need not limit slew rate. The following will explain how to use the methods available with the NE/SE5539.

LEAD AND LAG-LEAD COMPENSATION

A useful method for compensating the device for closed-loop gains below seven is to use lag-lead and lead networks as shown in Figure 1. The lead network is primarily concerned with compensating for loss of phase margin caused by distributed board capacitance and input capacitance, while lag-lead is mainly for optimizing transient response. Lead compensation modifies the feedback network and adds a zero to the overall transfer function. This increases the phase, but does not greatly change the gain magnitude. This zero improves the phase margin.

To determine components, it can be shown that the optimal conditions for amplifier stability occur when:

$$(R1)(C_{DIST}) = (R_F)(C_{LEAD}) \tag{1}$$

However, when the stability criteria is obtained, it should be noted that the actual bandwidth of the closed-loop amplifier will be reduced. Based on using a double-sided copper-clad printed circuit board with a distributed capacitance of 3.5pF and a unity gain configuration, C_{LEAD} would be 3.5pF. Another way of stating the relationship between the distributed capacitance closed-loop gain and the lead compensation capacitor is:

$$C_{LEAD} = C_{DIST}\frac{R1}{R_F} \tag{2}$$

When bandwidth is of primary concern, the lead compensation will usually be adequate. For closed-loop gains less than seven, lag-lead compensation is necessary for stability.

If transient response is also a factor in design, a lag-lead compensation network may be necessary (Reference Figure 1). For practical applications, the following equations can be used to determine proper lag-lead components:

$$\frac{R_F}{R1 / R_{LAG}} \geqslant 7 \tag{4}$$

Therefore,

$$R_{LAG} \leqslant \frac{R_F}{7 - R_F/R1} \tag{5}$$

Using the above equation will insure a closed-loop gain of seven above the network break frequency. C_{LAG} may now be approximated using:

$$W_{LAG} \cong \frac{2\pi(GBW)}{10}\ \text{Rad/Sec} \tag{6}$$

$$W_{LAG} = \frac{\pi(GBW)}{5}\ \text{Rad/Sec} \tag{7}$$

where

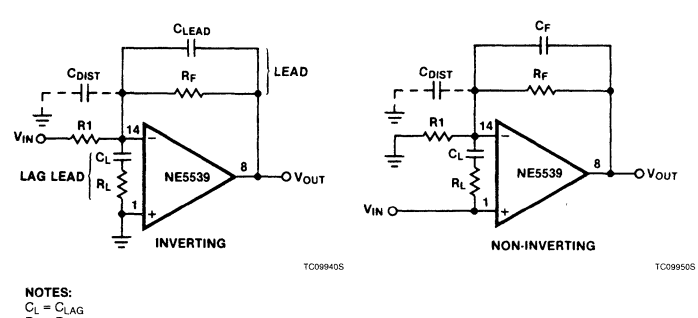
$$W_{LAG} = \frac{1}{(R_{LAG})(C_{LAG})} \tag{8}$$

therefore,

$$\frac{\pi(GBW)}{5} = \frac{1}{(R_{LAG})(C_{LAG})} \tag{9}$$

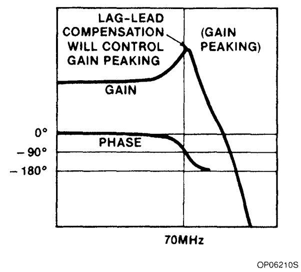
and

$$C_{LAG} = \frac{5}{\pi R_{LAG}(GBW)} \tag{10}$$

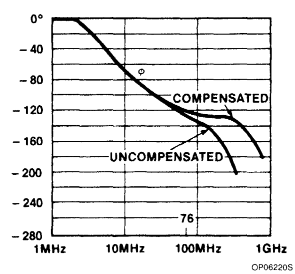


NOTES:
$C_L = C_{LAG}$
$R_L = R_{LAG}$

Figure 1. Standard Lag-Lead Compensation



a. Closed-Loop Inverting Gain of Seven Gain-Phase Response (Uncompensated)



b. Open-Loop Phase

Figure 2

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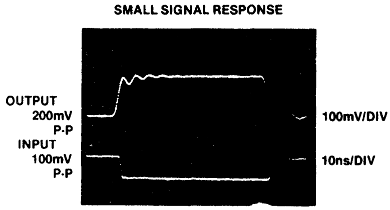


Figure 3. Compensated Pulse Response

This method adds a pole and zero to the transfer function of the device, causing the actual open-loop gain and phase curve to be reshaped, thus creating a progressive improvement above the critical frequency where phase changes rapidly. (Near 70MHz, see Figures 2a and 2b.) But also, the lag-lead network can be adjusted to optimize gain peaking for transient responses. Therefore, rise time, overshoot, and settling time can be changed for various closed-loop gains. The result of using this technique is shown for a pulse amplifier in Figure 3.

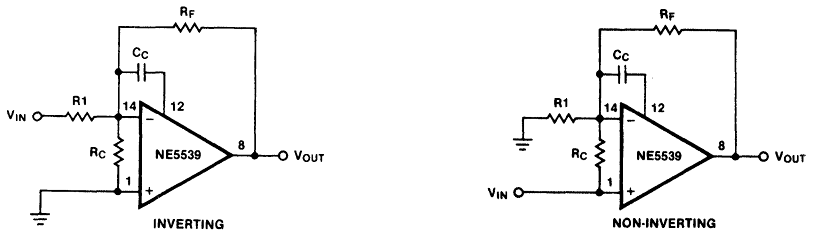


Figure 4. Pin 12 Compensation

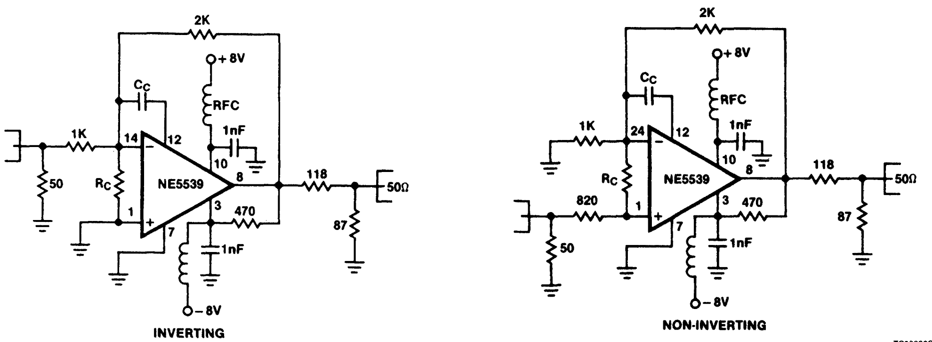
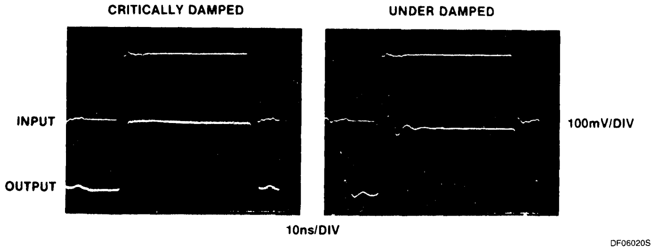


Figure 5. Pulse Response Test Circuits

Compensation Techniques for Use with the NE/SE5539 AN140



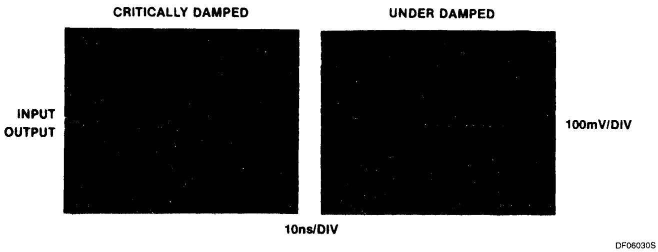
RISE TIME - 2.4ns
$R_C = 118\Omega$ — $C_C = 4.6pF$
PROPAGATION DELAY = 2.3ns

(a)

RISE TIME - 2.1ns
$R_C = 200$ — $C_C = 5.4pF$
PROPAGATION DELAY = 2.3ns

(b)

Figure 6. Small Signal Response — Non-Inverting



RISE TIME - 5.3ns
$R_C = 226\Omega$ — $C_C = 2.3pF$
PROPAGATION DELAY = 5.1ns

(a)

RISE TIME - 3.3ns
$R_C = 460$ — $C_C = 2.0pF$
PROPAGATION DELAY = 4.5ns

(b)

Figure 7. Small Signal Response — Inverting

USING PIN 12 COMPENSATION

An alternate method of external compensation is obtained by use of the NE/SE5539 frequency compensation pin. The circuits in Figure 4 show the correct way to use this pin. As can be seen, this method saves the use of one capacitor as compared to standard lag-lead and lead compensation as shown in Figure 1.

But, most importantly, both methods are equally effective; i.e., a good wide-band amplifier below 17dB, with control over ringing and overshoot. For example, inverting and non-inverting amplifier circuits using Pin 12 are shown in Figure 5. The corresponding pulse response for each circuit is shown in Figures 6 and 7 for the network values recommended. As shown by the response photos, the overshoot and settling time can be controlled by adjusting R_C and C_C. In damping the overshoot, rise time is slightly decreased. Also, the non-inverting configuration (Figure 6) gives a very fast response time compared to the inverting mode.

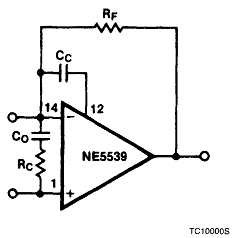


Figure 8. C_O Will Reduce Output Offset and Noise

If it is important to reduce output offset voltage and noise, an additional capacitor, C_O, can be added in series with the resistor (R_C) across the inputs. This should be a large value to block DC but not affect the benefits of the compensation components at high frequencies. A value of $0.01\mu F$ as shown in Figure 8 is sufficient.

INTERNAL CHARACTERISTICS OF THE NE/SE5539

In order to better understand the compensation procedure, a detailed discussion of the amplifier follows.

The complete amplifier schematic is shown in Figure 9. To clarify the effect of the compensation pin, the schematic is split into five main parts as shown in Figure 10.

Each segment in Figure 10 is defined as follows: starting from the non-inverting input, Section A_1 is the amplification from the input to the base of transistor Q_4. A_2 is from the base of Q_4 to the summation point at the collector of Q_3. Furthermore, A_3 represents the gain from the non-inverting input to the summation point via the common emitter side of Q_2 and Q_3. Finally, B_F is the feedback factor of the positive feedback loop from the collector of Q_3 to the base of Q_4.

From Figure 10, it can be seen that the total gain (A_T) is:

$$A_T = \frac{A_1 A_2}{1 - (B_F A_2)} + A_3 (1 + B_F A_2)$$

Each term in this equation plays a role at different frequencies to determine the total transfer function of the device. Of particular importance is the pole in A_3 (near 340MHz) which causes a roll-off of 12dB/octave and loss of phase margin just before unity gain. This can be seen in the Bode plot in Figure 11a. To overcome this pole, a capacitor and resistor are connected as shown in Figures 12a and 12b. The compensation pin is connected to the emitter of Q_5, which is in an emitter-follower configuration. Therefore, a reactance connected to Pin 12 acts essentially as if it were connected at the base of Q_5. Since the capacitor is connected here, it is now a component of B_F and a zero is added to the transfer function. The resistor across the input pins controls overall gain and causes A_T to cross 0dB at a lower frequency; the capacitor in the feedback loop controls phase shift and gain peaking.

To further explain, Bode plots of open-loop response using varying capacitor values and corresponding pulse responses are shown in Figures 13a through 13f. The changes in gain and phase can readily be seen, as is the effect on bandwidth.

11

Compensation Techniques for Use with the NE/SE5539 AN140

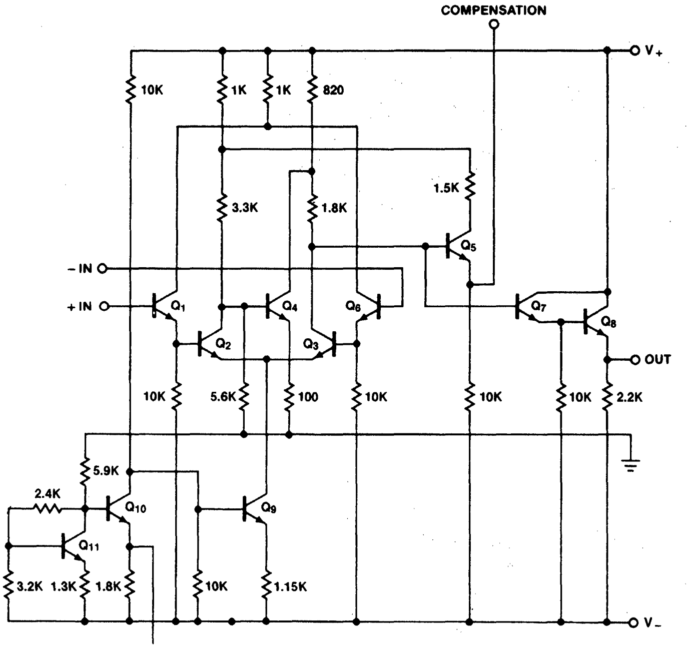


Figure 9. Complete Schematic of NE/SE5539

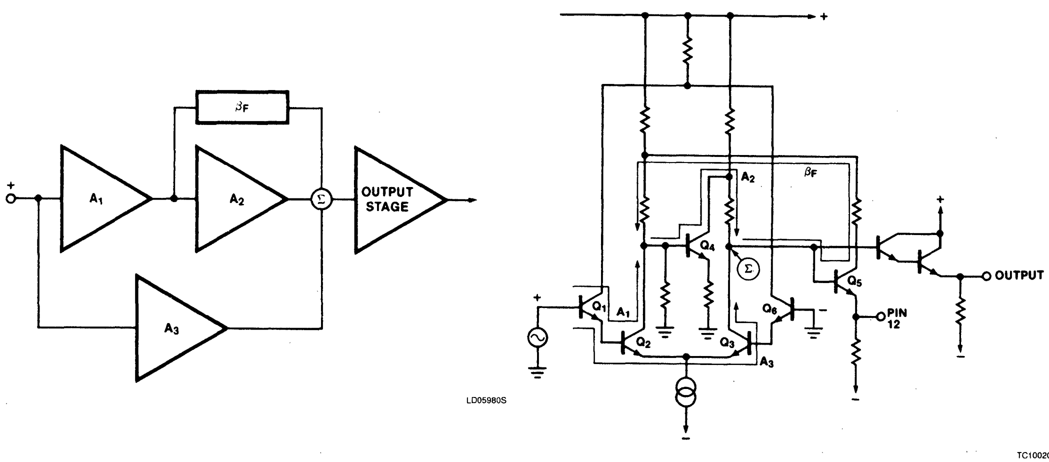
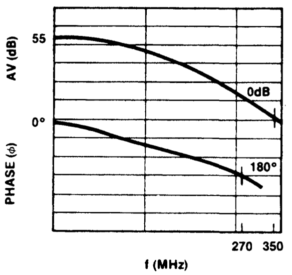


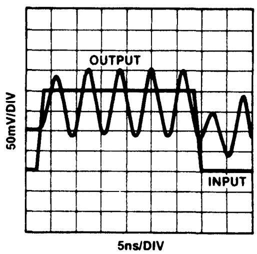
Figure 10. Internal Sections

COMPUTER ANALYSIS

The open-loop and pulse response plots were generated using an IBM 370 computer and SPICE, a general-purpose circuit simulation program. Each transistor in the part is mathematically modeled after actual device parameters, which were measured in the laboratory. These models are then combined with the resistors and voltage sources through node numbers so that the computer knows where each is connected.



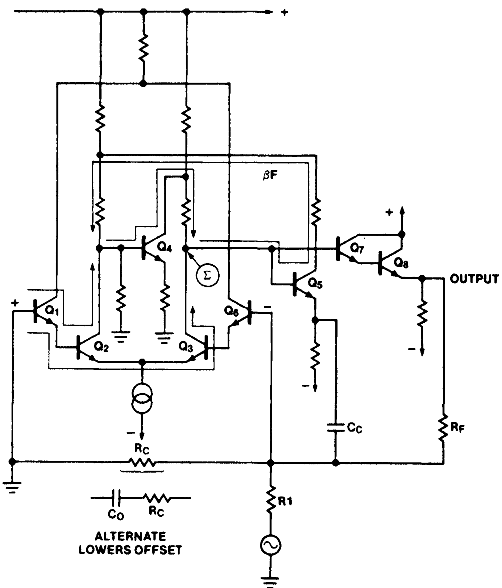
a. Open-Loop Gain — No Compensation (Computer Simulation)



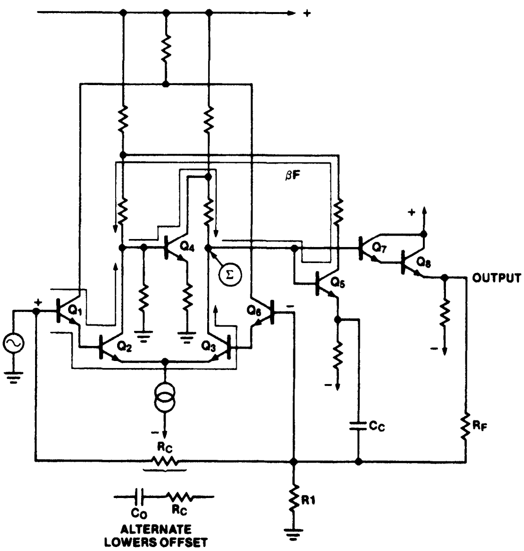
b. Closed-Loop Non-Inverting Response — No Compensation (Computer Simulation — Oscillation is Evident)

Figure 11

To indicate the accuracy of this system, the actual open-loop gain is compared to the computer plots in Figures 14 and 15. The real payoff for this system is that once a credible simulation is achieved, any outside circuit can be modeled around the op amp. This would be used to check for feasibility before breadboarding in the lab. The internal circuit can be treated like a black box and the outside circuit program altered to whatever application the user would like to examine.



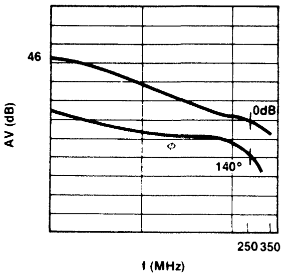
a. Pin 12 Compensation Showing Internal Connections — Inverting



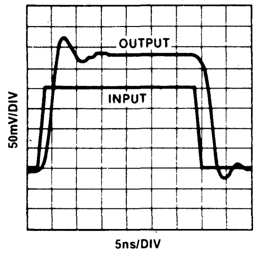
b. Pin 12 Compensation Showing Internal Connections — Non-Inverting

Figure 12

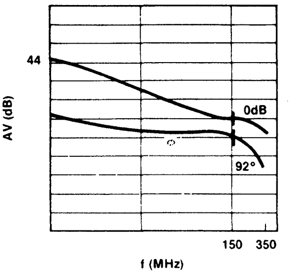
Compensation Techniques for Use with the NE/SE5539 AN140



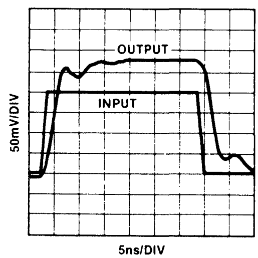
a. Open-Loop Pin 12 Compensation — $R_C = 200\Omega$, $C_C = 1pF$, (Computer Simulation)



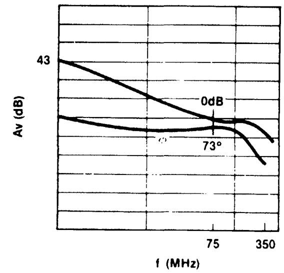
b. Closed-Loop Non-Inverting Pulse Response — $R_C = 200\Omega$, $C_C = 1pF$, $A_V = 3$ (Computer Simulation — Underdamped)



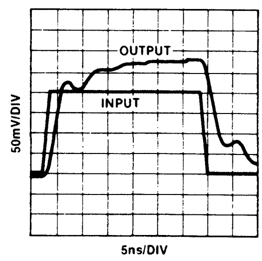
c. Open-Loop Pin 12 Compensation — $R_C = 200\Omega$, $C_C = 2pF$ (Computer Simulation)



d. Closed-Loop Non-Inverting Pulse Response — $R_C = 200\Omega$, $C_C = 2pF$, $Av = 3$ (Computer Simulation — Critically-Damped)



e. Open-Loop Pin 12 Compensation — $R_C = 200\Omega$, $C_C = 3pF$, (Computer Simulation)



f. Closed-Loop Non-Inverting Pulse Response — $R_C = 200\Omega$, $C_C = 3pF$, $A_V = 3$ (Computer Simulation — Overdamped)

Figure 13

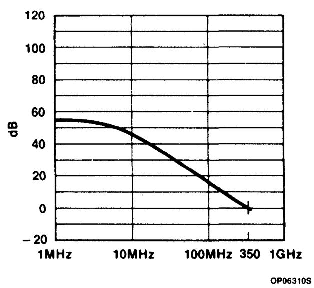


Figure 14. Actual Open-Loop Gain Measured in Lab

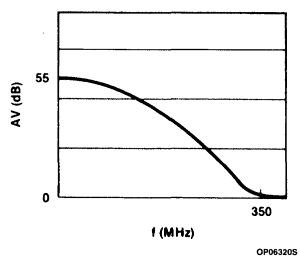


Figure 15. Computer-Generated Open-Loop Gain

1. J. Millman and C. C. Halkias: *Integrated Electronics: Analog and Digital Circuits and Systems,* McGraw-Hill Book Company, New York, 1972.

2. A. Vladimirescu, Kaihe Zhang, A. R. Newton, D. O. Peterson, A. Sanquiovanni-Vincentelli: "Spice Version 2G," University of California, Berkeley, California, August 10, 1981.

3. Signetics: *Analog Data Manual 1983,* Signetics Corporation, Sunnyvale, California 1983.

Signetics

NE5592
Video Amplifier

Product Specification

Linear Products

DESCRIPTION

The NE5592 is a dual monolithic, two-stage, differential output, wideband video amplifier. It offers a fixed gain of 400 without external components and an adjustable gain from 400 to 0 with one external resistor. The input stage has been designed so that with the addition of a few external reactive elements between the gain select terminals, the circuit can function as a high-pass, low-pass, or band-pass filter. This feature makes the circuit ideal for use as a video or pulse amplifier in communications, magnetic memories, display, video recorder systems, and floppy disk head amplifiers.

FEATURES

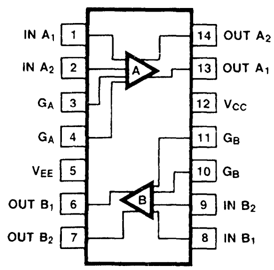
- **110MHz unity gain bandwidth**
- **Adjustable gain from 0 to 400**
- **Adjustable pass band**
- **No frequency compensation required**
- **Wave shaping with minimal external components**

APPLICATIONS

- **Floppy disk head amplifier**
- **Video amplifier**
- **Pulse amplifier in communications**
- **Magnetic memory**
- **Video recorder systems**

PIN CONFIGURATION

D, N Packages



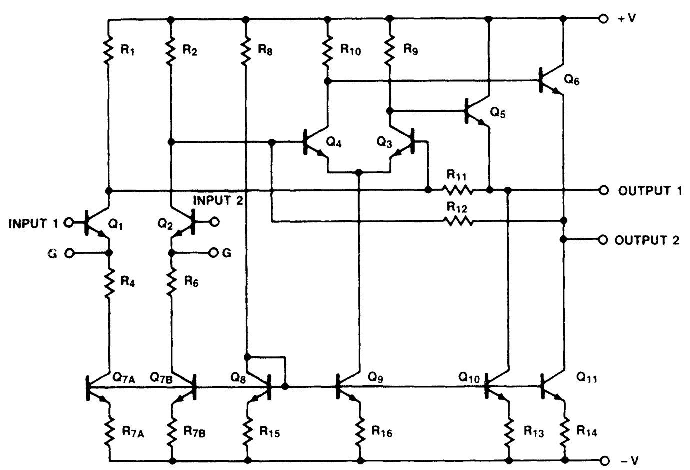
TOP VIEW

CD13050S

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
14-Pin Plastic DIP	0 to 70°C	NE5592N
14-Pin SO package	0 to 70°C	NE5592D

EQUIVALENT CIRCUIT



LD08030S

11

Video Amplifier NE5592

ABSOLUTE MAXIMUM RATINGS $T_A = 25°C$, unless otherwise specified.

SYMBOL	PARAMETER	RATING	UNIT
V_{CC}	Supply voltage	±8	V
V_{IN}	Differential input voltage	±5	V
V_{CM}	Common mode Input voltage	±6	V
I_{OUT}	Output current	10	mA
T_A	Operating temperature range NE5592	0 to +70	°C
T_{STG}	Storage temperature range	−65 to +150	°C
$P_{D\ MAX}$	Maximum power dissipation, $T_A = 25°C$ (still air)[1] D package N package	 1.03 1.48	 W W

NOTE:
1. Derate above 25°C at the following rates:
   D package 8.3mW/°C
   N package 11.9mW/°C

DC ELECTRICAL CHARACTERISTICS $T_A = +25°C$, $V_{SS} = \pm 6V$, $V_{CM} = 0$, unless otherwise specified. Recommended operating supply voltage is $V_S = \pm 6.0V$, and gain select pins are connected together.

SYMBOL	PARAMETER	TEST CONDITIONS	LIMITS			UNITS
			Min	Typ	Max	
A_{VOL}	Differential voltage gain	$R_L = 2k\Omega$, $V_{OUT} = 3V_{P-P}$	400	480	600	V/V
R_{IN}	Input resistance		3	14		kΩ
C_{IN}	Input capacitance			2.5		pF
I_{OS}	Input offset current			0.3	3	μA
I_{BIAS}	Input bias current			5	20	μA
	Input noise voltage	BW 1kHz to 10MHz		4		$nV/\sqrt{Hz}$
V_{IN}	Input voltage range		±1.0			V
CMRR	Common-mode rejection ratio	$V_{CM} \pm 1V$, $f < 100kHz$ $V_{CM} \pm 1V$, $f = 5MHz$	60	93 87		dB dB
PSRR	Supply voltage rejection ratio	$\Delta V_S = \pm 0.5V$	50	85		dB
	Channel separation	$V_{OUT} = 1V_{P-P}$; $f = 100kHz$ (output referenced) $R_L = 1k\Omega$	65	70		dB
V_{OS}	Output offset voltage gain select pins open	$R_L = \infty$ $R_L = \infty$		0.5 0.25	1.5 0.75	V V
V_{CM}	Output common-mode voltage	$R_L = \infty$	2.4	3.1	3.4	V
V_{OUT}	Output differential voltage swing	$R_L = 2k\Omega$	3.0	4.0		V
R_{OUT}	Output resistance			20		Ω
I_{CC}	Power supply current (total for both sides)	$R_L = \infty$		35	44	mA

Video Amplifier NE5592

DC ELECTRICAL CHARACTERISTICS $V_{SS} = \pm 6V$, $V_{CM} = 0$, $0°C \leqslant T_A \leqslant 70°C$, unless otherwise specified. Recommended operating supply voltage is $V_S = \pm 6.0V$, and gain select pins are connected together.

SYMBOL	PARAMETER	TEST CONDITIONS	LIMITS Min	LIMITS Typ	LIMITS Max	UNITS
A_{VOL}	Differential voltage gain	$R_L = 2k\Omega$, $V_{OUT} = 3V_{P-P}$	350	430	600	V/V
R_{IN}	Input resistance		1	11		kΩ
I_{OS}	Input offset current				5	μA
I_{BIAS}	Input bias current				30	μA
V_{IN}	Input voltage range		±1.0			V
CMRR	Common-mode rejection ratio	$V_{CM} \pm 1V$, $f < 100kHz$ $R_S = \phi$	55			dB
PSRR	Supply voltage rejection ratio	$\Delta V_S = \pm 0.5V$	50			dB
	Channel separation	$V_{OUT} = 1V_{P-P}$; $f = 100kHz$ (output referenced) $R_L = 1k\Omega$		70		dB
V_{OS}	Output offset voltage gain select pins connected together gain select pins open	 $R_L = \infty$ $R_L = \infty$			 1.5 1.0	 V V
V_{OUT}	Output differential voltage swing	$R_L = 2k\Omega$	2.8			V
I_{CC}	Power supply current (total for both sides)	$R_L = \infty$			47	mA

AC ELECTRICAL CHARACTERISTICS $T_A = +25°C$, $V_{SS} = \pm 6V$, $V_{CM} = 0$, unless otherwise specified. Recommended operating supply voltage $V_S = \pm 6.0V$. Gain select pins connected together.

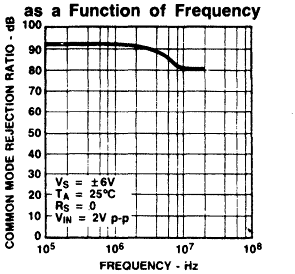
SYMBOL	PARAMETER	TEST CONDITIONS	LIMITS Min	LIMITS Typ	LIMITS Max	UNITS
BW	Bandwidth	$V_{OUT} = 1V_{P-P}$		25		MHz
t_R	Rise time			15	20	ns
t_{PD}	Propagation delay	$V_{OUT} = 1V_{P-P}$		7.5	12	ns

11

Video Amplifier NE5592

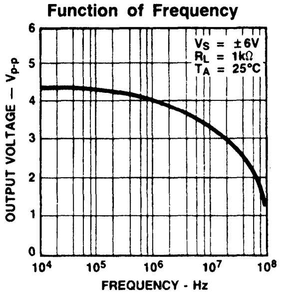
TYPICAL PERFORMANCE CHARACTERISTICS

Common-Mode Rejection Ratio as a Function of Frequency



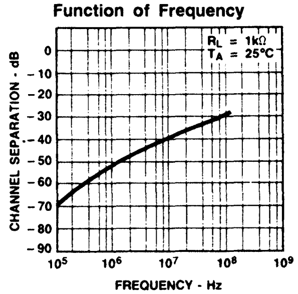
OP18580S

Output Voltage Swing as a Function of Frequency



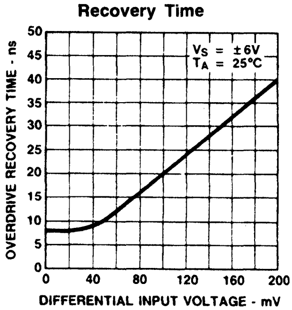
OP18590S

Channel Separation as a Function of Frequency



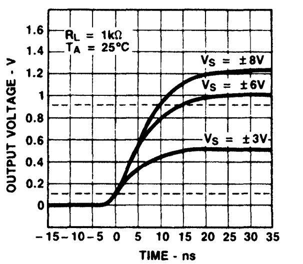
OP18600S

Differential Overdrive Recovery Time



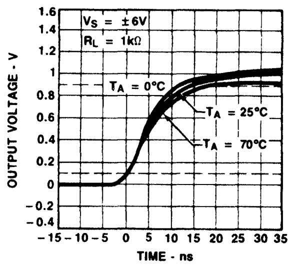
OP18610S

Pulse Response as a Function of Supply Voltage



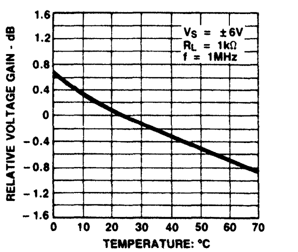
OP18620S

Pulse Response as a Function of Temperature



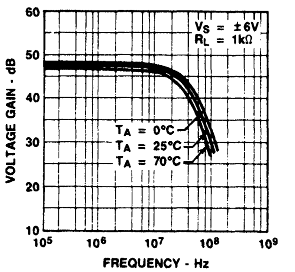
OP18630S

Voltage Gain as a Function of Temperature



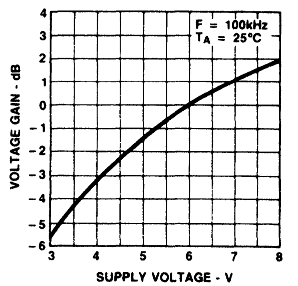
OP18640S

Gain vs Frequency as a Function of Temperature



OP18650S

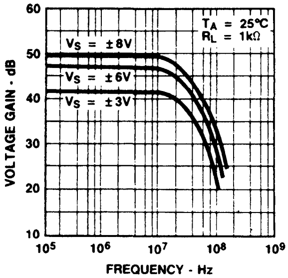
Voltage Gain as a Function of Supply Voltage



OP18660S

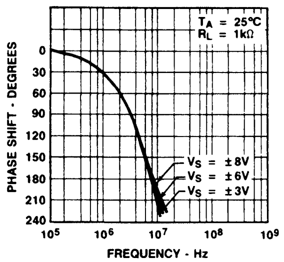
TYPICAL PERFORMANCE CHARACTERISTICS (Continued)

Gain vs Frequency as a Function of Supply Voltage



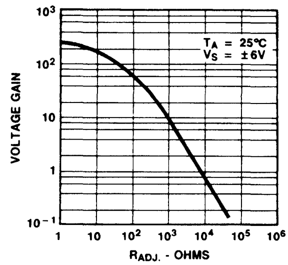
OP18670S

Phase vs Frequency as a Function of Supply Voltage



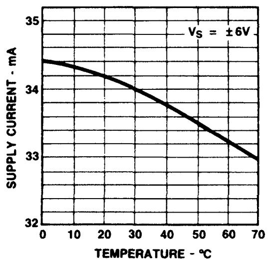
OP18680S

Voltage Gain as a Function of R_{ADJ}



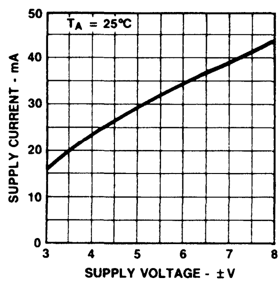
OP18690S

Supply Current as a Function of Temperature



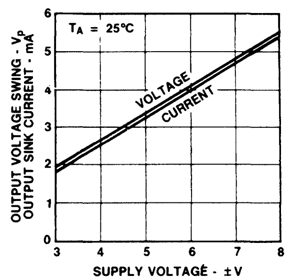
OP18700S

Supply Current as a Function of Supply Voltage



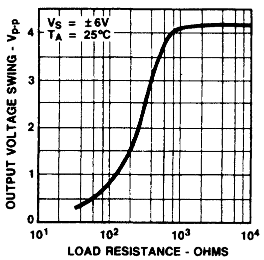
OP18710S

Output Voltage Swing and Sink Current as a Function of Supply Voltage



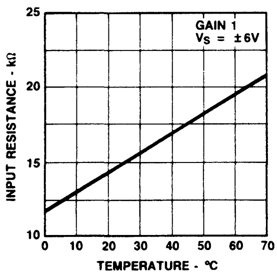
OP18720S

Output Voltage Swing as a Function of Load Resistance



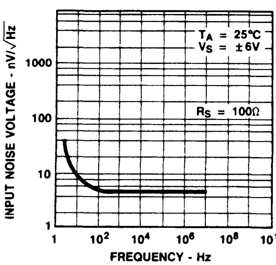
OP18730S

Input Resistance as a Function of Temperature



OP18740S

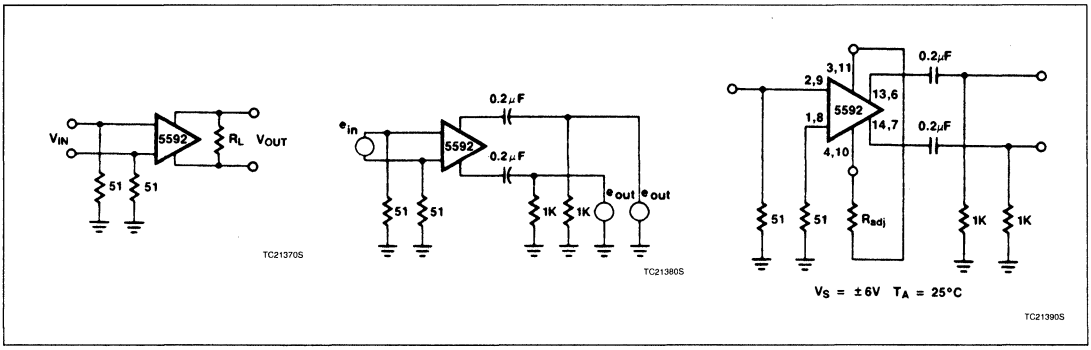
Input Noise Voltage as a Function of Frequency



OP18750S

Video Amplifier NE5592

TEST CIRCUITS $T_A = 25°C$, unless otherwise specified.



Signetics

NE/SA/SE592 Video Amplifier

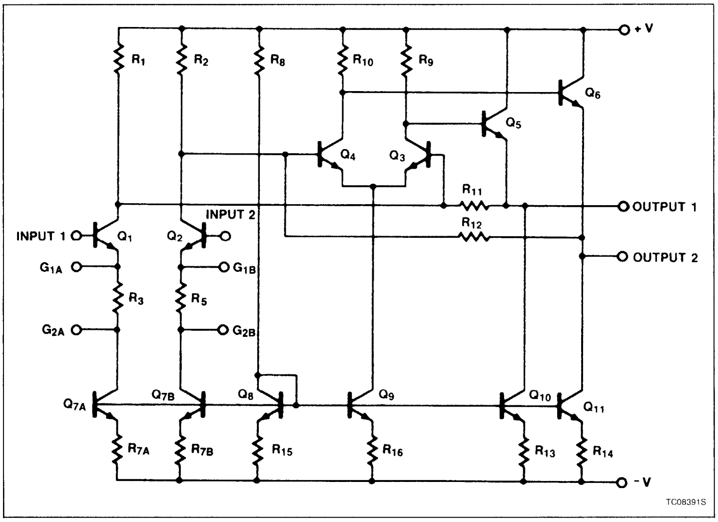
Product Specification

Linear Products

DESCRIPTION

The NE/SA/SE592 is a monolithic, two-stage, differential output, wideband video amplifier. It offers fixed gains of 100 and 400 without external components and adjustable gains from 400 to 0 with one external resistor. The input stage has been designed so that with the addition of a few external reactive elements between the gain select terminals, the circuit can function as a high-pass, low-pass, or band-pass filter. This feature makes the circuit ideal for use as a video or pulse amplifier in communications, magnetic memories, display, video recorder systems, and floppy disk head amplifiers. Now available in an 8-pin version with fixed gain of 400 without external components and adjustable gain from 400 to 0 with one external resistor.

EQUIVALENT CIRCUIT



FEATURES

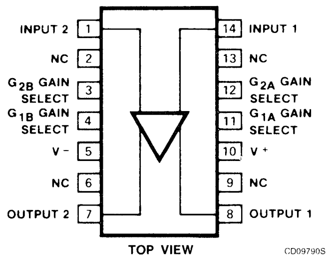
- **120MHz unity gain bandwidth**
- **Adjustable gains from 0 to 400**
- **Adjustable pass band**
- **No frequency compensation required**
- **Wave shaping with minimal external components**
- **MIL-STD processing available**

APPLICATIONS

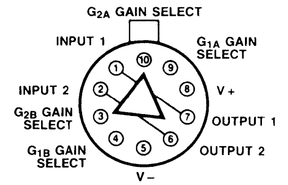
- **Floppy disk head amplifier**
- **Video amplifier**
- **Pulse amplifier in communications**
- **Magnetic memory**
- **Video recorder systems**

PIN CONFIGURATIONS

D, F, N Packages

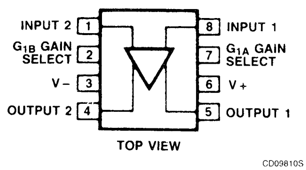


H Package*



NOTES:
Pin 5 connected to case.
*Metal cans (H) not recommended for new designs.

D, F, N, Packages



11

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
14-Pin Plastic DIP	0 to +70°C	NE592N14
14-Pin Cerdip	0 to +70°C	NE592F14
14-Pin Cerdip	-55°C to +125°C	SE592F14
14-Pin SO	0 to +70°C	NE592D14
8-Pin Plastic DIP	0 to +70°C	NE592N8
8-Pin Cerdip	-55°C to +125°C	SE592F8
8-Pin Plastic DIP	-40°C to +85°C	SA592N8
8-Pin SO	0 to +70°C	NE592D8
8-Pin SO	-40°C to +85°C	SA592D8
10-Lead Metal Can	0 to +70°C	NE592H
10-Lead Metal Can	-55°C to +125°C	SE592H

NOTE:
N8, N14, D8 and D14 package parts also available in "High" gain version by adding "H" before package designation, i.e., NE592HD8.

ABSOLUTE MAXIMUM RATINGS $T_A = +25°C$, unless otherwise specified.

SYMBOL	PARAMETER	RATING	UNIT
V_{CC}	Supply voltage	±8	V
V_{IN}	Differential input voltage	±5	V
V_{CM}	Common-mode input voltage	±6	V
I_{OUT}	Output current	10	mA
T_A	Operating ambient temperature range SE592 NE592	 -40 to +85 0 to +70	 °C °C
T_{STG}	Storage temperature range	-65 to +150	°C
$P_{D\ MAX}$	Maximum power dissipation, $T_A = 25°C$ (still air)[1] F-14 package F-8 package D-14 package D-8 package H package N-14 package N-8 package	 1.17 0.79 0.98 0.79 0.83 1.44 1.17	 W W W W W W W

NOTE:
1. Derate above 25°C at the following rates:
   F-14 package at 9.3mW/°C
   F-8 package at 6.3mW/°C
   D-14 package at 7.8mW/°C
   D-8 package at 6.3mW/°C
   H package at 6.7mW/°C
   N-14 package at 11.5mW/°C
   N-8 package at 9.3mW/°C

Video Amplifier NE/SA/SE592

DC ELECTRICAL CHARACTERISTICS

$T_A = +25°C$, $V_{SS} = \pm 6V$, $V_{CM} = 0$, unless otherwise specified. Recommended operating supply voltages $V_S = \pm 6.0V$. All specifications apply to both standard and high gain parts unless noted differently.

SYMBOL	PARAMETER	TEST CONDITIONS	NE/SA592			SE592			UNIT
			Min	Typ	Max	Min	Typ	Max	
A_{VOL}	Differential voltage gain, standard part Gain 1[1] Gain 2[2, 4]	$R_L = 2k\Omega$, $V_{OUT} = 3V_{P-P}$	 250 80	 400 100	 600 120	 300 90	 400 100	 500 110	 V/V V/V
	High gain part		400	500	600				V/V
R_{IN}	Input resistance Gain 1[1] Gain 2[2, 4]		 10	 4.0 30		 20	 4.0 30		 kΩ kΩ
C_{IN}	Input capacitance[2]	Gain 2[4]		2.0			2.0		pF
I_{OS}	Input offset current			0.4	5.0		0.4	3.0	μA
I_{BIAS}	Input bias current			9.0	30		9.0	20	μA
V_{NOISE}	Input noise voltage	BW 1kHz to 10MHz		12			12		μV_{RMS}
V_{IN}	Input voltage range		±1.0			±1.0			V
CMRR	Common-mode rejection ratio Gain 2[4] Gain 2[4]	 $V_{CM} \pm 1V$, $f < 100kHz$ $V_{CM} \pm 1V$, $f = 5MHz$	 60 	 86 60		 60 	 86 60		 dB dB
PSRR	Supply voltage rejection ratio Gain 2[4]	$\Delta V_S = \pm 0.5V$	50	70		50	70		dB
V_{OS}	Output offset voltage Gain 1 Gain 2[4] Gain 3[3]	 $R_L = \infty$ $R_L = \infty$ $R_L = \infty$		 0.35	 1.5 1.5 0.75		 0.35	 1.5 1.0 0.75	 V V V
V_{CM}	Output common-mode voltage	$R_L = \infty$	2.4	2.9	3.4	2.4	2.9	3.4	V
V_{OUT}	Output voltage swing differential	$R_L = 2k\Omega$	3.0	4.0		3.0	4.0		V
R_{OUT}	Output resistance			20			20		Ω
I_{CC}	Power supply current	$R_L = \infty$		18	24		18	24	mA

NOTES:
1. Gain select Pins G_{1A} and G_{1B} connected together.
2. Gain select Pins G_{2A} and G_{2B} connected together.
3. All gain select pins open.
4. Applies to 10- and 14-pin versions only.

DC ELECTRICAL CHARACTERISTICS

$V_{SS} = \pm 6V$, $V_{CM} = 0$, $0°C \leq T_A \leq 70°C$ for NE592; $-40°C \leq T_A \leq 85°C$ for SA592, $-55°C \leq T_A \leq 125°C$ for SE592, unless otherwise specified. Recommended operating supply voltages $V_S = \pm 6.0V$. All specifications apply to both standard and high gain parts unless noted differently.

SYMBOL	PARAMETER	TEST CONDITIONS	NE/SA592			SE592			UNIT
			Min	Typ	Max	Min	Typ	Max	
A_{VOL}	Differential voltage gain, standard part Gain 1[1] Gain 2[2, 4]	$R_L = 2k\Omega$, $V_{OUT} = 3V_{P-P}$	 250 80		 600 120	 200 80		 600 120	 V/V V/V
	High gain part		400	500	600				V/V
R_{IN}	Input resistance Gain 2[2, 4]		8.0			8.0			kΩ
I_{OS}	Input offset current				6.0			5.0	μA
I_{BIAS}	Input bias current				40			40	μA
V_{IN}	Input voltage range		±1.0			±1.0			V
CMRR	Common-mode rejection ratio Gain 2[4]	$V_{CM} \pm 1V$, $f < 100kHz$	50			50			dB
PSRR	Supply voltage rejection ratio Gain 2[4]	$\Delta V_S = \pm 0.5V$	50			50			dB
V_{OS}	Output offset voltage Gain 1 Gain 2[4] Gain 3[3]	 $R_L = \infty$ $R_L = \infty$ $R_L = \infty$			 1.5 1.5 1.0			 1.5 1.2 1.0	 V V V
V_{OUT}	Output voltage swing differential	$R_L = 2k\Omega$	2.8			2.5			V
I_{CC}	Power supply current	$R_L = \infty$			27			27	mA

NOTES:
1. Gain select Pins G_{1A} and G_{1B} connected together.
2. Gain select Pins G_{2A} and G_{2B} connected together.
3. All gain select pins open.
4. Applies to 10- and 14-pin versions only.

AC ELECTRICAL CHARACTERISTICS

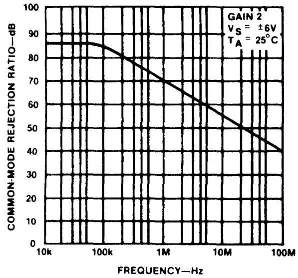
$T_A = +25°C$, $V_{SS} = \pm 6V$, $V_{CM} = 0$, unless otherwise specified. Recommended operating supply voltages $V_S = \pm 6.0V$. All specifications apply to both standard and high gain parts unless noted differently.

SYMBOL	PARAMETER	TEST CONDITIONS	NE/SA592			SE592			UNIT
			Min	Typ	Max	Min	Typ	Max	
BW	Bandwidth Gain 1[1] Gain 2[2, 4]			 40 90			 40 90		 MHz MHz
t_R	Rise time Gain 1[1] Gain 2[2, 4]	$V_{OUT} = 1V_{P-P}$		 10.5 4.5	 12		 10.5 4.5	 10	 ns ns
t_{PD}	Propagation delay Gain 1[1] Gain 2[2, 4]	$V_{OUT} = 1V_{P-P}$		 7.5 6.0	 10		 7.5 6.0	 10	 ns ns

NOTES:
1. Gain select Pins G_{1A} and G_{1B} connected together.
2. Gain select Pins G_{2A} and G_{2B} connected together.
3. All gain select pins open.
4. Applies to 10- and 14-pin versions only.

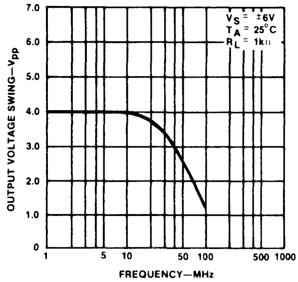
TYPICAL PERFORMANCE CHARACTERISTICS

Common-Mode Rejection Ratio as a Function of Frequency



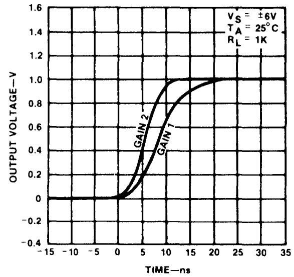
OP04421S

Output Voltage Swing as a Function of Frequency



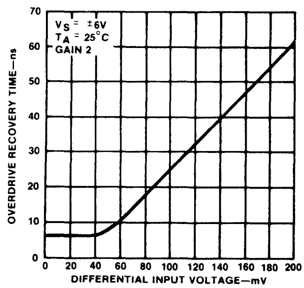
OP04430S

Pulse Response



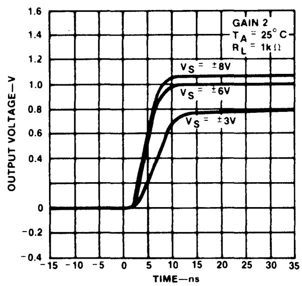
OP04440S

Differential Overdrive Recovery Time



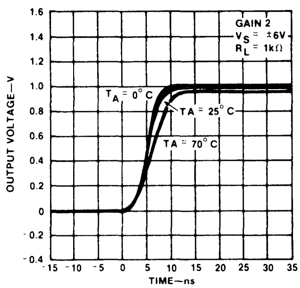
OP04450S

Pulse Response as a Function of Supply Voltage



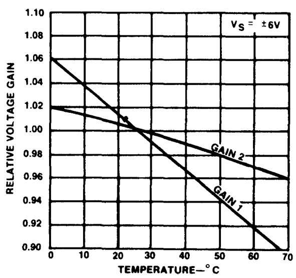
OP04460S

Pulse Response as a Function of Temperature



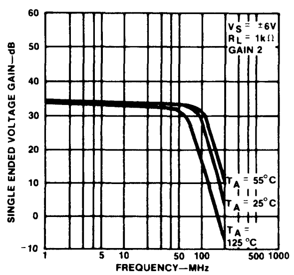
OP04470S

Voltage Gain as a Function of Temperature



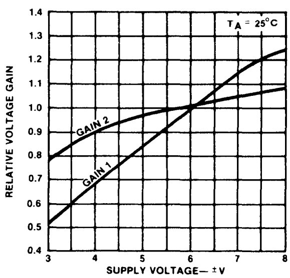
OP04480S

Gain vs Frequency as a Function of Temperature



OP04490S

Voltage Gain as a Function of Supply Voltage

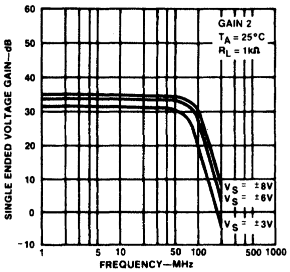


OP04500S

11

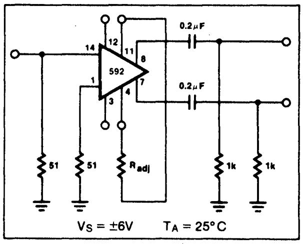
TYPICAL PERFORMANCE CHARACTERISTICS (Continued)

Gain vs Frequency as a Function of Supply Voltage



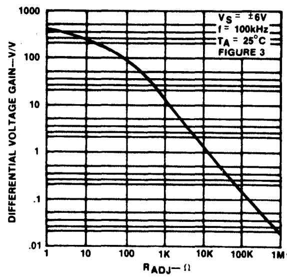
OP04510S

Voltage Gain Adjust Circuit



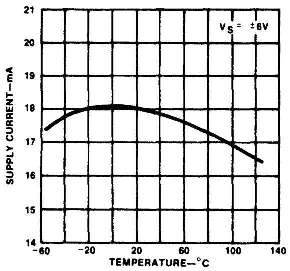
OP04521S

Voltage Gain as a Function of RADJ (Figure 3)



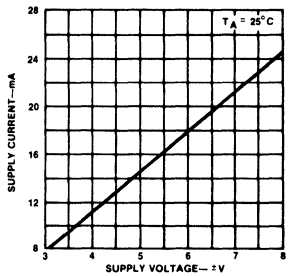
OP04530S

Supply Current as a Function of Temperature



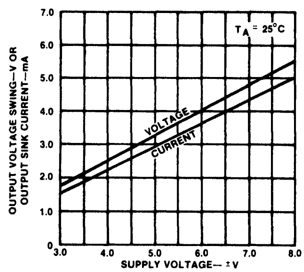
OP04540S

Supply Current as a Function of Supply Voltage



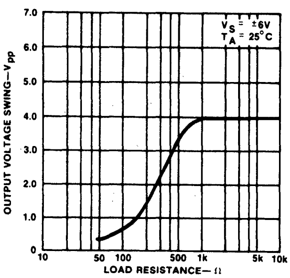
OP04550S

Output Voltage and Current Swing as a Function of Supply Voltage



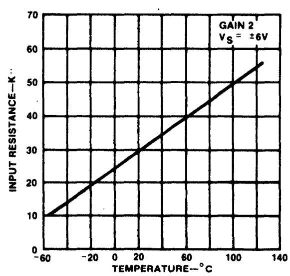
OP04560S

Output Voltage Swing as a Function of Load Resistance



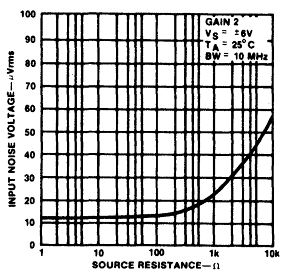
OP04570S

Input Resistance as a Function of Temperature



OP04580S

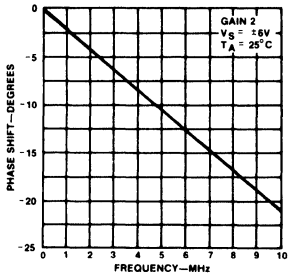
Input Noise Voltage as a Function of Source Resistance



OP04590S

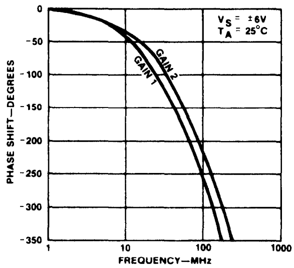
TYPICAL PERFORMANCE CHARACTERISTICS (Continued)

Phase Shift as a Function of Frequency



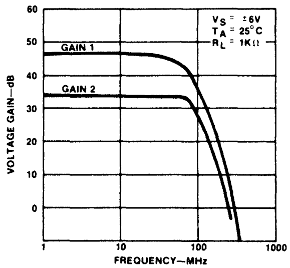
OP04600S

Phase Shift as a Function of Frequency



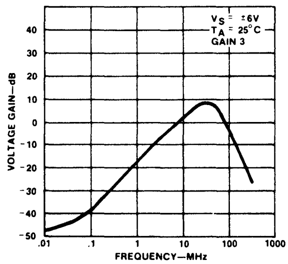
OP04610S

Voltage Gain as a Function of Frequency



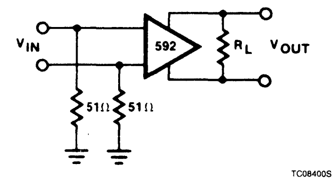
OP04620S

Voltage Gain as a Function of Frequency (All Gain Select Pins Open)

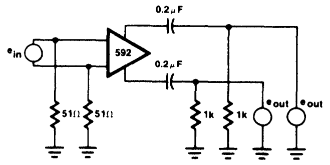


OP04630S

TEST CIRCUITS T_A = 25°C, unless otherwise specified.

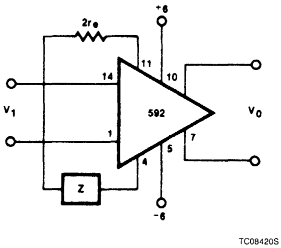


TC08400S



TC08410S

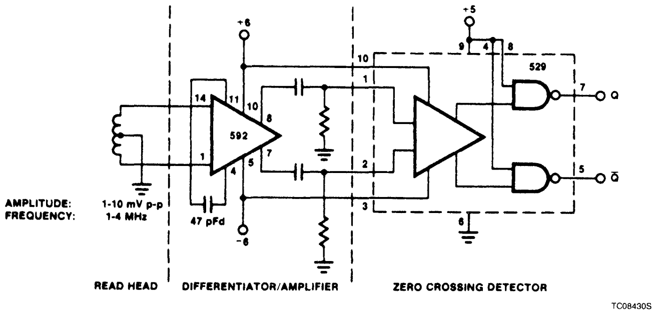
TYPICAL APPLICATIONS



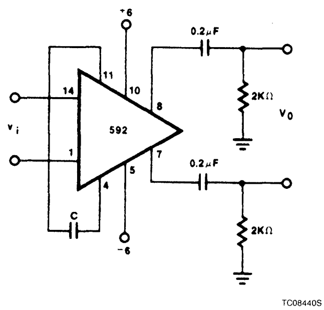
NOTE:

$$\frac{V_0(s)}{V_1(s)} \cong \frac{1.4 \times 10^4}{Z(s) + 2r_e} \cong \frac{1.4 \times 10^4}{Z(s) + 32}$$

Basic Configuration



Disc/Tape Phase-Modulated Readback Systems

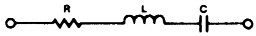
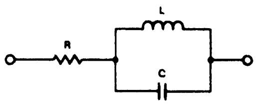


NOTE:
For frequency $F_1 << \frac{1}{2}\pi\,(32)\,C$

$$V_0 \cong 1.4 \times 10^4 C \frac{dVi}{dT}$$

Differentiation with High Common-Mode Noise Rejection

FILTER NETWORKS

Z NETWORK	FILTER TYPE	$\frac{V_O(s)}{V_1(s)}$ TRANSFER FUNCTION
	LOW PASS	$\frac{1.4 \times 10^4}{L}\left[\frac{1}{s + R/L}\right]$
	HIGH PASS	$\frac{1.4 \times 10^4}{R}\left[\frac{s}{s + 1/RC}\right]$
	BAND PASS	$\frac{1.4 \times 10^4}{L}\left[\frac{s}{s^2 + R/L\, s + 1/LC}\right]$
	BAND REJECT	$\frac{1.4 \times 10^4}{R}\left[\frac{s^2 + 1/LC}{s^2 + 1/LC + s/RC}\right]$

TC08422S

NOTES:
In the networks above, the R value used is assumed to include $2r_e$, or approximately 32Ω.
$S = j\omega$
$\omega = 2\pi f$

Signetics

Linear Products

AN141
Using the NE/SA/SE592 Video Amplifier

Application Note

VIDEO AMPLIFIER PRODUCTS

NE/SA/SE592 Video Amplifier

The 592 is a two-stage differential output, wide-band video amplifier with voltage gains as high as 400 and bandwidths up to 120MHz.

Three basic gain options are provided. Fixed gains of 400 and 100 result from shorting together gain select pins $G_{1A} - G_{1B}$ and $G_{2A} - G_{2B}$, respectively. As shown by Figure 1, the emitter circuits of the differential pair return through independent current sources. This topology allows no gain in the input stage if all gain select pins are left open. Thus, the third gain option of tying an external resistance across the gain select pins allows the user to select any desired gain from 0 to 400V/V. The advantages of this configuration will be covered in greater detail under the filter application section.

Three factors should be pointed out at this time:

1. The gains specified are differential. Single-ended gains are one-half the stated value.
2. The circuit 3dB bandwidths are a function of and are inversely proportional to the gain settings.
3. The differential input impedance is an inverse function of the gain setting.

In applications where the signal source is a transformer or magnetic transducer, the input bias current required by the 592 may be passed directly through the source to ground. Where capacitive coupling is to be used, the base inputs must be returned to ground through a resistor to provide a DC path for the bias current.

Due to offset currents, the selection of the input bias resistors is a compromise. To reduce the loading on the source, the resistors should be large, but to minimize the output DC offset, they should be small — ideally 0Ω. Their maximum value is set by the maximum allowable output offset and may be determined as follows:

1. Define the allowable output offset (assume 1.5V).
2. Subtract the maximum 592 output offset (from the data sheet). This gives the output offset allowed as a function of input offset currents (1.5V – 1.0V = 0.5V).
3. Divide by the circuit gain (assume 100). This refers the output offset to the input.
4. The maximum input resistor size is:

$$R_{MAX} = \frac{\text{Input Offset Voltage}}{\text{Max Input Offset Current}} \quad (1)$$

$$= \frac{0.005V}{5\mu A}$$

$$= 1.00k\Omega$$

Of paramount importance during the design of the NE592 device was bandwidth. In a monolithic device, this precludes the use of PNP transistors and standard level-shifting techniques used in lower frequency devices. Thus, without the aid of level shifting, the output common-mode voltage present on the NE592 is typically 2.9V. Most applications, therefore, require capacitive coupling to the load.

Filters

As mentioned earlier, the emitter circuit of the NE592 includes two current sources.

Since the stage gain is calculated by dividing the collector load impedance by the emitter impedance, the high impedance contributed by the current sources causes the stage gain to be zero with all gain select pins open. As shown by the gain vs. frequency graph of Figure 2, the overall gain at low frequencies is a negative 48dB.

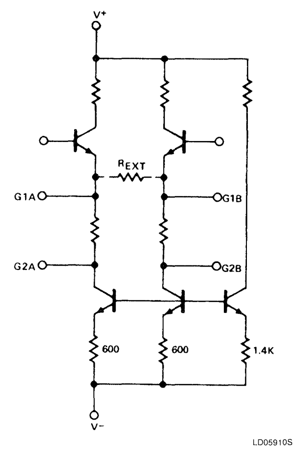
Higher frequencies cause higher gain due to distributed parasitic capacitive reactance. This reactance in the first stage emitter circuit causes increasing stage gain until at 10MHz the gain is 0dB, or unity.

Referring to Figure 3, the impedance seen looking across the emitter structure includes small r_e of each transistor.

Any calculations of impedance networks across the emitters then must include this quantity. The collector current level is approximately 2mA, causing the quantity of 2 r_e to be approximately 32Ω. Overall device gain is thus given by

$$\frac{V_O(s)}{V_{IN}(s)} = \frac{1.4 \times 10^4}{Z_{(S)} + 32} \quad (2)$$

where $Z_{(S)}$ can be resistance or a reactive impedance. Table 2 summarizes the possible configurations to produce low, high, and bandpass filters. The emitter impedance is made to vary as a function of frequency by using capacitors or inductors to alter the frequency response. Included also in Table 2 is the gain calculation to determine the voltage gain as a function of frequency.



NOTE:
All resistor values are in ohms.

Figure 1. 592 Input Structure

Table 1. Video Amplifier Comparison File

PARAMETER	NE/SA/SE592	733
Bandwidth (MHz)	120	120
Gain	0,100,400	10,100,400
R_{IN} (k)	4 – 30	4 – 250
V_{P-P} (Vs)	4.0	4.0

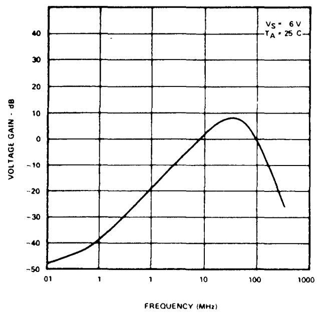
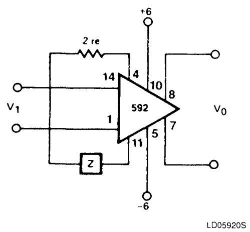


Figure 2. Voltage Gain as a Function of Frequency (All Gain Select Pins Open)



NOTE:

$$\frac{V_0(s)}{V_1(s)} = \frac{1.4 \times 10^4}{Z(s) + 2re} = \frac{1.4 \times 10^4}{Z(s) + 32}$$

Figure 3. Basic Gain Configuration for NE592, N14

Differentiation

With the addition of a capacitor across the gain select terminals, the NE592 becomes a differentiator. The primary advantage of using the emitter circuit to accomplish differentiation is the retention of the high common mode noise rejection. Disc file playback systems rely heavily upon this common-mode rejection for proper operation. Figure 4 shows a differential amplifier configuration with transfer function.

Disc File Decoding

In recovering data from disc or drum files, several steps must be taken to precondition the linear data. The NE592 video amplifier, coupled with the 8T20 bidirectional one-shot, provides all the signal conditioning necessary for phase-encoded data.

When data is recorded on a disc, drum or tape system, the readback will be a Gaussian shaped pulse with the peak of the pulse corresponding to the actual recorded transition point. This readback signal is usually $500\mu V_{P-P}$ to $3mV_{P-P}$ for oxide coated disc files and 1 to $20mV_{P-P}$ for nickel-cobalt disc files. In order to accurately reproduce the data stream originally written on the disc memory, the time of peak point of the Gaussian readback signal must be determined.

The classical approach to peak time determination is to differentiate the input signal. Differentiation results in a voltage proportional to the slope of the input signal. The zero-crossing point of the differentiator, therefore, will occur when the input signal is at a peak. Using a zero-crossing detector and one-shot, therefore, results in pulses occurring at the input peak points.

A circuit which provides the preconditioning described above is shown in Figure 5. Readback data is applied directly to the input of the first NE592. This amplifier functions as a wide-band AC-coupled amplifier with a gain of 100. The NE592 is excellent for this use because of its high phase linearity, high gain and ability to directly couple the unit with the readback head. By direct coupling of readback head to amplifier, no matched terminating resistors are required and the excellent common-mode rejection ratio of the amplifier is preserved. DC components are also rejected because the NE592 has no gain at DC due to the capacitance across the gain select terminals.

The output of the first stage amplifier is routed to a linear phase shift low-pass filter. The filter is a single-stage constant K filter, with a characteristic impedance of 200Ω. Calculations for the filter are as follows:

$L = {}^{2R}/_{\omega_C}$

where

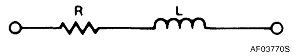
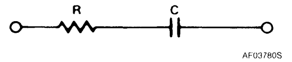
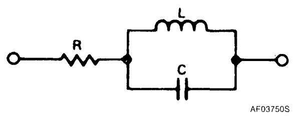
R = characteristic impedance (Ω)

$C = {}^{1}/_{\omega_C}$

where

ω_C = cut-off frequency (radians/sec)

Table 2. Filter Networks

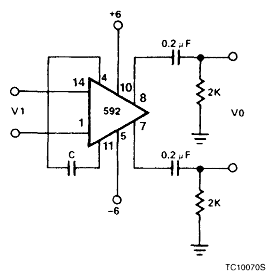
Z NETWORK	FILTER TYPE	$\frac{V_0(s)}{V_1(s)}$ TRANSFER FUNCTION
	LOW PASS	$\frac{1.4 \times 10^4}{L}\left[\frac{1}{s + R/L}\right]$
	HIGH PASS	$\frac{1.4 \times 10^4}{R}\left[\frac{s}{s + 1/RC}\right]$
	BAND PASS	$\frac{1.4 \times 10^4}{L}\left[\frac{s}{s^2 + R/Ls + 1/LC}\right]$
	BAND REJECT	$\frac{1.4 \times 10^4}{R}\left[\frac{s^2 + 1/LC}{s^2 + 1/LC + s/RC}\right]$

NOTES:

In the networks above, the R value used is assumed to include 2 r_e, or approximately 32Ω.

$S = j\Omega$

$\Omega = 2\pi f$



NOTES:

For frequency $F_1 << 1/2\pi(32)C$

$V_0 \cong 1.4 \times 10^4 C \frac{d_{vi}}{dT}$

All resistor values are in ohms.

Figure 4. Differential with High Common-Mode Noise Rejection

Using the NE/SA/SE592 Video Amplifier AN141

The second NE592 is utilized as a low noise differentiator/amplifier stage. The NE592 is excellent in this application because it allows differentiation with excellent common-mode noise rejection.

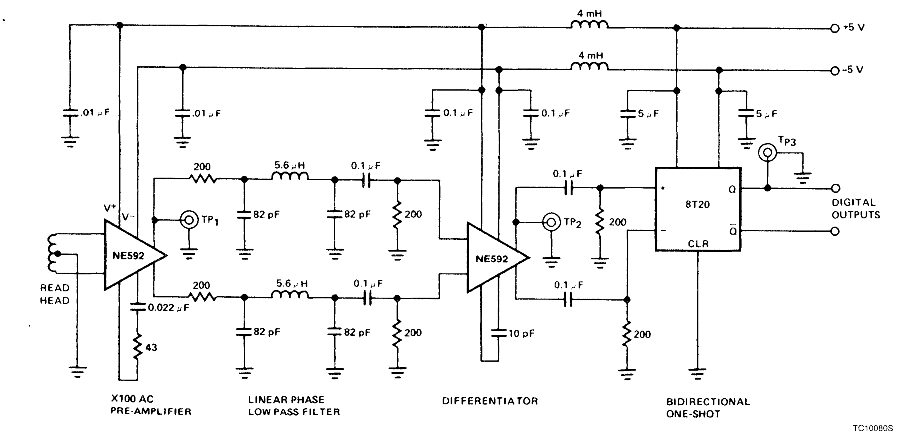
The output of the differentiator/amplifier is connected to the 8T20 bidirectional monostable unit to provide the proper pulses at the zero-crossing points of the differentiator.

The circuit in Figure 5 was tested with an input signal approximating that of a readback signal. The results are shown in Figure 7.

Automatic Gain Control

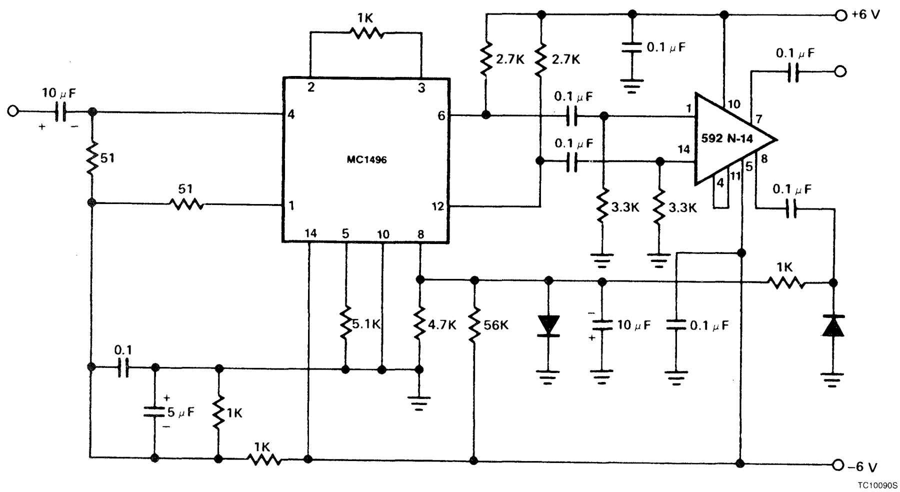
The NE592 can also be connected in conjunction with a MC1496 balanced modulator to form an excellent automatic gain control system.

The signal is fed to the signal input of the MC1496 and RC-coupled to the NE592. Unbalancing the carrier input of the MC1496 causes the signal to pass through unattenuated. Rectifying and filtering one of the NE592 outputs produces a DC signal which is proportional to the AC signal amplitude. After filtering; this control signal is applied to the MC1496 causing its gain to change.



NOTE:
All resistor values are in ohms

Figure 5. 5MHz Phase-Encoded Data Read Circuitry



NOTE:
All resistor values are in ohms

Figure 6. Wide-band AGC Amplifier

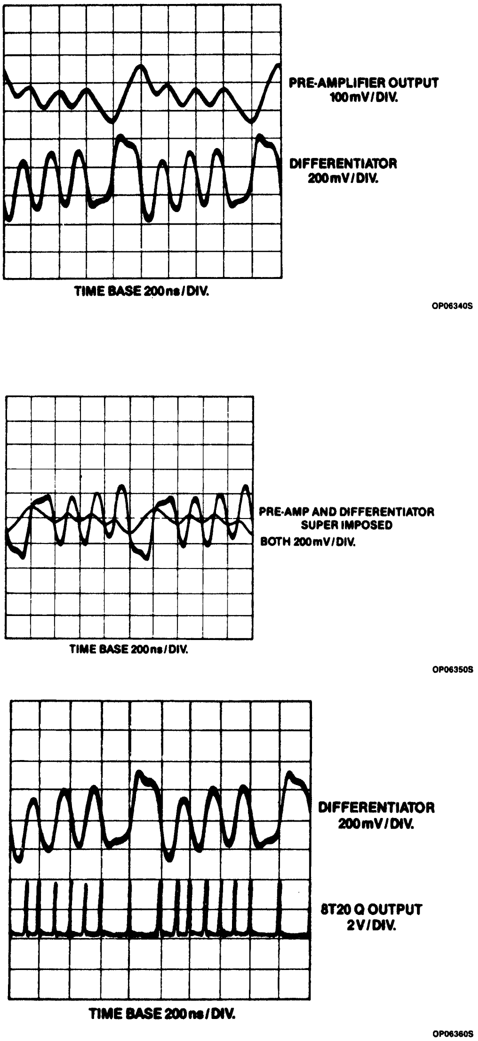


Figure 7. Test Results of Disc File Decoder Circuit

Signetics

μA733/733C
Differential Video Amplifier

Product Specification

Linear Products

DESCRIPTION

The 733 is a monolithic differential input, differential output, wide-band video amplifier. It offers fixed gains of 10, 100, or 400 without external components, and adjustable gains from 10 to 400 by the use of an external resistor. No external frequency compensation components are required for any gain option. Gain stability, wide bandwidth, and low phase distortion are obtained through use of the classic series-shunt feedback from the emitter-follower outputs to the inputs of the second stage. The emitter-follower outputs provide low output impedance, and enable the device to drive capacitive loads. The 733 is intended for use as a high-performance video and pulse amplifier in communications, magnetic memories, display and video recorder systems.

FEATURES

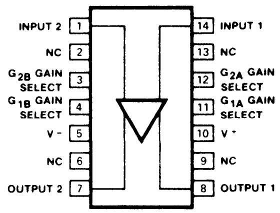
- **120MHz bandwidth**
- **250kΩ input resistance**
- **Selectable gains of 10, 100, and 400**
- **No frequency compensation required**
- **MIL-STD-883A, B, C available**

APPLICATIONS

- **Video amplifier**
- **Pulse amplifier in communications**
- **Magnetic memories**
- **Video recorder systems**

PIN CONFIGURATION

F, N Packages



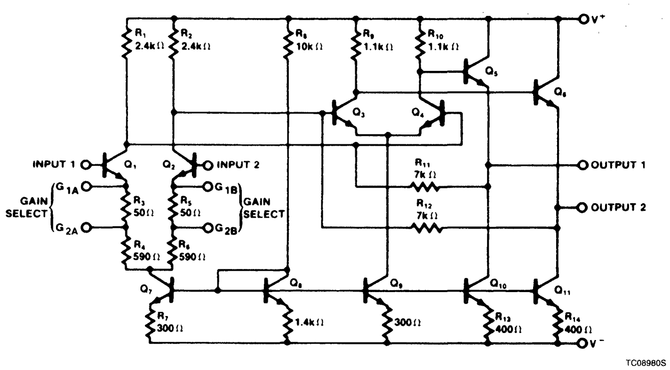
TOP VIEW

CD09960S

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE	ORDER CODE
14-Pin Ceramic DIP	−55°C to +125°C	μA733F
14-Pin Plastic DIP	−55°C to +125°C	μA733N
14-Pin Plastic DIP	0 to +70°C	μA733CN
14-Pin Ceramic DIP	0 to +70°C	μA733CF

CIRCUIT SCHEMATIC



TC08980S

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{DIFF}	Differential input voltage	±5	V
V_{CM}	Common-mode input voltage	±6	V
V_{CC}	Supply voltage	±8	V
I_{OUT}	Output current	10	mA
T_J	Junction temperature	+150	°C
T_{STG}	Storage temperature range	-65 to +150	°C
T_A	Operating ambient temperature range μA733C μA733	 0 to +70 -55 to +125	 °C °C
$P_{D\ MAX}$	Maximum power dissipation, 25°C ambient temperature (still-air)[1] F package N package	 1190 1420	 mW mW

NOTE:
1. The following derating factors should be applied above 25°C:
F package at 9.5mW/°C
N package at 11.4mW/°C.

DC ELECTRICAL CHARACTERISTICS

T_A = +25°C, V_S = ±6V, V_{CM} = 0, unless otherwise specified. Recommended operating supply voltages V_S = ±6.0V.

SYMBOL	PARAMETER	TEST CONDITIONS	μA733C			μA733			UNIT
			Min	Typ	Max	Min	Typ	Max	
	Differential voltage gain Gain 1[2] Gain 2[2] Gain 3[3]	R_l = 2kΩ, V_{OUT} = 3V_{P-P}	 250 80 8	 400 100 10	 600 120 12	 300 90 9	 400 100 10	 500 110 11	 V/V V/V V/V
BW	Bandwidth Gain 1[1] Gain 2[2] Gain 3[3]			 40 90 120			 40 90 120		 MHz MHz MHz
t_R	Rise time Gain 1[1] Gain 2[2] Gain 3[3]	V_{OUT} = 1V_{P-P}		 10.5 4.5 2.5	 12		 10.5 4.5 2.5	 10	 ns ns ns
t_{PD}	Propagation delay Gain 1[1] Gain 2[2] Gain 3[3]	V_{OUT} = 1V_{P-P}		 7.5 6.0 3.6	 10		 7.5 6.0 3.6	 10	 ns ns ns
R_{IN}	Input resistance Gain 1[2] Gain 2[2] Gain 3[3]		 10	 4.0 30 250		 20	 4.0 30 250		 kΩ kΩ kΩ
	Input capacitance[2]	Gain 2		2.0			2.0		pF
I_{OS}	Input offset current			0.4	5.0		0.4	3.0	μA
I_{BIAS}	Input bias current			9.0	30		9.0	20	μA
V_{NOISE}	Input noise voltage	BW = 1kHz to 10MHz		12			12		μV_{RMS}
V_{IN}	Input voltage range		±1.0			±1.0			V
CMRR	Common-mode rejection ratio Gain 2 Gain 2	 V_{CM} = ±1V, f ≤ 100kHz V_{CM} = ±1V, f = 5MHz	 60	 86 60		 60	 86 60		 dB dB
SVRR	Supply voltage rejection ratio Gain 2	ΔV_S = ±0.5V	50	70		50	70		dB

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Differential Video Amplifier — μA733/733C

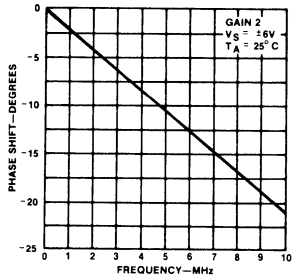
DC ELECTRICAL CHARACTERISTICS (Continued) $T_A = +25°C$, $V_S = \pm 6V$, $V_{CM} = 0$, unless otherwise specified. Recommended operating supply voltages $V_S = \pm 6.0V$.

SYMBOL	PARAMETER	TEST CONDITIONS	μA733C Min	μA733C Typ	μA733C Max	μA733 Min	μA733 Typ	μA733 Max	UNIT
	Output offset voltage Gain 1[1] Gain 2 and 3[2, 3]	$R_L = \infty$		 0.6 0.35	 1.5 1.5		 0.6 0.35	 1.5 1.0	 V V
V_{CM}	Output common-mode voltage	$R_L = \infty$	2.4	2.9	3.4	2.4	2.9	3.4	V
	Output voltage swing, differential	$R_L = 2k\Omega$	3.0	4.0		3.0	4.0		V_{P-P}
I_{SINK}	Output sink current		2.5	3.6		2.5	3.6		mA
R_{OUT}	Output resistance			20			20		Ω
I_{CC}	Power supply current	$R_L = \infty$		18	24		18	24	mA
THE FOLLOWING SPECIFICATIONS APPLY OVER TEMPERATURE			**$0°C \leq T_A \leq 70°C$**			**$-55°C \leq T_A \leq 125°C$**			
	Differential voltage gain Gain 1[1] Gain 2[2] Gain[3]	$R_l = 2k\Omega$, $V_{OUT} = 3V_{P-P}$	 250 80 8		 600 120 12	 200 80 8		 600 120 12	 V/V V/V V/V
R_{IN}	Input resistance Gain 2[2]		 8			 8			 kΩ
I_{OS}	Input offset current				6			5	μA
I_{BIAS}	Input bias current				40			40	μA
V_{IN}	Input voltage range		±1.0			±1.0			V
CMRR	Common-mode rejection ratio Gain 2	 $V_{CM} = \pm V$, $F \leq 100kHz$	 50			 50			 dB
SVRR	Supply voltage rejection ratio Gain 2	 $\Delta V_S = \pm 0.5V$	 50			 50			 dB
V_{OS}	Output offset voltage Gain 1[1] Gain 2 and 3[2, 3]	$R_L = \infty$			 1.5 1.5			 1.5 1.2	 V V
V_{DIFF}	Output voltage swing, differential	$R_L = 2k\Omega$	2.8			2.5			V_{P-P}
I_{SINK}	Output sink current		2.5			2.2			mA
I_{CC}	Power supply current	$R_L \pm \infty$			27			27	mA

NOTES:
1. Gain select pins G_{1A} and G_{1B} connected together.
2. Gain select pins G_{2A} and G_{2B} connected together.
3. All gain select pins open.

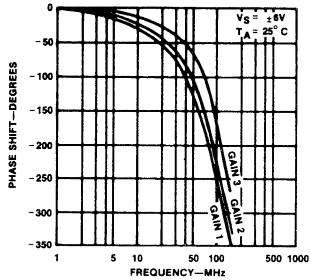
TYPICAL PERFORMANCE CHARACTERISTICS

Phase Shift as a Function of Frequency



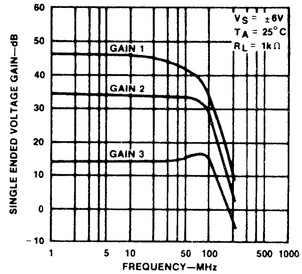
OP05620S

Phase Shift as a Function of Frequency



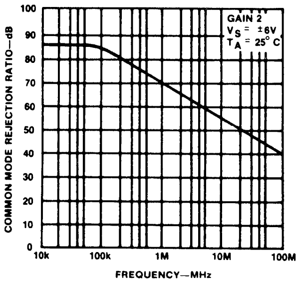
OP05630S

Voltage Gain as a Function of Frequency



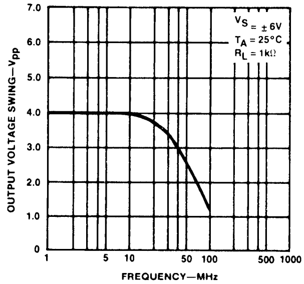
OP05640S

Common Mode Rejection Ratio as a Function of Frequency



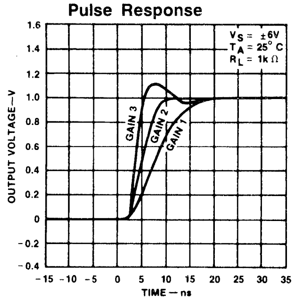
OP05650S

Output Voltage Swing as a Function of Frequency



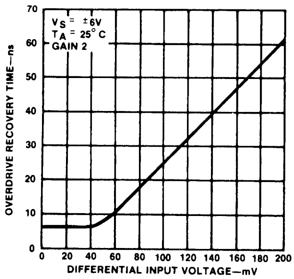
OP05660S

Pulse Response



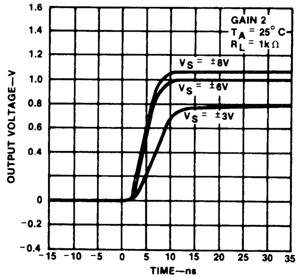
OP05670S

Differential Overdrive Recovery Time



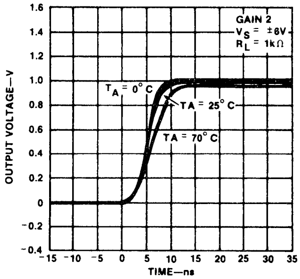
OP05680S

Pulse Response as a Function of Supply Voltage



OP05690S

Pulse Response as a Function of Temperature

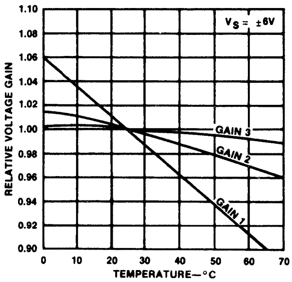


OP05700S

Differential Video Amplifier μA733/733C

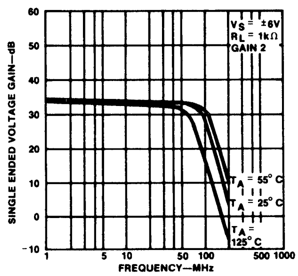
TYPICAL PERFORMANCE CHARACTERISTICS (Continued)

Voltage Gain as a Function of Temperature



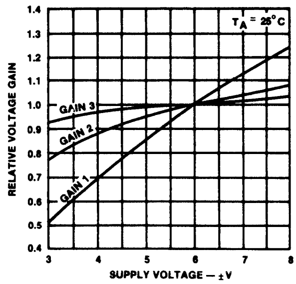
OP05710S

Gain vs Frequency as a Function of Temperature



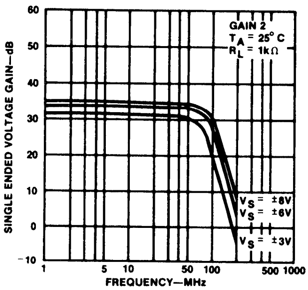
OP05720S

Voltage Gain as a Function of Supply Voltage



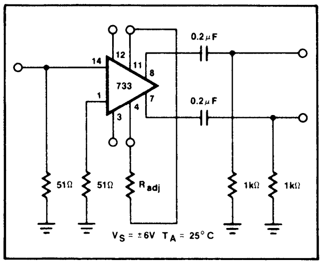
OP05730S

Gain vs Frequency as a Function of Supply Voltage



OP05740S

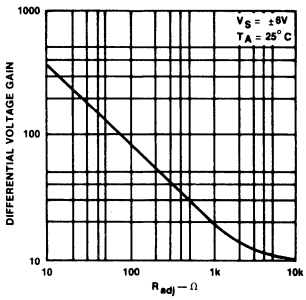
Voltage Gain Adjust Circuit



(Pin numbers apply to K Package)

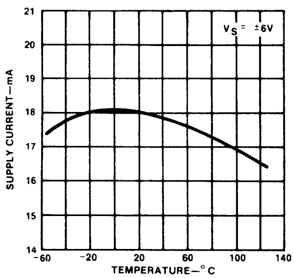
OP05750S

Voltage Gain as a Function of RADJ (Figure 3)



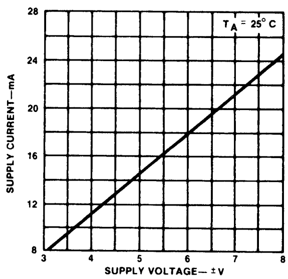
OP05760S

Supply Current as a Function of Temperature



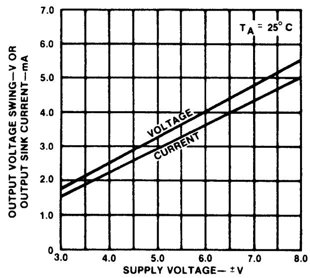
OP05770S

Supply Current as a Function of Supply Voltage



OP05780S

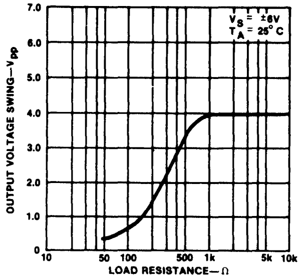
Output Voltage and Current Swing as a Function of Supply Voltage



OP05790S

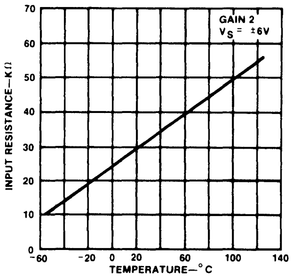
TYPICAL PERFORMANCE CHARACTERISTICS (Continued)

Output Voltage Swing as a Function of Load Resistance



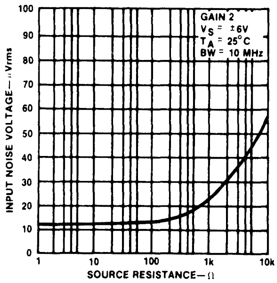
OP05800S

Input Resistance as a Function of Temperature



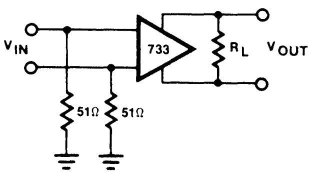
OP05810S

Input Noise Voltage as a Function of Source Resistance

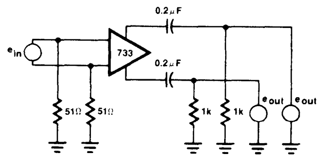


OP04590S

TEST CIRCUITS T_A = 25°C, unless otherwise specified.



TC08990S



TC09000S

11

Signetics

Section 12
Vertical Deflection

Linear Products

INDEX

12

Signetics

TDA2653A
Vertical Deflection

Product Specification

Linear Products

DESCRIPTION

The TDA2653A is a monolithic integrated circuit for vertical deflection in video monitors and large screen color television receivers, e.g. 30AX and PIL-S4 systems.

FEATURES

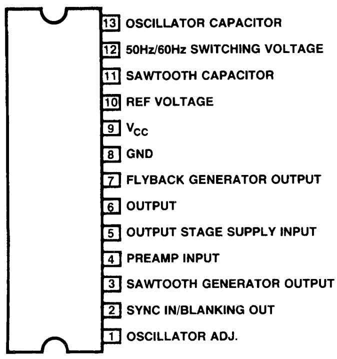
- **Oscillator; switch capability for 50Hz/60Hz operation**
- **Synchronization circuit**
- **Blanking pulse generator with guard circuit**
- **Sawtooth generator with buffer stage**
- **Preamplifier with fed-out inputs**
- **Output stage with thermal and short-circuit protection**
- **Flyback generator**
- **Voltage stabilizer**

APPLICATIONS

- **Video monitor**
- **Television receiver**

PIN CONFIGURATION

U Package



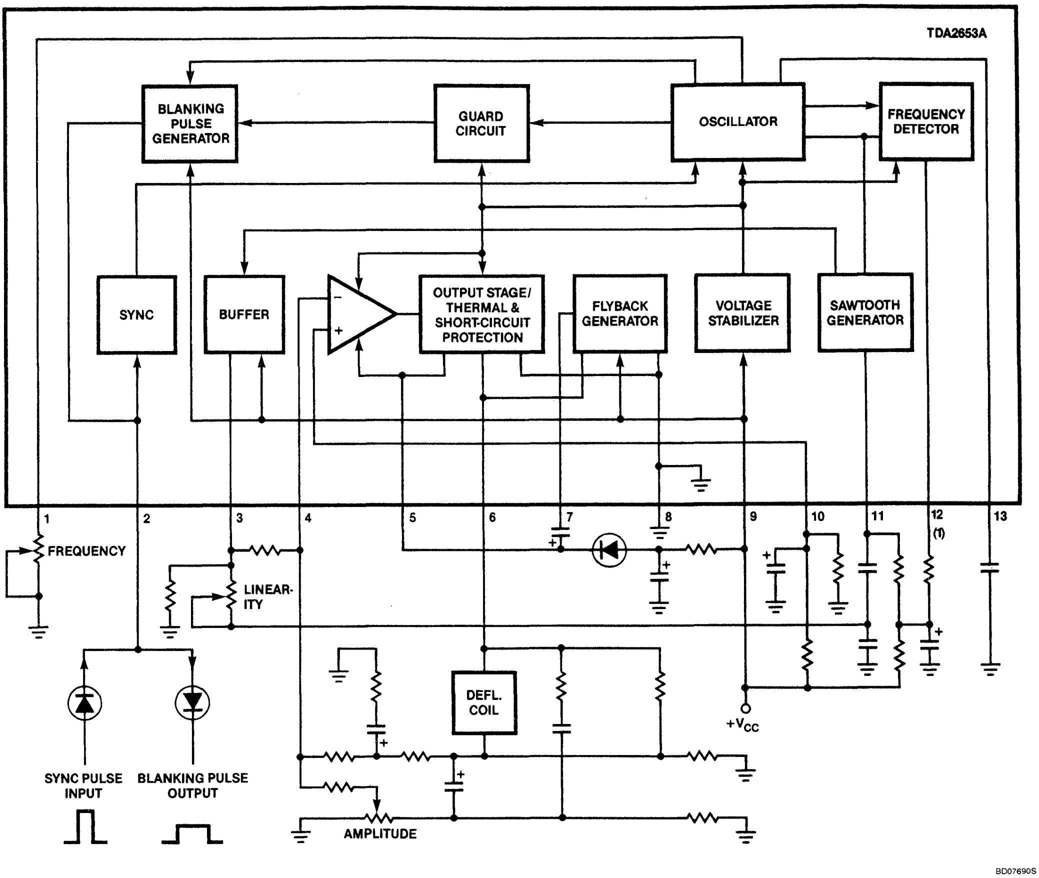
TOP VIEW

CD15070S

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
13-Pin Plastic SIP power package (SOT-141B)	−20°C to +85°C	TDA2653AU

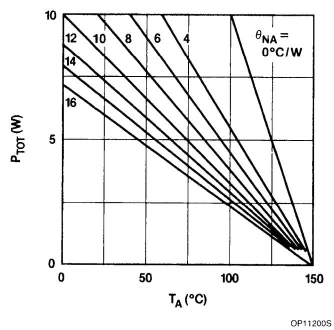
BLOCK DIAGRAM



NOTE:
1. Condition for Pin 12: LOW voltage level = 50Hz; HIGH voltage level = 60Hz.

Vertical Deflection TDA2653A

PIN NO.	DESCRIPTION
1, 13	Oscillator The oscillator frequency is determined by a potentiometer at Pin 1 and a capacitor at Pin 13.
2	Sync input/blanking output Combination of sync input and blanking output. The oscillator has to be synchronized by a positive-going pulse between 1V and 12V. The integrated frequency detector delivers a switching level at Pin 12. The blanking pulse amplitude is 20V with a load of 1mA.
3	Sawtooth generator output The sawtooth signal is fed via a buffer stage to Pin 3. It delivers the signal which is used for linearity control, and drive of the preamplifier. The sawtooth is applied via a shaping network to Pin 11 (linearity) and via a resistor to Pin 4 (preamplifier).
4	Preamplifier input The DC voltage is proportional to the output voltage (DC feedback). The AC voltage is proportional to the sum of the buffered sawtooth voltage at Pin 3 and the voltage, with opposite polarity, at the feedback resistor (AC feedback).
5	Positive supply of output stage This supply is obtained from the flyback generator. An electrolytic capacitor between Pins 7 and 5, and a diode between Pins 5 and 9 have to be connected for proper operation of the flyback generator.
6	Output of class-B power stage The vertical deflection coil is connected to this pin, via a series connection of a coupling capacitor and a feedback resistor, to ground.
7	Flyback generator output An electrolytic capacitor has to be connected between Pins 7 and 5 to complete the flyback generator.
8	Negative supply (ground) Negative supply of output stage and small signal part.
9	Positive supply The supply voltage at this pin is used to supply the flyback generator, voltage stabilizer, blanking pulse generator and buffer stage.
10	Reference voltage of preamplifier External adjustment and decoupling of reference voltage of the preamplifier.
11	Sawtooth capacitor This sawtooth capacitor has been split to realize linearity control.
12	50Hz/60Hz switching level This pin delivers a LOW voltage level for 50Hz and a HIGH voltage level for 60Hz. The amplitudes of the sawtooth signals can be made equal for 50Hz and 60Hz with these levels.



NOTE:
θ_{HA} includes θ_{MBH} which is expected when heatsink compound is used. $\theta_{JMB} \leqslant$ 5°C/W.

Figure 1. Total Power Dissipation

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
$V_9 = V_{CC}$	Supply voltage (Pin 9)	40	V
V_5	Supply voltage output stage (Pin 5)	58	V
Voltages			
V_3	Pin 3	7	V
V_{13}	Pin 13	7	V
$V_{4;\ 10}$	Pins 4 and 10	24	V
V_6	Pin 6	58	V
$-V_6$		0	V
$V_{7;\ 11}$	Pins 7 and 11	40	V
Currents			
I_1	Pin 1	0	mA
$-I_1$		1	mA
$\pm I_2$	Pin 2	10	mA
IP_3	Pin 3	0	mA
$-I_3$		5	mA
I_7	Pin 7	1.2	A
$-I_7$		1.5	A
I_{11}	Pin 11	50	mA
$-I_{11}$		1	mA
I_{12}	Pin 12	3	mA
$-I_{12}$		0	mA
T_{STG}	Storage temperature range	−25 to +150	°C
T_A	Operating ambient temperature range	−20 to limiting value	°C

NOTES:
1. Pins 5, 6 and 8: internally limited by the short-circuit protection circuit.
2. Total power dissipation: internally limited by the thermal protection circuit.

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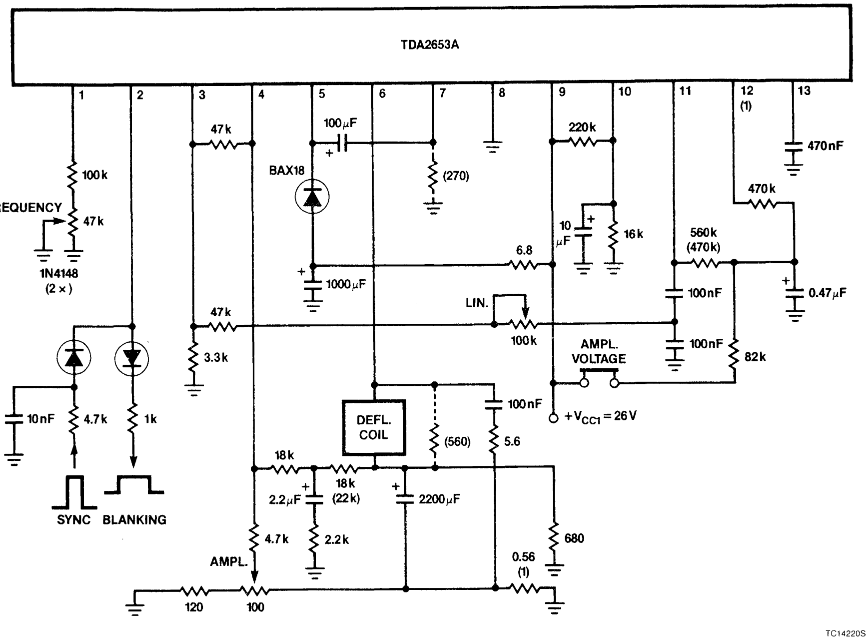
DC ELECTRICAL CHARACTERISTICS T^A = 25°C, unless otherwise specified.

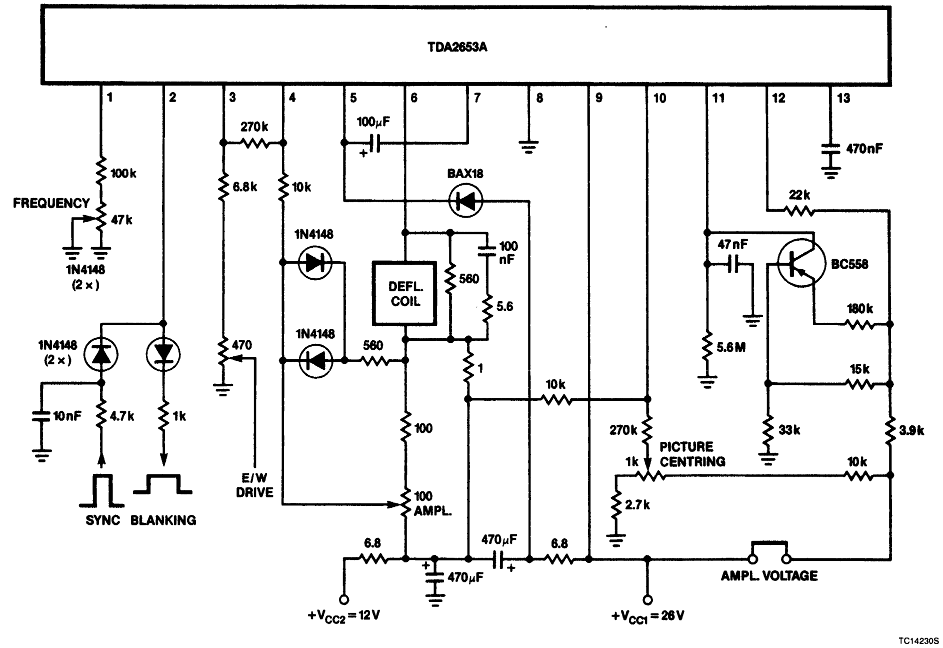
SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
$V_9 = V_{CC}$	Supply voltage	9		30	
	Output voltage				
V_6	at $-I_6 = 1.1A$	$V_5-2.2$	$V_5-1.9$		V
V_6	at $I_6 = 1.1A$		1.3	1.6	V
V_7	Flyback generator output voltage at $-I_6 = 1.1A$		$V_{CC}-2.2$		V
$\pm I_6$	Peak output current			1.2	A
$\pm I_7$	Flyback generator peak current			1.2	A
Feedback					
$-I_{4,\ 10}$	Input quiescent current		0.1		μA
Synchronization					
V_2	Sync input pulse	1		12	V
	Tracking range		28		%
Oscillator/sawtooth generator					
V_1	Oscillator frequency control input voltage	6		9	V
V_3	Sawtooth generator output voltage	0		V_{CC-1}	V
V_{11}		0		V_{CC-2}	V
$-I_3$	Sawtooth generator output current	0		4	mA
		−2			μA
I_{11}				+30	mA
$(\Delta f/f)/\Delta T_{CASE}$	Oscillator temperature dependency T_{CASE} = 20 to 100°C		10^4		°C
$(\Delta f/f)/\Delta V_S$	Oscillator voltage dependency V_S = 10 to 30V		4×10^4		V^{-1}
Blanking pulse generator					
V_2	Output voltage at $V_S = 24V$; $I_2 = 1mA$		18.5		V
$-I_2$	Output current			3	mA
R_2	Output resistance		410		Ω
t_B	Blanking pulse duration at 50Hz sync		1.4 ± 0.07		ms
50Hz/60Hz switch capability					
V_{12}	Saturation voltage; LOW voltage level		1		V
I_{12}	Output leakage current		1		μA



NOTES:

1. Condition for Pin 12: LOW voltage level = 50Hz; HIGH voltage level = 60Hz.
2. The values given in parentheses and the dotted components are valid for the PIL-S4 system.

Figure 2. Typical Vertical Deflection Circuit for 30AX System (26V)



NOTES:
1. Condition for Pin 12: LOW voltage level = 50Hz; HIGH voltage level = 60Hz.
2. V_{CC1} = 26V, V_{CC2} = 12V in Quasi-bridge Connection.

Figure 3. Typical Vertical Deflection Circuit for 30AX System

Data Measured in Figures 2 and 3

SYMBOL	PARAMETER		30AX SYSTEM (26V) (Figure 2)	30AX SYSTEM (26 V/12V) (Figure 3)	PIL-S4 SYSTEM (Figure 2)
V_{S1}	System supply voltages	typ	26	26	26V
V_{S2}		typ		12	−V
I_{S1}	System supply currents	typ	315	330	195mA
I_{S2}		typ		−35	−mA
V_{6-8}	Output voltage	typ	14	14.6	13.5V
V_{6-8}	Output voltage (peak value)	typ	42	42	49V
$I_{6(P-P)}$	Deflection current (peak-to-peak value)	typ	2.2	2.2	1.32A
t_{FL}	Flyback time	typ	1	0.9	1.1ms
P_{TO}	Total power dissipation per package	typ	4.1	4	3W
		max	4.8	4.8	3.4W[1]
f	Oscillator frequency unsynchronized	typ	46.5	46.5	46.5Hz

NOTE:
1. Calculated with $\Delta V_S = +5\%$ and $\Delta R_{YOKE} = -7\%$.

Signetics

TDA3654
Vertical Deflection Output Circuit

Product Specification

Linear Products

DESCRIPTION

The TDA3654 is a full-performance vertical deflection output circuit in a 9-lead, single in-line encapsulation. The circuit is intended for direct drive of the deflection coils and it can be used for a wide range of 90° and 110° deflection systems.

The TDA3654 is provided with a guard circuit which blanks the picture tube screen in case of absence of the deflection current.

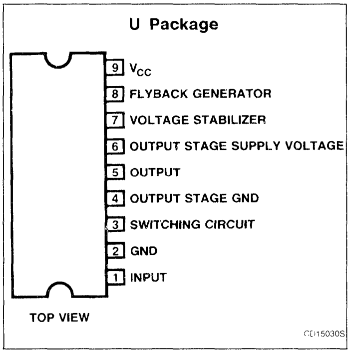
FEATURES

- **Direct drive to the deflection coils**
- **90° and 110° deflection system**
- **Internal blanking guard circuit**
- **Internal voltage stabilizer**

APPLICATIONS

- **Video monitors**
- **TV receivers**

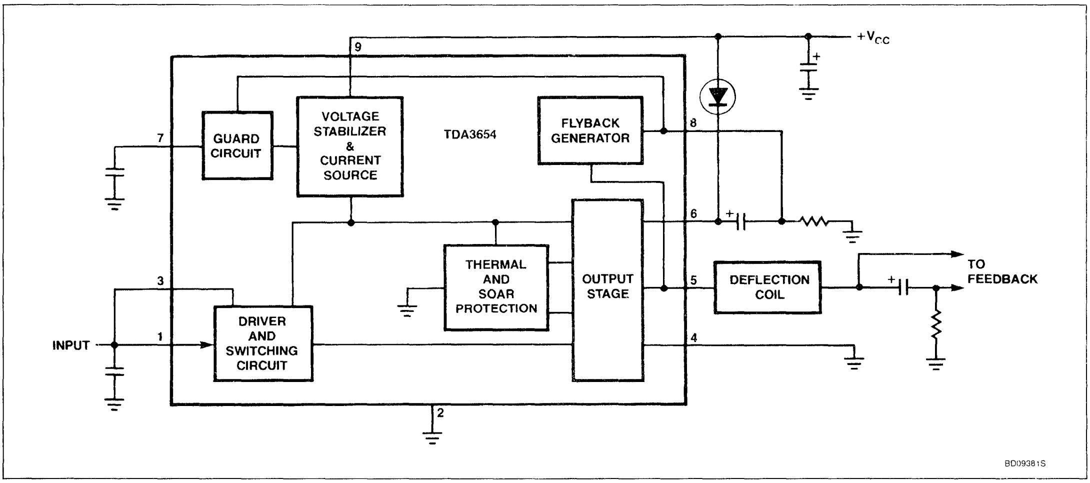
PIN CONFIGURATION



ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
9-Pin Plastic SIP (SOT-131B)	−25°C to +60°C	TDA3654U
9-Pin Plastic SIP (SOT-157B)	−25°C to +60°C	TDA3654AU

BLOCK DIAGRAM



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ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
Voltages			
V_{5-4}	Output voltage	60	V
V_{9-4}	Supply voltage	40	V
V_{6-4}	Supply voltage output stage	60	V
V_{1-2}	Input voltage	V_{9-4}	V
V_{3-2}	Input voltage switching circuit	V_{9-4}	V
V_{7-2}	External voltage at Pin 7	5.6	V
Currents			
$\pm I_{5RM}$	Repetitive peak output current	1.5	A
$\pm I_{5SM}$	Non-repetitive peak output current[1]	3	A
I_{8RM}	Repetitive peak output current of flyback generator	+1.5 −1.6	A A
$\pm I_{8SM}$	Non-repetitive peak output current of flyback generator[1]	3	A
Temperatures			
T_{STG}	Storage temperature range	−65 to +150	°C
T_A	Operating ambient temperature range (see Figure 2)	−25 to +60	°C
T_J	Operating junction temperature range	−25 to +150	°C
θ_{JMB}	Thermal resistance	4	°C/W

NOTE:
1. Pins 2 and 4 are externally connected to ground.

DC AND AC ELECTRICAL CHARACTERISTICS $T_A = 25°C$, supply voltage (V_{9-4}) = 26V, unless otherwise stated.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Supply					
V_{9-4}	Supply voltage, Pin 9[2]	10		40	V
V_{6-4}	Supply voltage output stage			60	V
I_6+I_9	Supply current, Pins 6 and 9[3]	35	55	85	mA
I_4	Quiescent current[4]	25	40	65	mA
TC	Variation of quiescent current with temperature		−0.04		mA/°C
Output current					
$I_{5(P-P)}$	Output current, Pin 5 (peak-to-peak)		2.5	3	A
$+I_{8(P-P)}$ $-I_{8(P-P)}$	Output current flyback generator, Pin 8		1.25 1.35	1.5 1.6	A A
Output voltage					
V_{5-4}	Peak voltage during flyback			60	V
 $V_{6-5(SAT)}$ $V_{5-6(SAT)}$ $V_{6-5(SAT)}$ $V_{5-6(SAT)}$	Saturation voltage to supply at $I_5 = -1.5A$ at $I_5 = 1.5A$[5] at $I_5 = -1.2A$ at $I_5 = 1.2A$[5]		 2.5 2.5 2.2 2.3	 3.2 3.2 2.7 2.8	 V V V V
 $V_{5-4(SAT)}$ $V_{5-4(SAT)}$	Saturation voltage to ground at $I_5 = 1.2A$ at $I_5 = 1.5A$		 2.2 2.5	 2.7 3.2	 V V
Flyback generator					
 $V_{9-8(SAT)}$ $V_{8-9(SAT)}$ $V_{9-8(SAT)}$ $V_{8-9(SAT)}$	Saturation voltage at $I_8 = -1.6A$ at $I_8 = 1.5A$[5] at $I_8 = -1.3A$ at $I_8 = 1.2A$[5]		 1.6 2.3 1.4 2.2	 2.1 3 1.9 2.7	 V V V V
$-I_8$	Leakage current at Pin 8		5	100	μA
V_{5-9}	Flyback generator active IF	4			V
Input					
I_1	Input current, Pin 1, for $I_5 = 1.5A$		0.33	0.55	mA
V_{1-2}	Input voltage during scan, Pin 1		2.35	3	V
I_3	Input current, Pin 3, during scan[6]	0.03			mA
V_{3-2}	Input voltage, Pin 3, during scan[6]	0.8		V_{9-4}	V
V_{1-2}	Input voltage, Pin 1, during flyback			250	mV
V_{3-2}	Input voltage, Pin 3, during flyback			250	mV
Guard circuit					
V_{7-2}	Output voltage, Pin 7, $R_L = 100k\Omega$[9]	4.1	4.5	5.5	V
V_{7-2}	Output voltage, Pin 7, at $I_L = 0.5mA$[9]	3.4	3.9	5.1	V
R_{I7}	Internal series resistance of Pin 7	0.95	1.35	1.7	kΩ
V_{8-2}	Guard circuit activates[7]			1.0	V
General data					
T_J	Thermal protection activation range	158	175	192	°C

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Vertical Deflection Output Circuit — TDA3654

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) $T_A = 25°C$, supply voltage $(V_{9-4}) = 26V$, unless otherwise stated.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Thermal resistance					
θ_{JMB}	From junction to mounting base		3.5	4	°C/W
P_{TOT}	Power dissipation		see Figure 2		
G_O	Open-loop gain at 1kHz[8]		33		dB
f_R	Frequency response, −3dB[10]		60		kHz

NOTES:
1. Non-repetitive duty factor 3.3%.
2. The maximum supply voltage should be chosen so that during flyback the voltage at Pin 5 does not exceed 60V.
3. When V_{5-4} is 13V and no load at Pin 5.
4. See Figure 3.
5. Duty cycle, d = 5% or d = 0.05.
6. When Pin 3 is driven separately from Pin 1.
7. During normal operation the voltage V_{8-2} may not be lower than 1.5V.
8. $R_L = 8\Omega$; $I_L = 125mA_{RMS}$
9. If guard circuit is active.
10. With a 22pF capacitor between Pins 1 and 5.

FUNCTIONAL DESCRIPTION

Output Stage and Protection Circuits

The output stage consists of two Darlington configurations in class B arrangement. Each output transistor can deliver 1.5A maximum and the V_{CEO} is 60V. Protection of the output stage is such that the operation of the transistors remains well within the SOA area in all circumstances at the output pin (Pin 5). This is obtained by the cooperation of the thermal protection circuit, the current-voltage detector, and the short-circuit protection. Special measures in the internal circuit layout give the output transistors extra solidity; this is illustrated in Figure 4, where typical SOA curves of the lower output transistors are given. The same curves also apply for the upper output device. The supply for the output stage is fed to Pin 6 and the output stage ground is connected to Pin 4.

Driver and Switching Circuit

Pin 1 is the input for the driver of the output stage. The signal at Pin 1 is also applied to Pin 3 which is the input of a switching circuit (Pins 1 and 3 are externally connected). This switching circuit rapidly turns off the lower output stage when the flyback starts, and therefore, allows a quick start of the flyback generator. The maximum required input signal for the maximum output current peak-to-peak value of 3A is only 3V; the sum of the currents in Pins 1 and 3 is then maximum 1mA.

Flyback Generator

During scan, the capacitor between Pins 6 and 8 is charged to a level which is dependent on the value of the resistor at Pin 8 (see Block Diagram). When the flyback starts and the voltage at the output pin (Pin 5) exceeds the supply voltage, the flyback generator is activated.

The supply voltage is then connected in series, via Pin 8, with the voltage across the capacitor during the flyback period. This implies that during scan the supply voltage can be reduced to the required scan voltage plus saturation voltage of the output transistors.

The amplitude of the flyback voltage can be chosen by changing the value of the external resistor at Pin 8. It should be noted that the application is chosen such that the lowest voltage at Pin 8 is > 1.5V during normal operation.

Guard Circuit

When there is no deflection current, for any reason, the voltage at Pin 8 becomes less than 1V and the guard circuit will produce a DC voltage at Pin 7. This voltage can be used to blank the picture tube so that the screen will not burn in.

Voltage Stabilizer

The internal voltage stabilizer provides a stabilized supply of 6V to drive the output stage, so the drive current is not affected by supply voltage variations.

Vertical Deflection Output Circuit | TDA3654

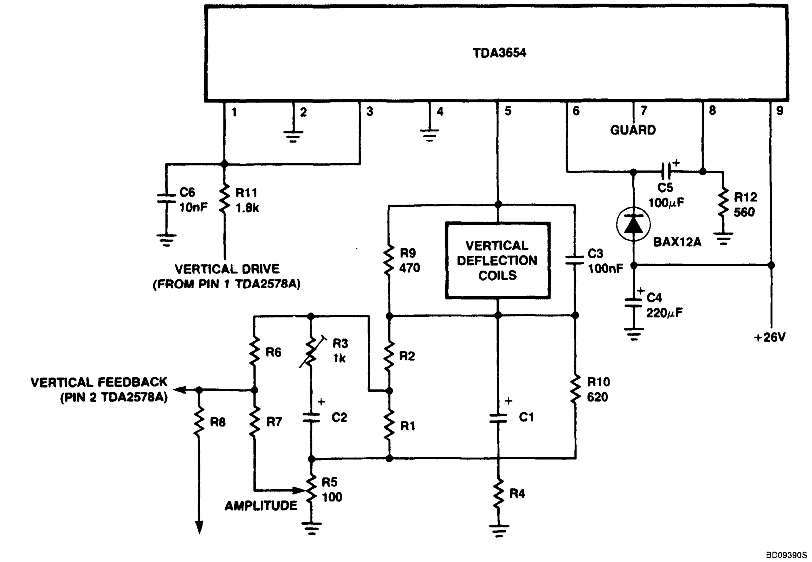

Figure 1. Application Diagram

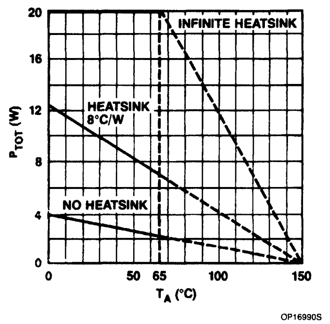

Figure 2. Power Derating Curve

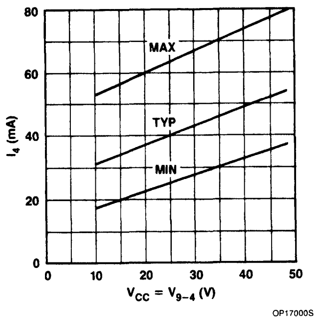

Figure 3. Quiescent Current as a Function of the Supply Voltage

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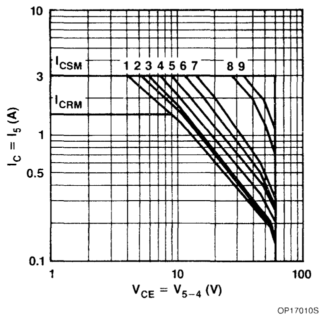
Vertical Deflection Output Circuit

TDA3654

CURVE	t_P	δ	PEAK JUNCTION TEMPERATURE
1	DC		150°C
2	10ms	0.5	150°C
3	10ms	0.25	150°C
4	1ms	0.5	150°C
5	1ms	0.25	150°C
6	1ms	0.05	150°C
7	1ms	0.05	180°C
8	0.2ms	0.1	150°C
9	0.2ms	0.1	180°C



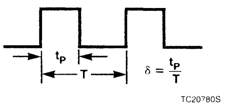


Figure 4. Typical SOA of Lower Output Transistor

Signetics

Section 13
SMPS for TV/Monitor

Linear Products

INDEX

Signetics

TDA2582
Control Circuit For Power Supplies

Product Specification

Linear Products

DESCRIPTION

The TDA2582 is a monolithic integrated circuit for controlling power supplies which are provided with the drive for the horizontal deflection stage.

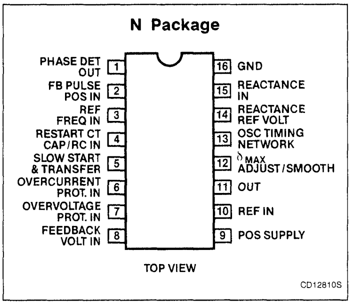
FEATURES

- **Voltage-controlled horizontal oscillator**
- **Phase detector**
- **Duty factor control for the negative-going transient of the output signal**
- **Duty factor increases from zero to its normal operation value**
- **Adjustable maximum duty factor**
- **Overvoltage and overcurrent protection with automatic restart after switch-off**
- **Counting circuit for permanent switch-off when n-times overcurrent or overvoltage is sensed**
- **Protection for open-reference voltage**
- **Protection for too-low supply voltage**
- **Protection against loop faults**
- **Positive tracking of duty factor and feedback voltage when the feedback voltage is smaller than the reference voltage minus 1.5V**
- **Normal and "smooth" remote ON/OFF possibility**

APPLICATIONS

- **Video monitors**
- **Power supplies**

PIN CONFIGURATION



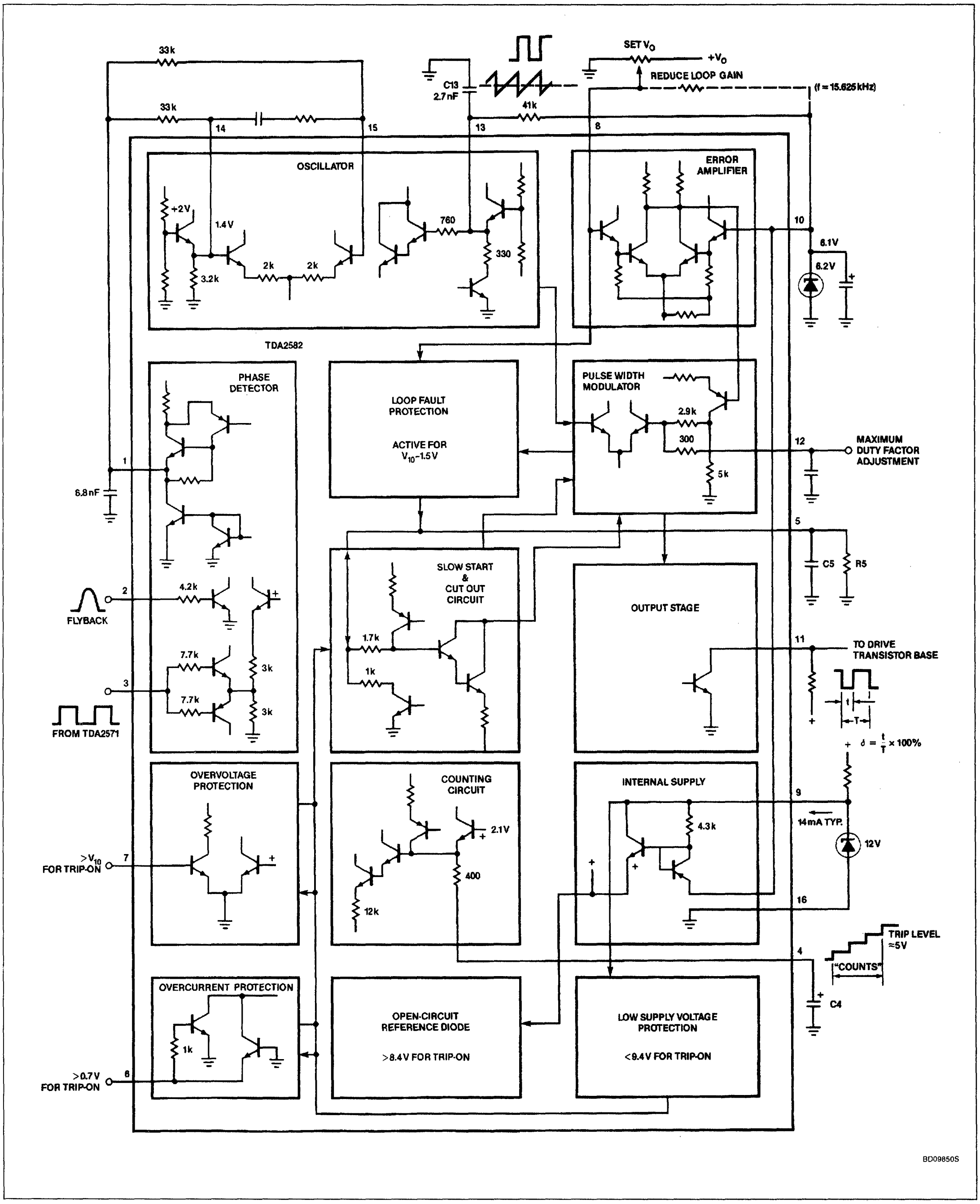
ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
16-Pin Plastic DIP (SOT-38)	-25°C to +80°C	TDA2582N

ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{9-16}	Supply voltage at Pin 9	14	V
V_{11-16}	Voltage at Pin 11	0 to 14	V
I_{11M}	Output current (peak value)	40	mA
P_{TOT}	Total power dissipation	280	mW
T_{STG}	Storage temperature	-65 to +150	°C
T_A	Operating ambient temperature	-25 to +80	°C

BLOCK DIAGRAM



DC ELECTRICAL CHARACTERISTICS V_{CC} = 12V; $V_{10\text{-}16}$ = 6.1V; T_A = 25°C, measured in Figure 3.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
V_{9-16}	Supply voltage range	10	12	14	V
V_{9-16}	Protection voltage too-low supply voltage	8.6	9.4	9.9	V
I_9	Supply current at $\delta = 50\%$		14		mA
I_9	Supply current during protection		14		mA
I_9	Minimum required supply current[1]			17	mA
P	Power consumption		170		mW
Required input signals					
V_{10-16}	Reference voltage[2]	5.6	6.1	6.6	V
$\lvert Z_{8-16} \rvert$	Feedback input impedance		200		kΩ
V_{10-16}	High reference voltage protection: threshold voltage	7.9	8.4	8.9	V
 $V_{3-16(P-P)}$ I_{3M} $\pm I_3$	Horizontal reference signal (square-wave or differentiated; negative transient is reference) voltage-driven (peak-to-peak value) current-driven (peak value) switching-level current	 5 −1		 12 1.5 100	 V mA μA
V_{2-16}	Flyback pulse or differential deflection current	1		5	V
I_{2M}	Flyback pulse current (peak value)			1.5	mA
$-V_{6-16}$ $+V_{6-16}$	Overcurrent protection:[3] threshold voltage	600 640	640 680	695 735	mV mV
V_{7-16}	Overvoltage protection: ($V_{REF} = V_{10-16}$) threshold voltage	$V_{REF}-130$	$V_{REF}-60$	$V_{REF}-0$	mV
V_{4-16}	Remote-control voltage; switch-off[4]	5.6			V
V_{4-16}	Remote-control voltage; switch-on			4.5	V
V_{5-16}	'Smooth' remote control; switch-off[5]	4.5			V
V_{5-16}	'Smooth' remote control; switch-on			3	V
I_4	Remote-control switch-off current			1	mA
Delivered output signals					
$V_{11-16(P-P)}$	Horizontal drive pulse (loaded with a resistor of 560Ω to +12V peak-to-peak value	11.6			V
I_{11M}	Output current; peak value			40	mA
 V_{CESAT} V_{CESAT}	Saturation voltage of output transistor at I_{11} = 20mA at I_{11} = 40mA		 200	 400 525	 mV mV
δ	Duty factor of output pulse[6]	0		98 ±0.8	%
I_4	Charge current for capacitor on Pin 4		110		μA
I_5	Charge current for capacitor on Pin 5		120		μA
I_{10}	Supply current for reference	0.6	1	1.45	mA

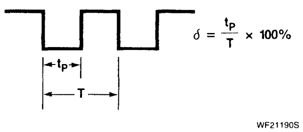
Control Circuit For Power Supplies TDA2582

DC ELECTRICAL CHARACTERISTICS (Continued) $V_{CC} = 12V$; $V_{10\text{-}16} = 6.1V$; $T_A = 25°C$, measured in Figure 3.

SYMBOL	PARAMETER	LIMITS			UNIT
		Min	Typ	Max	
Oscillator					
	Temperature coefficient		0.0003	0.0004	$°C^{-1}$
	Relative frequency deviation for V_{10-16} changing from 5.6 to 6.6V		−1.4	−2	%
	Oscillator frequency spread (with fixed external components)			3	%
	Frequency control sensitivity at Pin 15 $f_{NOM} = 15.625kHz$		5		kHz/V
Phase control loop					
	Loop gain of APC-system (automatic phase control)[7]		5		kHz/μs
Δf	Catching range (f_{NOM} = 15,625kHz)	1300		2100	Hz
t	Phase relation between negative transient of sync pulse and middle of flyback		1		μs
Δt	Tolerance of phase relation			±0.4	μs

NOTES:

1. This value refers to the minimum required supply current that will start all devices under the following conditions: $V_{9-16} = 10V$; $V_{10-16} = 6.2V$; $\delta = 50\%$.
2. Voltage obtained via an external reference diode. Specified voltages do not refer to the nominal voltages of reference diodes.
3. This spread is inclusive temperature rise of the IC due to warming up. For other ambient temperatures the values must be corrected by using a temperature coefficient of typical − 1.85mV/°C.
4. See application information Pin 4.
5. See application information Pin 5.
6. The duty factor is specified as follows: $\delta = \frac{t_P}{T} \times 100\%$ (see Figure 1). After switch-on, the duty factor rises gradually from 0% to the steady value. The relationship between V_{8-16} and the duty factor is given in Figure 6 and the relationship between V_{12-16} and the duty factor is shown in Figure 8.
7. For component values, see Block Diagram.



NOTE:

$\delta = \frac{t_P}{T} \times 100\%$.

Figure 1

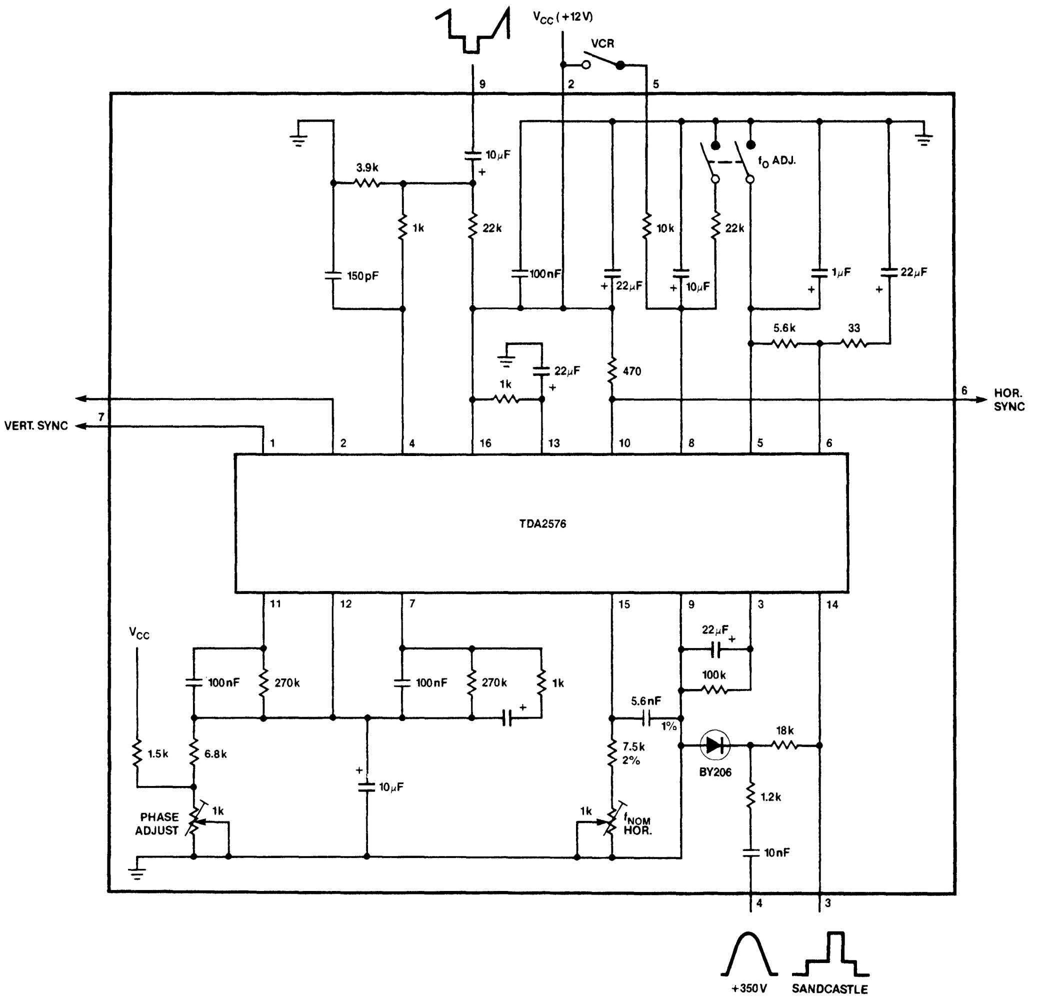


Figure 2a. Lead 6 (Pin 10) of Circuit TDA2576 Connected to Lead 2 (Pin 14) of Circuit TDA2582

13

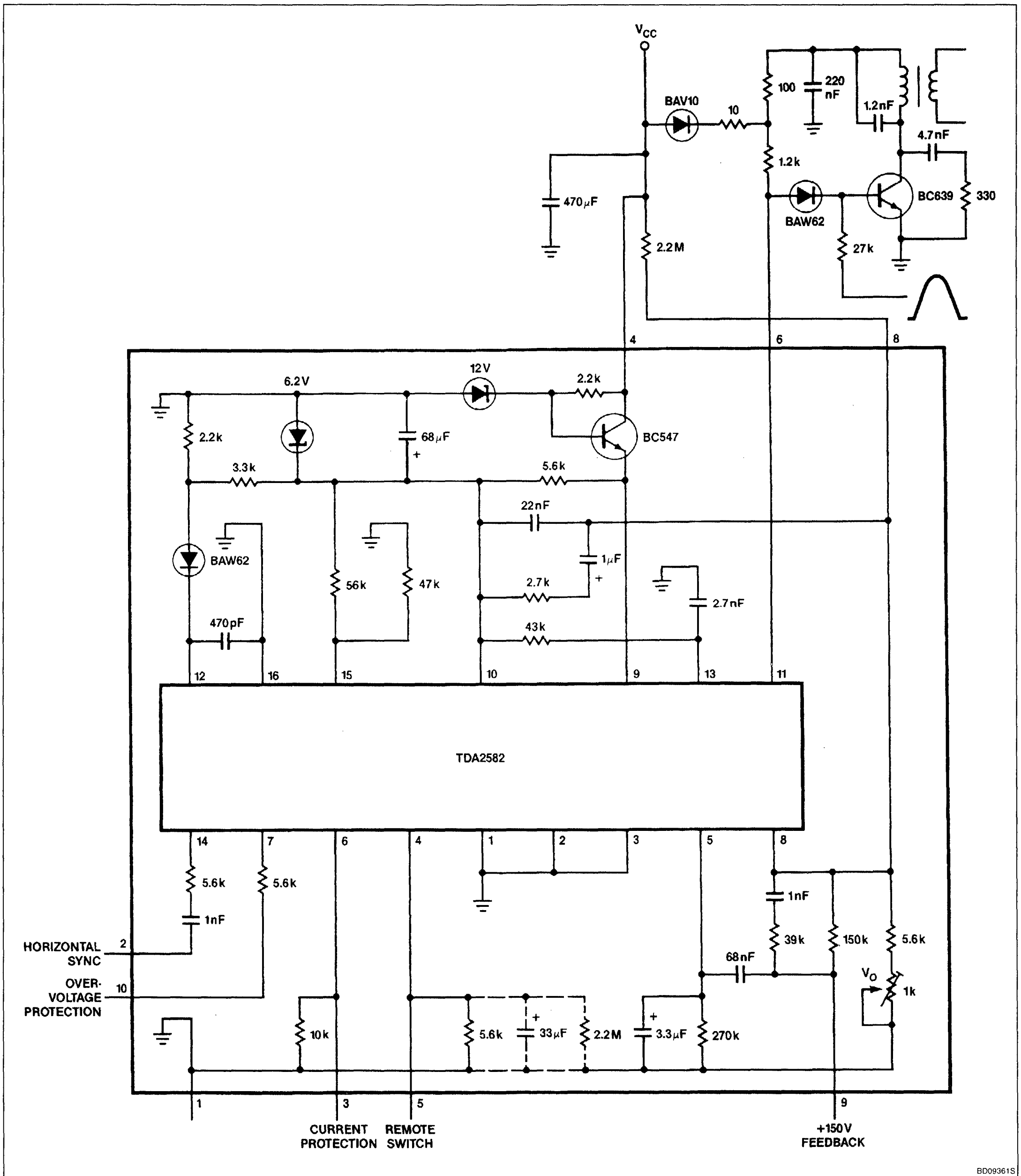
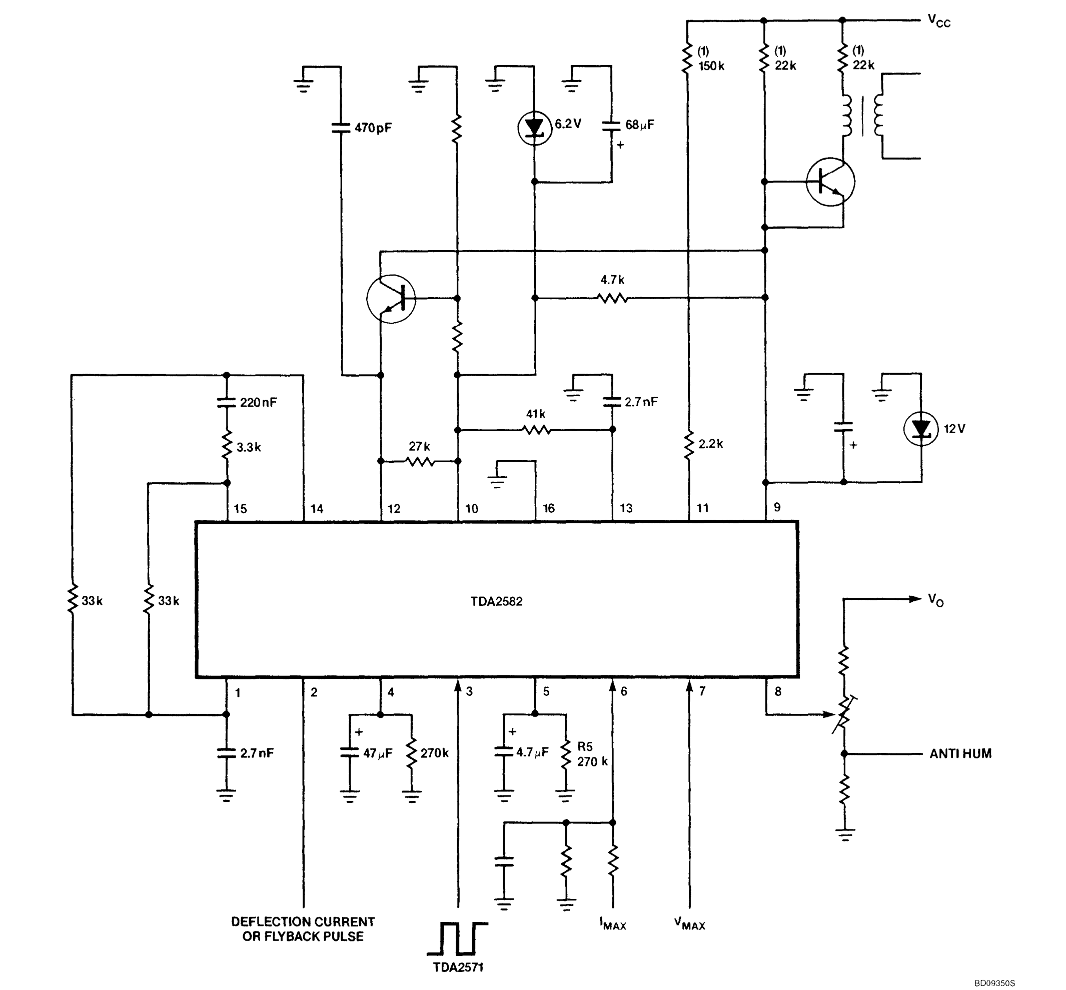


Figure 2b. Lead 6 (Pin 10) of Circuit TDA2576 Connected to Lead 2 (Pin 14) of Circuit TDA2582



NOTE:
1. Values depend on V_{CC}.

Figure 3. Circuit Diagram

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APPLICATION INFORMATION

The function is described beside the corresponding pin number.

1 Phase Detector Output — The output circuit consists of a bidirectional current source which is active for the time that the signal on Pin 2 exceeds 1V.

The current values are chosen such that the correct phase relation is obtained when the output signal of the TDA2571 is applied to Pin 3.

With a resistor of 2 × 33kΩ and a capacitor of 2.7nF, the control steepness is 0.55V/μs (Figure 3).

2 Flyback Pulse Input — The signal applied to Pin 2 is normally a flyback pulse with a duration of about 12μs. However, the phase detector system also accepts a signal derived by differentiating the deflection current by means of a small toroidal core (pulse duration > 3μs).

The toroidal transformer in Figure 4a is for obtaining a pulse representing the midflyback from the deflection current. The connection of the picture phase information is shown in Figure 4b.

3 Reference Frequency Input — The input circuit can be driven directly by the square-wave output voltage from Pin 8 of the TDA2571.

The negative-going transient switches the current source connected to Pin 1 from positive to negative.

The input circuit is made such that a differentiated signal of the square-wave from the TDA2571 is also accepted (this enables power line isolation). The input circuit switching level is about 3V and the input impedance is about 8kΩ.

4 Restart Count Capacitor/Remote-Control Input —

Counting

An external capacitor (C4 = 47μF) is connected between Pins 4 and 16. This capacitor controls the characteristics of the protection circuits as follows.

If the protection circuits are required to operate, e.g., overcurrent at Pin 6, the duty factor will be set to zero, thus turning off the power supply.

After a short interval (determined by the time constant on Pin 5), the power supply will be restarted via the slow-start circuit.

If the fault condition has cleared, then normal operation will be resumed. If the fault condition is persistent, the duty factor of the pulses is again reduced to zero and the protection cycle is repeated.

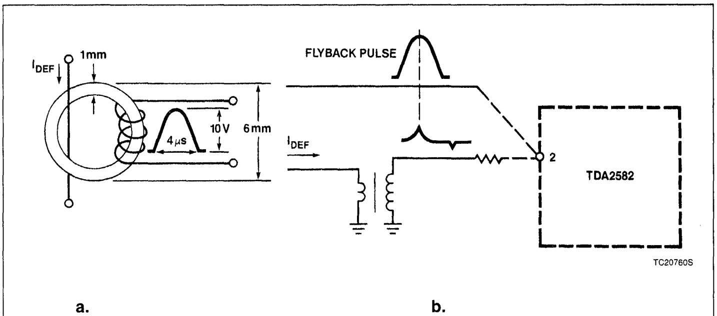

Figure 4

The number of times this action is repeated (n) for a persisting fault condition is now determined by: n = C4/C5.

Remote Control Input

For this application, the capacitor on Pin 4 has to be replaced by a resistor with a value between 4.7 and 18kΩ. When the externally-applied voltage $V_{4\text{-}16} > 5.6V$, the circuit switches off; switching on occurs when $V_{4\text{-}16} < 4.5V$ and the normal starting-up procedure is followed. Pin 4 is internally connected to an emitter-follower, with an emitter voltage of 1.5V.

5 Slow-Start and Transfer Characteristics for Low Feedback Voltage —

Slow-Start

An external shunt capacitor (C5 = 4.7μF) and resistor (R5 = 270kΩ) are connected between Pins 5 and 16. The network controls the rate at which the duty factor increases from zero to its steady-state value after switch-on. It provides protection against surges in the power transistor.

Transfer Characteristic for Low Feedback Voltages

The duty factor transfer characteristic for low feedback voltages can be influenced by R5.

The transfer for three different resistor values is given in Figure 6.

'Smooth' Remote ON/OFF

The ON/OFF information should be applied to Pin 5 via a high-ohmic resistor; a high OFF-level gives a slow rising voltage at Pin 5, which results in a slowly decreasing duty factor.

6 Overcurrent Protection Input — A voltage proportional to the current in the power switching device is applied to the integrated circuit between Pins 6 and 16. The circuit trips on both positive and negative polarity. When the tripping level is reached, the output pulse is immediately blocked and the starting circuit is activated again.

7 Over voltage Protection Input — When the voltage applied to this pin exceeds the threshold level, the protection circuit will operate.

The tripping level is about the same as the reference voltage on Pin 10.

8 Feedback Voltage Input — The control loop input is applied to Pin 8. This pin is internally connected to one input of a differential amplifier, functioning as an amplitude comparator, the other input of which is connected to the reference source on Pin 10.

Under normal operating conditions, the voltage on Pin 8 will be about equal to the reference voltage on Pin 10. For further information refer to Figures 6 and 7.

9 12V Positive Supply — The maximum voltage that may be applied is 14V. Where this is derived from an unstabilized supply rail, a regulator diode (12V) should be connected between Pins 9 and 16 to ensure that the maximum voltage does not exceed 14V. When the voltage on this pin falls below a minimum of 8.6V (typically 9.4V), the protection circuit will switch off the power supply.

10 Reference Input — An external reference diode must be connected between this pin and Pin 16.

The reference voltage must be between 5.6 and 6.6V. The IC delivers about 1mA into the external regulator diode. When the external load on the regulator diode approaches this current, replenishment of the current can be obtained by connecting a suitable resistor between Pins 9 and 10. A higher reference-voltage value up to 7.5V is allowed when use is made of a duty factor limiting resistor < 27kΩ between Pins 12 and 16.

11 Output — An external resistor determines the output current fed into the base of the driver transistor. The output circuit uses an NPN transistor with 3 series-connected clamping diodes to the internal 12V supply rail. This provides a low-impedance in the "ON" state, that is, with the drive transistor turned off.

12 Maximum Duty-Factor Adjustment/ Smoothing

Maximum Duty-Factor Adjustment

Pin 12 is connected to the output voltage of the amplitude comparator (V_{10-8}). This voltage is internally connected to one input of a differential amplifier, the other input of which is connected to the sawtooth voltage of the horizontal oscillator. A high voltage on Pin 12 results in a low duty factor. This enables the maximum duty factor to be adjusted by limiting the voltage by connecting Pin 12 to the emitter of an NPN transistor used as a voltage source.

Figure 8 plots the maximum duty factor as a function of the voltage applied to Pin 12. If some spread is acceptable, the maximum duty factor can also be limited by connecting a resistor from Pin 12 to Pin 16. A resistor of 12kΩ limits the maximum duty factor to about 50%. This application also reduces the total IC gain.

Smoothing

Any double pulsing of the IC due to circuit layout can be suppressed by connecting a capacitor of about 470pF between Pins 12 and 16.

13 Oscillator Timing Network — The timing network comprises a capacitor between Pins 13 and 16, and a resistor between Pin 13 and the reference voltage on Pin 10.

The charging current for the capacitor (C13) is derived from the voltage reference diode connected to Pin 10 and discharged via an internal resistor of about 330Ω.

14 Reactance-Stage Reference Voltage — This pin is connected to an emitter-follower which determines the nominal reference voltage for the reactance stage (1.4V for reference voltage $V_{10-16} = 6.1V$). Free-running frequency is obtained when Pins 14 and 15 are short-circuited.

15 Reactance-Stage Input — The output voltage of the phase detector (Pin 1) is connected to Pin 15 via a resistor. The voltage applied to Pin 15 shifts the upper level of the voltage sensor of the oscillator, thus changing the oscillator frequency and phase. The time-constant network is connected between Pins 14 and 15. Control sensitivity is typically 5kHz/V.

16 Negative Supply (Ground)

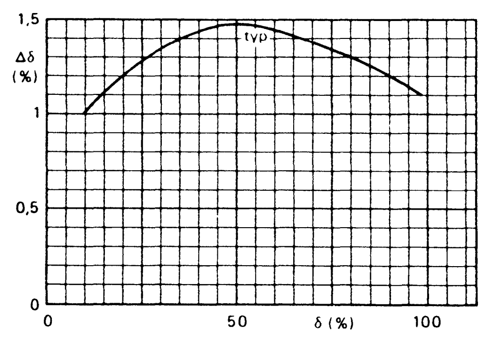

Figure 5. Duty Factor Change as a Function of Initial Duty Factor; at 1mV Error Amplifier Input Change; $\Delta V_{8-10(P-P)} = 1mV$

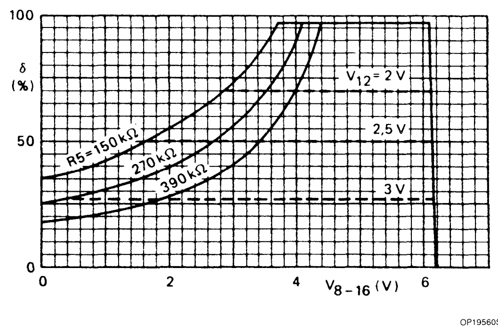

Figure 6. Duty Factor of Output Pulses as a Function of Feedback Input Voltage (V_{8-16}) With R5 as a Parameter and V_{12-16} as a Limiting Value; $V_{10-16} = 6.1V$

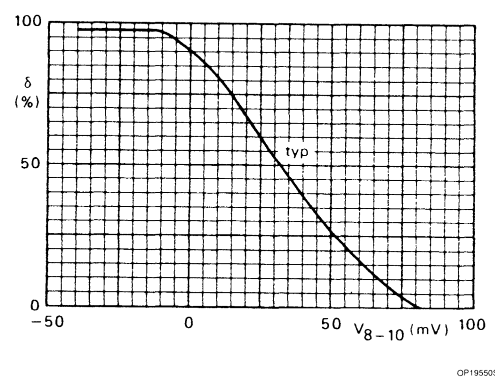

Figure 7. Duty Factor of Output Pulses as a Function of Error Amplifier Input (V_{8-10}); $V_{10-16} = 6.1V$

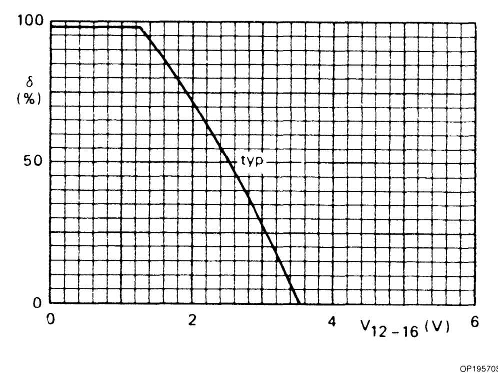

Figure 8. Maximum Duty Factor Limitation as a Function of the Voltage Applied to Pin 12; $V_{10-16} = 6.1V$

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Signetics

TEA1039

Control Circuit for Switched-Mode Power Supply

Product Specification

Linear Products

DESCRIPTION

The TEA1039 is a bipolar integrated circuit intended for the control of a switched-mode power supply. Together with an external error amplifier and a voltage regulator (e.g., a regulator diode) it forms a complete control system. The circuit is capable of directly driving the SMPS power transistor in small SMPS systems.

FEATURES

- **Wide frequency range**
- **Adjustable input sensitivity**
- **Adjustable minimum frequency or maximum duty factor limit**
- **Adjustable overcurrent protection limit**
- **Supply voltage out-of-range protection**
- **Slow-start facility**

APPLICATIONS

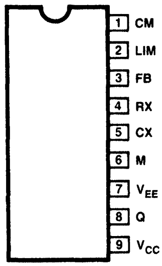
- **Home appliances**
- **Frequency regulation**
- **Flyback converters**
- **Forward converters**

ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
9-Pin Plastic SIP	−25°C to +125°C	TEA1039U

PIN CONFIGURATION

U Package

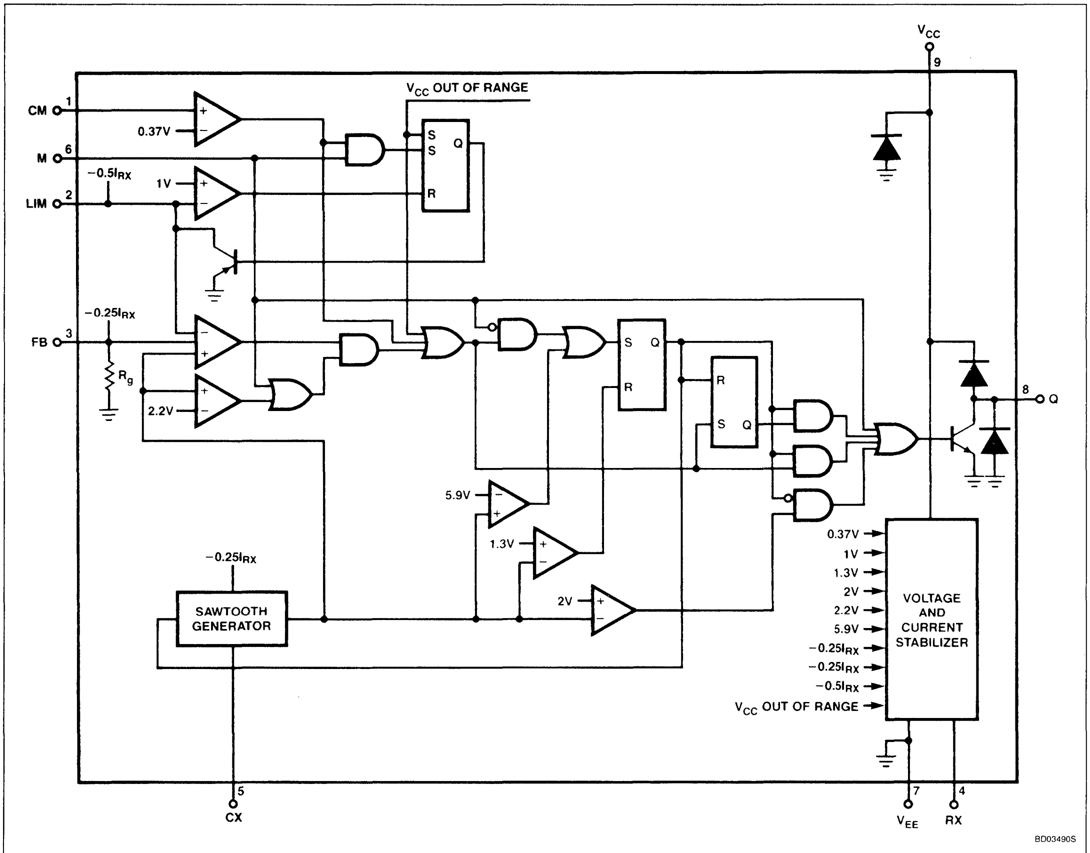


TOP VIEW

CD10370S

PIN NO.	SYMBOL	DESCRIPTION
1	CM	Overcurrent protection input
2	LIM	Limit setting input
3	FB	Feedback input
4	RX	External resistor connection
5	CX	External capacitor connection
6	M	Mode input
7	V_{EE}	Common
8	Q	Output
9	V_{CC}	Positive supply connection

BLOCK DIAGRAM



ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING	UNIT
V_{CC}	Supply voltage range, voltage source	-0.3 to +20	V
I_{CC}	Supply current range, current source	-30 to +30	mA
V_I	Input voltage range, all inputs	-0.3 to +6	V
I_I	Input current range, all inputs	-5 to +5	mA
V_{8-7}	Output voltage range	-0.3 to +20	V
 I_8 I_8	Output current range output transistor ON output transistor OFF	 0 to 1 -100 to +50	 A mA
T_{STG}	Storage temperature range	-65 to +150	°C
T_A	Operating ambient temperature range (see Figure 1)	-25 to +125	°C
F_D	Power dissipation (see Figure 1)	max. 2	W

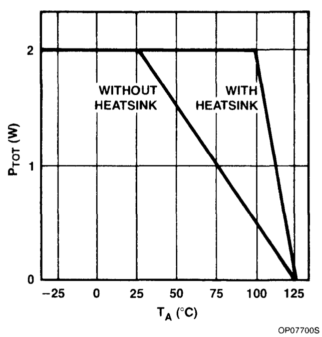


Figure 1. Power Derating Curve

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DC AND AC ELECTRICAL CHARACTERISTICS V_{CC} = 14, T_A = 25°C, unless otherwise specified.

SYMBOL	PARAMETER	MIN	TYP	MAX	UNIT
Supply V_{CC} (Pin 9)					
V_{CC}	Supply voltage, operating	11	14	20	V
	Supply current				
I_{CC}	at V_{CC} = 11V		7.5	11	mA
I_{CC}	at V_{CC} = 20V		9	12	mA
$\frac{\Delta I_{CC}/I_{CC}}{\Delta T}$	variation with temperature		-0.3		%/°C
	Supply voltage, internally limited				
V_{CC}	at I_{CC} = 30mA	23.5		28.5	V
$\Delta V_{CC}/\Delta T$	variation with temperature		18		mV/°C
V_{CCmin}	Low supply threshold voltage	9	10	11	V
$\Delta V_{CC}/\Delta T$	variation with temperature		-5		mV/°C
V_{CCmax}	High supply threshold voltage	21	23	24.6	V
$\Delta V_{CC}/\Delta T$	variation with temperature		10		mV/°C
Feedback input FB (Pin 3)					
$V_{3\ 7}$	Input voltage for duty factor = 0; M input open	0		0.3	V
$-I_{FB}$	Internal reference current		0.5 I_{RX}		mA
R_g	Internal resistor R_g		130		kΩ
Limit setting input LIM (Pin 2)					
$V_{2\ 7}$	Threshold voltage		1		V
$-I_{LIM}$	Internal reference current		0.25 I_{RX}		mA
Overcurrent protection input CM (Pin 1)					
$V_{1\ 7}$	Threshold voltage	300	370	420	mV
$\Delta V_{1\ 7}/\Delta T$	variation with temperature		0.2		mV/°C
t_{PHL}	Propagation delay, CM input to output		500		ns
Oscillator connections RX and CX (Pins 4 and 5)					
	Voltage at RX connection				
$V_{4\ 7}$	at $-I_4$ = 0.15 to 1mA	6.2	7.2	8.1	V
$\Delta V_{4\ 7}/\Delta T$	variation with temperature		2.1		mV/°C
V_{LS}	Lower sawtooth level		1.3		V
V_{FT}	Threshold voltage for output H to L transition in F mode		2		V
V_{FM}	Threshold voltage for maximum frequency in F mode		2.2		V
V_{HS}	Higher sawtooth level		5.9		V
$-I_{CX}$	Internal capacitor charging current, CX connection		0.25 I_{RX}		mA
f_{OSC}	Oscillator frequency (output pulse repetition frequency)	1		10^5	Hz
	Minimum frequency in F mode,				
$\Delta f/f$	initial deviation	-10		10	%
$\frac{\Delta f/f}{\Delta T}$	variation with temperature		0.034		%/°C
	Maximum frequency in F mode,				
$\Delta f/f$	initial deviation	-15		15	%
$\frac{\Delta f/f}{\Delta T}$	variation with temperature		-0.16		%/°C

DC AND AC ELECTRICAL CHARACTERISTICS (Continued) V_{CC} = 14, T_A = 25°C, unless otherwise specified.

SYMBOL	PARAMETER	MIN	TYP	MAX	UNIT
	Output LOW time in F mode,				
$\Delta t/t$	initial deviation	−15		15	%
$\frac{\Delta t/t}{\Delta T}$	variation with temperature		0.2		%/°C
	Pulse repetition frequency in D mode,				
$\Delta f/f$	initial deviation	−10		10	%
$\frac{\Delta f/f}{\Delta T}$	variation with temperature		0.034		%/°C
	Minimum output LOW time in D mode				
t_{OLmin}	at C_5 = 3.6nF		1		μs
$\frac{\Delta t/t}{\Delta T}$	variation with temperature		0.2		%/°C
Output Q (Pin 8)					
$V_{8\ 7}$	Output voltage LOW at I_8 = 100mA		0.8	1.2	V
$\Delta V_{8\ 7}/\Delta T$	variation with temperature		1.5		mV/°C
$V_{8\ 7}$	Output voltage LOW at I_8 = 1A		1.7	2.1	V
$\Delta V_{8\ 7}/\Delta T$	variation with temperature		−1.4		mV/°C

FUNCTIONAL DESCRIPTION

The TEA1039 produces pulses to drive the transistor in a switched-mode power supply. These pulses may be varied either in frequency (frequency regulation mode) or in width (duty factor regulation mode).

The usual arrangement is such that the transistor in the SMPS is ON when the output of the TEA1039 is HIGH, i.e., when the open-collector output transistor is OFF. The duty factor of the SMPS is the time that the output of the TEA1039 is HIGH divided by the pulse repetition time.

Supply V_{CC} (Pin 9)

The circuit is usually supplied from the SMPS that it regulates. It may be supplied either from its primary DC voltage or from its output voltage. In the latter case an auxiliary starting supply is necessary.

The circuit has an internal V_{CC} out-of-range protection. In the frequency regulation mode the oscillator is stopped; in the duty factor regulation mode the duty factor is made zero. When the supply voltage returns within its range, the circuit is started with the slow-start procedure.

When the circuit is supplied from the SMPS itself, the out-of-range protection also provides an effective protection against any interruption in the feedback loop.

Mode Input M (Pin 6)

The circuit works in the frequency regulation mode when the mode input M is connected to ground (V_{EE}, Pin 7). In this mode the circuit produces output pulses of a constant width but with a variable pulse repetition time.

The circuit works in the duty factor regulation mode when the mode input M is left open. In this mode the circuit produces output pulses with a variable width but with a constant pulse repetition time.

Oscillator Resistor and Capacitor Connections RX and CX (Pins 4 and 5)

The output pulse repetition frequency is set by an oscillator whose frequency is determined by an external capacitor C5 connected between the CX connection (Pin 5) and ground (V_{EE}, Pin 7), and an external resistor R4 connected between the RX connection (Pin 4) and ground. The capacitor C5 is charged by an internal current source, whose current level is determined by the resistor R4. In the frequency regulation mode these two external components determine the minimum frequency; in the duty factor regulation mode they determine the working frequency (see Figure 2). The output pulse repetition frequency varies less than 1% with the supply voltage over the supply voltage range.

In the frequency regulation mode the output is LOW from the start of the cycle until the voltage on the capacitor reaches 2V. The capacitor is further charged until its voltage reaches the voltage on either the feedback input FB or the limit setting input LIM, provided it has exceeded 2.2V. As soon as the capacitor voltage reaches 5.9V the capacitor is discharged rapidly to 1.3V and a new cycle is initiated (see Figures 3 and 4).

For voltages on the FB and LIM inputs lower than 2.2V, the capacitor is charged until this voltage is reached; this sets an internal maximum frequency limit.

In the duty factor regulation mode the capacitor is charged from 1.3V to 5.9V and discharged again at a constant rate. The output is HIGH until the voltage on the capacitor exceeds the voltage on the feedback input FB; it becomes HIGH again after discharge of the capacitor (see Figures 5 and 6). An internal maximum limit is set to the duty factor of the SMPS by the discharging time of the capacitor.

Feedback Input FB (Pin 3)

The feedback input compares the input current with an internal current source whose current level is set by the external resistor R4. In the frequency regulation mode, the higher the voltage on the FB input, the longer the external capacitor C5 is charged, and the lower the frequency will be. In the duty factor regulation mode external capacitor C5 is charged and discharged at a constant rate, the voltage on the FB input now determines the moment that the output will become LOW. The higher the voltage on the FB input, the longer the output remains HIGH, and the higher the duty factor of the SMPS.

Limit Setting Input LIM (Pin 2)

In the frequency regulation mode this input sets the minimum frequency, in the duty factor regulation mode it sets the maximum duty factor of the SMPS. The limit is set by an external resistor R2 connected from the LIM input to ground (Pin 7) and by an internal current source, whose current level is determined by external resistor R4.

A slow-start procedure is obtained by connecting a capacitor between the LIM input and ground. In the frequency regulation mode the frequency slowly decreases from f_{MAX} to the working frequency. In the duty factor regulation mode the duty factor slowly increases from zero to the working duty factor.

Overcurrent Protection Input CM (Pin 1)

A voltage on the CM input exceeding 0.37V causes an immediate termination of the output pulse. In the duty factor regulation mode the circuit starts again with the slow-start procedure.

Output Q (Pin 8)

The output is an open-collector NPN transistor, only capable of sinking current. It requires an external resistor to drive an NPN transistor in the SMPS (see Figures 7 and 8).

The output is protected by two diodes, one to ground and one to the supply.

At high output currents the dissipation in the output transistor may necessitate a heatsink. See the power derating curve (Figure 1).

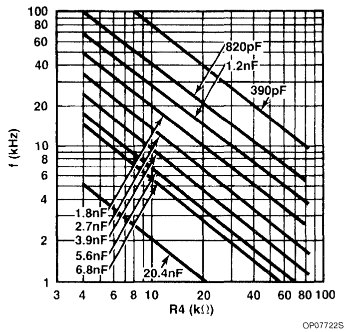
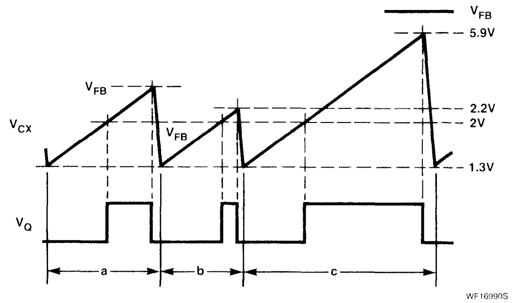


Figure 2. Minimum Pulse Repetition Frequency in the Frequency Regulation Mode, and Working Pulse Repetition Frequency in the Duty Factor Regulation Mode, as a Function of External Resistor R4 Connected Between RX and Ground with External Capacitor C5 Connected Between CX and Ground as a Parameter

Control Circuit for Switched-Mode Power Supply TEA1039



NOTES:
a. The voltages on inputs FB or LIM are between 2.2V and 5.9V. The circuit is in its normal regulation mode.
b. The voltage on input FB or input LIM is lower than 2.2V. The circuit works at its maximum frequency.
c. The voltages on inputs FB and LIM are higher than 5.9V. The circuit works at its minimum frequency.

Figure 3. Timing Diagram for the Frequency Regulation Mode Showing the Voltage on External Capacitor C5 Connected between CX and Ground and the Output Voltage as a Function of Time for Three Combinations of Input Signals

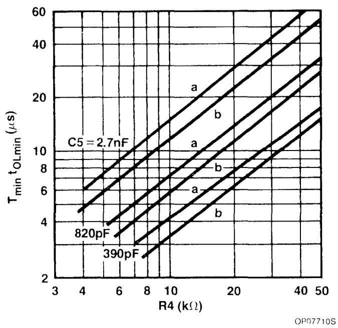
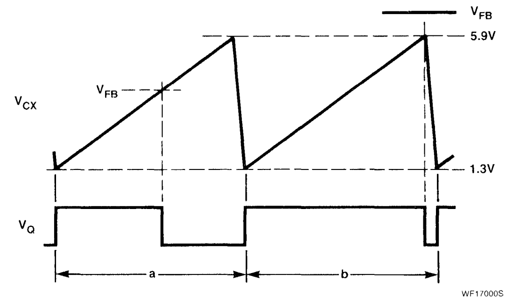


Figure 4. Minimum Output Pulse Repetition Time t_{MIN} (Curves a) and Minimum Output LOW Time t_{OLmin} (Curves b) in the Frequency Regulation Mode as a Function of External Resistor R4 Connected Between RX and Ground with External Capacitor C5 Connected Between CX and Ground as a Parameter



NOTES:
a. The voltages on inputs FB or LIM are below 5.9V. The circuit is in its normal regulation range.
b. The voltages on inputs FB and LIM are higher than 5.9V. The circuit produces its minimum output LOW time, giving the maximum duty factor of the SMPS.

Figure 5. Timing Diagram for the Duty Factor Regulation Mode Showing the Voltage on External Capacitor C5 Connected Between CX and Ground and the Output Voltage as a Function of Time for Two Combinations of Input Signals

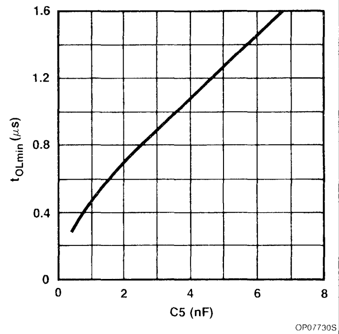
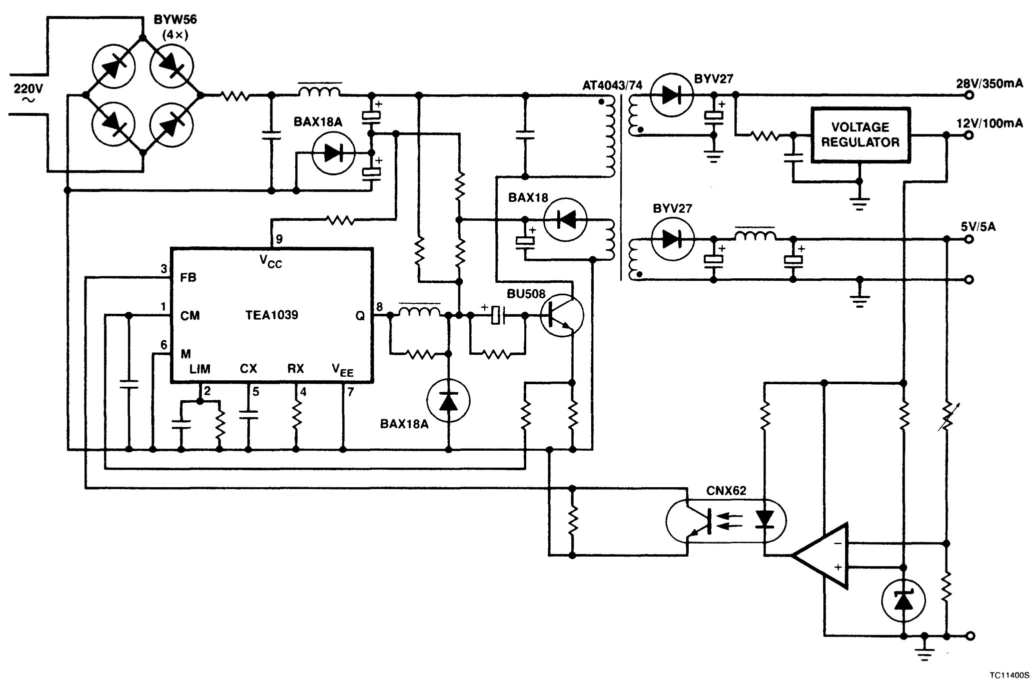
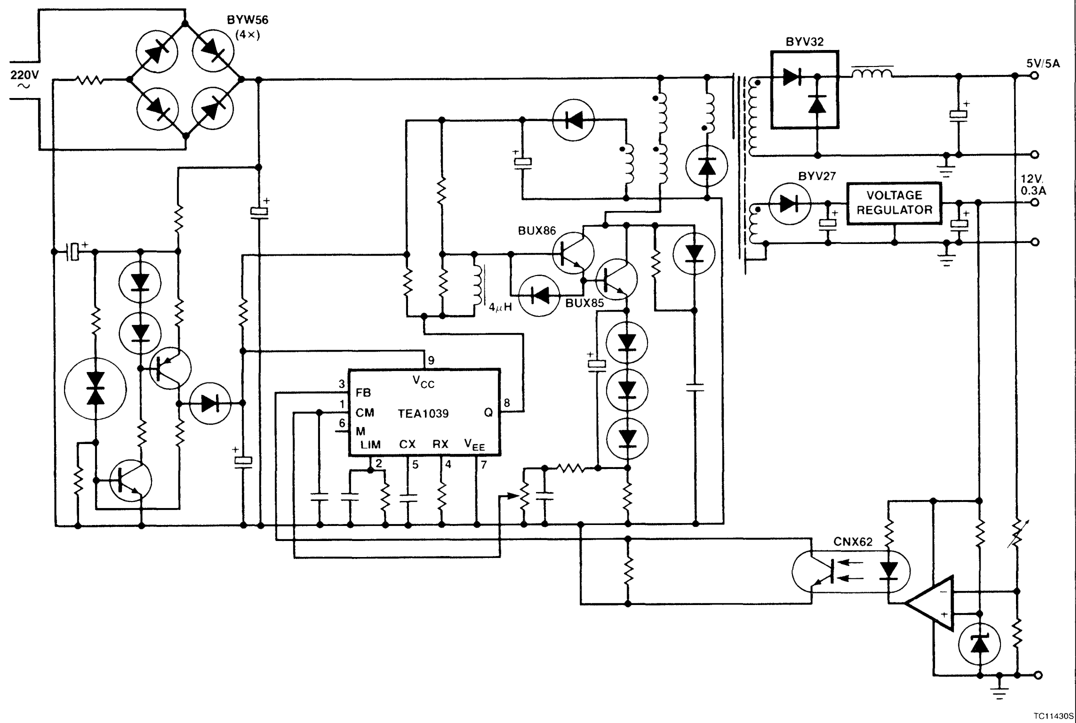


Figure 6. Minimum Output LOW Time t_{OLmin} in the Duty Factor Regulation Mode as a Function of External Capacitor C5 Connected Between CX and Ground. In This Mode the Minimum Output LOW Time is Independent of R4 for Values of R4 Between 4kΩ and 80kΩ



NOTE:
An Optocoupler CNX62 is Used for Voltage Separation.

Figure 7. Typical Application of the TEA1039 in a Variable-Frequency Flyback Converter Switched-Mode Power Supply



NOTE:
An Optocoupler CNX62 is Used for Voltage Separation.

Figure 8. Typical Application of the TEA21039 in a Fixed-Frequency Variable Duty Factor Forward Converter Switched-Mode Power Supply

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Section 14
Packaging Information

Linear Products

INDEX

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Substrate Design Guidelines for Surface-Mounted Devices

Linear Products

INTRODUCTION

SMD technology embodies a totally new automated circuit assembly process using a new generation of electronic components: surface-mounted devices (SMDs). Smaller than conventional components, SMDs are placed onto the surface of the substrate, not through it like leaded components. And from this, the fundamental difference between SMD assembly and conventional through-hole component assembly arises; SMD component positioning is relative, not absolute.

When a through-hole (leaded) component is inserted into a PCB, either the leads go through the holes, or they don't. An SMD, however, is placed onto the substrate surface, its position only relative to the solderlands, and placement accuracy is therefore influenced by variations in the substrate track pattern, component size, and placement machine accuracy.

Other factors influence the layout of SMD substrates. For example, will the board be a mixed-print (a combination of through-hole components and SMDs) or an all-SMD design? Will SMDs be on one side of the substrate or both? And there are process considerations, such as: what type of machine will place the components and how will they be soldered?

Using our expertise in the world of SMD technology, this section draws upon applied research in the area of substrate design and manufacture, and presents the basic guidelines to assist the designer in making the transition from conventional through-hole PCB assembly to SMD substrate manufacture.

Designing With SMD

SMD technology is penetrating rapidly into all areas of modern electronic equipment manufacture — in professional, industrial, and consumer applications. Boards are made with conventional print-and-etch PCBs, multilayer boards with thick film ceramic substrates, and with a host of new materials specially developed for SMD assembly.

However, before substrate layout can be attempted, footprints for all components must be defined. Such a footprint will include the combination of patterns for the copper solderlands, the solder resist, and, possibly, the solder paste. So the design of a substrate breaks down into two distinct areas: the SMD footprint definition, and the layout and track routing for SMDs on the substrate.

Each of these areas is treated individually; first, the general aspects of SMD technology, including substrate configurations, placement machines, and soldering techniques, are discussed.

Substrate Configurations

SMD substrate assembly configurations are classified as:

Type I — Total surface mount (all-SMD); substrates with no through-hole components at all. SMDs of all types (SM integrated circuits, discrete semiconductors, and passive devices) can be mounted either on one side, or both sides, of the substrate. See Figure 1a.

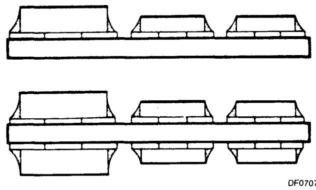
Type IIA — Double-sided mixed-print; substrates with both through-hole components and SMDs of all types on the top, and smaller SMDs (transistors and passives) on the bottom. See Figure 1b.

Type IIB — Underside attachment mixed-print; the top of the substrate is dedicated exclusively to through-hole components, with smaller SMDs (transistor and passives) on the bottom. See Figure 1c.

Although the all-SMD substrate will ultimately be the cheapest and smallest variation as there are no through-hole components, it's the mixed-print substrate that many manufacturers will be looking to in the immediate future, for this technique enjoys most of the advantages of SMD assembly and overcomes the problem of non-availability of some components in surface-mounted form.

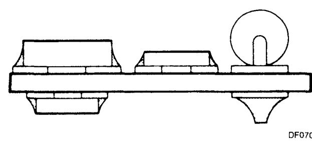
The underside attachment variation of the mixed-print (type IIB — which can be thought of as a conventional through-hole assembly with SMDs on the solder side) has the added advantages of only requiring a single-sided, print-and-etch PCB and of using the established wave soldering technique. The all-SMD and mixed-print assembly with SMDs on both sides require reflow or combination wave/reflow soldering, and, in most cases, a double-sided or multilayer substrate.

The relatively small size of most SMD assemblies compared with equivalent through-hole designs means that circuits can often be repeated several times on a single substrate. This multiple-circuit substrate technique (shown in Figure 2) further increases production efficiency.



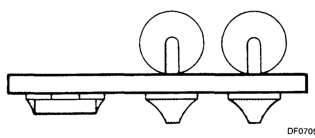
DF07070S

a. Type I — Total Surface-Mount (all-SMD) Substrates



DF07080S

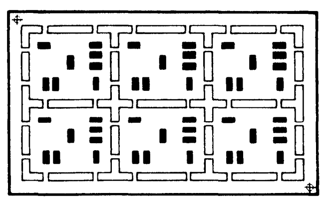
b. Type IIA — Mixed-Print (Double-Sided) Substrate



DF07090S

c. Type IIB — Mixed-Print (Underside Attachment) Substrate

Figure 1



DF07100S

Figure 2. Multiple-Circuit Substrate

Mixed Prints

The possibility of using a partitioned design should be investigated when considering the mixed-print substrate option. For this, part of the circuit would be an all-SMD substrate, and the remainder a conventional through-hole

Substrate Design Guidelines for Surface-Mounted Devices

PCB or mixed-print substrate. This allows the circuit to be broken down into, for example, high and low power sections, or high and low frequency sections.

Automated SMD Placement Machines

The selection of automated SMD placement machines for manufacturing requirements is an issue reaching far beyond the scope of this section. However, as a guide, the four main placement techniques are outlined. They are:

In-Line Placement — a system with a series of dedicated pick-and-place units, each placing a single SMD in a preset position on the substrate. Generally used for small circuits with few components. See Figure 3a.

Sequential Placement — a single pick-and-place unit sequentially places SMDs onto the substrate. The substrate is positioned below the pick-and-place unit using a computer-controlled X-Y moving table (a "software programmable" machine). See Figure 3b.

Simultaneous Placement — places all SMDs in a single operation. A placement module (or station), with a number of pick-and-place units, takes an array of SMDs from the packaging medium and simultaneously places them on the substrate. The pick and place units are guided to their substrate location by a program plate (a "hardware programmable" machine), or by software-controlled X-Y movement of substrate and/or pick-and-place units. See Figure 3c.

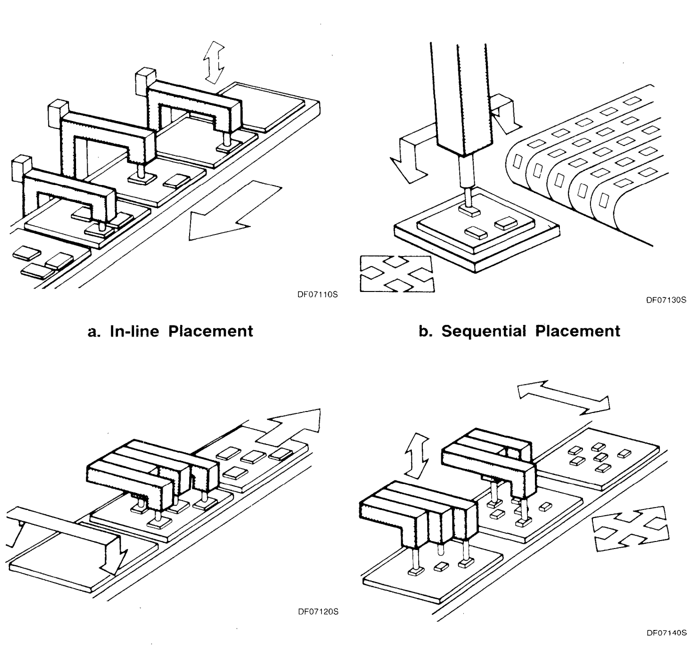
Sequential/Simultaneous Placement — a complete array of SMDs is transferred in a single operation, but the pick-and-place units within each placement module can place all devices simultaneously, or individually (sequentially). Positioning of the SMDs is software-controlled by moving the substrate on an X-Y moving table, by X-Y movement of the pick-and-place units, or by a combination of both. See Figure 3d.

All four techniques, although differing in detail, use the same two basic steps: picking the SMD from the packaging medium (tape, magazine, or hopper) and placing it on the substrate. In all cases, the exact location of each SMD must be programmed into the automated placement machine.

Soldering Techniques

The SMD-populated substrate is soldered by conventional wave soldering, reflow soldering, or a combination of both wave and reflow soldering. These techniques are covered at length in another publication entitled *SMD Soldering Techniques*, but, briefly, they can be described as follows:

Wave Soldering — the conventional method of soldering through-hole component assemblies where the substrate passes over a wave (or more often, two waves) of molten solder. This technique is favored for mixed-print assemblies with through-hole components on the top of the substrate, and SMDs on the bottom.



a. In-line Placement

b. Sequential Placement

c. Simultaneous Placement

d. Sequential/Simultaneous Placement

Figure 3

Reflow Soldering — a technique originally developed for thick-film hybrid circuits using a solder paste or cream (a suspension of fine solder particles in a sticky resin-flux base) applied to the substrate which, after component placement, is heated and causes the solder to melt and coalesce. This method is predominantly used for Type I (all-SMD) assemblies.

Combination Wave/Reflow Soldering — a sequential process using both the foregoing techniques to overcome the problems of soldering a double-sided mixed-print substrate with SMDs and through-hole components on the top, and SMDs only on the bottom. (Type IIB).

Footprint Definition

An SMD footprint, as shown in Figure 4, consists of:

- A pattern for the (copper) solderlands
- A pattern for the solder resist
- If applicable, a pattern for the solder cream.

The design for the footprint can be represented as a set of nominal coordinates and dimensions. In practice, the actual coordinates of each pattern will be distributed around these nominal values due to positioning and processing tolerances. Therefore, the coordinates are stochastic; the actual values form a probability distribution, with a mean value (the nominal value) and a standard deviation.

The coordinates of the SMD are also stochastic. This is due to the tolerances of the actual component dimensions and the positional errors of the automated placement machine.

The relative positions of solderland, solder resist pattern, and SMD, are not arbitrary. A number of requirements may be formulated concerning clearances and overlaps. These include:

- Limiting factors in the production of the patterns (for example, the spacing between solderlands or tracks has a minimum value)

Substrate Design Guidelines for Surface-Mounted Devices

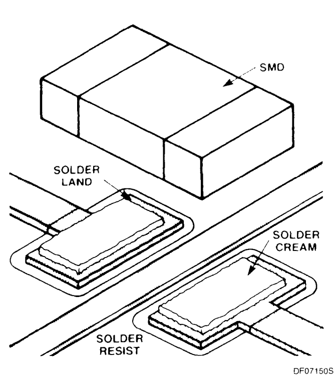


Figure 4. Component Lead, Solder Land, Solder Resist, and Solder Cream "Footprint"

- Requirements concerning the soldering process (for example, the solderlands must be free of solder resist)
- Requirements concerning the quality of the solder joint (for example, the solderland must protrude from the SMD metallization to allow an appropriate solder meniscus)

Mathematical elaboration of these requirements and substitution of values for all tolerances and other parameters lead to a set of inequalities that have to be solved simultaneously. To do this manually using worst-case design is not considered realistic. A better approach is to use a statistical analysis; although this requires a complex computer program, it can be done.

Such an approach may deliver more than one solution, and, if this is so, then the optimal solution must be determined. Optimization is achieved by setting the following objective — find the solution that:

- Minimizes the area occupied by the footprint
- Maximizes the number of tracks between adjacent solderlands.

The final SMD footprint design also depends on the soldering process to be used. The requirements for a wave-soldered substrate differ from those for a reflow-soldered substrate, so each is discussed individually.

Footprints for Wave Soldering

To determine the footprint of an SMD for a wave-soldered substrate, consider four main interactive factors:

- The component dimensions plus tolerances — determined by the component manufacturer
- The substrate metallization — positional tolerance of the solderland with respect to a reference point on the substrate
- The solder resist — positional tolerance of the solder resist pattern with respect to the same reference point
- The placement tolerance — the ability of an automated placement machine to accurately position the SMD on the substrate.

The coordinates of patterns and SMDs have to meet a number of requirements. Some of these have a general validity (the minimum overlap of SMD metallization and solderland) and available space for solder meniscus. Others are specifically required to allow successful wave soldering. One has to take into account factors like the "shadow effect" (missing of joints due to high component bodies), the risk of solder bridging, and the available space for a dot of adhesive.

The "Shadow Effect"

In wave soldering, the way in which the substrate addresses the wave is important. Unlike wave soldering of conventional printed boards where there are no component bodies to restrict the wave's freedom to traverse across the whole surface, wave soldering of SMD substrates is inhibited by the presence of SMDs on the solder-side of the board. The solder is forced around and over the SMDs as shown in Figure 5a, and the surface tension of the molten solder prevents its reaching the far end of the component, resulting in a dry-joint downstream of the solder flow. This is known as the "shadow effect."

The shadow effect becomes critical with high component bodies. However, wetting of the solderlands during wave soldering can be improved by enlarging each land as shown in Figure 5b. The extended substrate metallization makes contact with the solder and allows it to flow back and around the component metallization to form the joint.

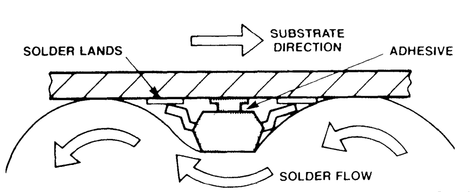
The use of the dual-wave soldering technique also partially alleviates this problem because the first, turbulent wave has sufficient upward pressure to force solder onto the component metallization, and the second, smooth wave "washes" the substrate to form good fillets of solder. Similarly, oil on the surface of the solder wave lowers the surface tension, (which lessens the shadow effect), but this technique introduces problems of contaminants in the solder when the oil decomposes.

Footprint Orientation

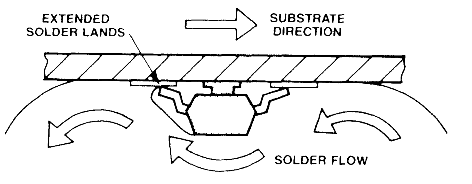
The orientation of SO (small outline) and VSO (very small outline) ICs is critical on wave-soldered substrates for the prevention of solder bridge formation. Optimum solder penetration is achieved when the central axis of the IC is parallel to the flow of solder as shown in Figure 6a. The SO package may also be transversely oriented, as shown in Figure 6b, but this is totally unacceptable for the VSO package.

Solder Thieves

Even with parallel mounted SO and VSO packages, solder bridges have a tendency to form on the leads downstream of the solder flow. The use of solder thieves (small squares of substrate metallization), shown in Figure 7 for a 40-pin VSO, further reduces the likelihood of solder-bridge formation.



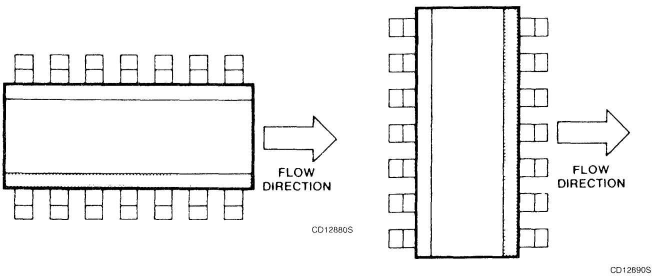
a. Surface Tension Can Prevent the Molten Solder From Reaching the Downstream End of the SMD, Known as the "Shadow Effect"



b. Extending the Solder Lands to Overcome the Shadow Effect

Figure 5

Substrate Design Guidelines for Surface-Mounted Devices



a. Parallel Orientation for SO and VSO Packages

b. Transverse Orientation for SO Packages Only

Figure 6

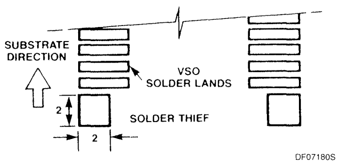


Figure 7. Example of Solder Thieves for VSO-40 Footprints (Dims in mm)

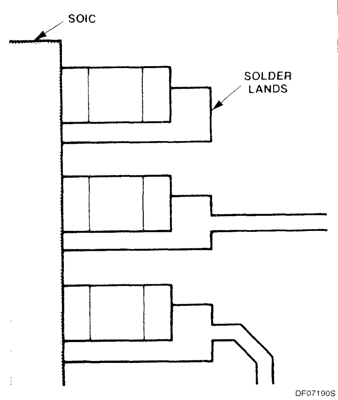


Figure 8. Misaligned Placement of SO Package Increases the Possibility of Solder Bridging

Placement Inaccuracy

Another major cause of solder bridges on SO ICs and plastic leaded chip carriers (PLCCs) is a slight misalignment as shown in Figure 8. The close spacing of the leads on these devices means that any inaccuracy in placement drastically reduces the space between adjacent pins and solderlands, thus increasing the chance of solder bridges forming.

Dummy Tracks for Adhesive Application

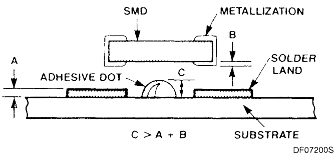
For wave soldering, an adhesive to affix components to the substrate is required. This is necessary to hold the SMDs in place between the placement operation and the soldering process (this technique is covered at length in another publication entitled *Adhesive Application and Curing*).

The amount of adhesive applied is critical for two reasons: first, the adhesive dot must be high enough to reach the SMD, and, second, there mustn't be too much adhesive which could foul the solderland and prevent the formation of a solder joint. The three parameters governing the height of the adhesive dot are shown in Figure 9. Although this diagram illustrates that the minimum requirement is $C > A + B$, in practice, $C > 2(A + B)$ is more realistic for the formation of a good strong bond.

Taking these parameters in turn, the substrate metallization height (A) can range from about 35μm for a normal print-and-etch PCB to 135μm for a plated through-hole board. And the component metallization height (B) (on 1206-size passive devices, for example) may differ by several tens of microns. Therefore, A + B can vary considerably, but it is desirable to keep the dot height (C) constant for any one substrate.

The solution to this apparent problem is to route a track under the device as shown in Figure 10. This will eliminate the substrate metallization height (A) from the adhesive dot-height criteria. Quite often, the high component density of SMD substrates necessitates the routing of tracks between solderlands, and, where it does not, a short dummy track should be introduced.

For bonding small outline (SO) ICs to the substrate, two dots of adhesive are sufficient for SO-8, -14, and -16 packages, but the SOL-20, -24, -28, and VSO-40 packages need three dots. The through-tracks (or dummy tracks) must be positioned beneath the IC accordingly to support the adhesive dots.



NOTES:
A = Substrate metallization height
B = SMD metallization height
C = Height of adhesive dot

Figure 9. Adhesive Dot Height Criteria

Footprints for Reflow Soldering

To determine the footprint of an SMD for a reflow-soldered substrate, there are now five interactive factors to consider: the four that affect the wave solder footprints (although the solder resist may be omitted), plus an additional factor relating to the solder cream application (the positional tolerance of the screen-printed solder cream with respect to the solderlands).

Solder Cream Application

In reflow soldering, the solder cream (or paste) is applied by pressure syringe dispensing or by screen printing. For industrial purposes, screen printing is the favored technique because it is much faster than dispensing.

Screen Printing

A stainless steel mesh coated with emulsion (except for the solderland pattern where cream is required) is placed over the substrate. A squeegee passes across the screen and forces solder cream through the uncoated areas of the mesh and onto the solderland. As a result, dots of solder cream of a given height and density (in mg/mm^2) are produced.

There is an optimum amount of solder cream for each joint. For example, the solder cream requirements for the C1206 SM capacitor are around 1.5mg per end; the SO IC requires between 0.5 and 0.75mg per lead.

The solder cream density, combined with the required amount of solder, makes a demand upon the area of the solderland (in mm^2). The footprint dimensions for the solder cream pattern are typically identical to those for the solderlands.

Substrate Design Guidelines for Surface-Mounted Devices

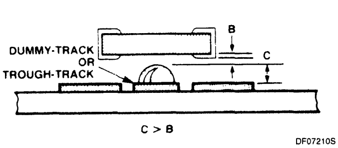


Figure 10. Through-Track or Dummy Track to Modify Dot Height Criteria

Floating

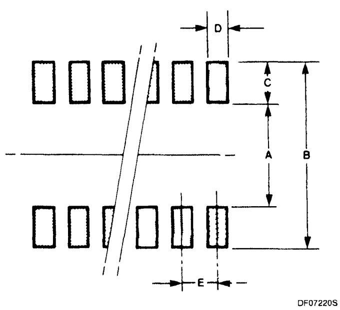
One phenomenon sometimes observed on reflow-soldered substrates is that known as "floating" (or "swimming"). This occurs when the solder paste reflows, and the force exerted by the surface tension of the now molten solder "pulls" the SMD to the center of the solderland.

When the solder reflows at both ends simultaneously, the swimming phenomenon results in the SMD self-centering on the footprint as the forces of surface tension fight for equilibrium. Although this effect can remove minor positional errors, it's not a dependable feature and cannot be relied upon. Components must always be positioned as accurately as possible.

Footprint Dimensions

The following diagrams (Fig. 11 to 19) show footprint dimensions for SO ICs, the VSO-40 package, PLCC packages, and the ra\ige of surface-mounted transistors, diodes, resistors, and capacitors. All dimensions given are based on the criteria discussed in these guidelines.

Please note — these footprints are based on our experience with both experimental and actual production substrates and are reproduced for guidance only. Research is constantly going on to cover all SMDs currently available and those planned for in the future, and data will be published when in it becomes available.



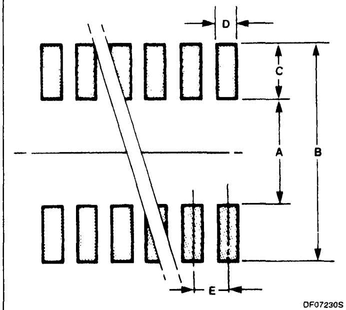
PACKAGE OUTLINE	INCHES A	B	C	D	E
SO-8, 14, 16	.155	.275	.060	.024	.050
SOL-16, 20, 24, 28	.310	.450	.070	.024	.050

PACKAGE OUTLINE	METRIC (mm) A	B	C	D	E
SO SMALL	4.0	7.0	1.5	.6	1.27
SO LARGE	7.8	11.4	1.8	.6	1.27

PACKAGE OUTLINE	METRIC (mm) A	B	C	D	E
SOL-8	9.0	13.2	2.1	.6	1.27

PACKAGE OUTLINE	INCHES A	B	C	D	E
SOL-8	.36	.528	.084	.024	.050

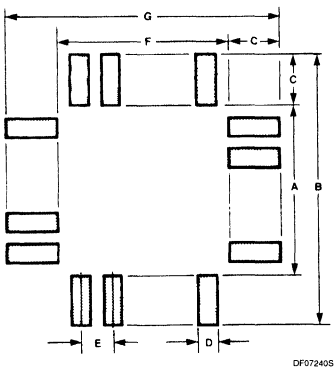
Figure 11. Footprints for SO ICs



PACKAGE OUTLINE	INCHES A	B	C	D	E
VSO-40	.32	.536	.108	.02	.030
VSO-56	.46	.676	.108	.02	.030

PACKAGE OUTLINE	METRIC (mm) A	B	C	D	E
VSO-40	8.0	13.4	2.7	.5	.762
VSO-56	11.5	16.9	2.7	.5	.75

Figure 12. Footprints for VSO ICs



PACKAGE OUTLINE	INCHES A	B	C	D	E	F	G
PLCC-20	.260	.440	.090	.024	.050	.260	.440
PLCC-28	.360	.540	.090	.024	.050	.360	.540
PLCC-44	.560	.740	.090	.024	.050	.560	.740
PLCC-52	.660	.840	.090	.024	.050	.660	.840
PLCC-68	.860	1.040	.090	.024	.050	.860	1.040
PLCC-84	1.060	1.240	.090	.024	.050	1.060	1.240
PLCC-32	.360	.540	.090	.024	.050	.460	.640

Figure 13. Footprints for PLCCs

14

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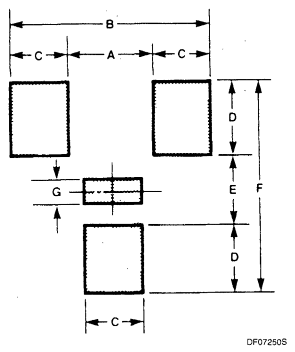
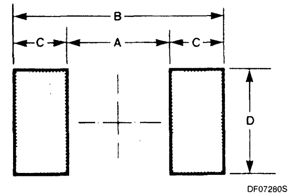


SOT-23	A	B	C	D	E	F
	INCHES					
Reflow	0.048	0.104	0.028	0.044	0.104	—
Wave	0.032	0.136	0.052	0.052	0.048	0.152

SOT-23	A	B	C	D	E	F
	METRIC (mm)					
Reflow	1.2	2.6	0.7	1.1	2.6	—
Wave	0.8	3.4	1.3	1.3	1.2	3.8

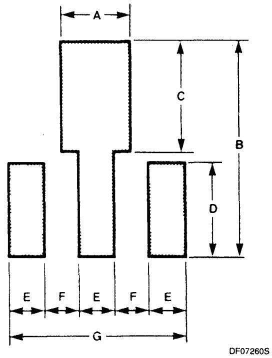
Figure 14. Footprints for SOT-23 Transistors



SOD-80	A	B	C	D
	INCHES			
Reflow	0.096	0.208	0.056	0.056
Wave	0.10	0.2	0.05	0.08

SOD-80	A	B	C	D
	METRIC (mm)			
Reflow	2.4	5.2	1.4	1.4
Wave	2.5	5.0	1.25	2.0

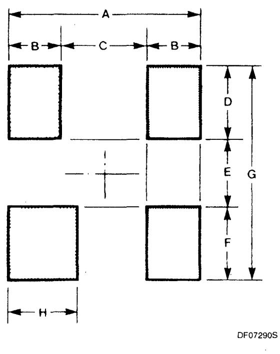
Figure 15. Footprints for SOD-80 Diodes



PACKAGE OUTLINE	A	B	C	D	E	F	G
	INCHES						
SOT-89	0.08	0.184	0.104	0.048	0.032	0.028	0.152

PACKAGE OUTLINE	A	B	C	D	E	F	G
	METRIC (mm)						
SOT-89	2.0	4.6	2.6	1.2	0.8	0.7	3.8

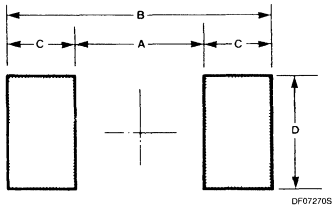
Figure 16. Footprints for Reflow-Soldered SOT-89 Transistors



PACKAGE OUTLINE	A	B	C	D	E	F	G	H
	INCHES							
SOT-143	0.104	0.028	0.048	0.036	0.044	0.036	0.116	0.044

PACKAGE OUTLINE	A	B	C	D	E	F	G	H
	METRIC (mm)							
SOT-143	2.6	0.7	1.2	0.9	1.1	0.9	2.0	1.1

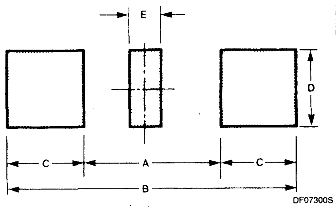
Figure 17. Footprints for Reflow-Soldered SOT-143 Transistors



CODE	SIZE	A	B	C	D
		INCHES			
C0805	0.08 × 0.05	0.032	0.136	0.052	0.056
R/C1206	0.128 × 0.064	0.072	0.184	0.056	0.068
C1210	0.128 × 0.1	0.072	0.184	0.056	0.104
C1808	0.18 × 0.08	0.112	0.248	0.068	0.084
C1812	0.18 × 0.128	0.112	0.248	0.068	0.132
C2220	0.228 × 0.2	0.16	0.296	0.068	0.204

CODE	SIZE	A	B	C	D
		METRIC (mm)			
C0805	2.0 × 1.25	0.8	3.4	1.3	1.4
R/C1206	3.2 × 1.6	1.8	4.6	1.4	1.7
C1210	3.2 × 2.5	1.8	4.6	1.4	2.6
C1808	4.5 × 2.0	2.8	6.2	1.7	2.1
C1812	4.5 × 3.2	2.8	6.2	1.7	3.3
C2220	5.7 × 5.0	4.0	7.4	1.7	5.1

Figure 18. Footprints for Reflow-Soldered Surface-Mounted Resistors and Ceramic Multilayer Capacitors



	SIZE	A	B	C	D	E
		INCHES				
C0805	0.08 × 1.05	0.048	0.144	0.048	0.048	0.016
R/C1206	0.128 × .064	0.08	0.192	0.056	0.056	0.020

CODE	SIZE	A	B	C	D	E
		METRIC (mm)				
C0805	2.0 × 1.25	1.2	3.6	1.2	1.2	0.4
R/C1206	3.2 × .6	2.0	4.8	1.4	1.4	0.5

Figure 19. Footprints for Wave-Soldered Surface-Mounted Resistors and Ceramic Multilayer Capacitors

Substrate Design Guidelines for Surface-Mounted Devices

Layout Considerations

Component orientation plays an important role in obtaining consistent solder-joint quality. The substrate layout shown in Figure 20 will result in significantly better solder joints than a substrate with SMD resistors and capacitors positioned parallel to the solder flow.

Component Pitch

The minimum component pitch is governed by the maximum width of the component and the minimum distance between adjacent components. When defining the maximum component width, the rotational accuracy of the placement machine must also be considered. Figure 21 shows how the effective width of the SMD is increased when the component is rotated with respect to the footprint by angle $\phi°$. (For clarity, the rotation is exaggerated in the illustration.)

The minimum permissible distance between adjacent SMDs is a figure based upon the gap required to avoid solder-bridging during the wave soldering process. Figure 22 shows how this distance and the maximum component width are combined to derive the basic expression for calculating the minimum pitch (F_{MIN}).

As a guide, the recommended minimum pitches for various combinations of two sizes of SMDs, the R/C1206 and C0805 (R or C designating resistor or capacitor respectively; the number referring to the component size), are given in Table 1. These figures are statistically derived under certain assumed boundary conditions as follows:

- Positioning error (Δp)± 0.3mm; (± 0.012")
- Pattern accuracy (Δq)± 0.3mm; (± 0.012")
- Rotational accuracy (ϕ)± 3°
- Component metallization/solderland overlap (M_{MIN}) 0.1mm (0.004") (Note this figure is only valid for wave soldering)
- The figure for the minimum permissible gap between adjacent components (G_{MIN}) is taken to be 0.5mm (0.020").

As these calculations are not based on worst-case conditions, but on a statistical analysis of all boundary conditions, there is a certain flexibility in the given data.

For example, it is possible to position R/C1206 SMDs on a 2.5mm pitch, but the probability of component placements occurring with G_{MIN} smaller than 0.5mm will increase; hence, the likelihood of solder-bridging also increases. Each application must be assessed on individual merit with regard to acceptable levels of rework, and so on.

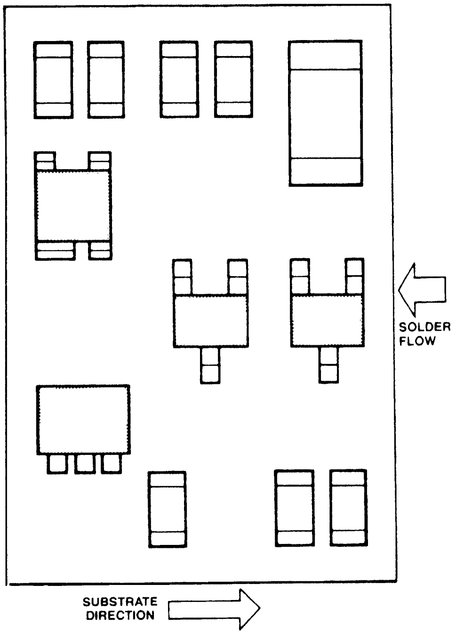
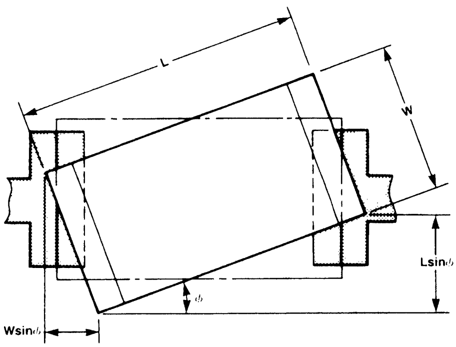


Figure 20. Recommended Component Orientation for Wave-Soldered Substrates



NOTES:
ϕ = Component rotation with respect to footprint
L sin ϕ = Effective increase in width
W sin ϕ = Effective increase in length

Figure 21. The Influence of Rotation of the SMD With Respect to the Footprint

Solderland/Via Hole Relationship

With reflow-soldered multilayer and double-sided, plated through-hole substrates, there must be sufficient separation between the via holes and the solderlands to prevent a solder well from forming. If too close to a solder joint, the via hole may suck the molten solder away from the component by capillary action; this results in insufficient wetting of the joint.

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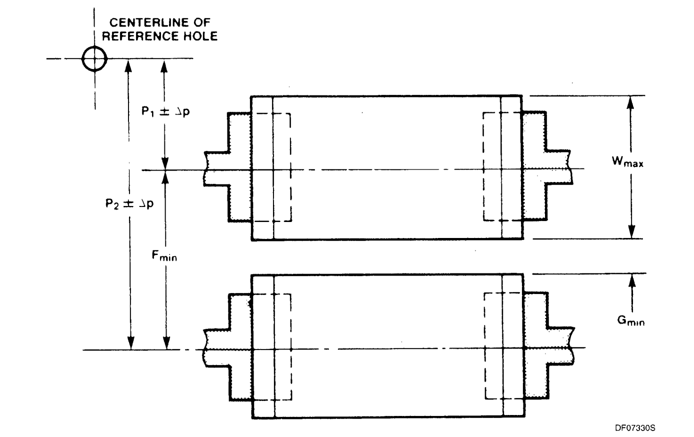
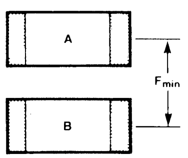
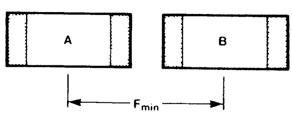
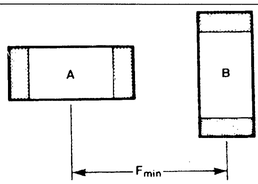
Substrate Design Guidelines for Surface-Mounted Devices



NOTES:
W_{MAX} = Maximum width of component
G_{MIN} = Minimum permissible gap
F_{MIN} = Minimum pitch
P_1 = Nominal position of component 1 (tolerance Δp)
P_2 = Nominal position of component 2 (tolerance Δp)
$F_{MIN} = W_{MAX} + 2\Delta p + G_{MIN}$

Figure 22. Criteria for Determining the Minimum Pitch of SMDs

Table 1. Recommended Pitch For R/C1206 and C0805 SMDs

Combination	Component A	Component B	
		R/C1206	C0805
	R/C1206 C0805	3.0 (0.12") 2.8 (0.112")	2.8 (0.112") 2.6 (0.0104")
	R/C1206 C0805	5.8 (0.232") 5.3 (0.212")	5.3 (0.212") 4.8 (0.192")
	R/C1206 C0805	4.1 (0.164") 3.6 (0.144")	3.7 (0.148") 3.0 (0.12")

Solderland/Component Lead Relationship

Of special consideration for mixed-print substrate layout is the location of leaded components with respect to the SMD footprints and the minimum distance between a protruding clinched lead and a conductor or SMD. Figure 23 shows typical configurations for R/C1206 SMDs mounted on the underside of a substrate with respect to the clinched leads of a leaded component. Minimum distances between the clinched lead ends and the SMDs or substrate conductors are 1mm (0.04") and 0.5 (0.02") respectively.

Placement Machine Restrictions

There are two ways of looking at the distribution of SMDs on the substrate: uniform SMD placement and non-uniform SMD placement. With nonuniform placement, center-to-center dimensions of SMDs are not exact multiples of a predetermined dimension as shown in Figure 24a, so the location of each is difficult to program into the machine.

Uniform placement uses a modular grid system with devices placed on a uniform center-to-center spacing. (For example, 2.5 (0.1") or 5mm (0.2") as shown in Figure 24b.) This placement has the distinct advantage of establishing a standard and enables the use of other automated placement machines for future production requirements without having to redesign boards.

Substrate Population

Population density of SMDs over the total area of the substrate must also be carefully considered, as placement machine limitations can create a "lane" or "zone" that restricts the total number of components which can be placed within that area on the substrate.

For example, on a hardware-programmable simultaneous placement machine (see Figure 3c), each pick-and-place unit within the placement module can only place a component on the substrate in a restricted lane (owing to

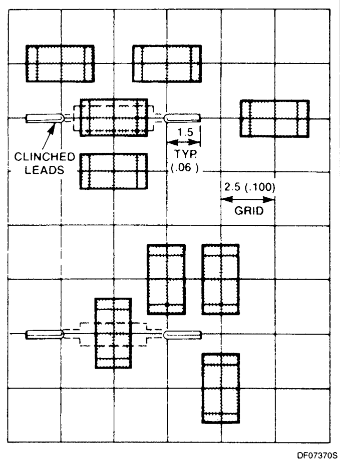
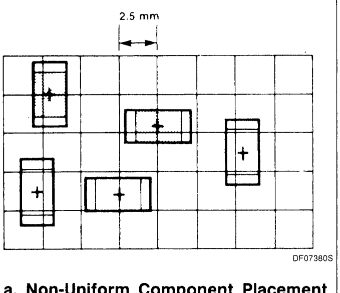
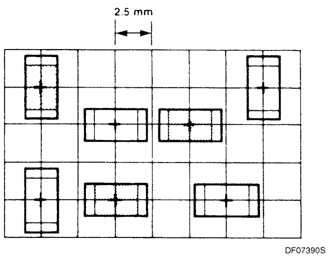


Figure 23. Location of R/C1206 SMDs on the Underside of a Mixed-Print Substrate with Respect to the Clinched Leads of Through-Hole Components (Dimensions in mm)

Substrate Design Guidelines for Surface-Mounted Devices



a. Non-Uniform Component Placement



b. Uniform Component Placement

Figure 24

adjacent pick-and-place units), typically 10 to 12mm (0.4" to 0.48") wide, as shown in Figure 25.

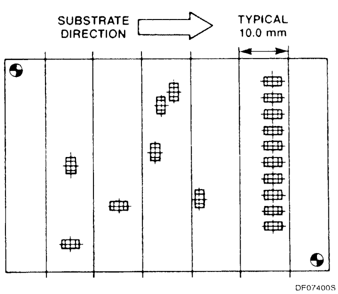


Figure 25. Substrate "Lanes" From Use of a Simultaneous Placement Machine

Placement of the 10 components in the lane on the right of the substrate shown will require a machine with 10 placement modules (or ten passes beneath a single placement module), an inefficient process considering that there are no more than three SMDs in any other lane.

Test Points

Siting of test points for in-circuit testing of SMD substrates presents problems owing to the fewer via holes, higher component densities, and components on both sides of SMD substrates. On conventional double-sided PCBs, the via holes and plated-through component lead-holes mean that most test-points are accessible from one side of the board. However, on SMD substrates, extra provision for test-points may have to be made on both sides of the substrate.

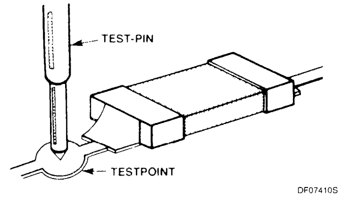
Figure 26a shows the recommended approach for positioning test-points in tracks close to components, and Figure 26b shows an acceptable (though not recommended) alternative where the solderland is extended to accommodate the test pin. This latter method avoids sacrificing too much board space, thus maintaining a high-density layout, but can introduce the problem of components moving ("floating") when reflow-soldered. The approach shown in Figure 26c is totally unacceptable since the pressure applied by the test pin can make an open-circuit soldered joint appear to be good, and, more importantly, the test pin can damage the metallization on the component, particularly with small SMDs.

CAD Systems for SMD Substrate Layout

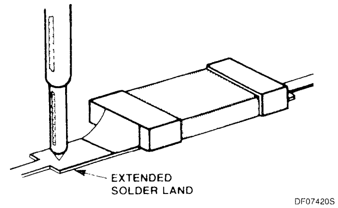
At present, about half of all PCBs are laid out using computer-aided design (CAD) techniques, and this proportion is expected to rise to over 90% by 1988. Of the many current CAD systems available for designing PCB layouts for conventional through-hole components and ICs in DIL packages, few are SMD-compatible, and systems dedicated exclusively to SMD substrate layout are still comparatively rare. There are two main reasons for this: some CAD suppliers are waiting for SMD technology to fully mature before updating their systems to cater to SMD-loaded substrates, and others are holding back until standard package outlines are fully defined.

However, updating CAD systems used for through-hole printed boards is not simply a case of substituting SMD footprints for conventional component footprints, since SMD-populated substrates impose far tougher restraints on PCB layout and require a total rethink of the layout programs. For example, systems must deal with higher component densities, finer track widths, devices on both sides of the substrate (possibly occupying corresponding positions on opposite sides), and even SMDs under conventional DILs on the same side of the substrate.

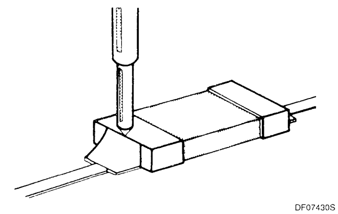
The amount of reworking that a program requires depends on whether it's an interactive (manual) system, or one with fully automatic routing and placement capabilities. For interactive systems, where the user positions the components and routes the tracks manually on-screen, program modifications will be minimal. Automatic systems, however, must contend with the stricter design rules for SMD substrate layout. For example, many auto-routing programs assume that every solderland is a plated through-hole and, therefore, can be used as a via hole. This is not applicable for SMD-populated substrates.



a. RECOMMENDED Test Point Location Close to an SMD



b. Acceptable Test Point Location



c. UNACCEPTABLE Test Point Location

Figure 26

CAD programs base the substrate layout on a regular grid. This method, analogous to drawing the layout on graph paper, must have the grid lines on a pitch that is no larger than the smallest component or feature (track width, pitch, and so on). For conventional DIL boards, this is typically 0.635mm (0.025"), but with the much smaller SMDs, a grid spacing of 0.0254mm (0.001") is required. Consequently, for the same area of substrate, a CAD system based on this finer grid requires

Substrate Design Guidelines for Surface-Mounted Devices

a resolution more than 600 times greater than that required for conventional-layout CAD systems.

To handle this, extra memory capacity can be added, or the allowable substrate area can be limited. In fact, the small size of SMDs, and the high-density layouts possible, generally result in a smaller substrate. However, high-density layout gives rise to additional complications not directly related to the SMD substrate design guidelines. Most CAD systems, for instance, cannot always completely route all interconnects, and some traces have to be routed manually. This can be particularly difficult with the fewer via holes and smaller component spacing of SMD boards.

Ideally, the CAD program should have a "tear-up and start again" algorithm that allows it to restart autorouting if a previous attempt reaches a position where no further traces can be routed before an acceptable percentage of interconnects (and this percentage must first be determined) have been made. This minimizes the manual reworking required.

CAE/CAD/CAM Interaction

Computer-aided production of printed boards has evolved from what was initially only a computer-aided manufacturing process (CAM — digitizing a manually-generated layout and using a photoplotter to produce the artwork) to fully-interactive computer-aided engineering, design, and manufacture using a common database. Figure 27 illustrates how this multi-dimensional interaction is particularly well-suited to SMD-populated substrate manufacture in its highly-automated environment of pick-and-place assembly machines and test equipment.

Using a fully-integrated system, linked by local area network to a central database, will make it possible to use the initial computer-aided engineering (CAE — schematic design, logic verification, and fault simulation) in the generation of the final test patterns at the end of the development process. These test patterns can then be used with the automatic test equipment (ATE) for functional testing of the finished substrates.

Such a system is particularly useful for testing SMD-populated substrates, as their high component density and fewer via-holes make in-circuit testing ("bed of nails" approach) difficult. Consequently, manufacturers are turning to functional testing as an alternative. These aspects are covered in another publication entitled *Functional Testing and Repair*.

Substrate Design Guidelines for Surface-Mounted Devices

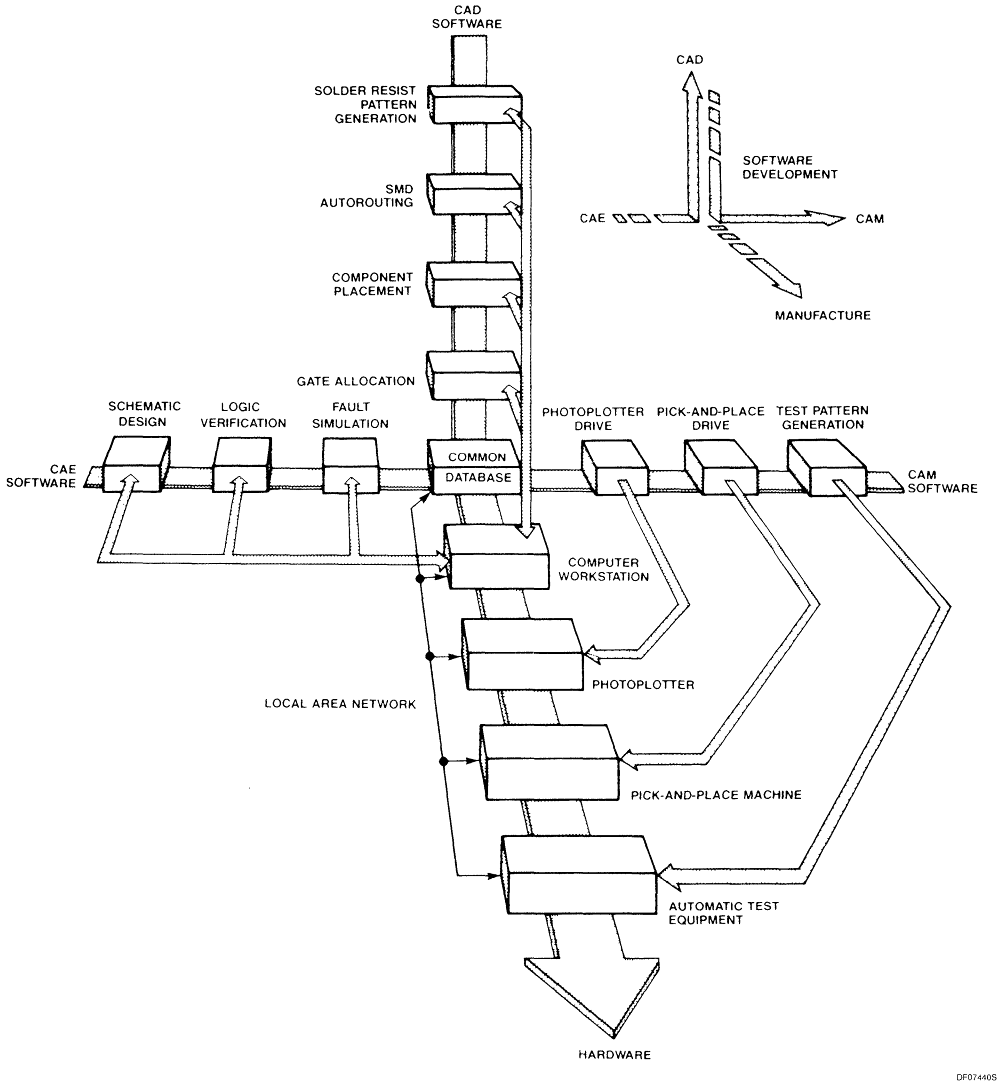


Figure 27. The Software-Hardware Interaction for the Computer-Aided Engineering, Design, and Manufacture of SMD Substrates

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Test and Repair

AN INTRODUCTION

The key questions that must be asked of any electronic circuit are "does it work, and will it continue to do so over a specified period of time?" Until zero-defect soldering is achieved, and all components are guaranteed serviceable by the vendors, manufacturers can only answer these questions by carrying out some form of test on the finished product.

The types of tests, and the depth to which they are carried out, are determined by the complexity of the circuit and the customer's requirements. The amount of rework to be performed on the circuit will depend on the results of these tests and the degree of reliability demanded. The criteria are true of all electronic assemblies, and the test engineer must formulate test schedules accordingly.

Substrates loaded with surface mounted devices (SMDs), however, pose additional problems to the test engineer. The devices are much smaller, and substrate population density is greater, leading to difficulty in accessing all circuit nodes and test points. Also SMD substrate layout designs often have fewer via and component lead holes, so test points may not all be on one side of the substrate and double-sided test fixtures become necessary.

To achieve the high throughput rates made possible by using highly automated SMD placement machines and volume soldering techniques, automatic testing becomes a necessity. Visual inspection of the finished substrate by trained inspectors can normally detect about 90% of defects. With the correct combination of automatic test equipment, the remainder can be eliminated. In this publication, we hope to provide the manufacturer with information to enable him to evaluate and select the best combination of test equipment and the most effective test methods for his product.

BARE-BOARD TESTING

Although SMD substrates will undoubtedly be smaller than conventional through-hole substrates and have less space between conductors, the principles of bare-board testing remain the same. Many of the testers already in use can, with little or no modification, be used for SMD substrates. As this is already a well-established and well-documented practice, it will not be discussed further in this publication, but it is recommended that bare-board testing always be used as the first step in assuring board integrity.

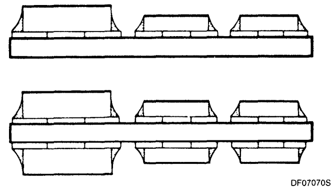
POST-ASSEMBLY TESTING

Testing densely populated substrates is no easy task, as the components may occupy both sides of the board and cover many of the circuit nodes (see Figure 1 for the three main types of SMD-populated substrates). Unlike conventional substrates, on which all test points are usually accessible from the bottom, SMD assemblies must be designed from the start with the siting of test points in mind. Probing SMD substrates is particularly difficult owing to the very close spacing of components and conductors.

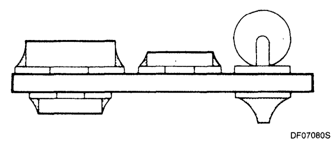
Mixed print or all-SMD assemblies with components on both sides further aggravate the testing problems, as not all test points are present on the same side of the board. Although two-sided test fixtures are feasible, they are expensive and require considerable time to build.

The application of a test probe to the top of an SMD termination could damage it, and probe pressure on a poor or open solder joint can force contact and thus allow a defective joint to be assessed as good. Figure 2a illustrates the recommended siting of test points close to SMD terminations, and Figure 2b shows an alternative, though not recommended, option. Here, problems could arise from reflow soldering (solder migrating from the joint) unless the test point area is separated from the solder land area with a stripe of solder resist. Excessive mechanical pressure caused by too many probes concentrated in a small area may also result in substrate damage.

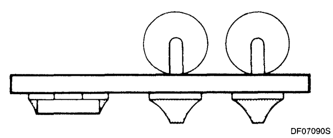
It is good practice for substrates to have test points on a regular grid so that conventional, rather than custom, testers may be used. If the substrate has tall components or heatsinks, the test points must be located far enough away to allow the probes to make good contact. All test points should be solder coated to provide good electrical contact. Via holes may also be used as test points, but the holes must be filled with solder to prevent the probe from sticking.



a. Type I — Total Surface Mount (All-SMD) Substrates



b. Type IIA — Mixed Print (Double-Sided) Substrate



c. Type IIB — Mixed Print (Underside Attachment) Substrate

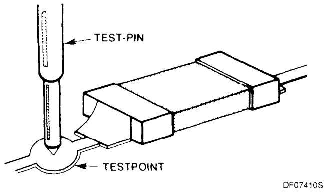
Figure 1

AUTOMATIC TEST EQUIPMENT (ATE)

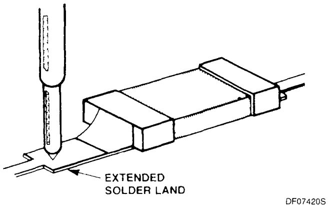
As manufacturers strive to increase production, the question becomes not whether to use automatic test engineering (ATE), but which ATE system to use and how much to spend on it. Because of the rapid fall in price of computers, memories, and peripherals, today's low-cost ATE equals the performance of the high-cost equipment of just two or three years ago. For factory automation, manufacturers must consider many factors, such as production volume, product complexity, and availability of skilled personnel.

One question is whether the ATE system can be used not only for production testing but also for service and repair to reduce the high cost of keeping a substrate inventory in the field. Another is whether assembly and process-induced faults represent a significant percentage of production defects, rather than out-of-tolerance components. These questions need to be answered before deciding on the type of ATE system required.

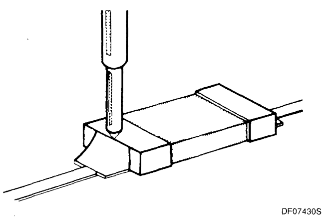
Test and Repair



a. Recommended Location of Test Points Close to SMDs



b. Acceptable, Though Not Recommended, Location of Test Points Close to SMDs



c. Unacceptable Location of Test Points Close to SMDs

Figure 2

Several systems are currently available to the manufacturer, including short-circuit testers, in-circuit testers, in-circuit analyzers, and functional testers. Figure 3 shows a bar-chart giving a comparison of percent fault detection and programming time for various ATE systems.

A loaded-board, short-circuit tester takes from two to six hours to program and its effective fault coverage is between 35% and 65%. It has the advantage of being operationally fast and comparatively inexpensive. On the negative side, however, it is limited to the detection of short-circuits and may require a double-sided, bed-of-nails test fixture (see Figure 4), which for SMD substrates may be expensive and take time to produce. Careful design can, however, often eliminate the need for double-sided test probe fixtures.

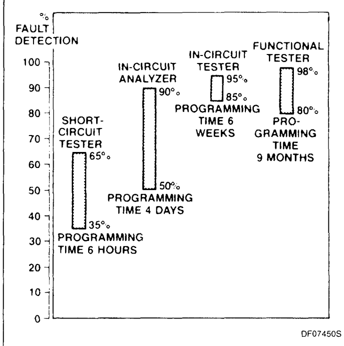


Figure 3. Bar Chart Showing a Comparison of Percent Fault Detection and Programming Time for Various ATE Systems

In-circuit testers power the assembly and check for open or short-circuits, circuit parameters, and can pinpoint defective components. They can provide around 90% fault coverage, but are more expensive than short-circuit testers and programming can take more than six weeks.

In-circuit analyzers are relatively simple to program and can detect manufacturing-induced faults in one third of the time required by an in-circuit tester. Fault coverage is between 50% and 90%. Because they do not power the assembly, they cannot detect digital logic faults, unlike an in-circuit tester or functional tester.

Functional testers, on the other hand, check the assembly's performance and simply make a go or no-go decision. Either the assembly performs its required function or it does not. They are much more expensive, but their fault coverage is between 80% and 98%. Their major disadvantages, apart from cost, are that they cannot locate defective components, and programming for a high-capacity system can take as long as nine months.

ATE Systems

An analysis of defects on a finished substrate will determine which combination of ATE will best meet the test requirements with regard to fault coverage and throughput rate.

If most defects are short-circuits, a loaded-board short-circuit tester, in tandem with an in-circuit tester, will pre-screen the substrate for short-circuits twice as fast as the in-circuit tester. This allows more time for the in-circuit tester to handle the more complex test requirements. This combination of ATE, instead of an in-circuit tester alone, improves the throughput rate.

Combining a short-circuit tester with a functional tester produces even more dramatic results. If most defects are manufacturing-produced shorts, the use of a short-circuit tester to relieve the functional tester of this task can increase throughput five-fold while maintaining a fault coverage of up to 98%.

If manufacturing faults and analog component defects are responsible for the majority of failures, a relatively low-cost, in-circuit analyzer can be used in tandem with an in-circuit tester or functional tester to reduce testing costs and improve throughput. The in-circuit analyzer is three times faster than an in-circuit tester in detecting manufacturing-induced faults, offers test and diagnostics usually within 10 seconds each, and is relatively simple to program. But because it is unpowered, an in-circuit analyzer cannot test digital logic faults; either an in-circuit tester or functional tester following the in-circuit analyzer must be used to locate this type of defect.

POLLUTED POWER SUPPLIES

Today's electronic components and the equipment used to test them are susceptible to electrical noise. Erroneous measurements on pass-or-fail tests could lower test throughput or, even more seriously, allow defective products to pass inspection. Semiconductor chips under test can also be damaged or destroyed as high-energy pulses or line-voltage surges stress the fine-line geometrics separating individual cells.

Noise pulses can be either in the normal (line-to-line) mode or common (line-to-ground) mode. Common-mode electrical noise poses a special threat to modern electronic circuitry since the safety ground line to which common-mode noise is referenced is often used as the system's logic reference point. Since parasitic capacitance exists between safety ground and the reference point, at high frequencies these points are essentially tied together, allowing noise to directly enter the system's logic.

MANUAL REPAIR

The repair of SMD-populated substrates will entail either the resoldering of individual joints and the removal of shorts or the replacement of defective components.

The reworking of defective joints will invariably involve the use of a manual soldering iron. Bits are commercially available in a variety of shapes, including special hollow bits used for desoldering and for the removal of solder bridges. The criteria for the inspec-

Test and Repair

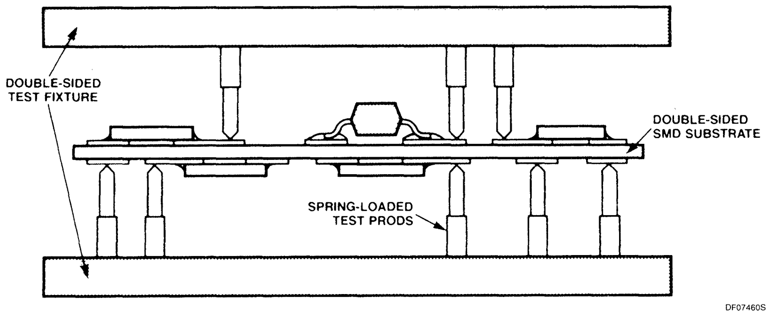


Figure 4. Double-Sided, Bed-of-Nails Test Fixture

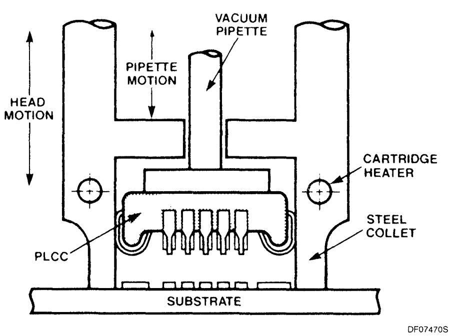


Figure 5. Heated Collet for the Removal and Replacement of Multi-Leaded SMDs (a PLCC is Shown Here)

tion of reworked soldered joints are the same as those for machine soldering.

Special care must be taken when reworking or replacing electrostatic sensitive devices. Soldering irons should be well grounded via a safety resistor of minimum 100kΩ. The ground connection to the soldering iron should be welded rather than clamped. This is because oxidation occurs beneath the clamp, thus isolating the ground connection. Voltage spikes caused by the switching of the iron can be avoided by using either continuously-powered irons, or irons that switch only at zero voltage on the AC sine curve.

To remove defective leadless SMDs, a variety of soldering iron bits are available that will apply the correct amount of heat to both ends of the component simultaneously and allow it to be removed from the substrate. If the substrate has been wave soldered, an adhesive will have been used, and the bond can be broken by twisting the bit. Any adhesive residue must then be removed. The same tool is then used to place and solder the new component, using either solder cream or resin-cored solder.

When a multi-leaded component, such as a plastic leaded chip carrier (PLCC), has to be removed, a heated collet can be used (see Figure 5). The collet is positioned over the PLCC, heat is applied to the leads and solder lands automatically until the solder reflows. The collet, complete with the PLCC, is then raised by vacuum. Solder cream is then re-applied to the solder lands by hand. No adhesive is required in this operation.

The collet is positioned over the replacement PLCC, which is held in place by the slight spring pressure of the PLCC leads against the walls of the collet. The collet, complete with PLCC, is then raised pneumatically and positioned over the solder lands.

Using air pressure, the center pin of the collet then pushes the PLCC into contact with the substrate where it is maintained with the correct amount of force. Heat is then applied through the walls of the collet to reflow the solder paste. The center pin maintains pressure on the PLCC until the solder has solidified, then the center pin is raised and the replacement is complete.

Another method, well-suited to densely populated SMD substrates, uses a stream of heated air, directed onto the SMD terminations. Once the solder has been reflowed, the component can be removed with the aid of tweezers. While the hot air is being directed onto the component, cooler air is played onto the bottom of the substrate to protect it from heat damage. During removal, the component should be twisted sideways slightly in order to break the surface tension of the solder and any adhesive bond between the component and the substrate. This prevents damage to the substrate when the component is lifted.

To fit a new component, the solder lands are first retinned and fluxed, the new component accurately placed, and the solder reflowed with hot air. Substituting superheated argon, nitrogen, or a mixture of nitrogen and hydrogen for the hot air stream removes any risk of contaminating or oxidizing the solder.

Focused infrared light has also been used successfully to reflow the solder on densely populated substrates.

In general, the equipment and procedures used for the replacement of PLCCs can be used for leadless ceramic chip carriers (LCCCs) and small-outline packages (SO ICs). SO ICs are somewhat easier to replace, as the leads are more accessible and only on two sides of the component.

Fluxing and Cleaning

INTRODUCTION

The adoption of mass soldering techniques by the electronics industry was prompted not only by economics, and a requirement for high throughput levels, but also by the need for a consistent standard of quality and reliability in the finished product unattainable by using manual methods. With surface-mounted device (SMD) assembly, this need is even greater.

The quality of the end-product depends on the measures taken during the design and manufacturing stages. The foundations of a high-quality electronic circuit are laid with good design, and with correct choice of components and substrate configuration. It is, however, at the manufacturing stage where the greatest number of variables, both with respect to materials and techniques, have to be optimized to produce high-quality soldering, a prerequisite for reliability.

Of the two most commonly-used soldering techniques, wave and reflow, wave soldering is by far the most widely used and understood. Many factors influence the outcome of the soldering operation, some relating to the soldering process itself, and others to the condition of components and substrate to which they are to be attached. These must be collectively assessed to ensure high-quality soldering.

One of the most important, most neglected, and least understood of these processes is the choice and application of flux. This section outlines the fluxing options available, and discusses the various cleaning techniques that may be required, for SMD substrate assembly.

FLUXES

Populating a substrate involves the soldering of a variety of terminations simultaneously. In one operation, a mixture of tinned copper, tin/lead-or gold-plated nickel-iron, palladium-silver, tin/lead-plated nickel-barrier, and even materials like Kovar, each possessing varying degrees of solderability, must be attached to a common substrate using a single solder alloy.

It is for this reason that the choice of the flux is so important. The correct flux will remove surface oxides, prevent reoxidization, help to transfer heat from source to joint area, and leave non-corrosive, or easily removable corrosive residues on the substrate. It will also improve wettability of the solder joint surfaces.

The wettability of a metal surface is its ability to promote the formation of an alloy at its interface with the solder to ensure a strong, low-resistance joint.

However, the use of flux does not eliminate the need for adequate surface preparation. This is very important in the soldering of SMD substrates, where any temptation to use a highly-active flux in order to promote rapid wetting of ill-prepared surfaces should be avoided because it can cause serious problems later when the corrosive flux residues have to be removed. Consequently, optimum solderability is an essential factor for SMD substrate assembly.

Flux is applied before the wave soldering process, and during the reflow soldering process (where flux and solder are combined in a solder cream). By coating both bare metal and solder, flux retards atmospheric oxidization which would otherwise be intensified at soldering temperature. In the areas where the oxide film has been removed, a direct metal-to-metal contact is established with one low-energy interface. It is from this point of contact that the solder will flow.

Types of Flux

There are two main characteristics of flux. The first is efficacy—its ability to promote wetting of surfaces by solder within a specified time. Closely related to this is the activity of the flux, that is, its ability to chemically clean the surfaces.

The second is the corrosivity of the flux, or rather the corrosivity of its residues remaining on the substrate after soldering. This is again linked to the activity; the more active the flux, the more corrosive are its residues.

Although there are many different fluxes available, and many more being developed, they fall into two basic categories; those with residues soluble in organic liquids, and those with residues soluble in water.

Organic Soluble Fluxes

Most of the fluxes soluble in organic liquids are based on colophony or rosin (a natural product obtained from pine sap that has been distilled to remove the turpentine content). Solid colophony is difficult to apply to a substrate during machine soldering, so it is dissolved in a thinning agent, usually an alcohol. It has a very low efficacy, and hence limited cleaning power, so activators are added in varying quantities to increase it. These take the form of either organic acids, or organic salts that are chemically active at soldering temperatures. It is therefore convenient to classify the colophony-based fluxes by their activator content.

Non-Activated Rosin (R) Flux

These fluxes are formed from pure colophony in a suitable solvent, usually isopropanol or ethyl alcohol. Efficacy is low and cleaning action is weak. Their uses in electronic soldering are limited to easily-wettable materials with a high level of solderability. They are used mainly on circuits where no risk of corrosion can be tolerated, even after prolonged use (implanted cardiac pacemakers, for example). Their flux residues are noncorrosive and can remain on the substrate, where they will provide good insulation.

Rosin, Mildly-Activated (RMA) Flux

These fluxes are also composed of colophony in a solvent, but with the addition of activators, either in the form of di-basic organic acids (such as succinc acid), or organic salts (such as dimethylammonium chloride or diethylammonium chloride). It is customary to express

the amount of added activator as mass percent of the chlorine ion on the colophony content, as the activator-to-colophony ratio determines the activity, and, hence, the corrosivity. In the case of RMA activated with organic salts, this is only some tenths of one percent.

When organic acids are used, a higher percentage of activator must be added to produce the same efficacy as organic salts, so frequently both salts and acids are added. The cleaning action of RMA fluxes is stronger than that of the R type, although the corrosivity of the residues is usually acceptable. These residues may be left on the substrate as they form a useful insulating layer on the metal surfaces. This layer can, however, impede the penetration of test probes at a later stage.

Rosin, Activated (RA) Flux

The RA fluxes are similar to the RMA fluxes, but contain a higher proportion of activators. They are used mainly when component or substrate solderability is poor and corrosion-risk requirements are less stringent. However, as good solderability is considered essential for SMD assembly, highly-activated rosin fluxes should not be necessary. The removal of

Fluxing and Cleaning

flux residues is optional and usually dependent upon the working environment of the finished product and the customer's requirements.

Water-Soluble Fluxes

The water-soluble fluxes are generally used to provide high fluxing activity. Their residues are more corrosive and more conductive than the rosin-based fluxes, and, consequently, must always be removed from the finished substrate. Although termed water soluble, this does not necessarily imply that they contain water; they may also contain alcohols or glycols. It is the flux residues that are water soluble. The usual composition of a water-soluble flux is shown below.

1. A chemically-active component for cleaning the surfaces.
2. A wetting agent to promote the spreading of flux constituents.
3. A solvent to provide even distribution.
4. Substances such as glycols or water-soluble polymers to keep the activator in close contact with the metal surfaces.

Although these substances can be dissolved in water, other solvents are generally used, as water has a tendency to spatter during soldering. Solvents with higher boiling points, such as ethylene glycol or polyethylene glycol are preferred.

Water-Soluble Fluxes With Inorganic Salts

These are based on inorganic salts such as zinc chloride, or ammonium chloride, or inorganic acids such as hydrochloric. Those with zinc or ammonium chloride must be followed by very stringent cleaning procedures as any halide salts remaining on the substrate will cause severe corrosion. These fluxes are generally used for non-electrical soldering. Although the hydrazine halides are among the best active fluxing agents known, they are highly suspect from a health point of view and are therefore no longer used by flux manufacturers.

Water-Soluble Fluxes With Organic Salts

These fluxes are based on organic hydrohalides such as dimethylammonium chloride, cyclo hexalamine hydrochloride, and aniline hydrochloride, and also on the hydrohalides of organic acids. Fluxes with organic halides usually contain vehicles such as glycerol or polyethylene glycol, and non-ionic surface-active agents such as nonylphenol polyoxyethylene. Some of the vehicles, such as the polyethylene glycols, can degrade the insulation resistance of epoxy substrate material and, by rendering the substrate hydrophilic, make it susceptible to electrical leakage in high-humidity environments.

Water-Soluble Fluxes With Organic Acids

Based on acids such as lactic, melonic, or citric, these fluxes are used when the presence of any halide is prohibited. However, their fluxing action is weak, and high acid concentrations have to be used. On the other hand, they have the advantage that the flux residues can be left on the substrate for some time before washing without the risk of severe corrosion.

Solder Creams

For reflow soldering, both the solder and the flux are applied to the substrate before soldering and can be in the form of solder creams (or pastes), preforms, electro-deposit, or a layer of solder applied to the conductors by dipping. For SMD reflow soldering, solder cream is generally used.

Solder cream is a suspension of solder particles in flux to which special compounds have been added to improve the rheological properties. The shape of the particles is important and normally spherical particles are used, although non-spherical particles are now being added, particularly in very fine-line soldering.

In principle, the same fluxes are used in solder creams as for wave soldering. However, due to the relatively large surface area of the solder particles (which can oxidize), more effective fluxing is required and, in general, solder creams contain a higher percentage of activators than the liquid fluxes. The drying of the solder paste during preheating (after component placement) is an important stage as it reduces any tendency for components to become displaced during soldering.

Flux Selection

Choosing an appropriate flux is of prime importance to the soldering system for the production of high-quality, reliable joints. When solderability is good, a mildly-activated flux will be adequate, but when solderability is poorer, a more effective, more active flux will be required. The choice of flux, moreover, will be influenced by the cleaning facilities available, and if, in fact, cleaning is even feasible.

With water-soluble fluxes, aqueous cleaning of the substrate after soldering is mandatory. If thorough cleaning is not carried out, severe problems may arise in the field, due to corrosion or short circuits caused by too low a surface resistance of the conductive residues.

For rosin-based fluxes, the need for cleaning will depend on the activity of the flux. Mildly-activated rosin residues can, in most cases, remain on the substrate where they will afford protection and insulation. In practice, for the great majority of electronic circuits, the choice will be between an RA or an RMA rosin-based flux.

Application of Flux

Three basic factors determine the method of applying flux: the soldering process (wave or reflow), the type of substrate being processed (all-SMD or mixed print), and the type of flux.

For wave soldering, the flux must be applied in liquid form before soldering. While it is possible to apply the flux at a separate fluxing station, with the high throughput rates demanded to maximize the benefits of SMD technology, today's wave-soldering machines incorporate an integral fluxing station prior to the preheat stage. This enables the preheat stage to be used to dry the flux as well as preheat the substrate to minimize thermal shock.

The most commonly-used methods of applying flux for wave soldering are by foam, wave, or spray.

Foam Fluxing

Foam flux is generated by forcing low-pressure clean air through an aerator immersed in liquid flux (see Figure 1). The fine bubbles produced by the aerator are guided to the surface by a chimney-shaped nozzle. The substrates are passed across the top of the nozzle so that the solder side comes in contact with the foam and an even layer of flux is applied. As the bubbles burst, flux penetrates any plated-through holes in the substrate.

Wave Fluxing

A double-sided wave can also be used to apply flux, where the washing action of the wave deposits a layer of flux on the solder side of the substrate (see Figure 2). Wave-height control is essential and a soft, wipe-off brush should be incorporated on the exit side of the fluxing station to remove excess flux from the substrate.

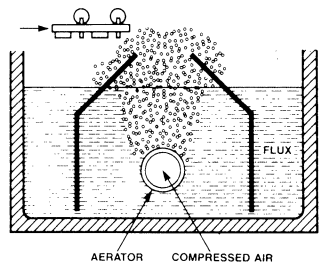


Figure 1. Schematic Diagram of Foam Fluxer

Fluxing and Cleaning

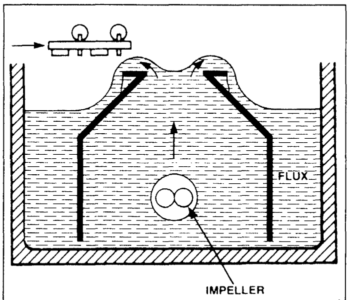


Figure 2. Schematic Diagram of Wave Fluxer

Spray Fluxing

Several methods of spray fluxing exist; the most common involves a mesh drum rotating in liquid flux. Air is blown into the drum which, when passing through the fine mesh, directs a spray of flux onto the underside of the substrate (see Figure 3). Four parameters affect the amount of flux deposited: conveyor speed, drum rotation, air pressure, and flux density. The thickness of the flux layer can be controlled using these parameters, and can vary between 1 and 10μm.

The advantages and disadvantages of these three flux application techniques are outlined in Table 1.

Flux Density

One of the main control factors for fluxes used in machine soldering is the flux density. This provides an indication of the solids content of the flux, and is dependent on the nature of the solvents used. Automatic control systems, which monitor flux density and inject more solvent as required, are commercially available, and it is relatively simple to incorporate them into the fluxing system.

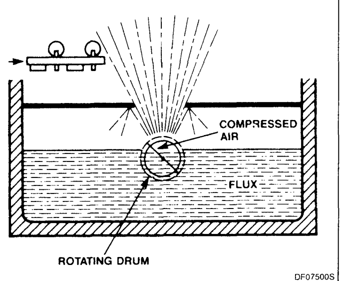


Figure 3. Schematic Diagram of Spray Fluxer

PREHEATING

Preheating the substrate before soldering serves several purposes. It dries the flux to evaporate most of the solvent, thus increasing the viscosity. If the viscosity is too low, the flux may be prematurely expelled from the substrate by the molten solder. This can result in poor wetting of the surfaces, and solder spatter.

Drying the flux also accelerates the chemical action of the flux on the surfaces, and so speeds up the soldering process. During the preheating stage, substrate and components are heated to between 80°C and 90°C (solvent-based fluxes) or to between 100°C and 110°C (water-based systems). This reduces the thermal shock when the substrate makes contact with the molten solder, and minimizes any likelihood of the substrate warping.

The most common methods of preheating are: convection heating with forced air, radiation heating using coils, infrared quartz lamps or heated panels, or a combination of both convection and radiation. The use of forced air has the added advantage of being more effective for the removal of evaporated solvent. Optimum preheat temperature and duration will depend on the nature and design of the substrate and the composition of the flux.

Figure 4 shows a typical method of preheat temperature control. The desired temperature is set on the control panel, and the microprocessor regulates preheater No. 1 to provide approximately 60% of the required heat. The IR detector scans the substrate immediately following No. 1 heater and reads the surface temperature. By taking into account the surface temperature, conveyor speed, and the thermal characteristics of the substrate, the microprocessor then calculates the amount of additional heat required to be provided by heater No. 2 in order to attain the preset temperature. In this way, each substrate will have the same surface temperature on reaching the solder bath.

POSTSOLDERING CLEANING

Now that worldwide efforts in both commercial and industrial electronics are converting old designs from conventional assembly to surface mounting, or a combination of both, it can also be expected that high-volume cleaning systems will convert from in-line aqueous cleaners to in-line solvent cleaners or in-line saponification systems (a technique that uses an alkaline material in water to react with the rosin so that it becomes water soluble). These systems may, however, become subject to environmental objections, and new governmental restrictions on the use of halogenated hydrocarbons.

The major reason for this is that the water-soluble flux residues, containing a higher concentration of activators, or showing hygroscopic behavior, are much more difficult to remove from SMD-populated substrates than rosin-based flux residues. This is primarily because the higher surface tension of water, compared to solvents, makes it difficult for the cleaning agents to penetrate beneath SMDs, especially the larger ones, with their greatly reduced off-contact distance (the distance between component and substrate).

Postsoldering cleaning removes any contamination, such as surface deposits, inclusions, occlusions, or absorbed matter which may degrade to an unacceptable level the chemical, physical, or electrical properties of the assembly. The types of contaminant on substrates that can produce either electrical or mechanical failure over short or prolonged periods are shown in Table 2.

All these contaminants, regardless of their origin, fall into one of two groups: polar and non-polar.

Polar Contaminants

Polar contaminants are compounds that dissociate into free ions which are very good conductors in water, quite capable of causing circuit failures. They are also very reactive with metals and produce corrosive reactions. It is essential that polar contaminants be removed from the substrates.

Non-Polar Contaminants

Non-polar contaminants are compounds that do not dissociate into free ions or carry an electrical current and are generally good insulators. Rosin is a typical example of a non-polar contaminant. In most cases, non-polar contamination does not contribute to corrosion or electrical failure and may be left on the substrate. It may, however, impede functional testing by probes and prevent good conformal coat adhesion.

Solvents

The solvents currently used for the post-soldering cleaning of substrates are normally organic based and are covered by three classifications: hydrophobic, hydrophillic, and azeotropes of hydrophobic/hydrophillic blends.

Azeotropic solvents are mixtures of two or more different solvents which behave like a single liquid insomuch that the vapor produced by evaporation has the same composition as the liquid, which has a constant boiling point between the boiling points of the two solvents that form the azeotrope. The basic ingredients of the azeotropic solvents are combined with alcohols and stabilizers. These stabilizers, such as nitromethane, are included to prevent corrosive reaction be-

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Fluxing and Cleaning

Table 1. Advantages and Disadvantages of Flux Application Methods

Method	Advantages	Disadvantages
Foam Fluxing	• Compatible with continuous soldering process • Foam crest height not critical • Suitable for mixed-print substrates	• Not all fluxes have good foaming capabilities • Losses throught evaporation may be appreciable • Prolonged preheating because of high boiling point of solvents
Wave Fluxing	• Can be used with any liquid flux • Compatible with continuous soldering process • Suitable for densely-populated mixed print	• Wave crest height is critical to ensure good contact with bottom of substrate without contaminating the top
Spray fluxing	• Can be used with most liquid fluxes • Short preheat time if appropriate alcohol solvents are used • Layer thickness is controllable	• High flux losses due to non-recoverable spray • System requires frequent cleaning

tween the metallization of the substrate and the basic solvents.

Hydrophobic solvents do not mix with water at concentrations exceeding 0.2%, and consequently have little effect on ionic contamination. They can be used to remove non-polar contaminants such as rosin, oils, and greases.

Hydrophillic solvents do mix with water and can dissolve both polar and non-polar contamination, but at different rates. To overcome these differences, azeotropes of the various solvents are formulated to maximize the dissolving action for all types of contamination.

Solvent Cleaning

Two types of solvent cleaning systems are in use today: batch and conveyorized systems, either of which can be used for high-volume production. In both systems, the contaminated substrates are immersed in the boiling solvents, and ultrasonic baths or brushes may also be used to further improve the cleaning capabilities.

The washing of rosin-based fluxes offers advantages and disadvantages. Washed substrates can usually be inserted into racks easier, as there will be no residues on their edges; test probes can make better contact without a rosin layer on the test points, and the removal of the residues makes it easier to visually examine the soldered joints. On the other hand, washing equipment is expensive, and so are the solvents, and some solvents present a health or environmental hazard if not correctly dealt with.

Aqueous Cleaning

For high-volume production, special machines have been developed in which the substrates are conveyor-fed through the various stages of spraying, washing, rinsing, and drying. The final rinse water is blown from the substrates to prevent any deposits from the water being left on the substrate.

Where water-soluble fluxes have been used in the soldering process, substrate cleaning is mandatory. For the rosin-based fluxes, it is optional, and is often at the discretion of the customer.

Conformal Coatings

A conformal, or protective coating on the substrate, applied at the end of processing, prevents or minimizes the effects of humidity and protects the substrate from contamination by airborne dust particles. Substrates that are to be provided with a conformal coating (dependent on the environmental conditions to which the substrate will be subjected) must first be washed.

Environmental and Ecological Aspects of Fluxes and Solvents

Fumes and vapors produced during soldering processes, or during cleaning, will not, under normal circumstances, present a health hazard, if relevant health and safety regulations are observed.

Fumes originating from colophony can cause respiratory problems, so an efficient fume-extraction system is essential. The extraction system must cover the fluxing, preheating, and soldering stations, remain operational for at least one hour after machine shutdown, and conform to local regulations. Today, the problem of noxious fumes is unlikely to concern the cleaning station, as all commercial systems are equipped to condense the vapors back into the system. In the future, however, it can be expected that a much lower degree of escape of noxious fumes from any system will be allowed, and all systems may have to be reviewed.

Certain fluxes, particularly some water-soluble ones, contain highly aggressive substances, and must not be allowed to come into contact with the skin or eyes. Any contamination should immediately be removed with plenty of clean, fresh water. Deionized water should also be readily available as an eye-wash. Should contamination occur, a qualified medical practitioner should be consulted. Protective clothing should be worn during cleaning or maintenance of the fluxing station.

Conclusion

SMD technology imposes tougher restraints on fluxing and cleaning of substrate assemblies. Traditionally, rosin-based fluxes have been used in electronic soldering where residues were considered "safe" and could be left on the board. However, increased SMD packing density, fine-line tracks, and more rigid specifications have resulted in changes to this basic philosophy.

There is now a demand for surfaces free from residues; test probes are more efficient when they do not have to penetrate rosin flux residues, and conformal coating and board inspection benefit from the absence of such residues.

Cleaning also poses problems for SMD substrates. The close proximity of component and substrate means that solvents cannot effectively clean beneath devices. Components must also be compatible with the cleaning process. They must, for example, be resistant to the solvents used and to the temperatures of the cleaning process. They must also be sealed to prevent cleaning fluids from entering the devices and degrading performance.

So, eliminating the need for cleaning is better than poor or incomplete cleaning. And in a well-balanced system, mildly-activated rosin-based fluxes, leaving only non-corrosive residues, can be successfully used for SMD substrate soldering without subsequent cleaning.

Much research into fluxes and solder creams is presently being done — for example, the production of synthetic resin, with qualities superior to colophony at a lower cost. Another area of research is that of solder creams with non-melting additives, such as lead or ceramic spheres, that increase the distance

Fluxing and Cleaning

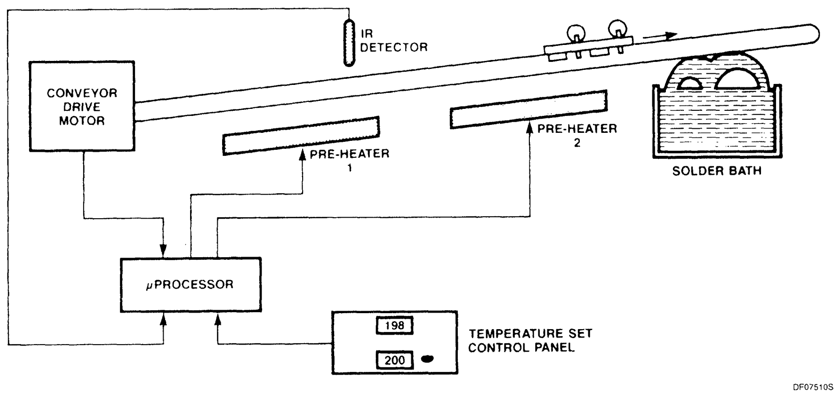

Figure 4. Schematic Diagram of a Typical Controlled Preheat System

Table 2. Substrate Contaminants

Contaminant	Origin
Organic compounds	Fluxes, solder mask
Inorganic insoluble compounds	Photo-resists, substrate processing
Organo-metallic compounds	Fluxes, substrate processing
Inorganic soluble compounds	Fluxes
Particle matter	Dust, fingerprints

between component and substrate, thus making it easier for cleaning fluids to penetrate beneath the component. It also increases the joint's ability to withstand thermal cycling.

Rosin-free and halide-free fluxes are also being developed with similar activities to conventional rosin-based fluxes. These new types will combine the "safety" of rosin fluxes with easier removal in conventional solvents. Using non-polar materials, ionizable or corrosive residues are eliminated, and the need for cleaning immediately after soldering is avoided.

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Thermal Considerations for Surface-Mounted Devices

Linear Products

INTRODUCTION

Thermal characteristics of integrated circuit (IC) packages have always been a major consideration to both producers and users of electronics products. This is because an increase in junction temperature (T_J) can have an adverse effect on the long-term operating life of an IC. As will be shown in this section, the advantages realized by miniaturization can often have trade-offs in terms of increased junction temperatures. **Some of the VARIABLES affecting T_J are controlled by the PRODUCER of the IC, while others are controlled by the USER and the ENVIRONMENT in which the device is used.**

With the increased use of Surface-Mount Device (SMD) technology, management of thermal characteristics remains a valid concern, not only because the SMD packages are much smaller, but also because the thermal energy is concentrated more densely on the printed wiring board (PWB). For these reasons, the designer and manufacturer of surface-mount assemblies (SMAs) must be more aware of all the variables affecting T_J.

POWER DISSIPATION

Power dissipation (P_D), varies from one device to another and can be obtained by multiplying V_{CC} Max by typical I_{CC}. Since I_{CC} decreases with an increase in temperature, maximum I_{CC} values are not used.

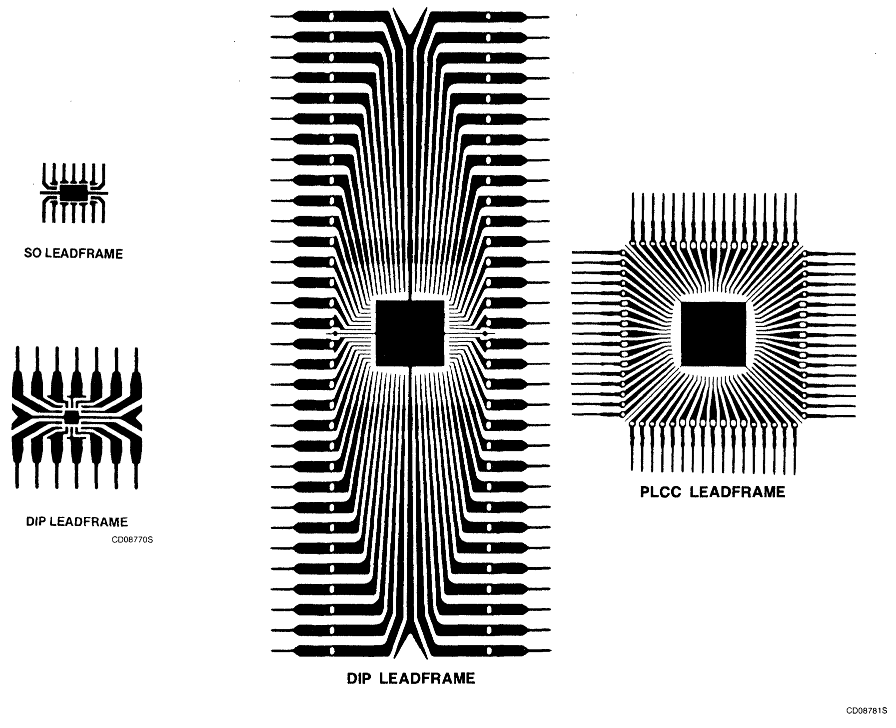
THERMAL RESISTANCE

The ability of the package to conduct this heat from the chip to the environment is expressed in terms of thermal resistance. The term normally used is Theta JA (θ_{JA}). θ_{JA} is often separated into two components: thermal resistance from the junction to case, and the thermal resistance from the case to ambient. θ_{JA} represents the total resistance to heat flow from the chip to ambient and is expressed as follows:

$$\theta_{JC} + \theta_{CA} = \theta_{JA}$$

JUNCTION TEMPERATURE (T_J)

Junction temperature (T_J) is the temperature of a powered IC measured by Signetics at the



a. SO-14 Leadframe Compared to a 14-Pin DIP Leadframe

b. PLCC-68 Leadframe Compared to a 64-Pin DIP Leadframe

Figure 1

Thermal Considerations for Surface-Mounted Devices

substrate diode. When the chip is powered, the heat generated causes the T_J to rise above the ambient temperature (T_A). T_J is calculated by multiplying the power dissipation of the device by the thermal resistance of the package and adding the ambient temperature to the result.

$$T_J = (P_D \times \theta_{JA}) + T_A$$

FACTORS AFFECTING θ_{JA}

There are several factors which affect the thermal resistance of any IC package. Effective thermal management demands a sound understanding of all these variables. Package variables include the leadframe design and materials, the plastic used to encapsulate the device, and, to a lesser extent, other variables such as the die size and die attach methods. Other factors that have a significant impact on the θ_{JA} include the substrate upon which the IC is mounted, the density of the layout, the air-gap between the package and the substrate, the number and length of traces on the board, the use of thermally-conductive epoxies, and external cooling methods.

PACKAGE CONSIDERATIONS

Studies with dual in-line plastic (DIP) packages over the years have shown the value of proper leadframe design in achieving minimum thermal resistance. SMD leadframes are smaller than their DIP counterparts (see Figures 1a and 1b). Because the same die is used in each of the packages, the die-pad, or flag, must be at least as large in the SO as in the DIP.

While the size and shape of the leads have a measurable effect on θ_{JA}, the design factors that have the most significant effect are the die-pad size and the tie-bar size. With design constraints caused by both miniaturization and the need to assemble packages in an automated environment, the internal design of an SMD is much different than in a DIP. However, the design is one that strikes a balance between the need to miniaturize, the need to automate the assembly of the package, and the need to obtain optimum thermal characteristics.

LEAD FRAME MATERIAL is one of the more important factors in thermal management. For years, the DIP leadframes were constructed out of Alloy-42. These leadframes met the producers' and users' specifications in quality and reliability. However, three to five years ago the leadframe material of DIPs was changed from Alloy-42 to Copper (CLF) in order to provide reduced θ_{JA} and extend the reliable temperature-operating range. While this change has already taken place for the DIP, it is still taking place for the SO package. Signetics began making 14-pin SO packages with CLF in April 1984 and completed conversion to CLF for all SO packages by 1985. As is shown in Figures 10 through 14, the change to CLF is producing dramatic results in the θ_{JA} of SO packages. All PLCCs are assembled with copper leadframes.

The MOLDING COMPOUND is another factor in thermal management. The compound used by Signetics and Philips is the same high purity epoxy used in DIP packages (at present, HC-10, Type II). This reduces corrosion caused by impurities and moisture.

OTHER FACTORS often considered are the die-size, die-attach methods, and wire bonding. Tests have shown that die size has a minor effect on θ_{JA} (see Figures 10 through 14).

While there is a difference between the thermal resistance of the silver-filled adhesive used for die attach and a gold silicon eutectic die attach, the thickness of this layer (1 – 2 mils) is so small it makes the difference insignificant.

Gold-wire bonding in the range of 1.0 to 1.3 mils does not provide a significant thermal path in any package.

In summary, the SMD leadframe is much smaller than in a DIP and, out of necessity, is designed differently; however, the SMD package offers an adequate θ_{JA} for all moderate power devices. Further, the change to CLF will reduce the θ_{JA} even more, lowering the T_J and providing an even greater margin of reliability.

SIGNETICS' THERMAL RESISTANCE MEASUREMENTS — SMD PACKAGES

The graphs illustrated in this application note show the thermal resistance of Signetics' SMD devices. These graphs give the relationship between θ_{JA} (junction-to-ambient) or θ_{JC} (junction-to-case) and the device die size. Data is also provided showing the difference between still air (natural convection cooling) and air flow (forced cooling) ambients. All θ_{JA} tests were run with the SMD device soldered to test boards. It is important to recognize that the test board is an essential part of the test environment and that boards of different sizes, trace layouts, or compositions may give different results from this data. Each SMD user should compare his system to the Signetics test system and determine if the data is appropriate or needs adjustment for his application.

Test Method

Signetics uses what is commonly called the TSP (temperature-sensitive parameter) method. This method meets MIL-STD 883C, Method 1012.1. The basic idea of this method is to use the forward voltage drop of a calibrated diode to measure the change in junction temperature due to a known power dissipation. The thermal resistance can be calculated using the following equation:

$$\theta_{JA} = \frac{\Delta T_J}{P_D} = \frac{T_J - T_A}{P_D}$$

Test Procedure

TSP Calibration

The TSP diode is calibrated using a constant-temperature oil bath and constant-current power supply. The calibration temperatures used are typically 25°C and 75°C and are measured to an accuracy of ±0.1°C. The calibration current must be kept low to avoid significant junction heating; data given here used constant currents of either 1.0mA or 3.0mA. The temperature coefficient (K-Factor) is calculated using the following equation:

$$K = \left.\frac{T_2 - T_1}{V_{F2} - V_{F1}}\right| I_F = \text{Constant}$$

Where: K = Temperature Coefficient (°C/mV)
T_2 = Higher Test Temperature (°C)
T_1 = Lower Test Temperature (°C)
V_{F2} = Forward Voltage at I_F and T_2
V_{F1} = Forward Voltage at I_F and T_1
I_F = Constant Forward Measurement Current
(See Figure 2)

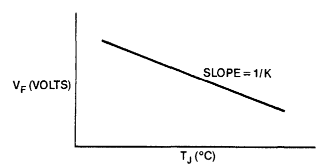


Figure 2. Forward Voltage — Junction Temperature Characteristics of a Semiconductor Junction Operating at a Constant Current. The K Factor is the Reciprocal of the Slope

Thermal Resistance Measurement

The thermal resistance is measured by applying a sequence of constant current and constant voltage pulses to the device under test. The constant current pulse (same current at which the TSP was calibrated) is used to measure the forward voltage of the TSP. The constant voltage pulse is used to heat the part. The measurement pulse is very short

Thermal Considerations for Surface-Mounted Devices

(less than 1% of cycle) compared to the heating pulse (greater than 99% of cycle) to minimize junction cooling during measurement. This cycle starts at ambient temperature and continues until steady-state conditions are reached. The thermal resistance can then be calculated using the following equation:

$$\theta_{JA} = \frac{\Delta T_J}{P_D} = \frac{K(V_{FA} - V_{FS})}{V_H \times I_H}$$

Where: V_{FA} = Forward Voltage of TSP at Ambient Temperature (mV)

V_{FS} = Forward Voltage of TSP at Steady-State Temperature (mV)

V_H = Heating Voltage (V)

I_H = Heating Current (A)

Test Ambient

θ_{JA} Tests

All θ_{JA} test data collected in this application note was obtained with the SMD devices soldered to either Philips SO Thermal Resistance Test Boards or Signetics PLCC Thermal Resistance Test Boards with the following parameters:

Board size — SO Small
1.12″ × 0.75″ × 0.059″
— SO Large:
1.58″ × 0.75″ × 0.059″
— PLCC:
2.24″ × 2.24″ × 0.062″

Board Material — Glass epoxy, FR-4 type with 1oz. sq.ft. copper solder coated

Board Trace Configuration — See Figure 3.

SO devices are set at 8 – 9mil stand-off and SO boards use one connection pin per device lead. PLCC boards generally use 2 – 4 connection pins regardless of device lead count. Figure 5 shows a cross-section of an SO part soldered to test board, and Figure 4 shows typical board/device assemblies ready for θ_{JA} Test.

The still-air tests were run in a box having a volume of 1 cubic foot of air at room temperature. The air-flow tests were run in a 4″ × 4″ cross-section by 26″ long wind tunnel with air at room temperature. All devices were soldered on test boards and held in a horizontal test position. The test boards were held in a Textool ZIF socket with 0.16″ stand-off. Figure 6 shows the air-flow test setup.

θ_{JC} Tests

The θ_{JC} test is run by holding the test device against an "infinite" heat sink (water-cooled block approximately 4″ × 7″ × 0.75″) to give a θ_{CA} (case-to-ambient) approaching zero. The copper heat sink is held at a constant temperature (≈20°C) and monitored with a thermocouple (0.040″ diameter sheath, grounded junction type K) mounted flush with heat-sink surface and centered below die in the test device. Figure 7 shows the θ_{JC} test mounting for a PLCC device.

SO devices are mounted with the bottom of the package held against the heat sink. This is achieved by bending the device leads straight out from the package body. Two small wires are soldered to the appropriate leads for tester connection. Thermal grease is used between the test device and heat sink to assure good thermal coupling.

PLCC devices are mounted with the top of the package held against the heat sink. A small spacer is used between the hold-down mechanism and PLCC bottom pedestal. Small hook-up wires and thermal grease are used as with the SO setup. Figure 7 shows the PLCC mounting.

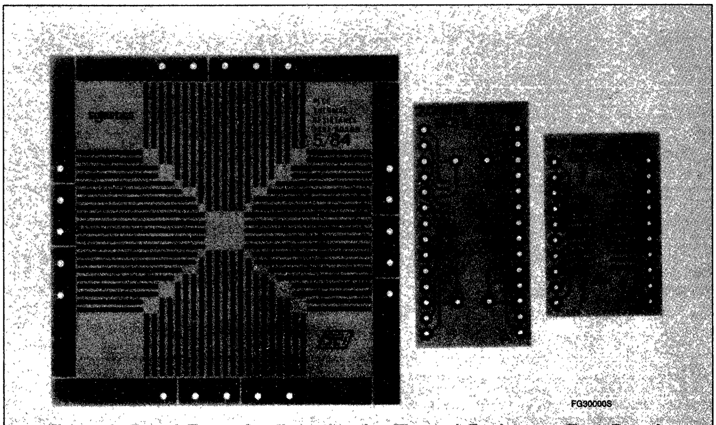


Figure 3. Board Trace Configuration for Thermal Resistance Test Boards

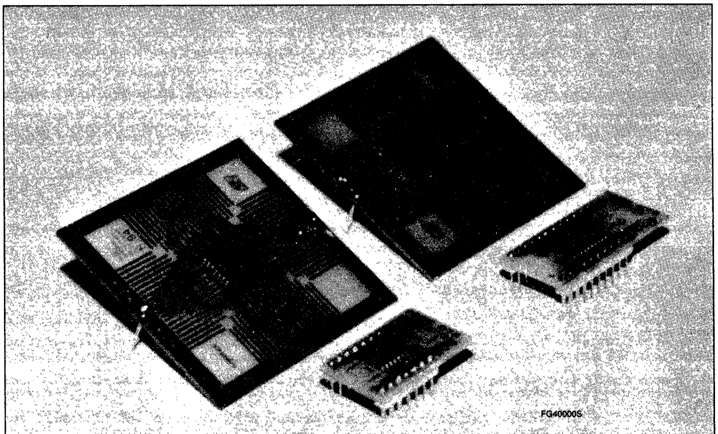


Figure 4. Device/Board Assemblies

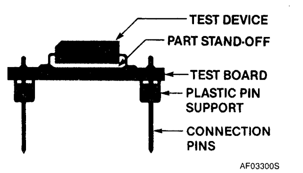
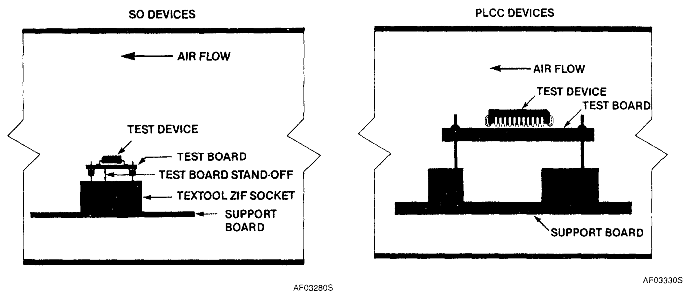


Figure 5. Cross-Section of Test Device Soldered to Test Board

Thermal Considerations for Surface-Mounted Devices



SO Devices **PLCC Devices**
Figure 6. Air-Flow Test Setup

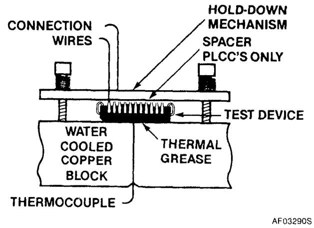


Figure 7. θ_{JC} Test Setup With PLCC Device

DATA PRESENTATION

The data presented in this application note was run at constant power dissipation for each package type. The power dissipation used is given under Test Conditions for each graph. Higher or lower power dissipation will have a slight effect on thermal resistance. The general trend of thermal resistance decreasing with increasing power is common to all packages. Figure 8 shows the average effect of power dissipation on SMD θ_{JA}.

Thermal resistance can also be affected by slight variations in internal leadframe design such as pad size. Larger pads give slightly lower thermal resistance for the same size die. The data presented represents the typical Signetics leadframe/die combinations with large die on large pads and small die on small pads. The effect of leadframe design is within the $\pm 15\%$ accuracy of these graphs.

SO devices are currently available in both copper or alloy 42 leadframes; however, Signetics is converting to copper only. PLCC devices are only available using copper leadframes.

The average lowering effect of air flow on SMD θ_{JA} is shown in Figure 9.

Thermal Calculations

The approximate junction temperature can be calculated using the following equation:

$$T_J = (\theta_{JA} \times P_D) + T_A$$

Where: T_J = Junction Temperature (°C)

θ_{JA} = Thermal Resistance Junction-to-Ambient (°C/W)

P_D = Power Dissipation at a T_J ($V_{CC} \times I_{CC}$) (W)

T_A = Temperature of Ambient (°C)

Example: Determine approximate junction temperature of SOL-20 at 0.5W dissipation using 10,000 sq. mil die and copper leadframe in still air and 200 LFPM air-flow ambients. Given $T_A = 30°C$,

1. Find θ_{JA} for SOL-20 using 10,000 sq. mil die and copper leadframe from typical θ_{JA} data — SOL-20 graph.

   Answer: 88°C/W @ 0.7W

2. Determine θ_{JA} @ 0.5W using Average Effect of Power Dissipation on AMD θ_{JA}, Figure 8.

   Percent change in Power

   $$= \frac{0.5W - 0.7W}{0.7W} \times 100$$

   $$= -28.6\%$$

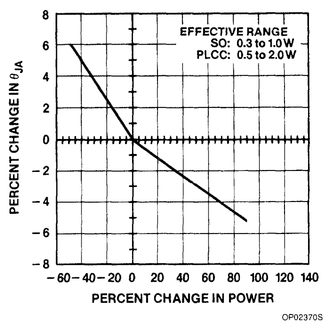


Figure 8. Average Effect of Power Dissipation on SMD θ_{JA}

From Figure 8:
28.6% change in power gives 3.5% increase in θ_{JA}

Answer:
88°C/W + (88 × 0.035)
= 91°C/W @ 0.5W

3. Determine θ_{JA} @ 0.5W in 200 LFPM air flow from Average Effect of Air Flow on SMD θ_{JA}, Figure 9.

From Figure 9: 200 LFPM air flow gives 14% decrease in θ_{JA}

Answer:
91°C/W – (91 × 0.14) = 78°C/W

4. Calculate approximate junction temperature

Answer:
T_J (still-air)
= (91°C/W × 0.5W) + 30
= 76°C
T_J (200 LFPM)
= (78°C/W × 0.5W) + 30
= 69°C

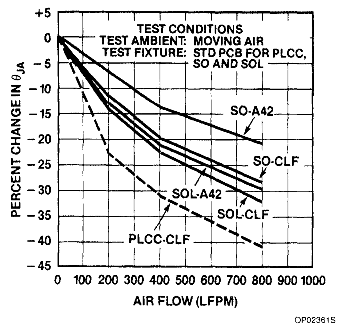


Figure 9. Average Effect of Air Flow on SMD θ_{JA}

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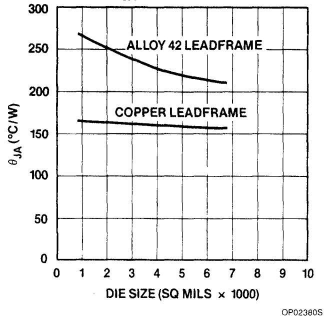
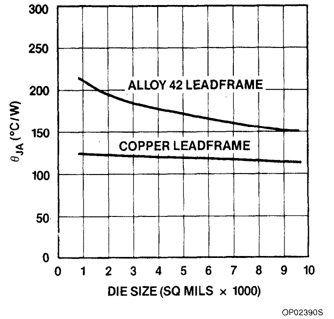
Thermal Considerations for Surface-Mounted Devices

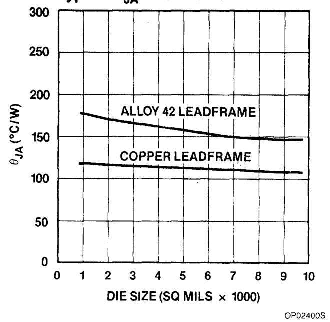
Typical θ_{JA} Data SO-8[1]



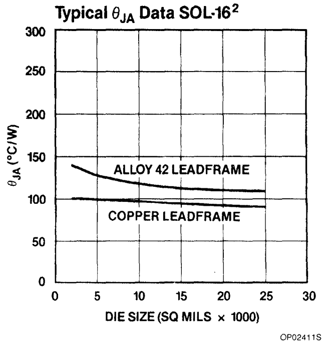
Typical θ_{JA} Data SO-14[1]



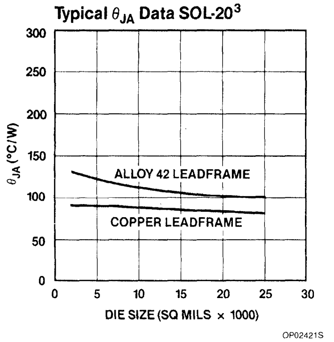
Typical θ_{JA} Data SO-16[1]



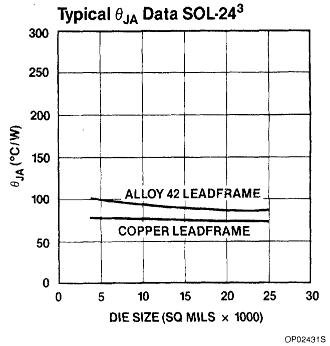
Typical θ_{JA} Data SOL-16[2]



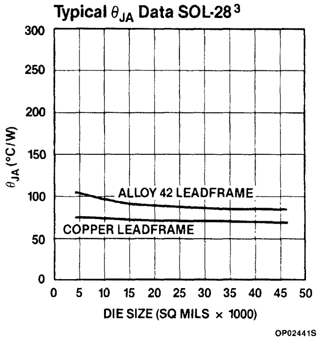
Typical θ_{JA} Data SOL-20[3]



Typical θ_{JA} Data SOL-24[3]



Typical θ_{JA} Data SOL-28[3]



NOTES:

1. TEST CONDITIONS:

Test ambient:	Still air
Power dissipation:	0.5W
Test fixture:	Philips PCB (1.12″ × 0.75″ × 0.059″)
Accuracy:	± 15%

2. TEST CONDITIONS:

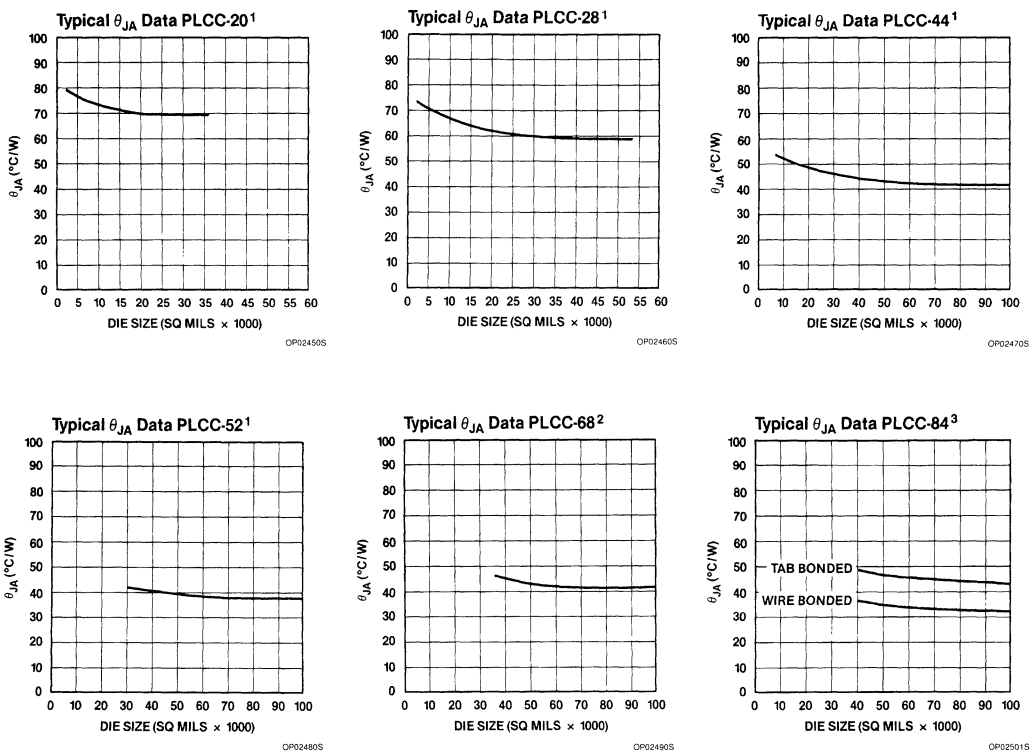
Test ambient:	Still air
Power dissipation:	0.5W
Test fixture:	Philips PCB (1.58″ × 0.75″ × 0.059″)
Accuracy:	± 15%

3. TEST CONDITIONS:

Test ambient:	Still air
Power dissipation:	0.7W
Test fixture:	Philips PCB (1.58″ × 0.75″ × 0.059″)
Accuracy:	± 15%

Figure 10. Typical SMD Thermal (θ_{JA}) Characteristics

Thermal Considerations for Surface-Mounted Devices



NOTES:

1. TEST CONDITIONS:
Test ambient: Still air
Power dissipation: 0.75W
Test fixture: Signetics PCB (2.24″ × 2.24″ × 0.062″)
Accuracy: ± 15%

2. TEST CONDITIONS:
Test ambient: Still air
Power dissipation: 1.0W
Test fixture: Signetics PCB (2.24″ × 2.24″ × 0.062″)
Accuracy: ± 15%

3. TEST CONDITIONS:
Test ambient Still air
Power dissipation: 1.5W
Test fixture: Signetics PCB (2.24″ × 2.24″ × 0.062″)
Accuracy: ± 15%

Figure 11. Typical SMD Thermal (θ_{JA}) Characteristics

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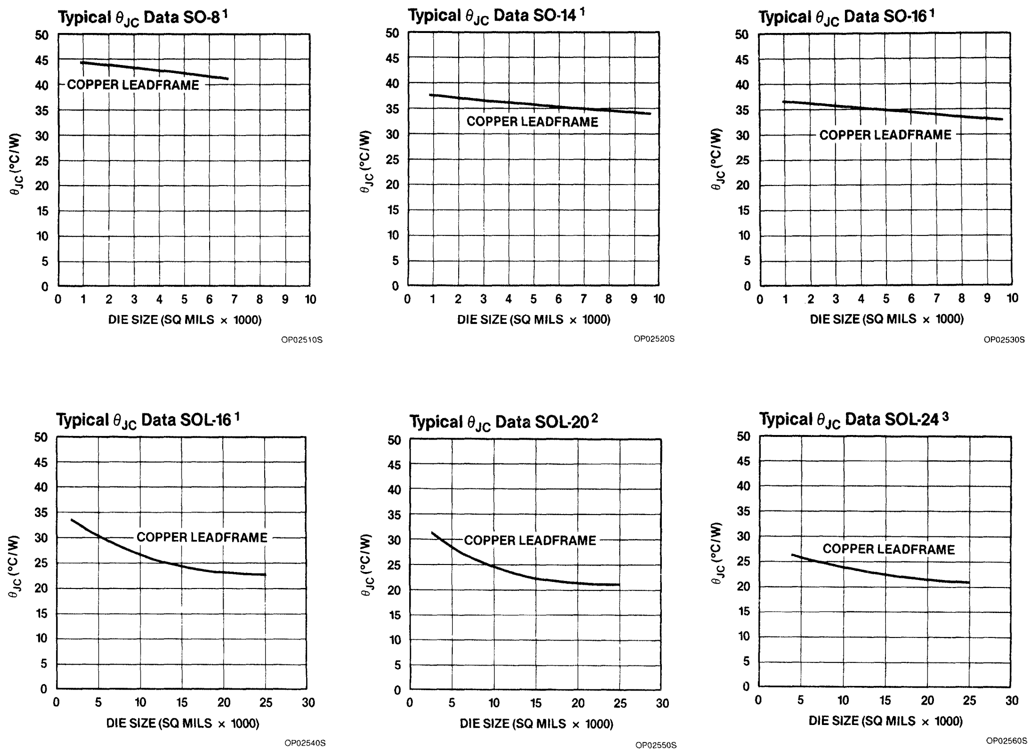
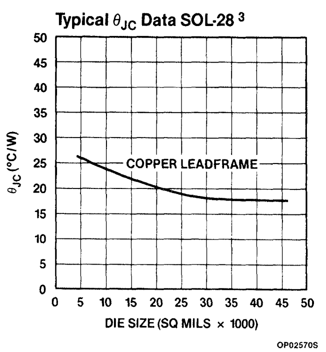
Thermal Considerations for Surface-Mounted Devices





NOTES:

1. TEST CONDITIONS:

Power dissipation:	0.5W
Test fixture:	"Infinite" heat sink
Accuracy:	± 15%

2. TEST CONDITIONS:

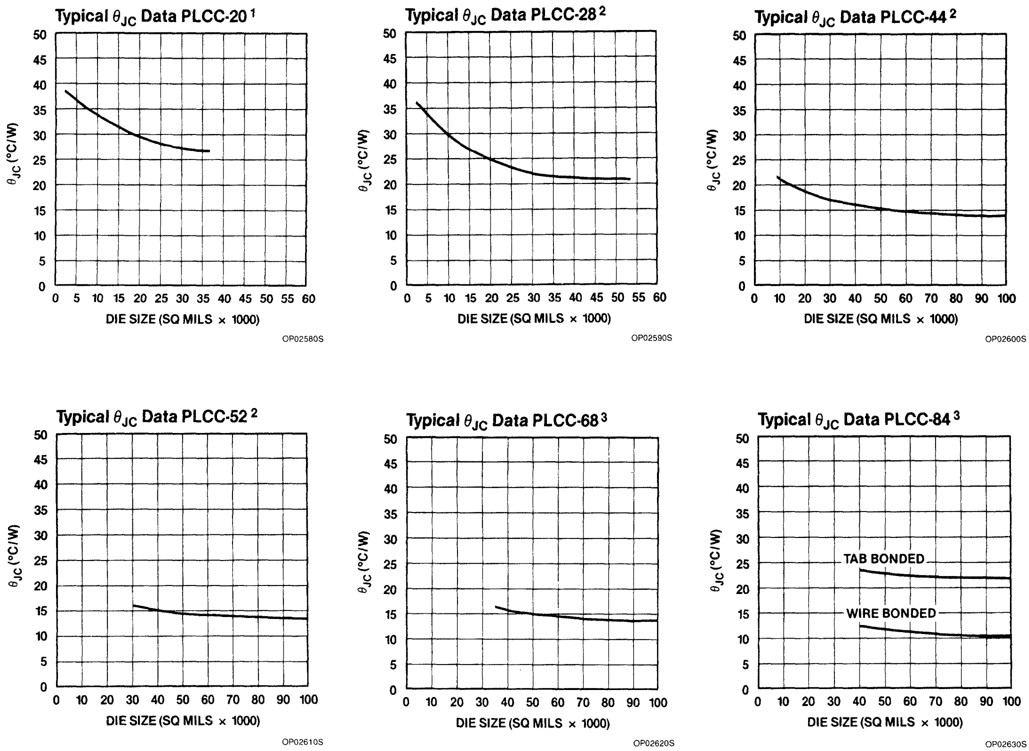
Power dissipation:	0.7W
Test fixture:	"Infinite" heat sink
Accuracy:	± 15%

3. TEST CONDITIONS:

Power dissipation:	1.0W
Test fixture:	"Infinite" heat sink
Accuracy:	± 15%

Figure 12. Typical SMD Thermal (θ_{JC}) Characteristics

Thermal Considerations for Surface-Mounted Devices



NOTES:

1. TEST CONDITIONS:

Power dissipation:	0.75W
Test fixture:	"Infinite" heat sink
Accuracy:	± 15%

2. TEST CONDITIONS:

Power dissipation:	1.0W
Test fixture:	"infinite" heat sink
Accuracy:	± 15%

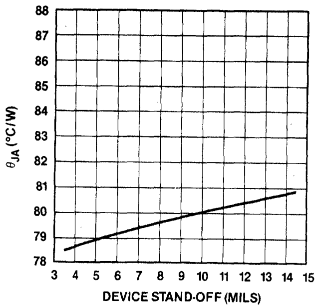
3. TEST CONDITIONS:

Power dissipation:	2.0W
Test fixture:	"Infinite" heat sink
Accuracy:	± 15%

Figure 13. Typical SMD Thermal (θ_{JC}) Characteristics

Thermal Considerations for Surface-Mounted Devices

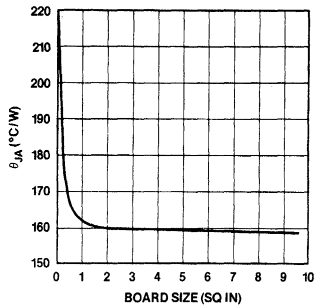
Effect of Device Stand-Off on SO θ_{JA} [1]



NOTES:
1. TEST CONDITIONS

Package type:	SOL-20 CLF
Die size:	11,322sq.mils
Test Ambient:	Still-air
Power dissipation:	0.75W
Test fixture:	Phillips PCB (1.58'' × 0.75'' × 0.059'')

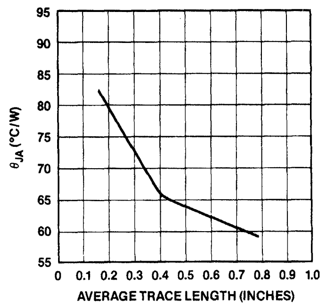
Effect of Board Size on SO θ_{JA} [2]



NOTES:
2. TEST CONDITIONS

Package type:	SOL-14 CLF
Die size:	5,040sq.mils
Test Ambient:	Still-air
Power dissipation:	0.6W
Test fixture:	0.062'' thick PCB with ''no traces'' 8 – 9mil stand-off

Effect of Trace Length on 28-Lead PLCC θ_{JA} [3]



NOTES:
3. TEST CONDITIONS

Package type:	PLCC-28 CLF
Die size:	10,445sq.mils
Test Ambient:	Still-air
Power dissipation:	1.0W
Test fixture:	Signetics PCB (2.24'' × 2.24'' × 0.062'') trace 27mil-wide 1oz sq.ft copper

Figure 14

Thermal Considerations for Surface-Mounted Devices

SYSTEM CONSIDERATIONS

With the increases in layout density resulting from surface mounting with much smaller packages, other factors become even more important. THE USER IS IN CONTROL OF THESE FACTORS.

One of the most obvious factors is the substrate material on which the parts are mounted. Environmental constraints, cost considerations, and other factors come into play when choosing a substrate. The choice is expanding rapidly, from the standard glass epoxy PWB materials and ceramic substrates to flexible circuits, injection-molded plastics, and coated metals. Each of these has its own thermal characteristics which must be considered when choosing a substrate material.

Studies have shown that the air gap between the bottom of the package and the substrate has an effect on θ_{JA}. The larger the gap, the higher the θ_{JA}. Using thermally conductive epoxies in this gap can slightly reduce the θ_{JA}.

It has long been recognized that external cooling can reduce the junction temperatures of devices by carrying heat away from both the devices and the board itself. Signetics has done several studies on the effects of external cooling on boards with SO packages. The results are shown in Figures 15 through 18.

The designer should avoid close spacing of high power devices so that the heat load is spread over as large an area as possible. Locate components with a higher junction temperature in the cooler locations on the PCBs.

The number and size of traces on a PWB can affect θ_{JA} since these metal lines can act as radiators, carrying heat away from the package and radiating it to the ambient. Although the chips themselves use the same amount of energy in either a DIP or an SO package, the increased density of a surface-mounted assembly concentrates the thermal energy into a smaller area.

It is evident that nothing is free in PWB layout. More heat concentrated into a smaller area makes it incumbent on the system designer to provide for the removal of thermal energy from his system.

Large conductor traces on the PCB conduct heat away from the package faster than small traces. Thermal vias from the mounting surface of the PCB to a large area ground plane in the PCB reduce the heat buildup at the package.

In addition to the package's thermal considerations, thermal management requires one to at least be aware of potential problems caused by mismatch in thermal expansion.

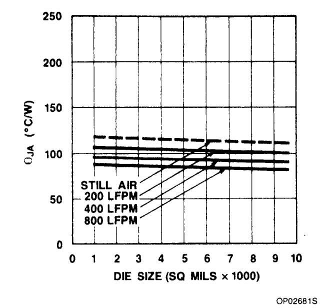


Figure 15. Results of Air Flow on θ_{JA} on SO-14 With Copper Leadframe

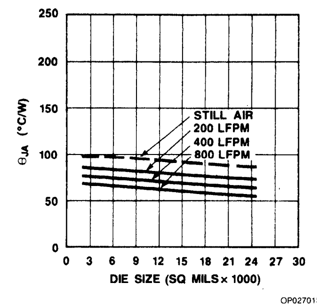


Figure 16. Results of Air Flow on θ_{JA} on SOL-16 With Copper Leadframe

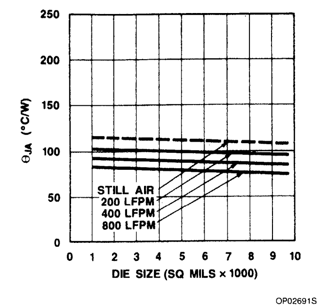


Figure 17. Results of Air Flow on θ_{JA} on SO-16 With Copper Leadframe

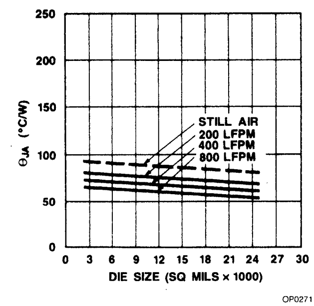


Figure 18. Results of Air Flow on θ_{JA} on SOL-20 With Copper Leadframe

The very nature of the SMD assembly, where the devices are soldered directly onto the surface, not through it, results in a very rigid structure. If the substrate material exhibits a different thermal coefficient of expansion (TCE) than the IC package, stresses can be set up in the solder joints when they are subjected to temperature cycling (and during the soldering process itself) that may ultimately result in failure.

Because some of the boards assembled will require the use of Leadless Ceramic Chip Carriers (LCCCs), TCE must be understood. As will be seen below, TCE is less of a problem with the commercial SMD packages with leads.

Take the example of a leadless ceramic chip carrier with a TCE of about 6×10^{-6}/°C soldered to a conventional glass-epoxy laminate with a TCE in the region of 16×10^{-6}/°C. This thermal expansion mismatch has been shown to fracture the solder joints during thermal cycling. Substrate materials with matched TCEs should be evaluated for these SMD assemblies to avoid problems caused by thermal expansion mismatch.

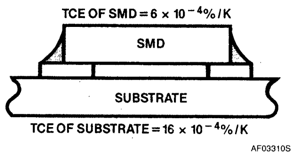
The stress level associated with thermal expansion and contraction of small SMDs such as capacitors and resistors, where the actual change in length is small, is normally rather low. However, as component sizes increase, stresses can increase substantially.

Thermal expansion mismatch is unlikely to cause too many problems in systems operating in benign environments; but, in harsher conditions, such as thermal cycling in military or avionic applications, the mechanical stresses set up in solder joints due to the different TCEs of the substrate and the component are likely to cause failure.

The basic problem is outlined in Figure 19. The leadless SMD is soldered to the substrate as shown, resulting in a very rigid structure. If the substrate material exhibits a different TCE from that of the SMD material, the amount of expansion for each will differ for any given increase in temperature. The soldered joint will have to accommodate this difference, and failure can ultimately result. The larger the component size, the higher the stress levels so that this phenomenon is at its

Thermal Considerations for Surface-Mounted Devices

most critical in applications requiring large LCCCs with high pin counts.



NOTE:
Data provided by N.V. Philips

Figure 19. The Basic Problem of Thermal Expansion Mismatch Is That the Substrate and Component May Each Have Different Thermal Coefficients of Expansion

To address this problem, three basic solutions are emerging. First, the use of leadless ceramic chip carriers can sometimes be avoided by using leaded devices; the leads can flex and absorb the stress. Second, when this solution is not feasible, the stresses can be taken up by inserting a compliant elastomeric layer between the ceramic package and the epoxy glass substrate. Third, TCE values of component and substrate can be matched.

USING LEADED DEVICES (SO, SOL, and PLCC)

The current evolution in commercial electronics includes the adoption of the commercial SMD packages, i.e., SO with gull-wing leads or the PLCC with rolled-under J-leads, rely on the compliance of the leads themselves to avoid any serious problems of thermal expansion mismatch. At elevated temperatures, the leads flex slightly and absorb most of the mechanical stress resulting from the thermal expansion differentials.

Similarly, leaded holders can be used with LCCCs to attach them to the substrate and thus absorb the stress.

Unfortunately, using a lead does not always ensure sufficient compliancy. The material from which the lead is made, and the way it is formed and soldered can adversely affect it. For example, improper soldering techniques, which cause excess solder to over-fill the bend of the gull-wing lead of an SO, can significantly reduce the lead's compliancy.

COMPLIANT LAYER

This approach introduces a compliant layer onto the interface surface of the substrate to absorb some of the stresses. A 50μm thick elastomeric layer is bonded to the laminate. To make contacts, carbon or metallic powders are introduced to form conductive stripes in the nonconductive elastomer material. Unfortunately, substrates using this technique are substantially more expensive than standard uncoated boards.

Another solution is to increase the compliancy of the solder joint. This is done by increasing the stand-off height between the underside of the component and the substrate. To do this, a solder paste containing lead or ceramic spheres which do not melt when the surrounding solder reflows, thus keeping the component above the substrate, can be used.

MATCHING TCE

There are two ways to approach this solution. The TCE of the substrate laminate material can be matched to that of the LCCC either by replacing the glass fibers with fibers exhibiting a lower TCE (composites such as epoxy-Kevlar®or polyimide-Kevlar and polyimide-quartz), or by using low TCE metals (such as Invar®, Kovar, or molybdenum).

This latter approach involves bonding a glass-polyimide or a glass-epoxy multilayer to the low TCE restraining core material. Typical of such materials are copper-Invar-copper, Alloy-42, copper-molybdenum-copper, and copper-graphite. These restraining-core constructions usually require that the laminate be bonded to both sides to form a balanced structure so that they will not warp or twist.

This inevitably means an increase in weight, which has always been a negative factor in this approach. However, the SMD substrate can be smaller and the components more densely packed, in many cases overcoming the weight disadvantages. On the positive side, the material's high thermal conductivity helps to keep the components cool. Moreover, copper-clad Invar lends itself readily to moisture-proof multilayering for the creation of ground and power planes and for providing good inherent EMI/RFI shielding.

Kevlar is lighter and widely used for substrates in military applications; but, it suffers from a serious drawback which, although overcome to a certain extent by careful attention to detail, can cause problems. The material, when laminated, can absorb moisture and chemical processing fluids around the edges. Thermal conductivity, machinability, and cost are not as attractive as for copper-clad Invar.

For the majority of commercial substrates, however, where the use of ceramic chip carriers in any quantity is the exception rather than the rule, and when adequate cooling is available, the mismatch of TCEs poses little or no problem. For these substrates, traditional FR-4 glass-epoxy and phenolic-paper will no doubt remain the most widely-used materials.

Although FR-4 epoxy-glass has been the traditional material for plated-through professional substrates, it is phenolic-paper laminate (FR-2) which finds the widest use in consumer electronics. While it is the cheapest material, it unfortunately has the lowest dimensional stability, rendering it unsuitable for the mounting of LCCCs.

SUBSTRATE TYPES

FR-4 glass-epoxy substrates are the most commonly used for commercial electronic circuits. They have the advantage of being cheap, machinable, and lightweight. Substrate size is not limited. On the negative side, they have poor thermal conductivity and a high TCE, between 13 and 17 $\times$ 10^{-6}/°C. This means they are a poor match to ceramic.

Glass polyimide substrates have a similar TCE range to glass-epoxy boards, but better thermal conductivity. They are, however, three to four times more expensive.

Polyimide Kevlar substrates have the advantage of being lightweight and not restricted in size. Conventional substrate processing methods can be used and its TCE (between 4 and 8), matches that of ceramic. Its disadvantages are that it is expensive, difficult to drill, and is prone to resin microcracking and water absorption.

Polyimide quartz substrates have a TCE between 6 and 12, making them a good match for LCCCs. They can be processed using conventional techniques, although drilling vias can be difficult. They have good dielectric properties and compare favorably with FR-4 for substrate size and weight.

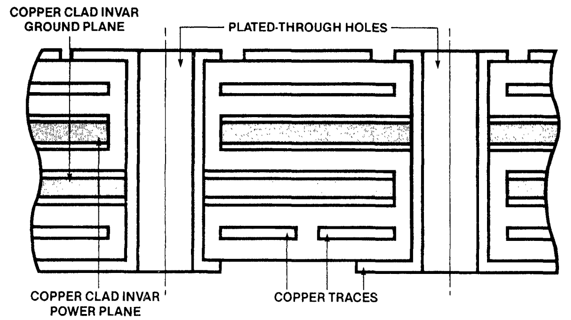
Alumina (ceramic) substrates are used extensively for high-reliability military applications and thick-film hybrids. The weight, cost, limited substrate size and inherent brittleness of alumina means that its use as a substrate material is limited to applications where these disadvantages are outweighed by the advantage of good thermal conductivity and a TCE that exactly matches that of LCCCs. A further limitation is that they require thick-film screening processing.

Copper-clad Invar substrates are the leading contenders for TCE control at present. It can be tailored to provide a selected TCE by varying the copper-to-Invar ratio. Figure 20 shows the construction of a typical multilayer substrate employing two cores providing the power and ground planes. Plated-through holes provide an integral board-to-board interconnection. The low TCE of the core dominates the TCE of the overall substrate,

Thermal Considerations for Surface-Mounted Devices

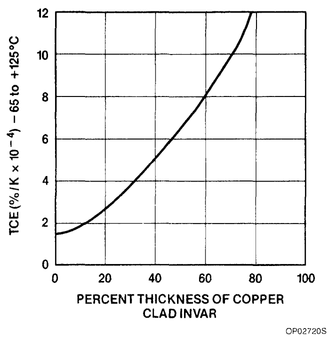
making it possible to mount LCCCs with confidence.

Because the TCE of copper is high, and that of Invar is low, the overall TCE of the substrate can be adjusted by varying the thickness of the copper layers. Figure 21 plots the TCE range of the copper-clad Invar as a function of copper thickness and shows the TCE range of each of several other materials to which the clad material can be matched. For example, if the TCE of Alumina is to be matched, then the core should have about 46% thickness of copper. When this material is used as a thermal mounting plane, it also acts as a heatsink.



NOTE:
Data provided by N.V. Philips

Figure 20. Section Through a Typical Multilayer Substrate Incorporating Copper-Clad Invar Ground and Power Planes, Interconnected via Plated-Through Holes



NOTE:
Data provided by N.V. Philips

Figure 21. The TCE Range of Copper-Clad Invar as a Function of Copper Thickness

Thermal Considerations for Surface-Mounted Devices

Table 1. Substrate Material Properties

SUBSTRATE MATERIAL	TCE (10^{-6}/°C)	THERMAL CONDUCTIVITY (W/m^{3}K)
Glass-epoxy (FR-4)	13 – 17	0.15
Glass polyimide	12 – 16	0.35
Polyimide Kevlar	4 – 8	0.12
Polyimide quartz	6 – 12	TBD
Copper-clad Invar	6.4 (typical)	165 (lateral) 16 (transverse)
Alumina	5 – 7	21
Compliant layer Substrate	See Notes	0.15 – 0.3

NOTES:
Compliant layer conforms to TCE of the LCCC and to base substrate material.
Data provided by N.V. Philips
KEVLAR® is a registered trademark of DU PONT.
INVAR® is a registered trademark of TEXAS INSTRUMENTS.

CONCLUSION

Thermal management remains a major concern of producers and users of ICs. The advent of SMD technology has made a thorough understanding of the thermal characteristics of both the devices and the systems they are used in mandatory. The SMD package, being smaller, does have a higher θ_{JA} than its standard DIP counterpart . . . even with copper leadframes. That is the major trade-off one accepts for package miniaturization. However, consideration of all the variables affecting IC junction temperatures will allow the user to take maximum advantage of the benefits derived from use of this technology.

Signetics

Package Outlines
For Prefixes ADC, AM, AU, CA, DAC, ICM, LF, LM, MC, NE, SA, SE, SG, μA, UC

Linear Products

INTRODUCTION

The following information applies to all packages unless otherwise specified on individual package outline drawings.

GENERAL

1. Dimensions shown are metric units (millimeters), except those in parentheses which are English units (inches).
2. Lead spacing shall be measured within this zone.
   a. Shoulder and lead tip dimensions are to centerline of leads.
3. Tolerances non-cumulative.
4. Thermal resistance values are determined by utilizing the linear temperature dependence of the forward voltage drop across the substrate diode in a digital device to monitor the junction temperature rise during known power application across V_{CC} and ground. The values are based upon 120mils square die for plastic packages and a 90mils square die in the smallest available cavity for hermetic packages. All units were solder-mounted to PC boards, with standard stand-off, for measurement.

PLASTIC ONLY

5. Lead material: Alloy 42 (Nickel/Iron Alloy), Olin 194 (Copper Alloy), or equivalents, solder-dipped.
6. Body material: Plastic (Epoxy)
7. Round hole in top corner denotes lead No. 1.
8. Body dimensions do not include molding flash.
9. SO packages/microminiature packages:
   a. Lead material: Alloy-42.
   b. Body material: Plastic (Epoxy).

HERMETIC ONLY

10. Lead material
    a. ASTM alloy F-15 (KOVAR) or equivalent — gold-plated, tin-plated, or solder-dipped.
    b. ASTM alloy F-30 (Alloy 42) or equivalent — tin-plated, gold-plated or solder-dipped.
    c. ASTM alloy F-15 (KOVAR) or equivalent — gold-plated.
11. Body Material
    a. Eyelet, ASTM alloy F-15 or equivalent — gold- or tin-plated, glass body.
    b. Ceramic with glass seal at leads.
    c. BeO ceramic with glass seal at leads.
    d. Ceramic with ASTM alloy F-30 or equivalent.
12. Lid Material
    a. Nickel- or tin-plated nickel, weld seal.
    b. Ceramic, glass seal.
    c. ASTM alloy F-15 or equivalent, gold-plated, alloy seal.
    d. BeO ceramic with glass seal.
13. Signetics symbol, angle cut, or lead tab denotes Lead No. 1.
14. Recommended minimum offset before lead bend.
15. Maximum glass climb 0.010 inches.
16. Maximum glass climb or lid skew is 0.010 inches.
17. Typical four places.
18. Dimension also applies to seating plane.

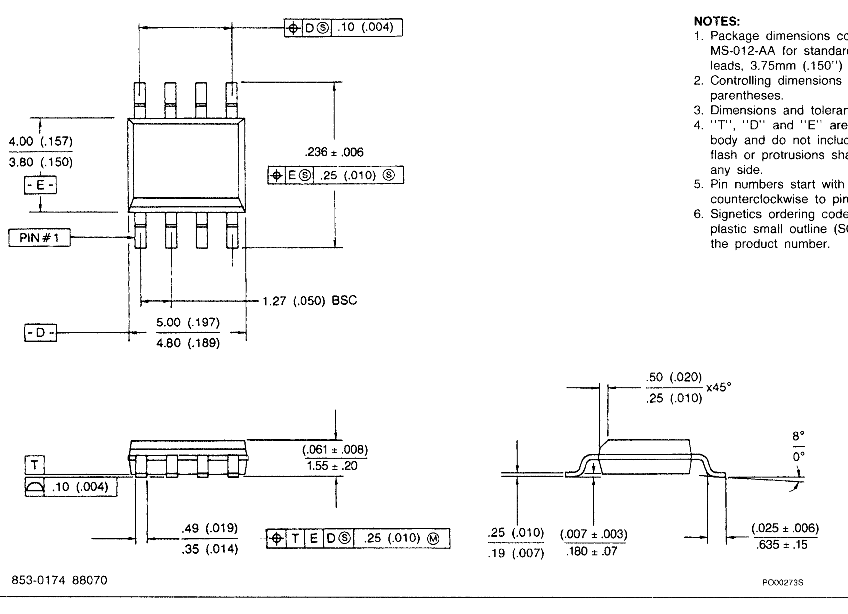
For Prefixes ADC, AM, AU, CA, DAC, ICM, LF, LM, MC, NE, SA, SE, SG, μA, UC

Package Outlines

PLASTIC PACKAGES

DESCRIPTION	PACKAGE CODE	θ_{JA}/θ_{JC} (°C/W)	PACKAGE TYPE
Standard Dual-in-Line Packages			
8-Pin	N	110/49	
14-Pin	N	90/46	TO-116/MO-001
16-Pin	N	90/46	MO-001
18-Pin	N	79/36	
20-Pin	N	79/35	
22-Pin	N	56/23	
24-Pin	N	58/30	MO-015
28-Pin	N	56/30	MO-015
Metal Headers			
4-Pin	E	100/20	TO-46 Header
4-Pin	E	150/25	TO-72 Header
8-Pin	H	150/25	TO-5 Header
10-Pin	H	150/25	TO-5/TO-100 Header, Short Can
10-Pin	H	150/25	TO-5/TO-100 Header, Tall Can
Cerdip Family			
8-Pin	FE	162/26	Dual-in-Line Ceramic
14-Pin	F	109/26	Dual-in-Line Ceramic
16-Pin	F	105/26	Dual-in-Line Ceramic
18-Pin	F	88/22	Dual-in-Line Ceramic
20-Pin	F	85/22	Dual-in-Line Ceramic
22-Pin	F	75/13	Dual-in-Line Ceramic
24-Pin	F	65/16	Dual-in-Line Ceramic
28-Pin	F	62/16	Dual-in-Line Ceramic
Laminated Ceramic, Side-Brazed Lead			
16-Pin	I	90/25	DIP Laminate

8-PIN PLASTIC SO (D PACKAGE)

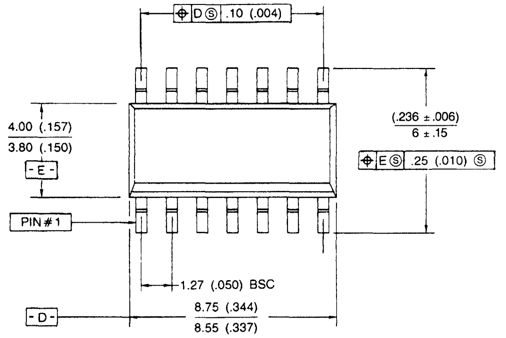


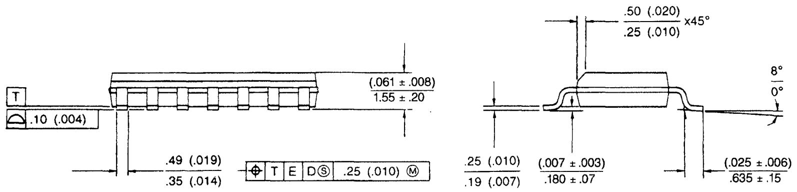
NOTES:
1. Package dimensions conform to JEDEC specification MS-012-AA for standard small outline (SO) package, 8 leads, 3.75mm (.150") body width (issue A, June 1985).
2. Controlling dimensions are in mm. Inch dimensions in parentheses.
3. Dimensions and tolerancing per ANSI Y14.5M- 1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .15mm (.006") on any side.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #8 when viewed from top.
6. Signetics ordering code for a product packaged in a plastic small outline (SO) package is the suffix D after the product number.

For Prefixes ADC, AM, AU, CA, DAC, ICM, LF, LM, MC, NE, SA, SE, SG, μA, UC

Package Outlines

14-PIN PLASTIC SO (D PACKAGE)



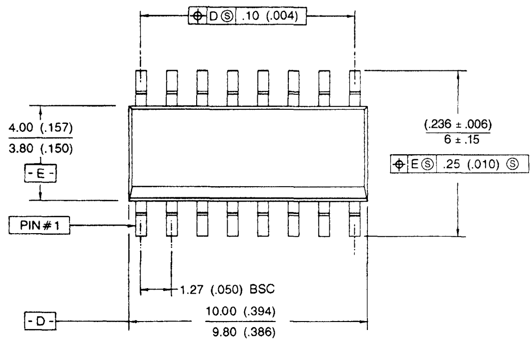


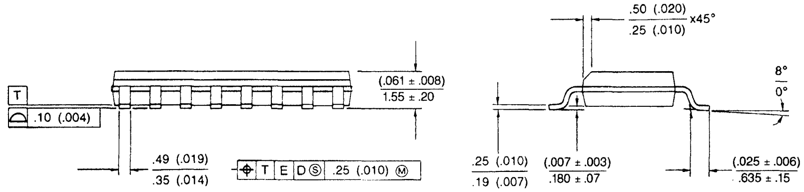
853-0175 88068

PO00263S

NOTES:
1. Package dimensions conform to JEDEC specification MS-012-AB for standard small outline (SO) package, 14 leads, 3.75mm (.150") body width (issue A, June 1985).
2. Controlling dimensions are in mm. Inch dimensions in parentheses.
3. Dimensions and tolerancing per ANSI Y14.5M- 1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .15mm (.006") on any side.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #14 when viewed from top.
6. Signetics ordering code for a product packaged in a plastic small outline (SO) package is the suffix D after the product number.

16-PIN PLASTIC SO (D PACKAGE)





853-0005 88069

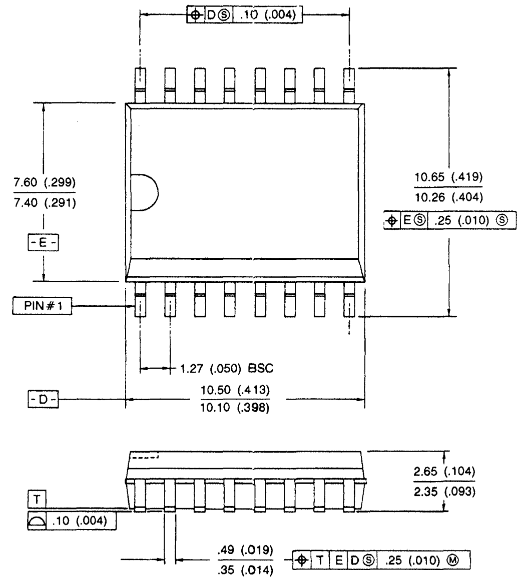
PO00252S

NOTES:
1. Package dimensions conform to JEDEC specification MS-012-AC for standard small outline (SO) package, 16 leads, 3.75mm (.150") body width (issue A, June 1985).
2. Controlling dimensions are in mm. Inch dimensions in parentheses.
3. Dimensions and tolerancing per ANSI Y14.5M- 1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .15mm (.006") on any side.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #16 when viewed from top.
6. Signetics ordering code for a product packaged in a plastic small outline (SO) package is the suffix D after the product number.

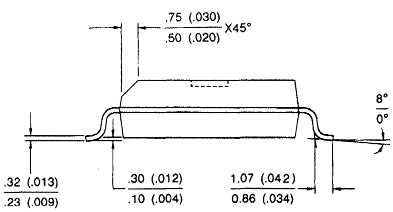
For Prefixes ADC, AM, AU, CA, DAC, ICM, LF, LM, MC, NE, SA, SE, SG, μA, UC

Package Outlines

16-PIN PLASTIC SOL (D PACKAGE)



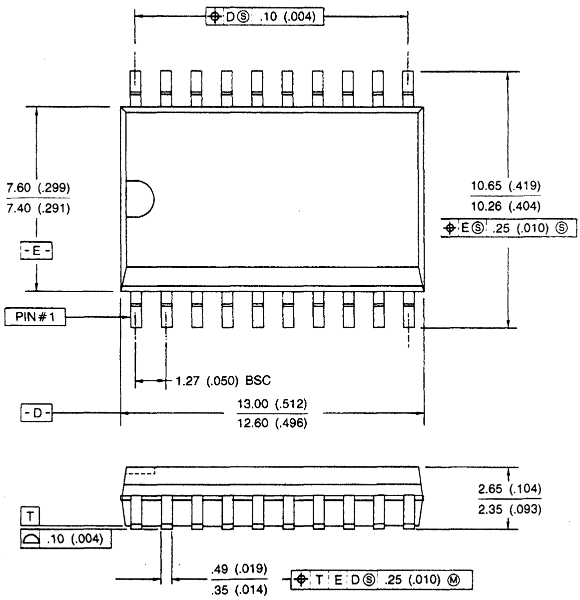
NOTES:
1. Package dimensions conform to JEDEC specification MS-013-AA for standard small outline (SO) package, 16 leads, 7.50mm (.300") body width (issue A, June 1985).
2. Controlling dimensions are in mm. Inch dimensions in parentheses.
3. Dimensions and tolerancing per ANSI Y14.5M- 1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .15mm (.006") on any side.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #16 when viewed from top.
6. Signetics ordering code for a product packaged in a plastic small outline (SO) package is the suffix D after the product number.



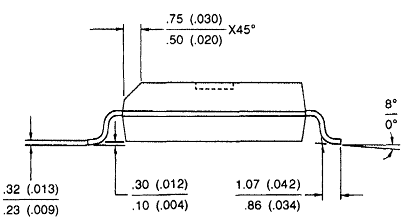
853-0171 81218

PO00241S

20-PIN PLASTIC SOL (D PACKAGE)



NOTES:
1. Package dimensions conform to JEDEC specification MS-013-AC for standard small outline (SO) package, 20 leads, 7.50mm (.300") body width (issue A, June 1985).
2. Controlling dimensions are in mm. Inch dimensions in parentheses.
3. Dimensions and tolerancing per ANSI Y14.5M- 1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .15mm (.006") on any side.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #20 when viewed from top.
6. Signetics ordering code for a product packaged in a plastic small outline (SO) package is the suffix D after the product number.



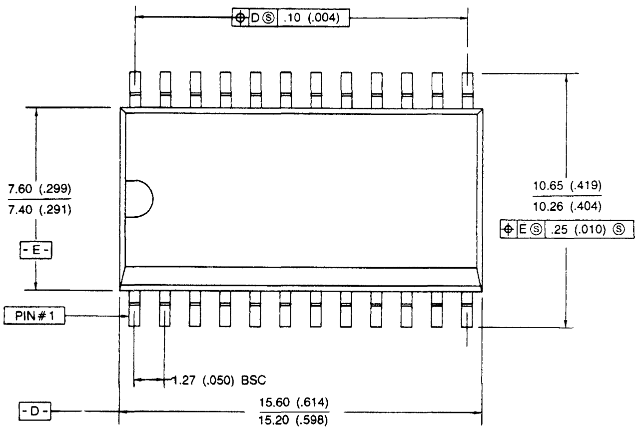
853-0172 81219

PO00231S

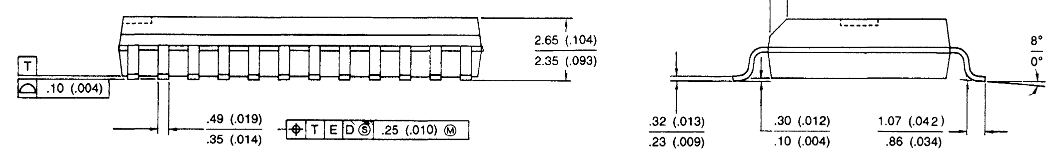
For Prefixes ADC, AM, AU, CA, DAC, ICM, LF, LM, MC, NE, SA, SE, SG, μA, UC

Package Outlines

24-PIN PLASTIC SOL (D PACKAGE)



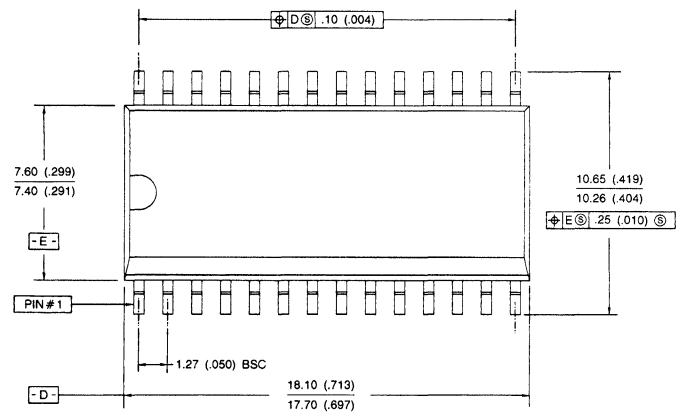
NOTES:
1. Package dimensions conform to JEDEC specification MS-013-AD for standard small outline (SO) package, 24 leads, 7.50mm (.300") body width (issue A, June 1985).
2. Controlling dimensions are in mm. Inch dimensions in parentheses.
3. Dimensions and tolerancing per ANSI Y14.5M- 1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .15mm (.006") on any side.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #24 when viewed from top.
6. Signetics ordering code for a product packaged in a plastic small outline (SO) package is the suffix D after the product number.



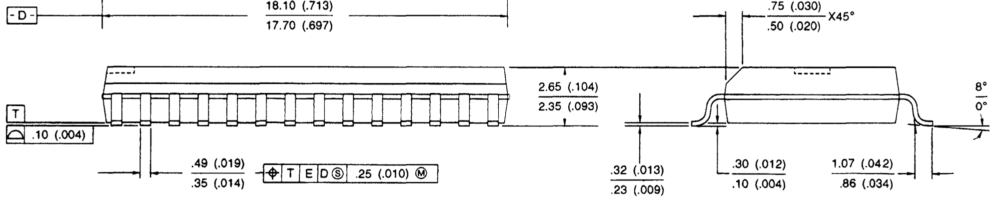
853-0173 81220

PO00221S

28-PIN PLASTIC SOL (D PACKAGE)



NOTES:
1. Package dimensions conform to JEDEC specification MS-013-AE for standard small outline (SO) package, 28 leads, 7.50mm (.300") body width (issue A, June 1985).
2. Controlling dimensions are in mm. Inch dimensions in parentheses.
3. Dimensions and tolerancing per ANSI Y14.5M- 1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .15mm (.006") on any side.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #28 when viewed from top.
6. Signetics ordering code for a product packaged in a plastic small outline (SO) package is the suffix D after the product number.



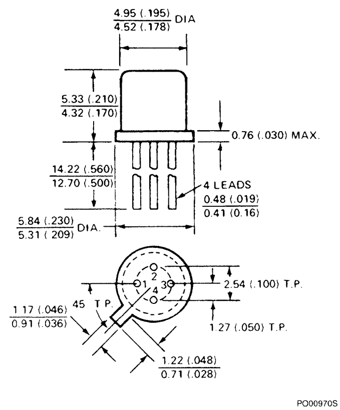
853-0006 81217

PO00212S

For Prefixes ADC, AM, AU, CA, DAC, ICM, LF, LM, MC, NE, SA, SE, SG, μA, UC

Package Outlines

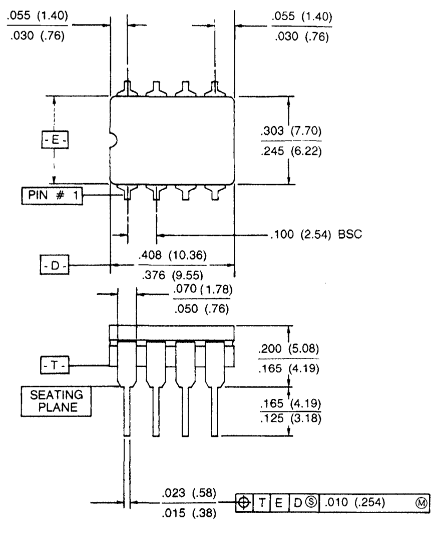
4-PIN HERMETIC TO-72 HEADER (E PACKAGE)

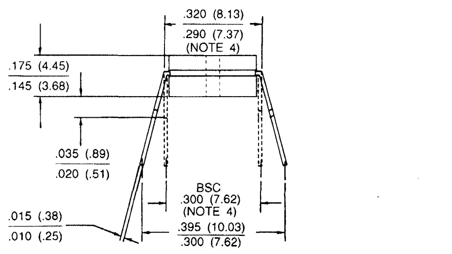


8-PIN CERDIP (FE PACKAGE)

NOTES:

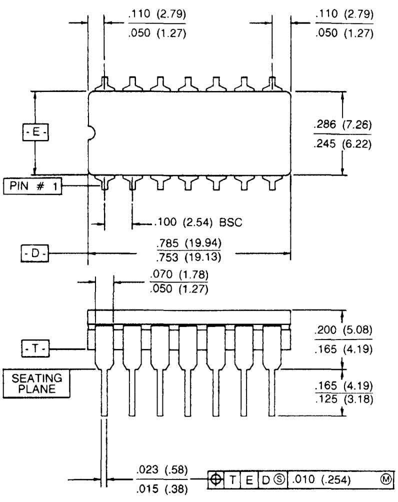
1. Controlling dimension: inches. Millimeters are shown in parentheses.
2. Dimensions and tolerancing per ANSI Y14.5M – 1982.
3. "T", "D", and "E" are reference datums on the body and include allowance for glass overrun and meniscus on the seal line, and lid to base mismatch.
4. These dimensions measured with the leads constrained to be perpendicular to plane T.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #8 when viewed from the top.

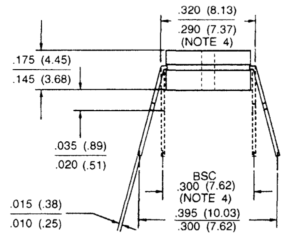




For Prefixes ADC, AM, AU, CA, DAC, ICM, LF, LM, MC, NE, SA, SE, SG, μA, UC — Package Outlines

14-PIN CERDIP (F PACKAGE)



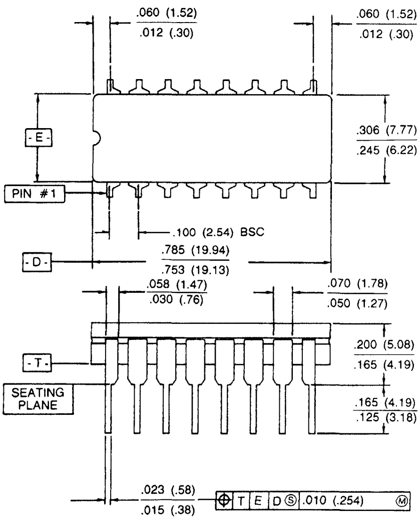


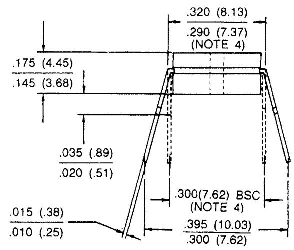
NOTES:
1. Controlling dimension: inches. Millimeters are shown in parentheses.
2. Dimensions and tolerancing per ANSI Y14.5M - 1982.
3. "T", "D", and "E" are reference datums on the body and include allowance for glass overrun and meniscus on the seal line, and lid to base mismatch.
4. These dimensions measured with the leads constrained to be perpendicular to plane T.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #14 when viewed from the top.

853-0581 81594

PO00440S

16-PIN CERDIP (F PACKAGE)





NOTES:
1. Controlling dimension: inches. Millimeters are shown in parentheses.
2. Dimensions and tolerancing per ANSI Y14.5M - 1982.
3. "T", "D", and "E" are reference datums on the body and include allowance for glass overrun and meniscus on the seal line, and lid to base mismatch.
4. These dimensions measured with the leads constrained to be perpendicular to plane T.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #16 when viewed from the top.

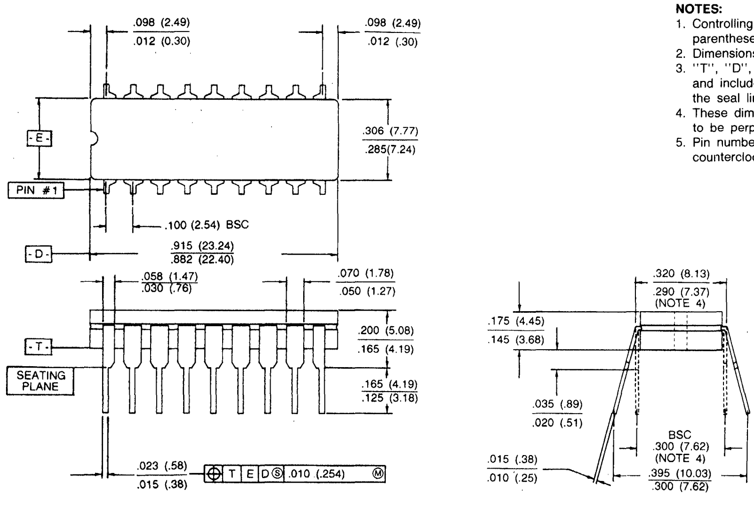
853-0582 81594

PO00490S

For Prefixes ADC, AM, AU, CA, DAC, ICM, LF, LM, MC, NE, SA, SE, SG, μA, UC

Package Outlines

18-PIN CERDIP (F PACKAGE)

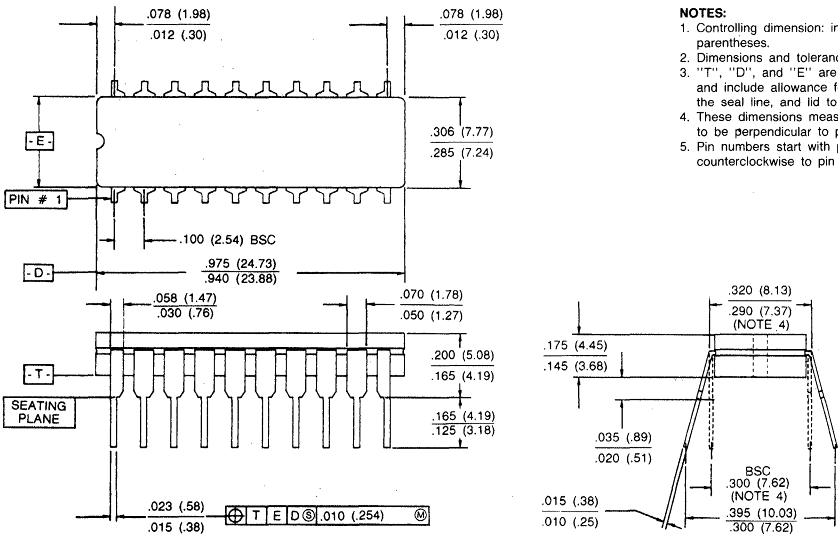


NOTES:
1. Controlling dimension: inches. Millimeters are shown in parentheses.
2. Dimensions and tolerancing per ANSI Y14.5M – 1982.
3. "T", "D", and "E" are reference datums on the body and include allowance for glass overrun and meniscus on the seal line, and lid to base mismatch.
4. These dimensions measured with the leads constrained to be perpendicular to plane T.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #18 when viewed from the top.

853-0583 81594

PO00470S

20-PIN CERDIP (F PACKAGE)



NOTES:
1. Controlling dimension: inches. Millimeters are shown in parentheses.
2. Dimensions and tolerancing per ANSI Y14.5M – 1982.
3. "T", "D", and "E" are reference datums on the body and include allowance for glass overrun and meniscus on the seal line, and lid to base mismatch.
4. These dimensions measured with the leads constrained to be perpendicular to plane T.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #20 when viewed from the top.

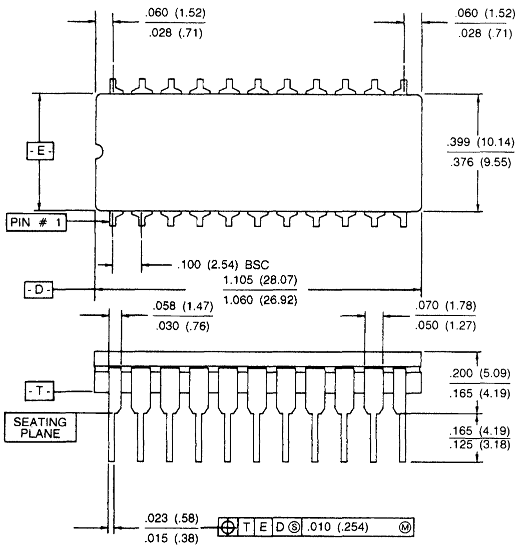
853-0584 81594

PO00430S

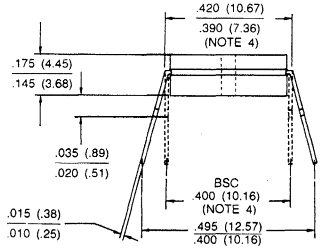
For Prefixes ADC, AM, AU, CA, DAC, ICM, LF, LM, MC, NE, SA, SE, SG, μA, UC

Package Outlines

22-PIN CERDIP (F PACKAGE)



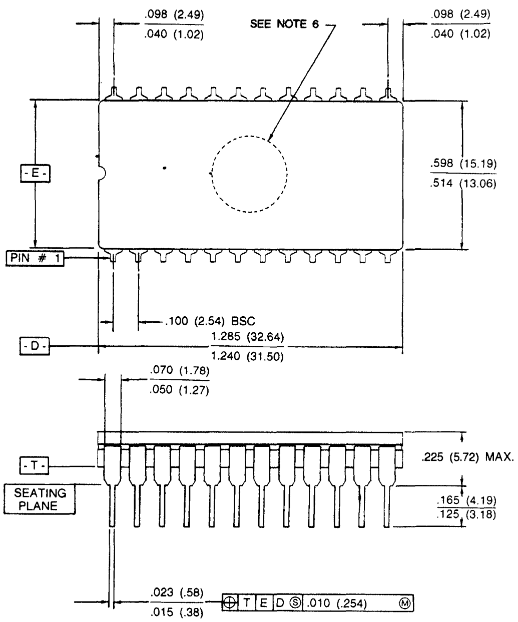
NOTES:
1. Controlling dimension: inches. Millimeters are shown in parentheses.
2. Dimensions and tolerancing per ANSI Y14.5M – 1982.
3. "T", "D", and "E" are reference datums on the body and include allowance for glass overrun and meniscus on the seal line, and lid to base mismatch.
4. These dimensions measured with the leads constrained to be perpendicular to plane T.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #22 when viewed from the top.



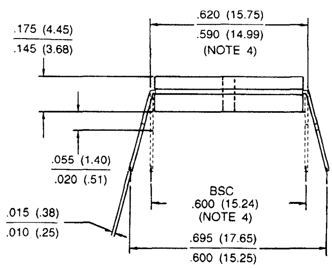
853-0585 81594

PO00420S

24-PIN CERDIP (F PACKAGE)



NOTES:
1. Controlling dimension: inches. Millimeters are shown in parentheses.
2. Dimensions and tolerancing per ANSI Y14.5M – 1982.
3. "T", "D", and "E" are reference datums on the body and include allowance for glass overrun and meniscus on the seal line, and lid to base mismatch.
4. These dimensions measured with the leads constrained to be perpendicular to plane T.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #24 when viewed from the top.



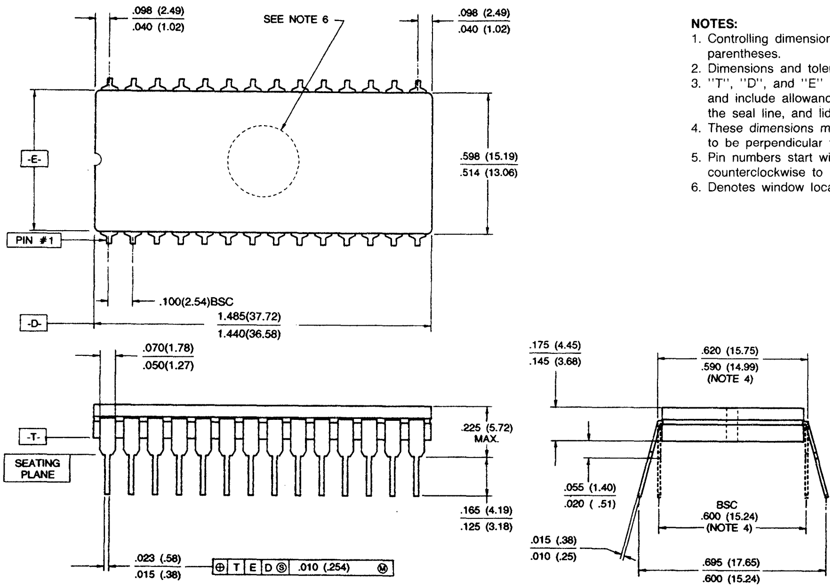
853-0588 84221

PO00464S

For Prefixes ADC, AM, AU, CA, DAC, ICM, LF, LM, MC, NE, SA, SE, SG, μA, UC

Package Outlines

28-PIN CERDIP (F PACKAGE)

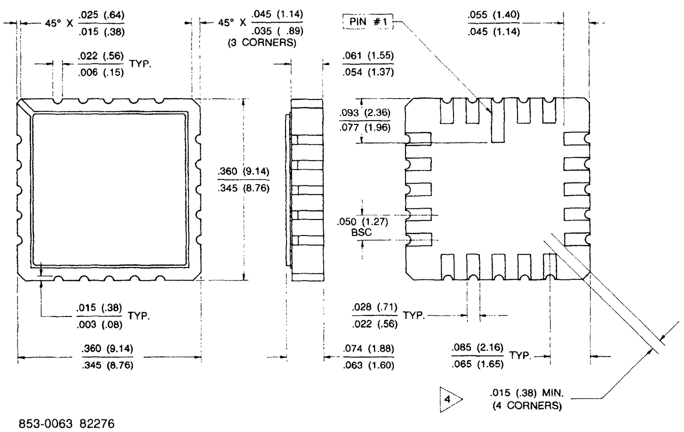


NOTES:
1. Controlling dimension: inches. Millimeters are shown in parentheses.
2. Dimensions and tolerancing per ANSI Y14.5M – 1982.
3. "T", "D", and "E" are reference datums on the body and include allowance for glass overrun and meniscus on the seal line, and lid to base mismatch.
4. These dimensions measured with the leads constrained to be perpendicular to plane T.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #28 when viewed from the top.
6. Denotes window location for EPROM Products.

853-0589 84000

PO00660S

20-PIN PGA (G PACKAGE)



NOTES:
1. Package dimensions conform to Mil-M-38510, outline NO. C – 2, 20 leads, square ceramic leadless chip carrier.
2. Controlling dimension: inches, millimeters are shown in parenthesis.
3. Dimension and tolerancing per ANSI Y14.5M – 1982.
4. This dimension represents the minimum spacing between the corner contact pads. These corner pads may have a .020 inch by 45 degree maximum chamfer to accomplish the .015 *minimum spacing.*
5. Pin numbers start with pin #1 and continue counterclockwise to pin #20 when viewed from the top.
6. Signetics order code for product packaged in a CCCL is the suffix "G" after the product number.

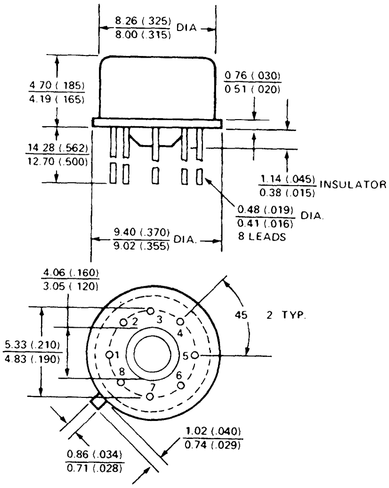
853-0063 82276

PO00870S

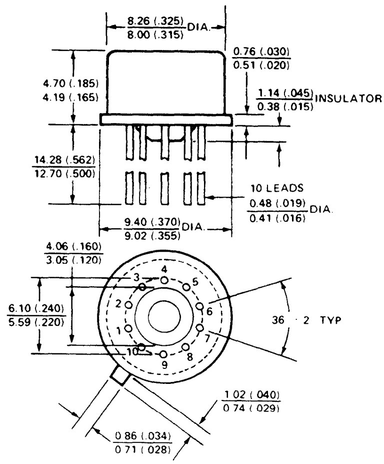
For Prefixes ADC, AM, AU, CA, DAC, ICM, LF, LM, MC, NE, SA, SE, SG, μA, UC

Package Outlines

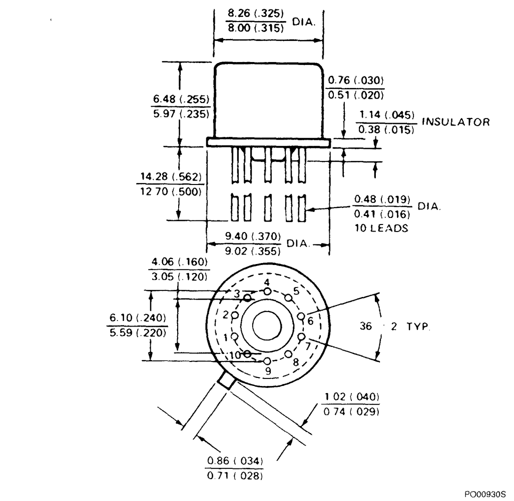
8-PIN HERMETIC TO-5 HEADER (H PACKAGE)



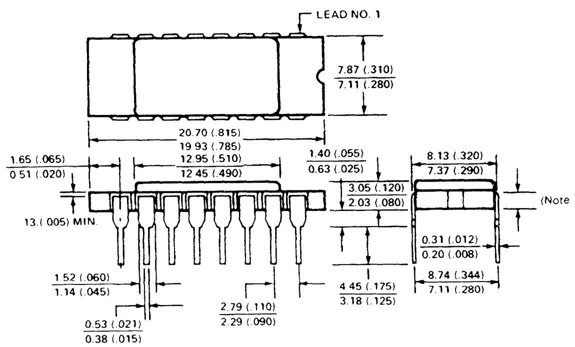
10-PIN HERMETIC TO-5/100 HEADER SHORT CAN (H PACKAGE)



10-PIN HERMETIC TO-5/100 HEADER TALL CAN (H PACKAGE)



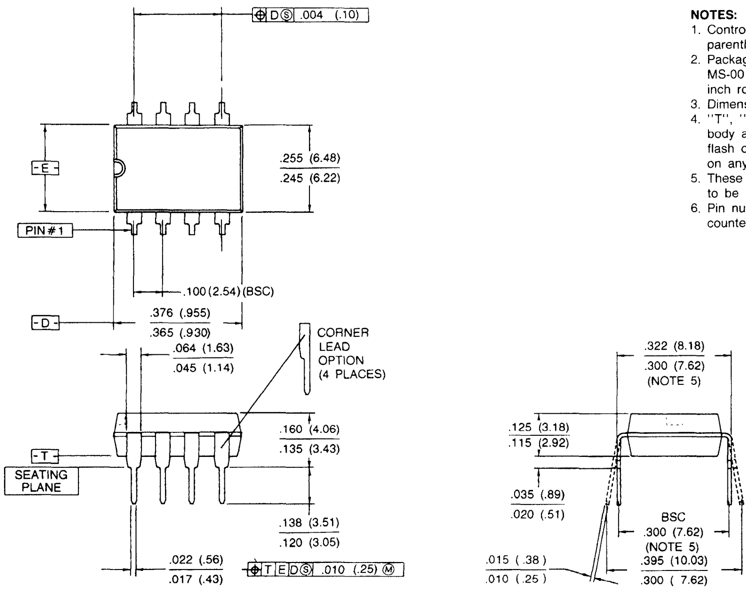
16-PIN HERMETIC SDIP (I PACKAGE)



For Prefixes ADC, AM, AU, CA, DAC, ICM, LF, LM, MC, NE, SA, SE, SG, μA, UC

Package Outlines

8-PIN PLASTIC PDIP (N PACKAGE)

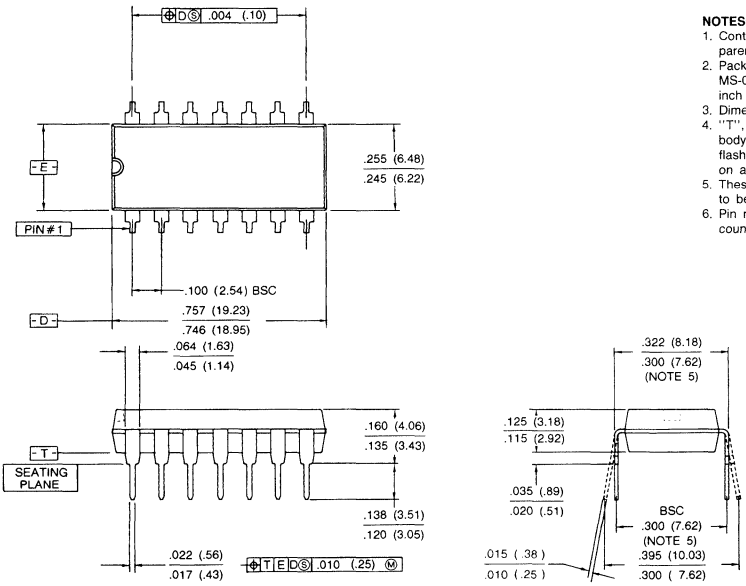


NOTES:
1. Controlling dimension: inches. Metric are shown in parentheses.
2. Package dimensions conform to JEDEC specification MS-001-AB for standard dual in-line (DIP) package .300 inch row spacing (PLASTIC) 8 leads (issue B. 7/85)
3. Dimensions and tolerancing per ANSI Y14. 5M-1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .010 inch (.25mm) on any side.
5. These dimensions measured with the leads constrained to be perpendicular to plane T.
6. Pin numbers start with pin #1 and continue counterclockwise to pin #8 when viewed from the top.

853-0404 81230

PO00312S

14-PIN PLASTIC DIP (N PACKAGE)



NOTES:
1. Controlling dimension: inches. Metric are shown in parentheses.
2. Package dimensions conform to JEDEC specification MS-001-AC for standard dual in-line (DIP) package .300 inch row spacing (PLASTIC) 14 leads (issue B. 7/85)
3. Dimensions and tolerancing per ANSI Y14. 5M-1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .010 inch (.25mm) on any side.
5. These dimensions measured with the leads constrained to be perpendicular to plane T.
6. Pin numbers start with pin #1 and continue counterclockwise to pin #14 when viewed from the top.

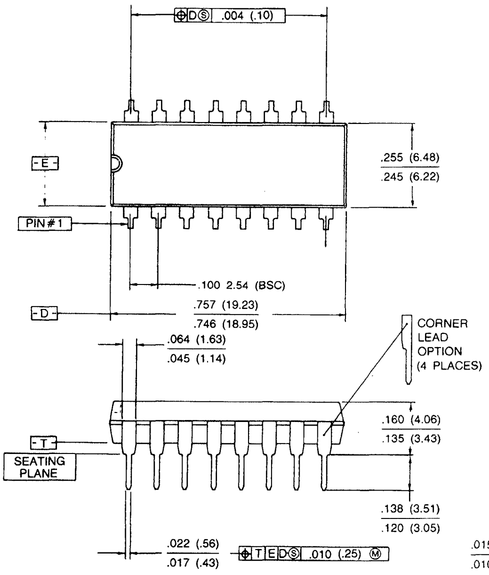
853-0405 81231

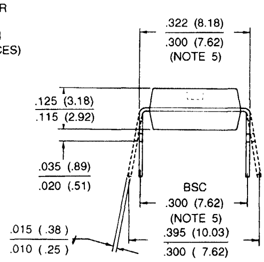
PO00330S

For Prefixes ADC, AM, AU, CA, DAC, ICM, LF, LM, MC, NE, SA, SE, SG, μA, UC

Package Outlines

16-PIN PLASTIC DIP (N PACKAGE)



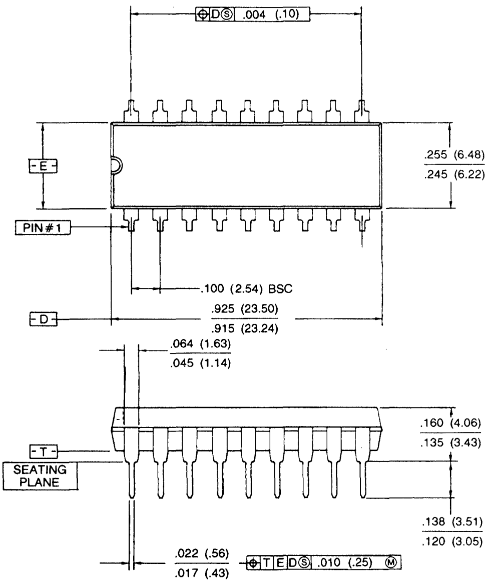


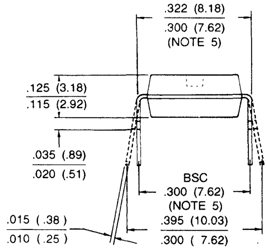
NOTES:
1. Controlling dimension: inches. Metric are shown in parentheses.
2. Package dimensions conform to JEDEC specification MS-001-AA for standard dual in-line (DIP) package .300 inch row spacing (PLASTIC) 16 leads (issue B. 7/85)
3. Dimensions and tolerancing per ANSI Y14. 5M-1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .010 inch (.25mm) on any side.
5. These dimensions measured with the leads constrained to be perpendicular to plane T.
6. Pin numbers start with pin #1 and continue counterclockwise to pin #16 when viewed from the top.

853-0406 81232

PO00280S

18-PIN PLASTIC DIP (N PACKAGE)





NOTES:
1. Controlling dimension: inches. Metric are shown in parentheses.
2. Package dimensions conform to JEDEC specification MS-001-AD for standard dual in-line (DIP) package .300 inch row spacing (PLASTIC) 18 leads (issue B. 7/85)
3. Dimensions and tolerancing per ANSI Y14. 5M-1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .010 inch (.25mm) on any side.
5. These dimensions measured with the leads constrained to be perpendicular to plane T.
6. Pin numbers start with pin #1 and continue counterclockwise to pin #18 when viewed from the top.

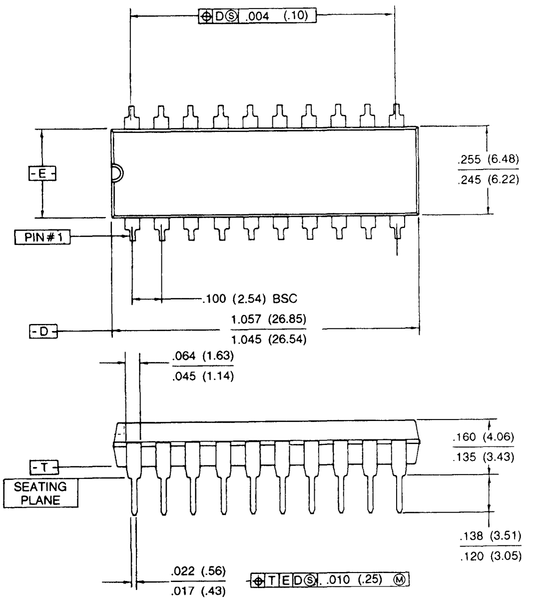
853-0407 81233

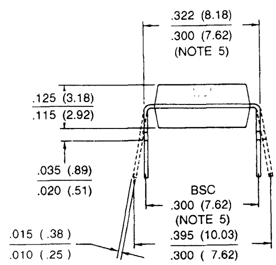
PO00300S

For Prefixes ADC, AM, AU, CA, DAC, ICM, LF, LM, MC, NE, SA, SE, SG, μA, UC

Package Outlines

20-PIN PLASTIC DIP (N PACKAGE)



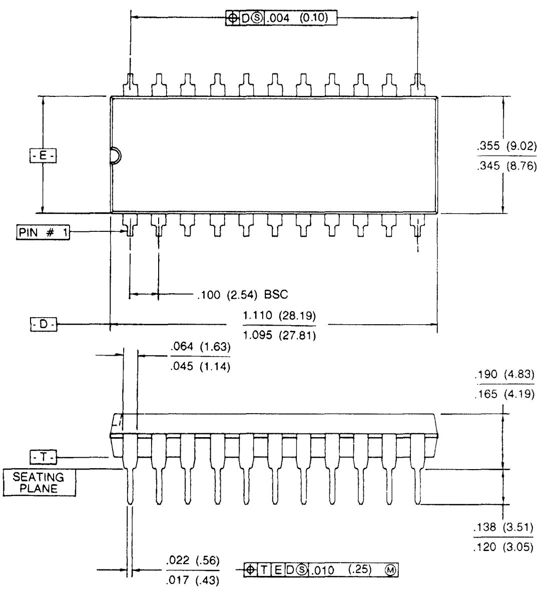


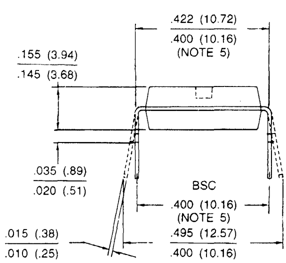
NOTES:
1. Controlling dimension: inches. Metric are shown in parentheses.
2. Package dimensions conform to JEDEC specification MS-001-AE for standard dual in-line (DIP) package .300 inch row spacing (PLASTIC) 20 leads (issue B. 7/85)
3. Dimensions and tolerancing per ANSI Y14. 5M-1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .010 inch (.25mm) on any side.
5. These dimensions measured with the leads constrained to be perpendicular to plane T.
6. Pin numbers start with pin #1 and continue counterclockwise to pin #20 when viewed from the top.

853-0408 81234

PO00290S

22-PIN PLASTIC DIP (N PACKAGE)





NOTES:
1. Controlling dimension: inches. Metric are shown in parentheses.
2. Package dimensions conform to JEDEC specification MS-010-AA for standard dual in-line (DIP) package .400 inch row spacing (PLASTIC) 22 leads (issue A. 7/85)
3. Dimensions and tolerancing per ANSI Y14. 5M-1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .010 inch (.25mm) on any side.
5. These dimensions measured with the leads constrained to be perpendicular to plane T.
6. Pin numbers start with pin #1 and continue counterclockwise to pin #22 when viewed from the top.

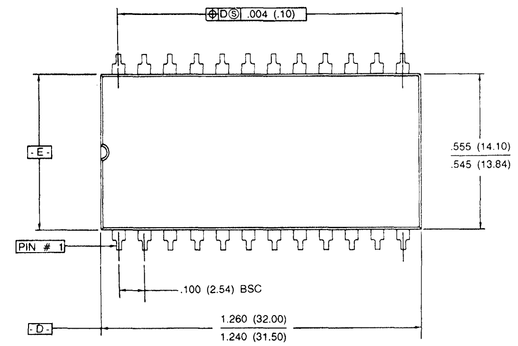
853-0409 81235

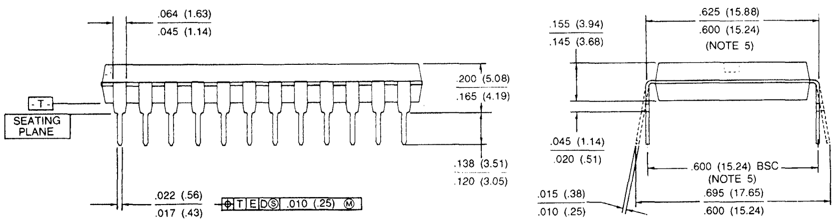
PO00341S

For Prefixes ADC, AM, AU, CA, DAC, ICM, LF, LM, MC, NE, SA, SE, SG, µA, UC

Package Outlines

24-PIN PLASTIC DIP (N PACKAGE)



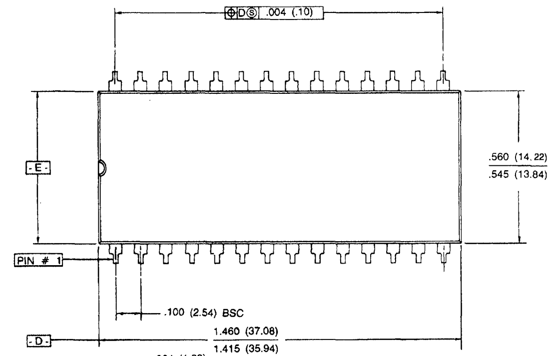


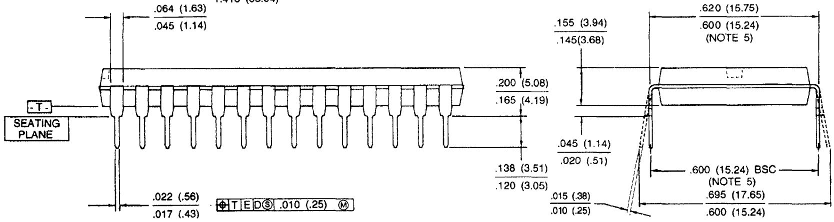
NOTES:
1. Controlling dimension: inches. Metric are shown in parentheses.
2. Package dimensions conform to JEDEC specification MS-011-AA for standard dual in-line (DIP) package .600 inch row spacing (PLASTIC) 24 leads (issue B. 7/85)
3. Dimensions and tolerancing per ANSI Y14. 5M-1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .010 inch (.25mm) on any side.
5. These dimensions measured with the leads constrained to be perpendicular to plane T.
6. Pin numbers start with pin #1 and continue counterclockwise to pin #24 when viewed from the top.

853-0412 81238

PO00361S

28-PIN PLASTIC DIP (N PACKAGE)





NOTES:
1. Controlling dimension: inches. Metric are shown in *parentheses.*
2. Package dimensions conform to JEDEC specification MS-011-AB for standard dual in-line (DIP) package .600 inch row spacing (PLASTIC) 28 leads (issue B. 7/85)
3. Dimensions and tolerancing per ANSI Y14. 5M-1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .010 inch (.25mm) on any side.
5. These dimensions measured with the leads constrained to be perpendicular to plane T.
6. Pin numbers start with pin #1 and continue counterclockwise to pin #28 when viewed from the top.

853-0413 84099

PO00372S

Package Outlines
For Prefixes HEF, OM, PCD, PCF, PNA, SAA, SAB, TDA, TDD, TEA

INTRODUCTION

Soldering

1. By hand

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

2. By dip or wave

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

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

3. Repairing soldered joints

The same precautions and limits apply as in (1) above.

SMALL OUTLINE (SO) PACKAGES

The Reflow Solder Technique

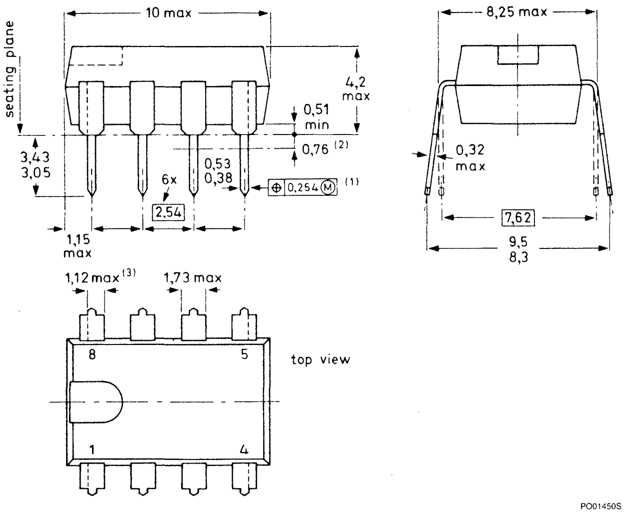
The preferred technique for mounting miniature components on hybrid thick or thin-film circuits is reflow soldering. Solder is applied to the required areas on the substrate by dipping in a solder bath or, more usually, by screen printing a solder paste. Components are put in place and the solder is reflowed by heating.

Solder pastes consist of very finely powdered solder and flux suspended in an organic liquid binder. They are available in various forms depending on the specification of the solder and the type of binder used. For hybrid circuit use, a tin-lead solder with 2 to 4% silver is recommended. The working temperature of this paste is about 220 to 230°C when a mild flux is used.

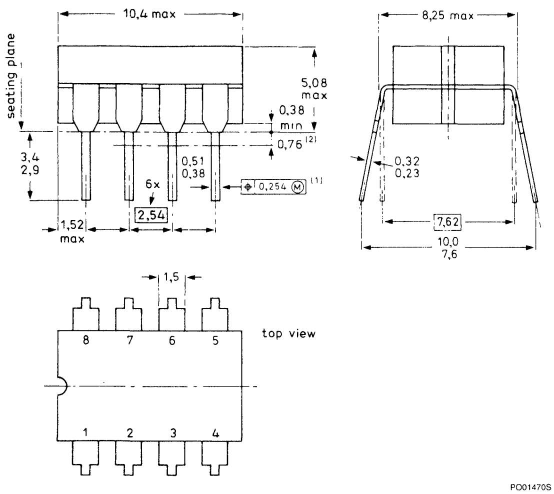
For printing the paste onto the substrate a stainless steel screen with a mesh of 80 to 105μm is used for which the emulsion thickness should be about 50μm. To ensure that sufficient solder paste is applied to the substrate, the screen aperture should be slightly larger than the corresponding contact area.

The contact pins are positioned on the substrate, the slight adhesive force of the solder paste being sufficient to keep them in place. The substrate is heated to the solder working temperature preferably by means of a controlled hot plate. The soldering process should be kept as short as possible: 10 to 15 seconds is sufficient to ensure good solder joints and evaporation of the binder fluid. After soldering, the substrate must be cleaned of any remaining flux.

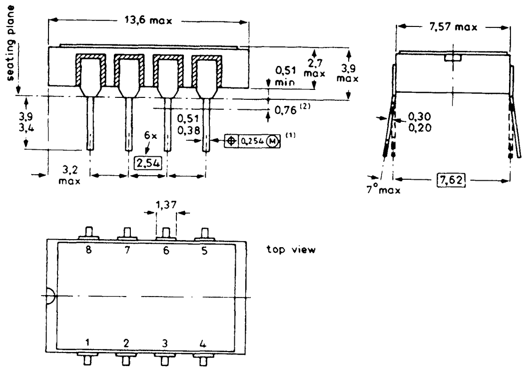
8-PIN PLASTIC (SOT-97A)



8-PIN CERDIP (SOT-151A)

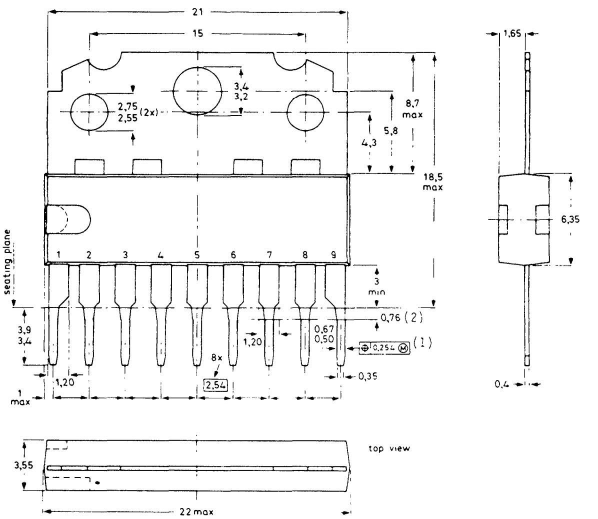


8-PIN METAL CERDIP (SOT-153B)



PO01030S

9-PIN PLASTIC SIP (SOT-110B)

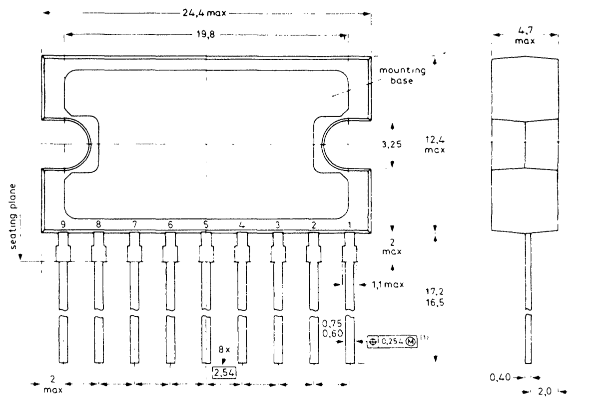


PO01040S

For Prefixes HEF, OM, PCD, PCF, PNA, SAA, SAB, TDA, TDD, TEA

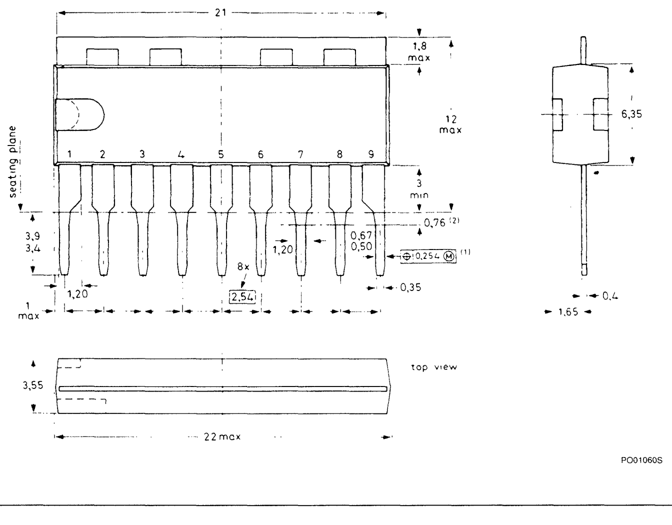
Package Outlines

9-PIN PLASTIC POWER SIP (SOT-131A, B)



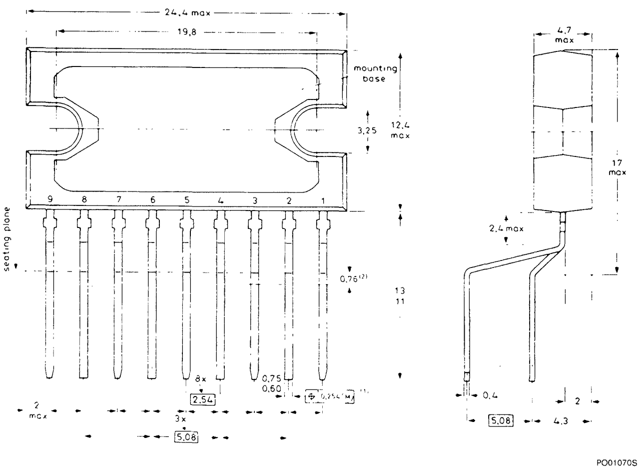
PO01050S

9-PIN PLASTIC SIP (SOT-142)

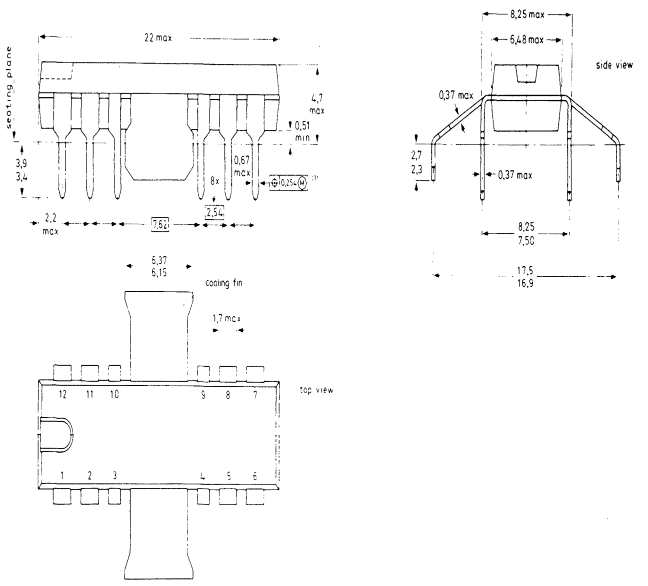


PO01060S

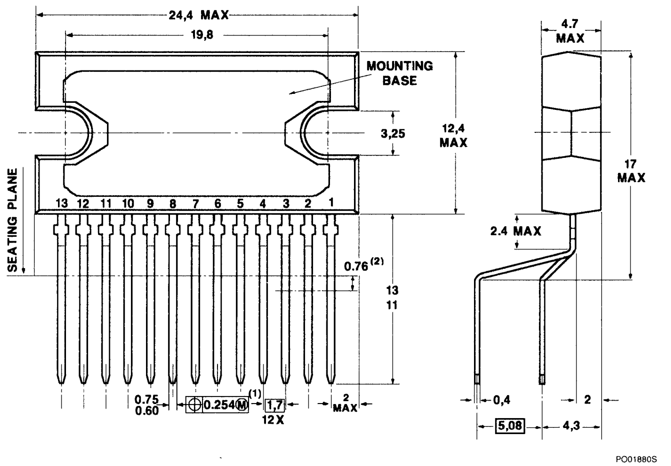
9-PIN PLASTIC POWER SIP-BENT-TO-DIP (SOT-157B)



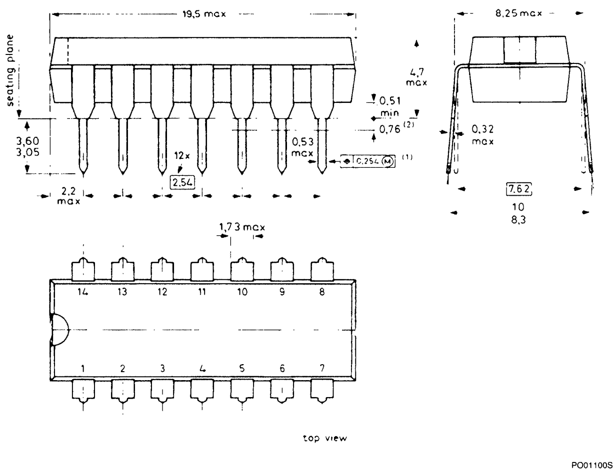
12-PIN PLASTIC DIP WITH METAL COOLING FIN (SOT-150)



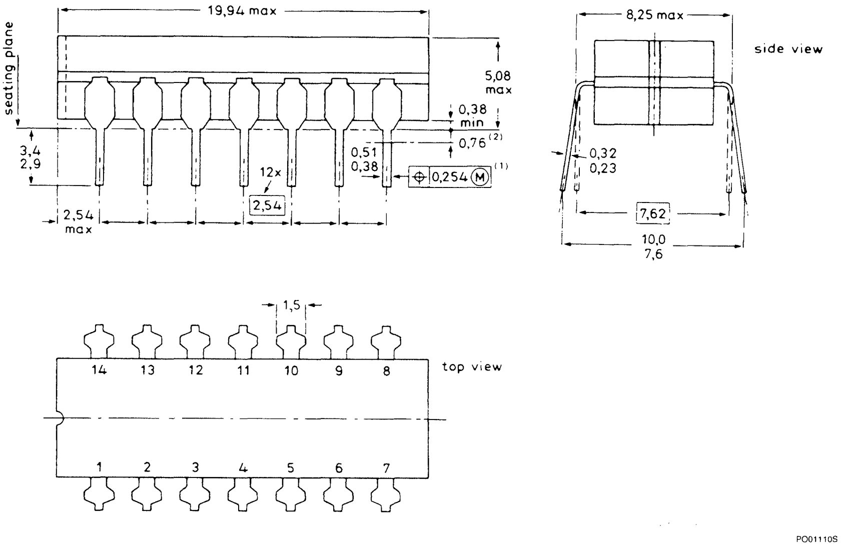
13-PIN PLASTIC POWER SIP-BENT-TO-DIP (SOT-141BA)



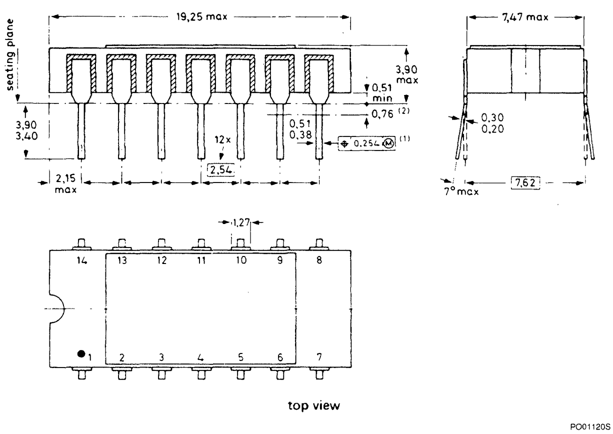
14-PIN PLASTIC DIP (SOT-27K, M, T)



14-PIN CERDIP (SOT-73A, B, C)



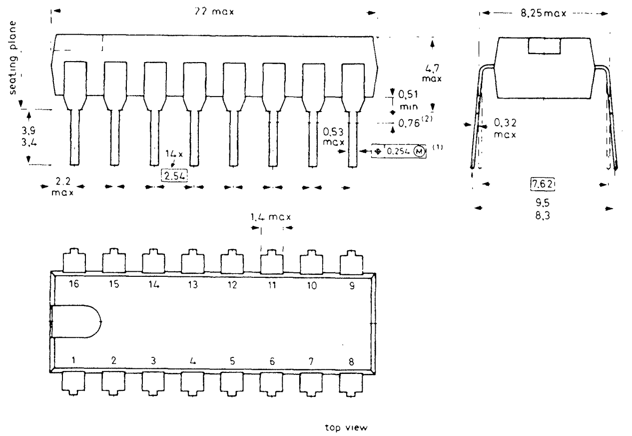
14-PIN METAL CERDIP (SOT-83B)



For Prefixes HEF, OM, PCD, PCF, PNA, SAA, SAB, TDA, TDD, TEA

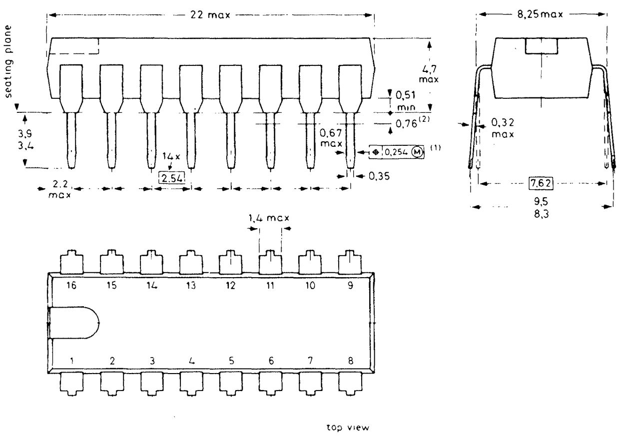
Package Outlines

16-PIN PLASTIC DIP (SOT-38)



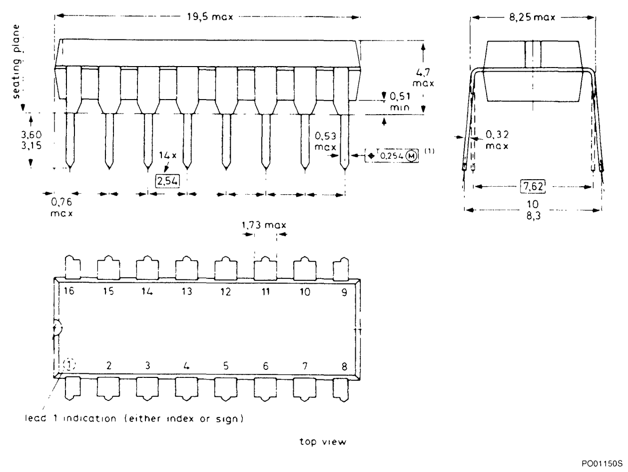
PO01130S

16-PIN PLASTIC DIP (SOT-38A)

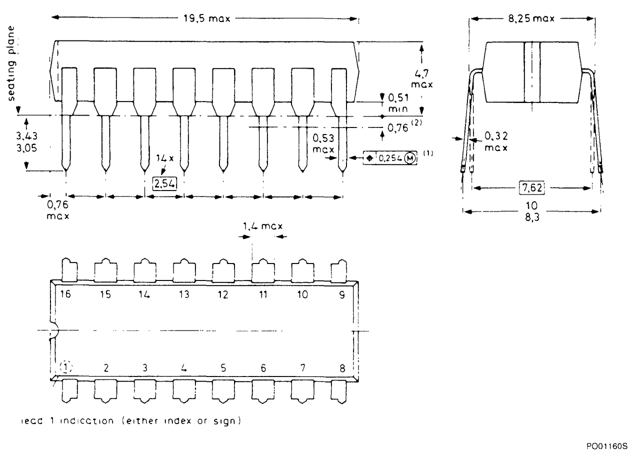


PO01140S

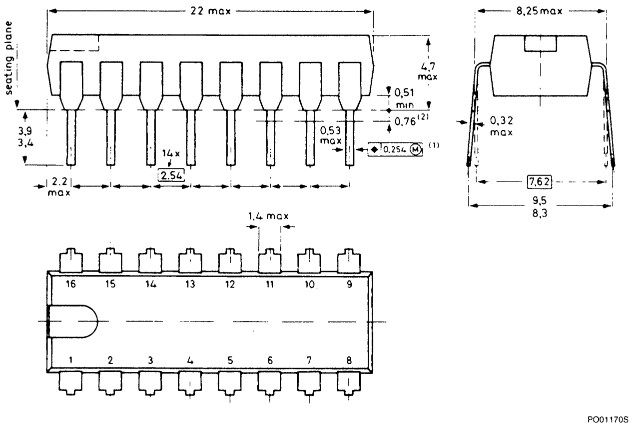
16-PIN PLASTIC DIP (SOT-38D, DE)



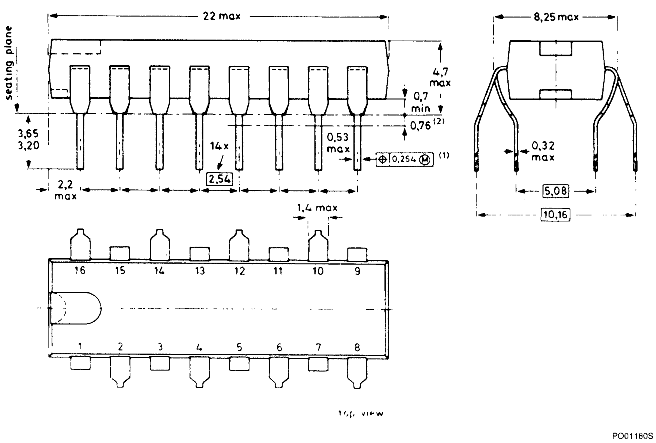
16-PIN PLASTIC DIP (SOT-38Z)



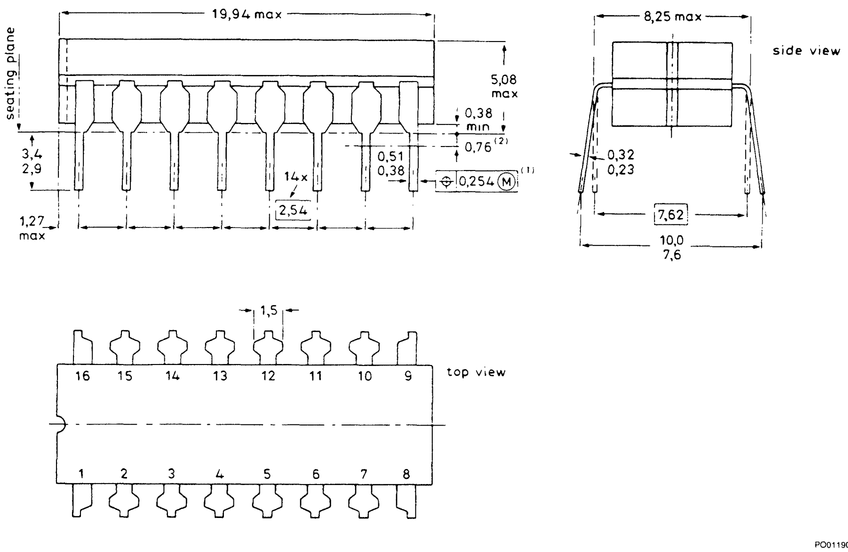
16-PIN PLASTIC DIP WITH INTERNAL HEAT SPREADER (SOT-38WE-2)



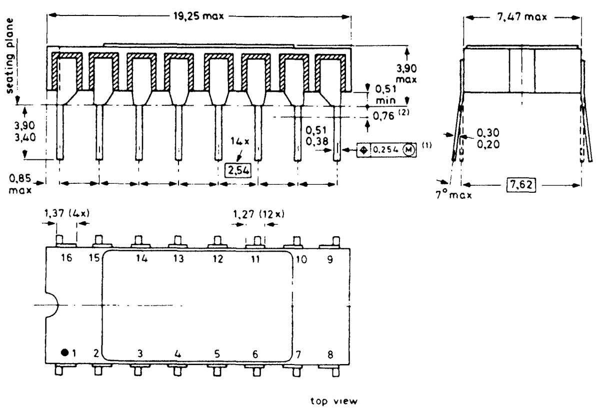
16-PIN PLASTIC QIP (SOT-58)



16-PIN CERDIP (SOT-74A, B, C)

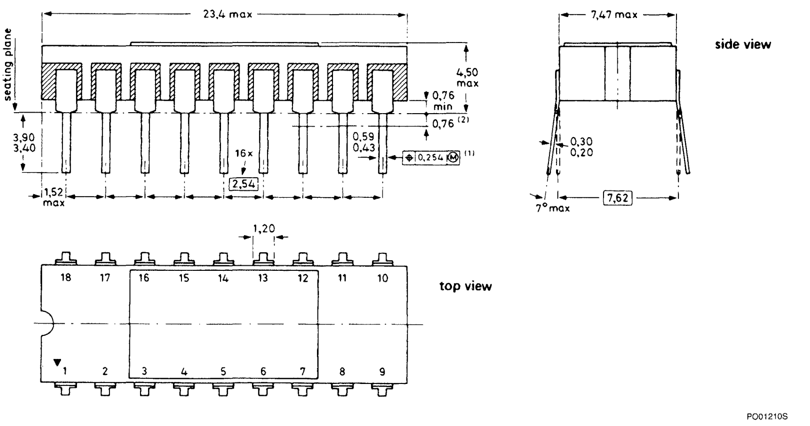


16-PIN METAL CERDIP (SOT-84B)

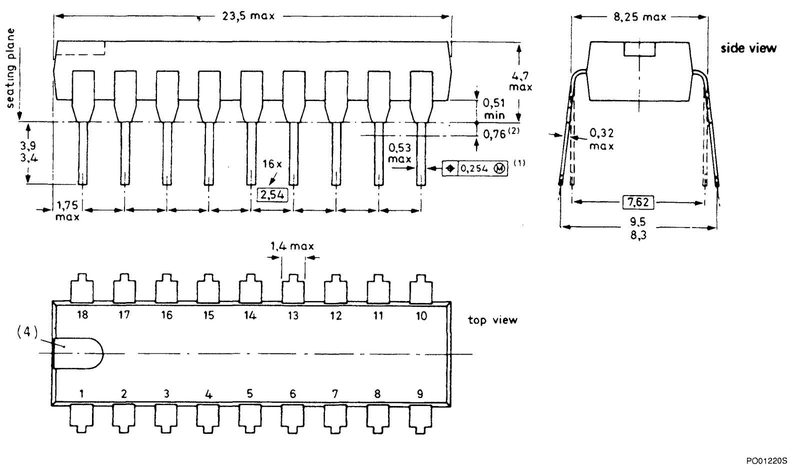


14

18-PIN METAL CERDIP (SOT-85B)



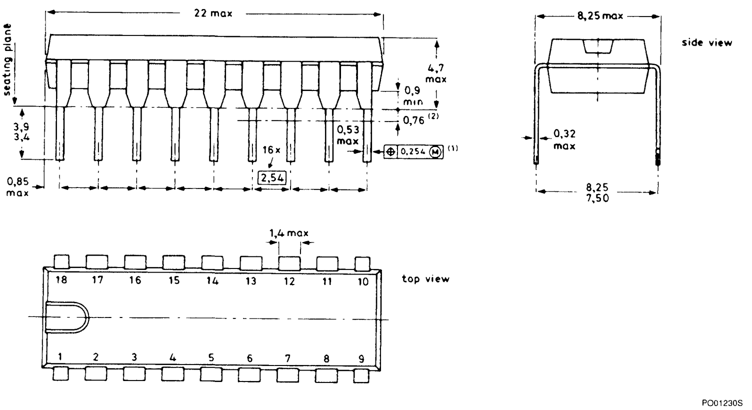
18-PIN PLASTIC DIP (SOT-102A)



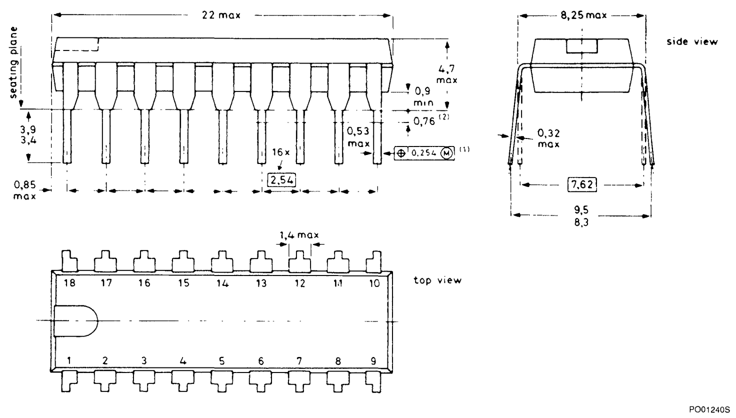
For Prefixes HEF, OM, PCD, PCF, PNA, SAA, SAB, TDA, TDD, TEA

Package Outlines

18-PIN PLASTIC DIP (SOT-102C)



18-PIN PLASTIC DIP (SOT-102CS)

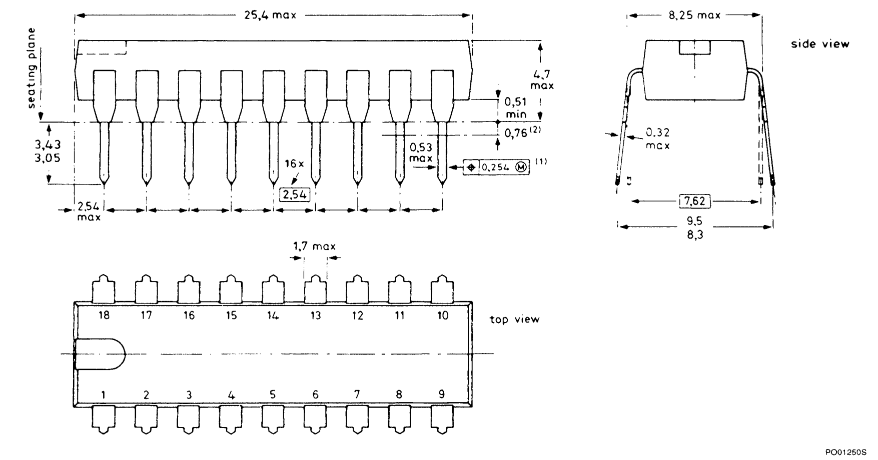


14

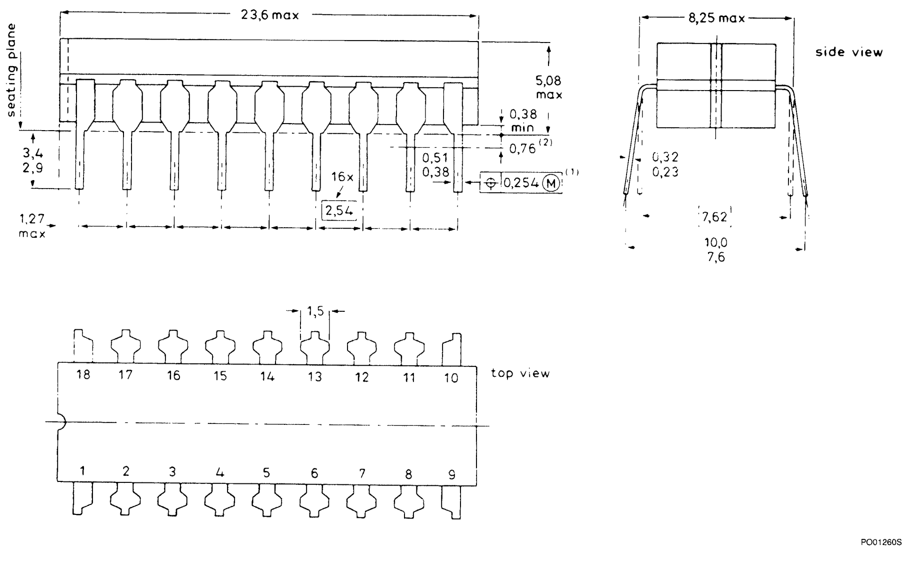
For Prefixes HEF, OM, PCD, PCF, PNA, SAA, SAB, TDA, TDD, TEA

Package Outlines

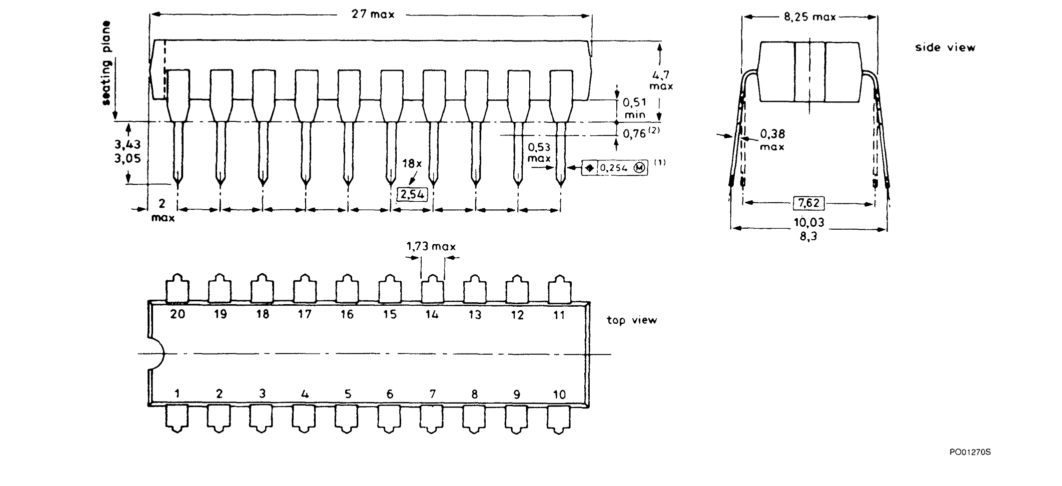
18-PIN PLASTIC DIP (SOT-102G)



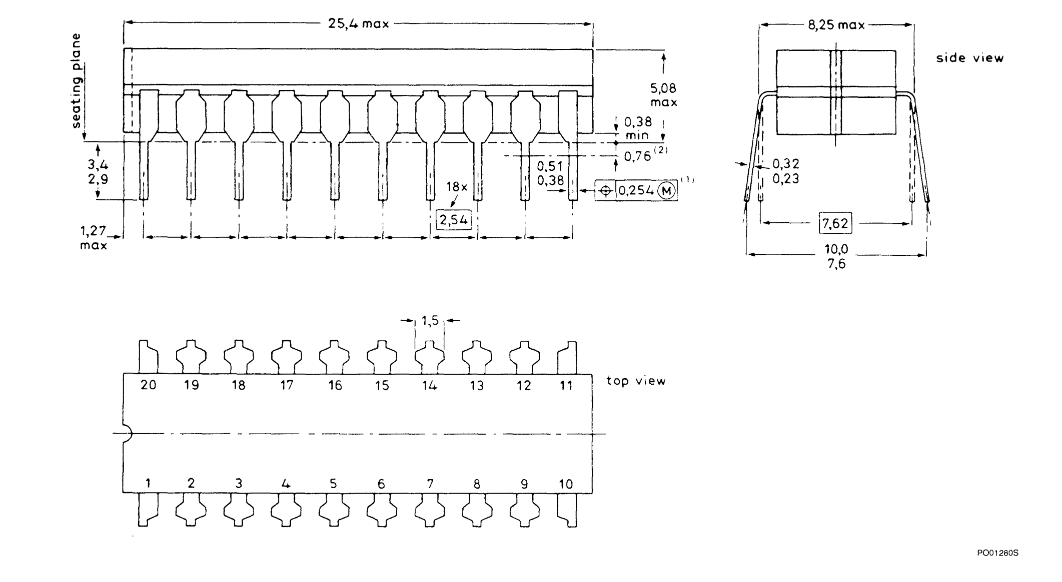
18-PIN CERDIP (SOT-133A, B)



20-PIN PLASTIC DIP (SOT-146)



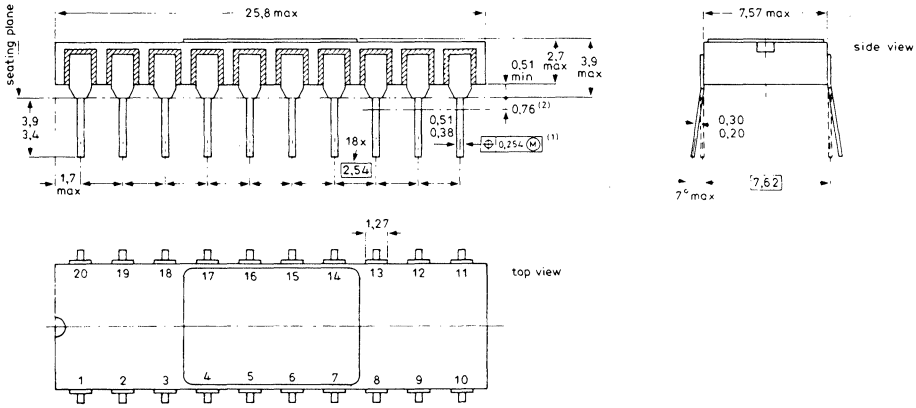
20-PIN CERDIP (SOT-152B, C)



For Prefixes HEF, OM, PCD, PCF, PNA, SAA, SAB, TDA, TDD, TEA

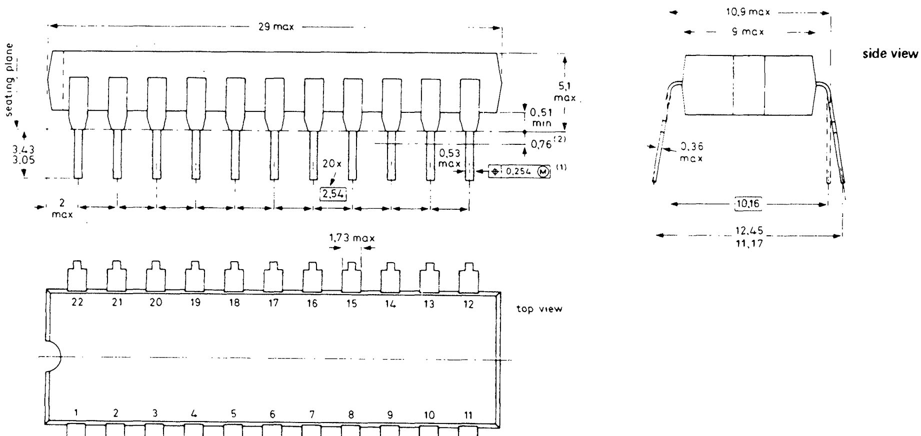
Package Outlines

20-PIN METAL CERDIP (SOT-154B)



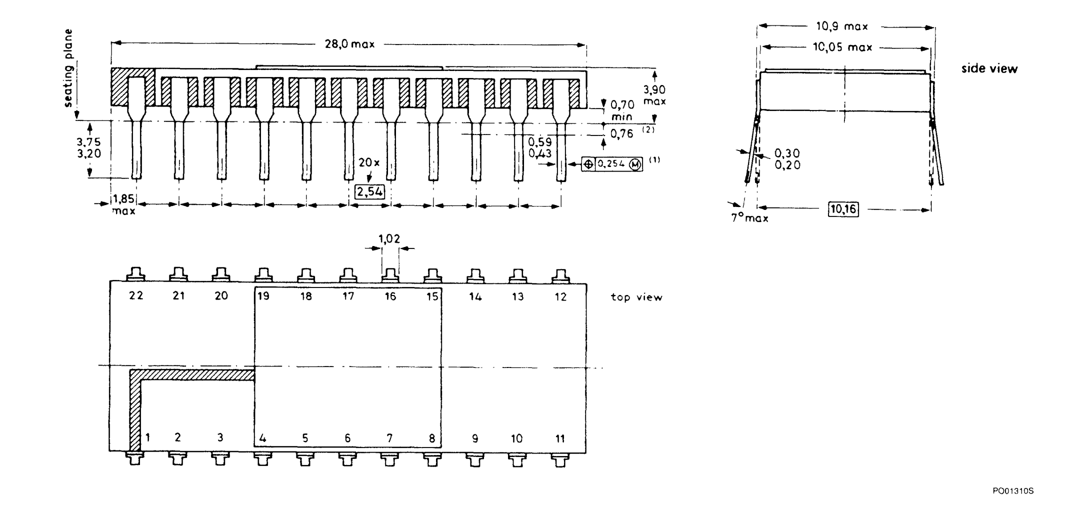
PO01290S

20-PIN PLASTIC DIP (SOT-116)

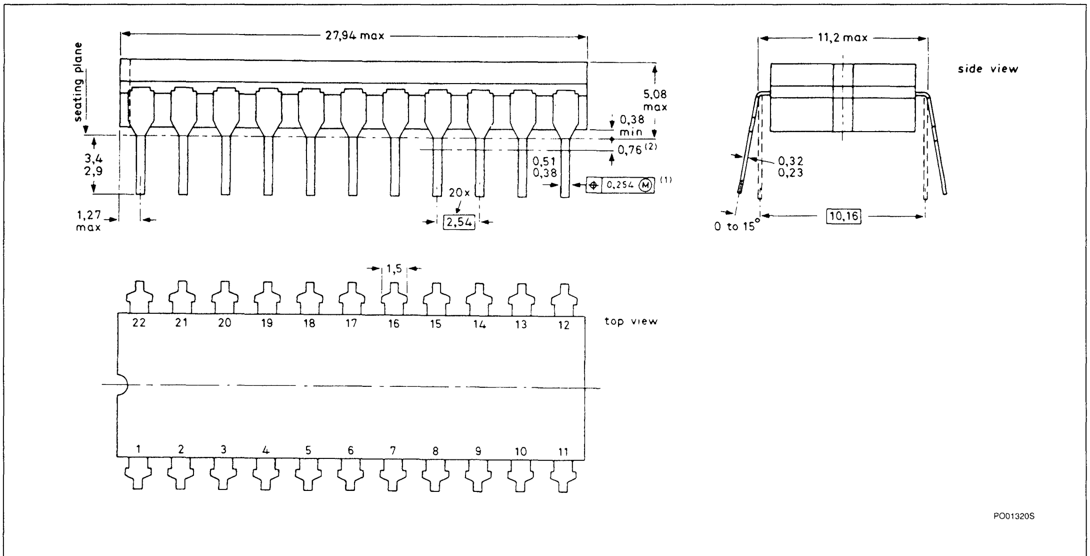


PO01300S

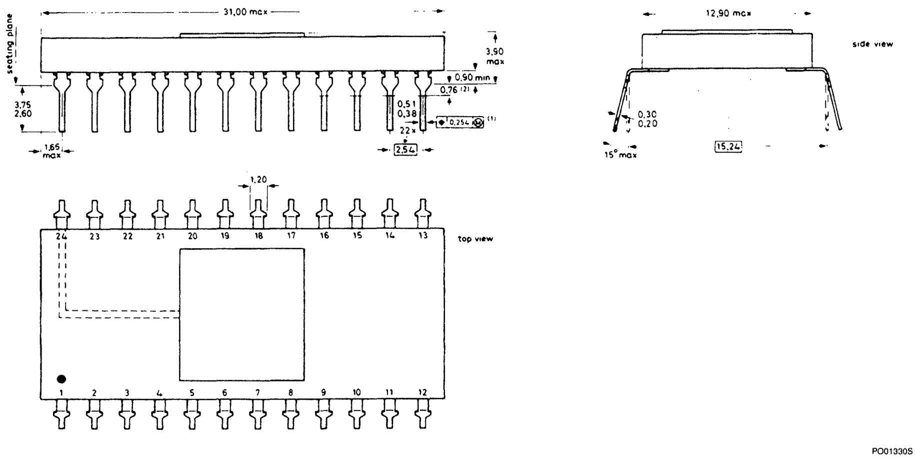
22-PIN METAL CERDIP (SOT-118B)



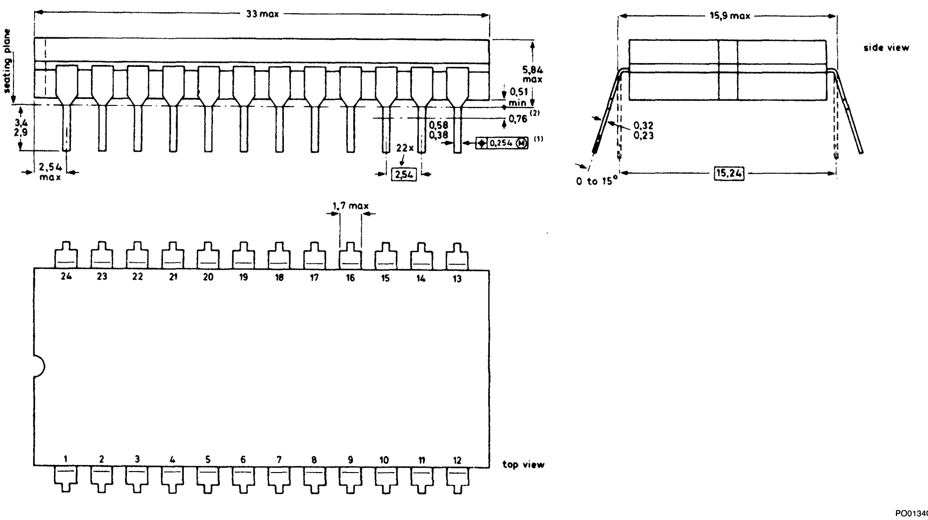
22-PIN CERDIP (SOT-134A)



24-PIN METAL CERDIP (SOT-86A)



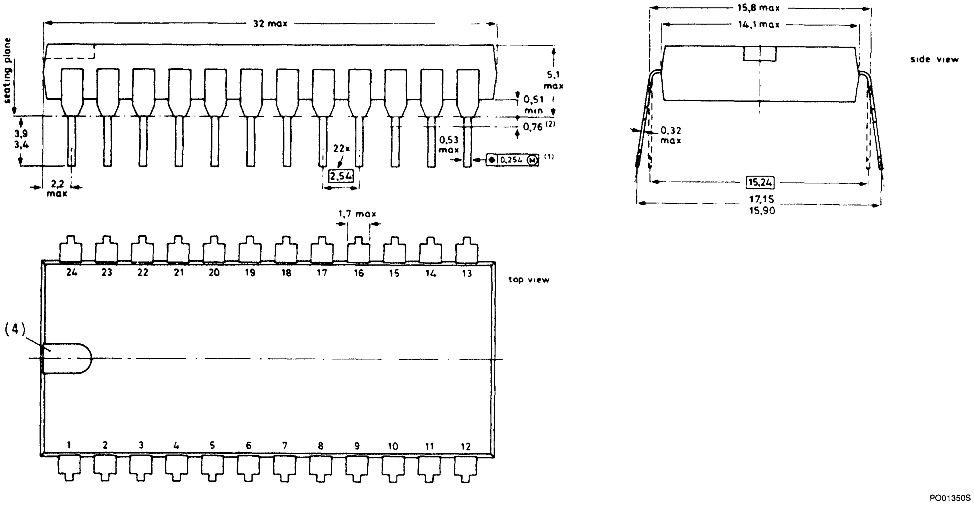
24-PIN CERDIP (SOT-94)



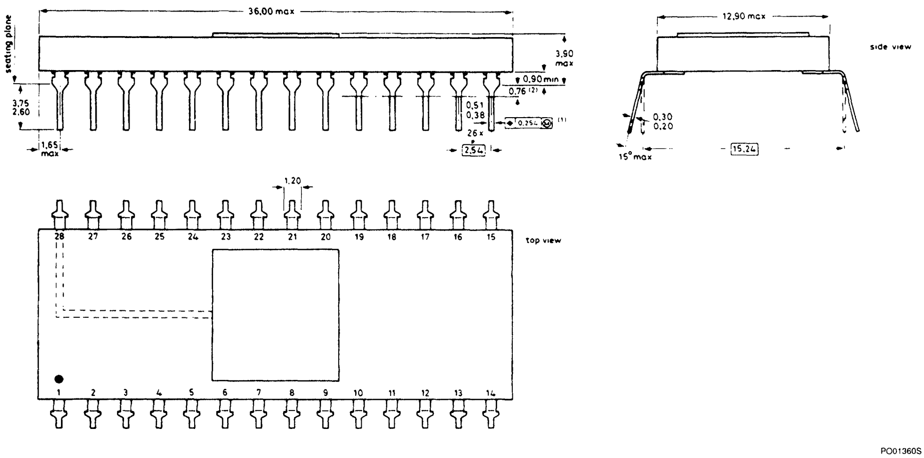
For Prefixes HEF, OM, PCD, PCF, PNA, SAA, SAB, TDA, TDD, TEA

Package Outlines

24-PIN PLASTIC DIP WITH INTERNAL HEAT SPREADER (SOT-101A, B)



28-PIN METAL CERDIP (SOT-87A)



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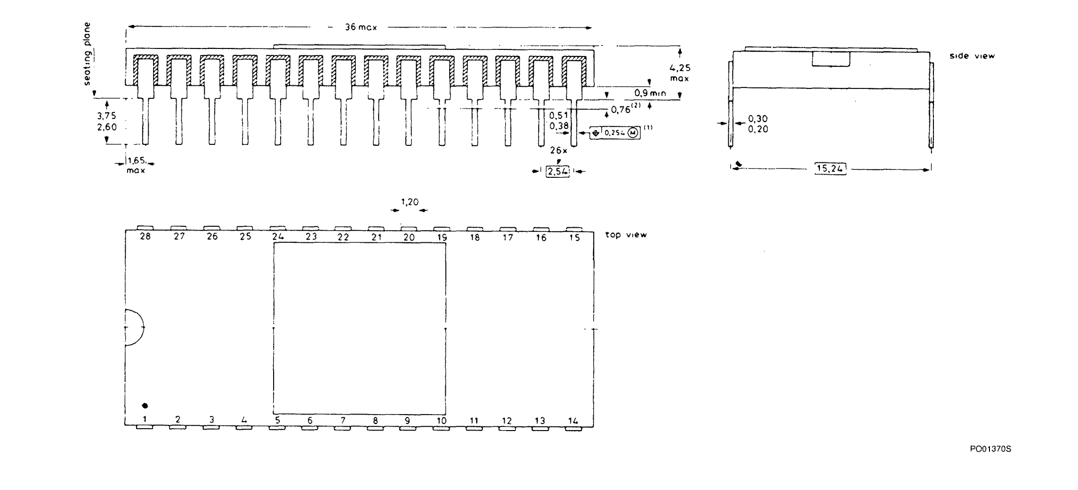
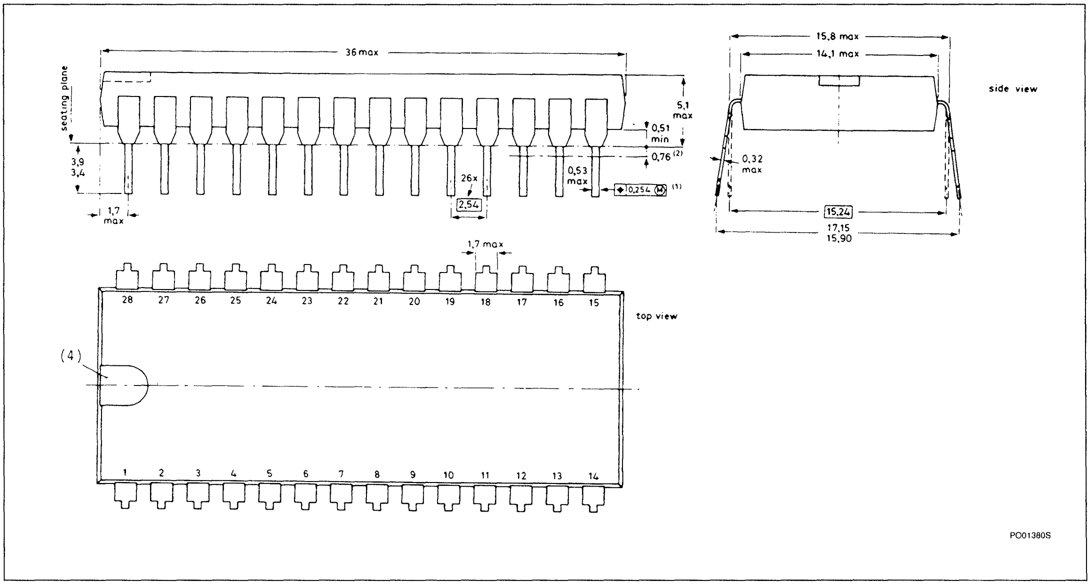
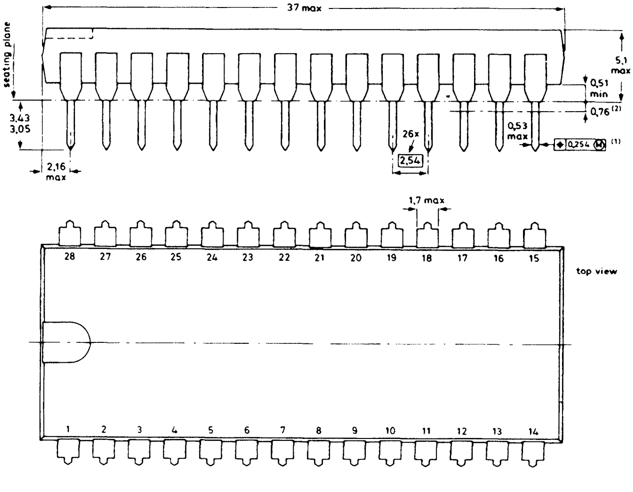
28-PIN METAL CERDIP (SOT-87B)

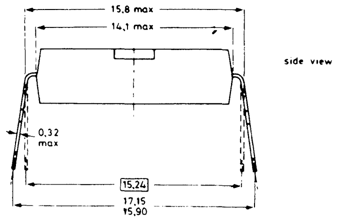


28-PIN PLASTIC DIP (SOT-117)

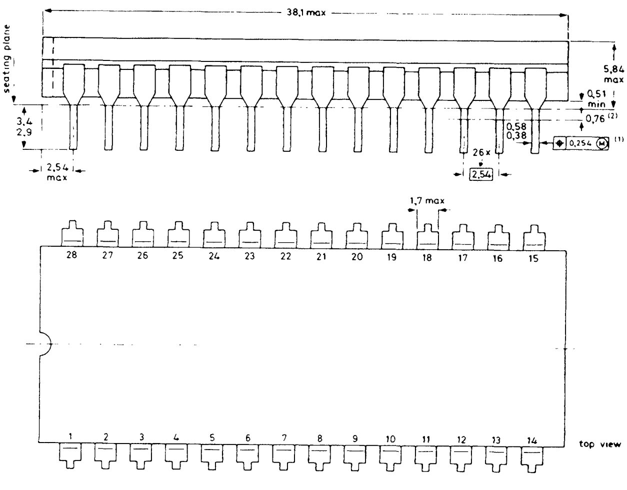


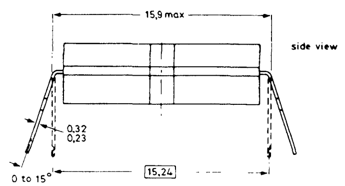
28-PIN PLASTIC DIP (SOT-117D)



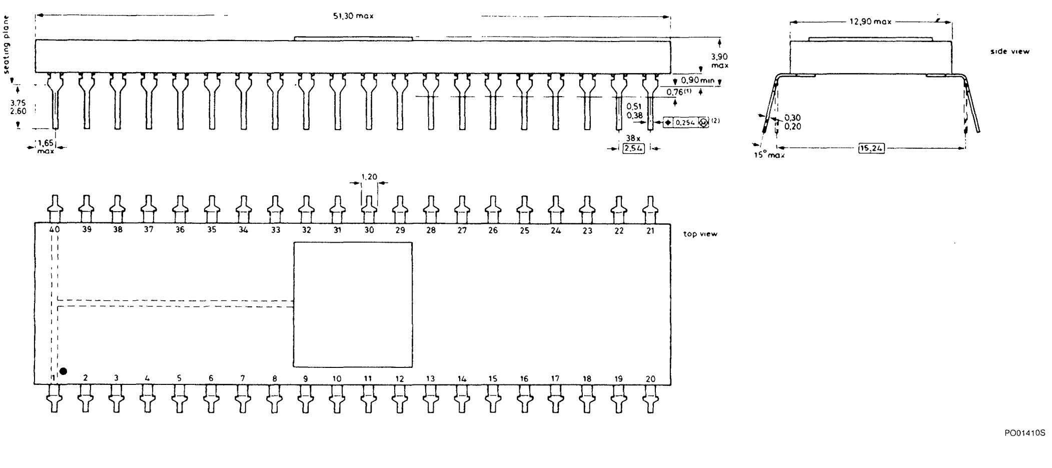


28-PIN CERDIP (SOT-135A)

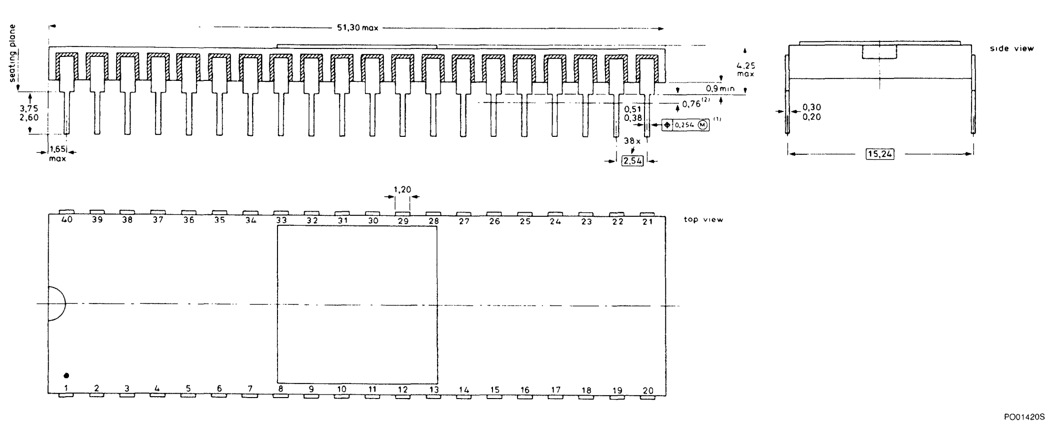




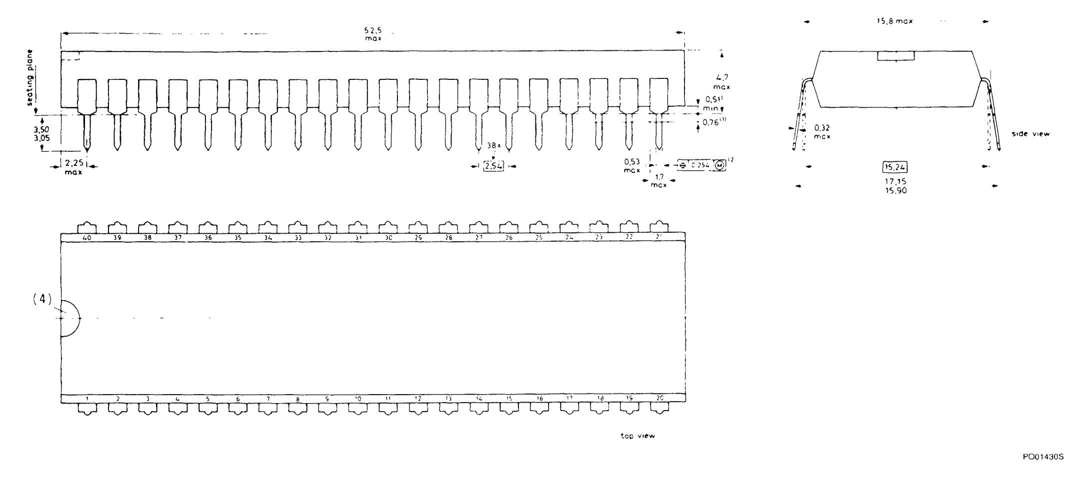
40-PIN METAL CERDIP (SOT-88)



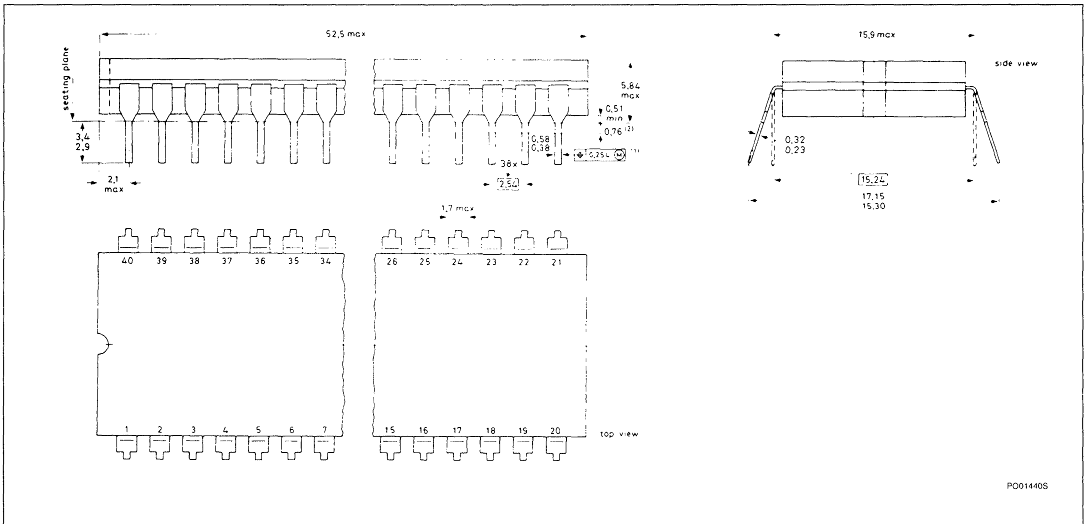
40-PIN METAL CERDIP (SOT-88B)



40-PIN PLASTIC DIP (SOT-129)



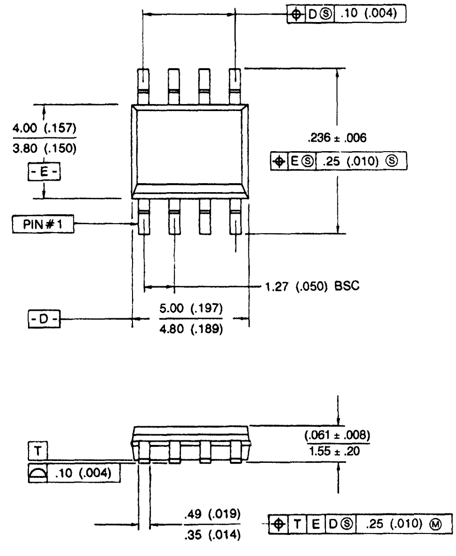
40-PIN CERDIP (SOT-145)

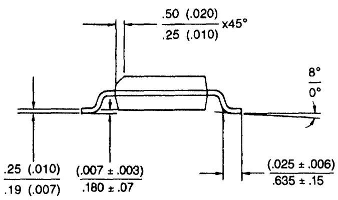


For Prefixes HEF, OM, PCD, PCF, PNA, SAA, SAB, TDA, TDD, TEA

Package Outlines

8-PIN PLASTIC SO (D PACKAGE) (SO-8, SOT-96A)



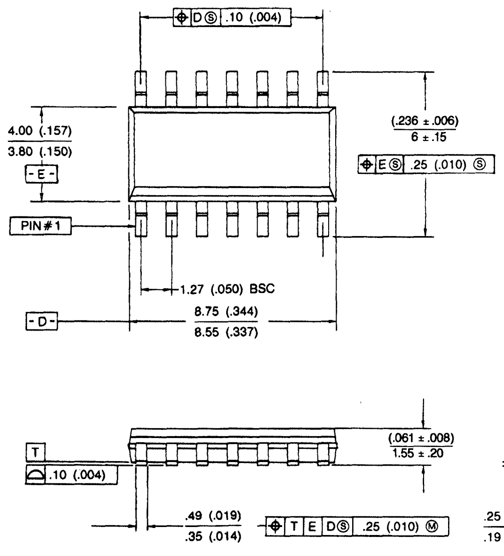


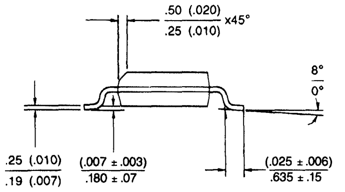
NOTES:
1. Package dimensions conform to JEDEC specification MS-012-AA for standard small outline (SO) package, 8 leads, 3.75mm (.150") body width (issue A, June 1985).
2. Controlling dimensions are in mm. Inch dimensions in parentheses.
3. Dimensions and tolerancing per ANSI Y14.5M-1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .15mm (.006") on any side.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #8 when viewed from top.
6. Signetics ordering code for a product packaged in a plastic small outline (SO) package is the suffix D after the product number.

853-0174 88070

PO00273S

14-PIN PLASTIC SO (D PACKAGE) (SO-14, SOT-108A)





NOTES:
1. Package dimensions conform to JEDEC specification MS-012-AB for standard small outline (SO) package, 14 leads, 3.75mm (.150") body width (issue A, June 1985).
2. Controlling dimensions are in mm. Inch dimensions in parentheses.
3. Dimensions and tolerancing per ANSI Y14.5M-1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .15mm (.006") on any side.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #14 when viewed from top.
6. Signetics ordering code for a product packaged in a plastic small outline (SO) package is the suffix D after the product number.

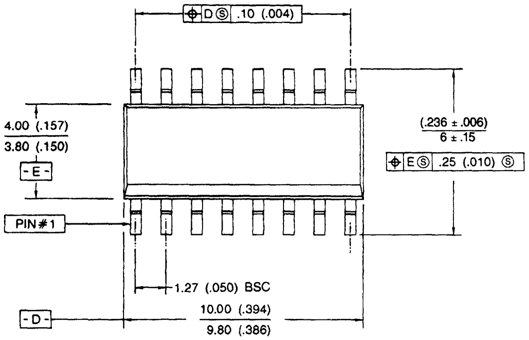
853-0175 88068

PO00263S

For Prefixes HEF, OM, PCD, PCF, PNA, SAA, SAB, TDA, TDD, TEA

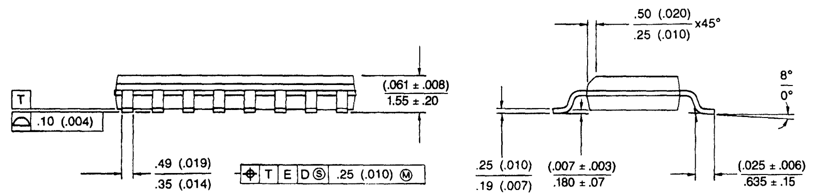
Package Outlines

16-PIN PLASTIC SO (D PACKAGE) (SO-16, SOT-109A)



NOTES:

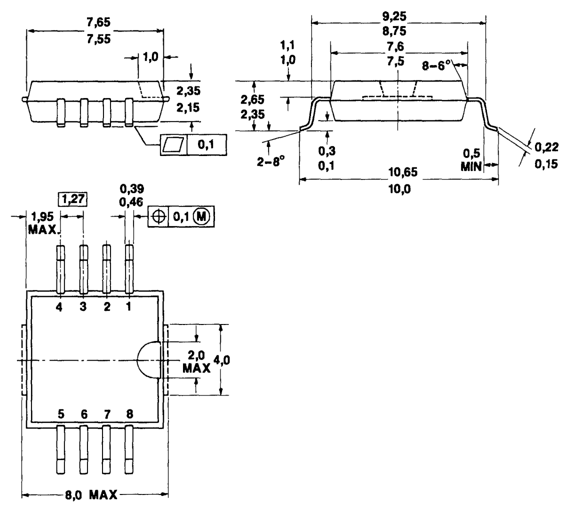
1. Package dimensions conform to JEDEC specification MS-012-AC for standard small outline (SO) package, 16 leads, 3.75mm (.150") body width (issue A, June 1985).
2. Controlling dimensions are in mm. Inch dimensions in parentheses.
3. Dimensions and tolerancing per ANSI Y14.5M-1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .15mm (.006") on any side.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #16 when viewed from top.
6. Signetics ordering code for a product packaged in a plastic small outline (SO) package is the suffix D after the product number.



853-0005 88069

PO00252S

8-PIN PLASTIC SOL (D Package) (SOL-8, SOT-176)



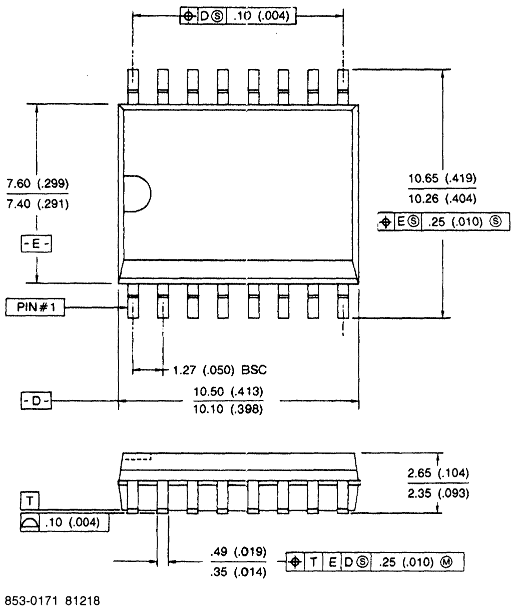
PO01890S

14

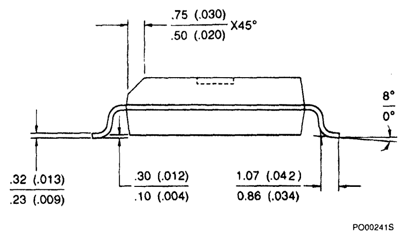
For Prefixes HEF, OM, PCD, PCF, PNA, SAA, SAB, TDA, TDD, TEA

Package Outlines

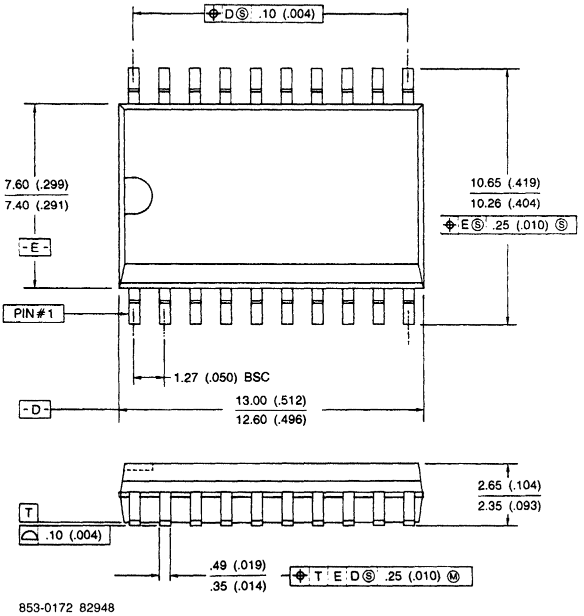
16-PIN PLASTIC SOL (D PACKAGE) (SOL-16, SOT-162A)



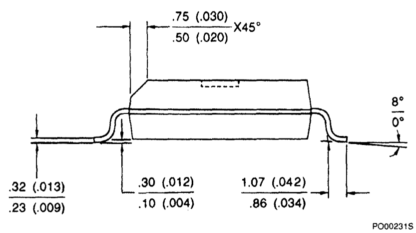
NOTES:
1. Package dimensions conform to JEDEC specification MS-013-AA for standard small outline (SO) package, 16 leads, 7.50mm (.300") body width (issue A, June 1985).
2. Controlling dimensions are in mm. Inch dimensions in parentheses.
3. Dimensions and tolerancing per ANSI Y14.5M-1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .15mm (.006") on any side.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #16 when viewed from top.
6. Signetics ordering code for a product packaged in a plastic small outline (SO) package is the suffix D after the product number.



20-PIN PLASTIC SOL (D PACKAGE) (SOL-20, SOT-163A)



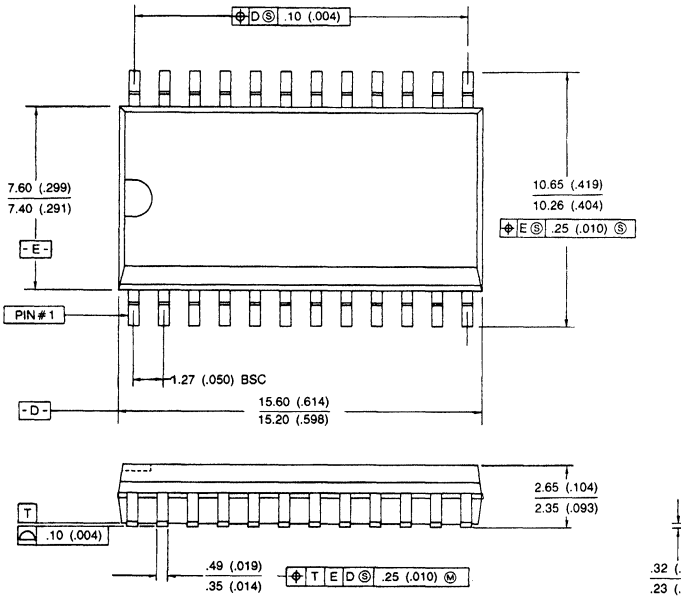
NOTES:
1. Package dimensions conform to JEDEC specification MS-013-AC for standard small outline (SO) package, 20 leads, 7.50mm (.300") body width (issue A, June 1985).
2. Controlling dimensions are in mm. Inch dimensions in parentheses.
3. Dimensions and tolerancing per ANSI Y14.5M-1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .15mm (.006") on any side.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #20 when viewed from top.
6. Signetics ordering code for a product packaged in a plastic small outline (SO) package is the suffix D after the product number.

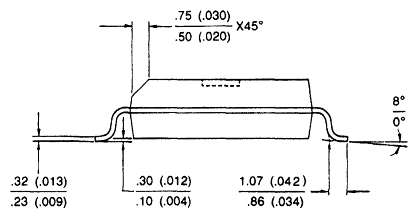


For Prefixes HEF, OM, PCD, PCF, PNA, SAA, SAB, TDA, TDD, TEA

Package Outlines

24-PIN PLASTIC SOL (D PACKAGE) (SOL-24, SOT-137A)



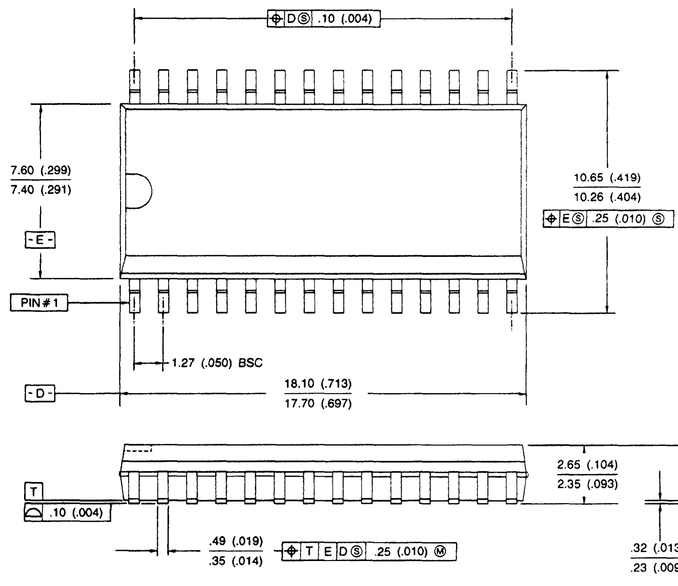


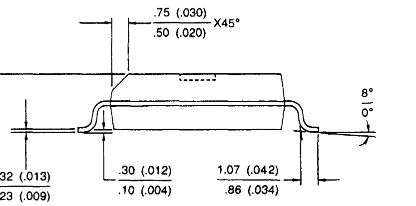
NOTES:
1. Package dimensions conform to JEDEC specification MS-013-AD for standard small outline (SO) package, 24 leads, 7.50mm (.300") body width (issue A, June 1985).
2. Controlling dimensions are in mm. Inch dimensions in parentheses.
3. Dimensions and tolerancing per ANSI Y14.5M-1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .15mm (.006") on any side.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #24 when viewed from top.
6. Signetics ordering code for a product packaged in a plastic small outline (SO) package is the suffix D after the product number.

853-0173 82949

PO00221S

28-PIN PLASTIC SOL (D PACKAGE) (SOL-28, SOT-136A)



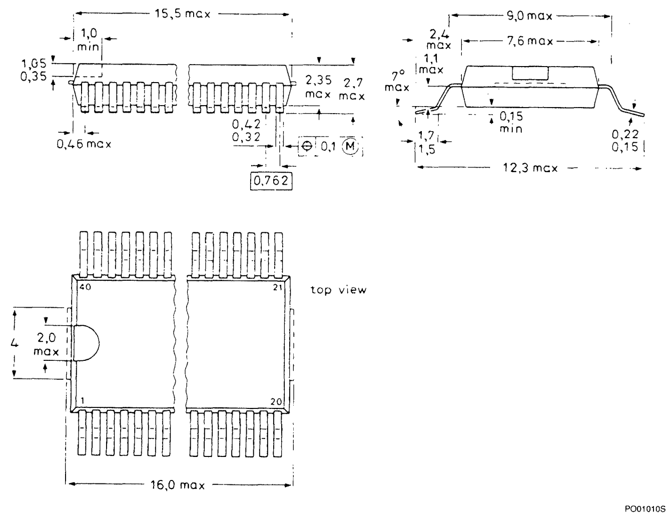


NOTES:
1. Package dimensions conform to JEDEC specification MS-013-AE for standard small outline (SO) package, 28 leads, 7.50mm (.300") body width (issue A, June 1985).
2. Controlling dimensions are in mm. Inch dimensions in parentheses.
3. Dimensions and tolerancing per ANSI Y14.5M-1982.
4. "T", "D" and "E" are reference datums on the molded body and do not include mold flash or protrusions. Mold flash or protrusions shall not exceed .15mm (.006") on any side.
5. Pin numbers start with pin #1 and continue counterclockwise to pin #28 when viewed from top.
6. Signetics ordering code for a product packaged in a plastic small outline (SO) package is the suffix D after the product number.

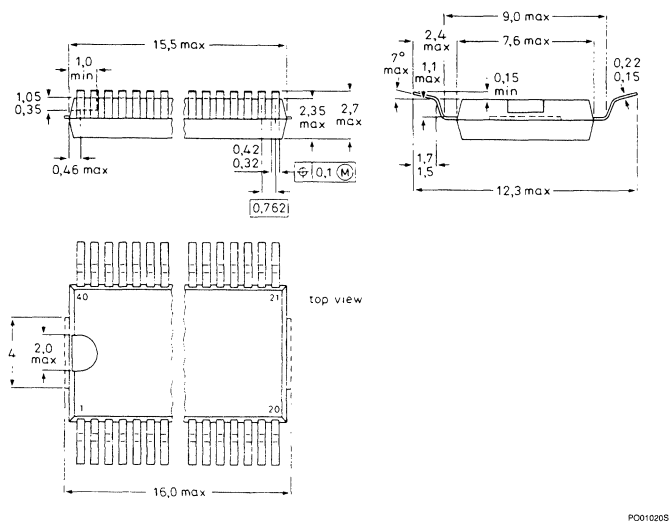
853-0006 81217

PO00212S

40-PIN PLASTIC SO (VSO-40, SOT-158A)



40-PIN PLASTIC SO (OPPOSITE BENT LEADS) (VSO-40, SOT-158B)



Signetics

Section 15
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Linear Products

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Signetics Company
811 E. Arques Avenue
P.O. Box 3409
Sunnyvale, California 94088-3409
Telephone 408/991-2000

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