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A MICROCIRCUIT TONE RECEIVER FOR USE
IN TELEPHONE SYSTEMS

by

Michael C.J. Cowpland, M. Eng.

A Thesis submitted to the Faculty of
Graduate Studies in partial fulfillment
of the requirements for the degree of
Doctor of Philosophy

Carleton University
Ottawa, Canada

September, 1973

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ABSTRACT

This thesis is concerned with a new linear tone receiver for telephone systems which makes optimum use of thick film and tantalum thin film microcircuits to achieve excellent performance with small cost and size. A new system configuration is used which eliminates the need for the group separation filtering used in previous designs by the use of a tightly controlled automatic gain control amplifier and cross-coupled group threshold control circuitry. The only remaining critical filtering is in the eight bandpass filters required for frequency detection. These are realised in a single laser trimmed tantalum thin film hybrid module which includes an eight amplifier silicon integrated circuit chip. This exceptional packing density is achieved by virtue of simplification of the filter section to the minimum, without sacrifice in performance.

In the process of the filter design a simplified optimisation procedure was evolved for minimising the 'gain sensitivity product,' which is a measure of filter stability. To reduce manufacturing cost and time, an accurate and rapid filter centre frequency measuring technique was developed which enables the frequency to be rapidly monitored using a period counter instead of slower sweep generator or phasemeter methods. This greatly enhances the speed and accuracy of the filter trimming. The tightly controlled automatic gain control amplifier uses a new approach of degeneratively adding control voltage feedthrough to enable fast response to be achieved without instability. A thick/thin film hybrid, which includes a silicon integrated circuit, is used to realise a quad alternating current level detector for use at the outputs of the bandpass filters. This unit has precisely controlled triggering and hysteresis voltages with

accuracy to better than 1%, enhancing the overall accuracy and stability of the complete receiver. Conventional discrete components and 741 type operational amplifiers are used for an untrimmed highpass pre-filter and the threshold control circuitry. The receiver occupies 25 square inches of single sided printed circuit board and has a low profile suitable for mounting on 1/2" centres. Dynamic range is adjustable from 20 to 46 dB. A single unregulated power supply is required. Power consumption is 800 mW.

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TABLE OF CONTENTS

	Page
Chapter I Introduction	1
Chapter II Subscriber Tone Dialing: Systems Considerations	5
2.0 Introduction	5
2.1 Frequency Allocations	6
2.2 Signal Power Levels	6
2.2.1 Dialing to Central Offices	6
2.2.2 Dialing to PBX's	10
2.2.3 End to End Signalling Levels	11
2.3 Frequency Tolerance Considerations	13
2.3.1 Tone Generator Tolerance	13
2.3.2 Carrier Induced Frequency Offset	13
2.4 Signal to Noise Ratio	13
2.5 Dial Tone	14
2.6 Common Mode Noise	16
2.7 Echoes	16
Chapter III Tone Receiver Detection System	17
3.0 Introduction	17
3.1 BTL, CO, PBX, and Key Telephone Tone Receivers	17
3.2 Bell System 403 Data Set	19
3.3 New Detection System Using Pivoting Threshold Control	20
3.3.1 Receiver Guard Action	21
3.3.2 Amplitude Ratio Control	24
3.3.3 Echo Protection	25
3.3.4 Output Format	25
Chapter IV AGC Amplifier Circuit	26
4.0 Introduction	26
4.1 Background	26
4.2 Description of Circuit Operation	28
4.3 Theoretical ML1314K Circuit Characteristics	32
4.3.1 Maximum Input Level	32
4.3.2 Input Impedance	33
4.3.3 Maximum Gain of Voltage Controlled Amplifier	33
4.3.4 Common Mode Range	34
4.3.5 Current Drain	34

	4.3.6	Control Voltage Law	34
	4.3.7	Pre-charge of C Store	34
	4.3.8	AC Output Voltage	35
	4.3.9	Attack Tone Burst Transient Response	36
	4.3.10	Recovery Time	37
	4.4	External Amplifier	37
	4.5	Measured Performance of AGC Amplifier	39
	4.5.1	Reproducibility	39
	4.5.2	Variations with Supply Voltage	39
	4.5.3	Temperature Variation of Output Level	41
	4.5.4	Distortion Levels with Continuous Sine Wave Input	41
	4.5.5	Tone Burst Response	42
	4.5.6	Regulation	45
	4.6	Fabrication	45
Chapter	V	The Eight Bandpass Filter Module	47
	5.0	Introduction	47
	5.1	Choice of Filter Type	48
	5.2	Gain Sensitivity Optimisation - Approximate Analysis	51
	5.2.1	Filter Types 1,2	52
	5.2.2	Filter Types 3,4	54
	5.2.3	Filter Types 5,6	55
	5.2.4	Summary of Approximate GSP Optimisation	56
	5.3	GSP Optimisation - Exact Computed Results	56
	5.4	Chosen Filter Configuration	58
	5.4.1	Design Values of Selected Filter	63
	5.5	Filter Amplifier SIC - M11302	64
	5.5.1	Computer Aided Analysis of the Filter Amplifier	65
	5.6	Computer Simulation of Complete Filter Amplifier Units	68
	5.7	Fabrication	69
	5.7.1	Process Outline	69
	5.7.2	Automatic Trimming Method for Frequency and Q Adjustment	70
	5.7.3	Special Layout Considerations	78
	5.8	Measured Performance	79
	5.8.1	Temperature Effects of Amplifier Alone	79
	5.8.2	Overall Filter Drift Vs. Temperature	80

Chapter	VI	Detector/Trigger Units - MH8802	82
	6.0	Introduction	82
	6.1	Circuit Requirements	82
	6.2	Description of Circuit Operation	83
		6.2.1 Detector Circuit	83
		6.2.2 Trigger Circuit	83
	6.3	Calculated Performance Characteristics	84
		6.3.1 Temperature Effects on the AC Input Voltage at which Triggering Occurs	84
		6.3.2 Time Delay Before Triggering	86
	6.4	Fabrication	86
	6.5	Measured Performance	92
		6.5.1 Linearity	92
		6.5.2 Temperature Drift	92
Chapter	VII	Support Circuitry for Complete Receiver	93
	7.0	Introduction	93
	7.1	Power Supply	93
	7.2	Input Stage	93
	7.3	Prefilter	95
		7.3.1 Dial Tone Rejection	96
		7.3.2 3 kHz Pre-emphasis	96
		7.3.3 Response at Signal Frequencies	98
	7.4	Trigger Threshold Voltage Control Circuitry	98
Chapter	VIII	Complete Tone Receiver	100
	8.0	Introduction	100
	8.1	Overall Performance Characteristics	100
		8.1.1 Signal Levels	102
		8.1.2 Overall Signal Recognition Bandwidth Vs. Temperature	103
		8.1.3 Voice Immunity Performance	104
		8.1.4 Error Rate Vs. Signal to Noise Ratio	105
	8.2	Fabrication of Complete Receiver	106
	8.3	Comparison with Other Receivers	107
Chapter	IX	Conclusions	111
Chapter	X	References	113

Appendices		Page
Appendix A	Program Used for GSP Computation	115
Appendix B	Data Used in Amplifier Stability Analysis Program	117
Appendix C	Data Used in Q Stability Analysis	120
Appendix D	Data Pertaining to Analysis of Prefilter Frequency Response	124
Appendix E	Microsystems International Limited Preliminary Data Sheet on the New Receiver	128

LIST OF ILLUSTRATIONS

Figure		Page
2.1	Frequency encoding matrix for tone dialing	5
2.2	Simplified system schematic for dialing to Central Offices	6
2.3	Insertion loss of 1300 Ω of cable for various gauges	9
2.4	Simplified system schematic for end to end signalling	11
2.5	C message filter weighting characteristic	15
3.1	Block schematic of BTL tone receiver	18
3.2	Block schematic of Bell System 403 tone receiver	19
3.3	Block schematic of the new tone receiver (MIL ME8800)	20
3.4	Receiver sensitivity reduction vs. signal to noise ratio for the limiter guard system (A), and the new system (B)	23
4.1	Feedback AGC system block schematic	26
4.2	Schematic of AGC amplifier used in the tone receiver	30
4.3	Complete AGC circuit including external amplification	38
4.4	Effects of supply voltage variation on levels in AGC circuit	40
4.5	Temperature variation of AGC amplifier output level	41
4.6	AGC amplifier output distortion vs. frequency	42
4.7(a)	Tone burst response of AGC amplifier	43
4.7(b)	Response of AGC amplifier to periodically attenuated input	43
4.8	Regulation of AGC amplifier output voltage	44
4.9	ML1314K AGC integrated circuit die (70 x 65 mil.)	46
5.1	Six positive gain, minimum complexity filter sections	50
5.2	Two true bandpass filter sections	51
5.3	Filters 1,2 Exact GSP vs. component ratios (a^2 = capacitor ratio, b^2 = resistor ratio)	59

Figure

5.4	Filters 3,4 Exact GSP vs. component ratios (a^2 = capacitor ratio, b^2 = resistor ratio)	60
5.5	Filters 5,6 Exact GSP vs. component ratios (a^2 = capacitor ratio, b^2 = resistor ratio)	61
5.6(a)	Filter configuration selected for use in the new receiver	62
5.6(b)	Normalised frequency response of bandpass filter section	62
5.7	Filter amplifier schematic	65
5.8	Computed gain error of filter amplifier vs. temperature: $R_x = R_y = 3 \text{ K}\Omega$	66
5.9	Computed amplifier frequency response: $R_x = R_y = 3 \text{ K}\Omega$	66
5.10	Predicted Q drift due to amplifier for filters 1,5	69
5.11	Substrates for the filter module	71
5.12	Eight amplifier IC used in the filter module (ML1312)	72
5.13	Complete filter module assembly (MH8802)	73
5.14	Schematic of eight bandpass filter module (MH8802)	74
5.15	External circuit for making filter oscillate	75
5.16	Filter oscillation frequency offset and amplitude vs. R_a	77
5.17	Layout geometry for R_x for good Q trimming resolution	78
5.18	Measured Q drift with temperature due to amplifier alone ($Q=15$)	79
5.19	Measured overall filter centre frequency drift vs. temperature	80
5.20	Measured overall filter Q drift vs. temperature ($Q = 15$)	81
6.1	Detector/trigger schematic	82
6.2	Quad detector/trigger module schematic (MH8803)	88
6.3	Substrates for the detector/trigger module (MH8803)	89
6.4	Quad detector/trigger IC chip (ML1313)	90

Figure

6.5	Complete MH8803 detector/trigger module assembly	91
6.6	Peak ac triggering and release voltages of MH8803 vs. V_t	92
6.7	Peak ac triggering & release voltages of MH8803 vs. $^{\circ}\text{C}$ ($V_t = 5\text{V}$)	92
7.1	Complete Tone Receiver Schematic	94
7.2	Frequency response of prefilter	97
8.1	Receiver waveforms in response to 40 mS input	101
8.2	Overall receiver recognition bandwidth vs. temperature	103
8.3	Simulations on ME8800 receiver vs. dynamic range	105
8.4	Error rate of ME8800 receiver vs. S/N ratio	106
8.5	Complete ME8800 tone receiver PCB	108
C1	Type 5 circuit for computer analysis of Q stability	121
C2	Type 1 circuit for computer analysis of Q stability	121
D1	Nodal circuit diagram for ME8800 prefilter used for KRON simulation	125

LIST OF TABLES

	Page
2A LC tone dial output levels	8
2B Maximum and minimum CO receive levels	10
2C Measured inter-office insertion loss at 1000 Hz (1969 Bell System survey)	12
2D Maximum and minimum end to end receive levels	12
2E C message weighting noise according to 1966 survey of the North American telephone network	14
4A Measurements on 15 randomly selected AGC microcircuits	40
5A Summary of Results of Approximate Analysis of GSP Optimisation	57
5B Amplifier transfer function vs. temperature	67
8A Receiver comparison data	110
A1 Univac Fortran V program for exact GSP computation	116
B1 Amplifier stability analysis data	118
B2 Summary of data used for NPN and PNP transistor models	119
C1 (a) Input data for SLIC analysis of effect of amplifier variations on Q and f_o	122
C1 (b) Input data for SLIC analysis of effect of amplifier variations on Q and f_o	123
D1 Input data for KRON simulation of ME8800 prefilter response	126
D2 Output data from KRON simulation of ME8800 prefilter response	127

GLOSSARY OF SYMBOLS AND ABBREVIATIONS

Q	filter quality factor = $1/\text{bandwidth}$
AGC	automatic gain control
SIC	silicon integrated circuit
mA	10^{-3} amps
mW	10^{-3} watts
dB	decibel
ac	alternating current
dc	direct current
CO	Central Office
PBX	Private Branch Exchange
BTL	Bell Telephone Laboratories
MIL	Microsystems International Limited
BNR	Bell Northern Research
A.T.& T. American Telephone & Telegraph Corporation	
GSP	gain sensitivity product
$\hat{\hat{\text{GSP}}}$	minimum GSP with unconstrained component value ratios
$\hat{\text{GSP}}$	minimum GSP with practical component value ratios
Hz	hertz
LC	inductor-capacitor
ft.	foot
VGE	variable gain element
tc _r	temperature coefficient of resistance
B _n	B _n of NPN SIC transistor
ppm	parts per million
mS	10^{-3} seconds

σ	standard deviation
mil	one-thousandth of one inch
PCB	printed circuit board
MOS	metal oxide silicon
rms	root mean square
SiC	silicon carbide
Vd	approximate voltage drop of an SIC base emitter junction ($\approx 0.65V$)
IF	intermediate frequency
Vp	the peak ac voltage of a sine wave
Vo	detector output voltage
Vt	detector threshold voltage
β_l	low frequency group signal level at AGC circuit output
β_h	high frequency group signal level at AGC circuit output
$\hat{\beta}_l$	nominal value of β_l
$\hat{\beta}_h$	nominal value of β_h
β_o	total AGC circuit output
β_n	noise level at AGC circuit output
S/N	signal to noise ratio
Δ	change from nominal level of detector output
$\Delta_{max.}$	maximum value of Δ accepted by receiver
Ψ	combined voltage gain of operational amplifiers 5.6 in Fig. 7.1

Chapter I

INTRODUCTION

Pushbutton telephone dialing was introduced by Bell Telephone Laboratories (BTL) to the North American telephone system in the early 1960's. (1.1) Since then it has been steadily replacing rotary dialing due to its advantages both to users and to operating companies. Japan and Belgium have recently introduced the same tone dialing system and many other countries will shortly be following in a general worldwide trend in this direction.

Although there are many conceivable modulator-demodulator systems used for pushbutton dialing, the approach taken in the present BTL system has the advantage of using voiceband signals. In contrast to the output of rotary dials, which is in 50V digital pulses, and an earlier tone dialing system tested by BTL which used a combination of tones and direct current (dc) level shifts, (1.2) these signals can be transmitted over any voice link and can be used to transmit data after the link has been established ('end to end' signalling). Thus, the ordinary telephone handset can also perform as a low speed data set and have applications in remote control, credit card activities, stock price information, and many other areas. Information can be returned to the telephone set by means of automatic voice response units -- which are now available from many sources. (1.3)

A drawback to the use of voiceband signals, however, is the possibility of 'talk-off,' i.e. speech signals simulating pushbutton signals. To get around this problem the BTL tone dialing system was carefully designed so that the probability of talk-off is very small. The key to the system is

that two simultaneous non-harmonically related frequencies are generated when each button is pushed. The receiver can then check that the amplitude ratio of the two frequencies is within expected limits and that no significant energy at non-valid frequencies is present. The resulting talk-off rate can then be as low as one simulation in ten-thousand calls for Central Office (CO) receivers.^(1.4)

The feature of end to end signalling has thus been provided for at the cost of somewhat increased system complexity. To the present day, however, very little use has been made of this feature mainly due to the high cost of available end to end receivers which require much greater sensitivity than CO receivers and are more expensive.

The objective of this thesis, which was initiated at Carleton University in April, 1971, was to create a high performance, general purpose receiver suitable for low cost volume production. This objective was achieved and the circuitry was adopted by Microsystems International Limited (MIL) for volume production in 1973. By designing the receiver to be adaptable for Central Office, Private Branch Exchange (PBX) or end to end use, market opportunities became available in the existing applications as well as the potentially large data transmission area.

The overall approach taken was to consider the most economical high performance realisation of a receiver using the technology available at MIL. Significant results of this work were:

- a new tone receiver detection system with reduced filtering requirements but with relatively high voice immunity*
- a high performance audio automatic gain control (AGC) circuit with rapid response and precise level control*

- optimisation analysis of second order minimum component count bandpass filter sections using "Gain Sensitivity Product" (GSP) and minimal area as performance criteria.
- a simplified automatic trimming technique for adjusting Q and centre frequency of the bandpass amplifiers resulting in a trim time of about 6 seconds per section*
- a precisely controlled a.c level detector with 0.5% accuracy.*

* Four patent applications (assigned to Microsystems International Limited) were filed on the items asterisked above.

Three silicon integrated circuits and two hybrid microcircuits were especially designed and fabricated for the new receiver. This helped to reduce the cost and size as well as improve the performance due to the superior temperature tracking qualities of hybrid microcircuit components. One of the circuits is an eight filter module which contains all the critical filtering of the receiver and provides a dramatic size reduction compared with discrete component filters.

Summaries of the contents of the following chapters are outlined below.

Chapter 2 deals with the overall system that tone receivers are required to operate in. The system parameters that are included were gathered from many sources and were used to provide part of the performance specification of the new receiver. Chapter 3 describes previous receivers designed by BTL and the system approach used in the new receiver.

Chapters 4, 5 and 6 describe in detail the three new microcircuit modules of the new receiver, the AGC circuit, the eight filter circuit, and the

quad detector circuit. Chapter 7 describes the support circuitry that completes the receiver. Chapter 8 describes the performance of the complete receiver. Chapters 9, 10 and 11 contain the conclusions, references and appendices.

Chapter II

SUBSCRIBER TONE DIALING - SYSTEMS CONSIDERATIONS

2.0 Introduction

The new tone receiver is intended for use in existing tone dialing telephone systems. Since no standard specification for tone receivers was available at the start of this work, it was necessary to derive appropriate objectives for the performance of the new receiver by study of the available literature. This chapter gives a brief background of the tone dialing system and derives appropriate performance limits for CO, PBX, and end to end signalling tone receiver applications.

Fig. 2.1 Frequency encoding matrix for tone dialing

2.1 Frequency Allocations

When any button on the pushbutton dial pad is pushed, two frequencies are simultaneously generated and sent down the telephone line. One of four low-group frequencies identifies the row of the button, and one of four high-group frequencies identifies the column of the button. The frequencies involved have been standardised throughout North America since 1960. Subsequently, at the Tokyo meeting of CCITT Study Group XI, they were recommended for adoption as an international standard for pushbutton telephones. They are summarised in Fig. 2.1.

2.2 Signal Power Levels

2.2.1 Dialing to Central Offices

Fig. 2.2 Simplified system schematic
for dialing to Central Offices

Fig. 2.2. shows a simplified schematic of the telephone system during tone dialing. Signals generated at the telephone set by the dial are transmitted via the subscriber loop to the tone receiver at the CO. The tone receiver demodulates the tones into digital information which is used to make appropriate connections in the switching matrix to set up the call.

The expected variation in tone signal amplitudes received at the CO can be calculated from a knowledge of dial output, and loop loss statistics. The standard tone dial used in the North American telephone system is the Meacham Bridge LC unit described in reference 1.1. A table of maximum short loop output levels and minimum long loop output levels of the standard LC tone dial is shown in Table 2A. A variation in the nominal dial output level vs. loop length is caused by regulation circuitry in the telephone set which attenuates output levels on short loops to partially compensate for the variable insertion loss of the subscriber loop. The values in Table 2A also include allowance for ± 3 dB tolerance on the nominal output levels of the dial. Thus, at 1300 Ω loop length, the nominal output level of the dial is high but since the minimum value is of interest, the 3 dB tolerance is subtracted from this to obtain the figures in the Table. Similarly, at 0 Ω loop length, the nominal output level of the dial is low but since the maximum value is of interest, the 3 dB tolerance is added to obtain the values in the Table. The ± 3 dB tolerance on the dial allows for both unit to unit variation and temperature effects over the -45°C to $+65^{\circ}\text{C}$ dial operating temperature range.

<u>Frequency Hz</u>	<u>1300Ω loop min. output dBV</u>	<u>0Ω loop max. output dBV</u>
697	-7.5	-5.8
770	-7.0	-5.3
852	-6.8	-5.1
941	-6.5	-4.8
1209	-4.7	-3.0
1336	-4.6	-2.9
1477	-4.5	-2.8
1633	-4.4	-2.7

(These figures include a ± 3 dB tolerance on the standard LC tone generator output; load = 900 Ω)

Table 2A - LC tone dial output levels

Subscriber Loop Insertion Loss

The maximum loop length that can normally be used is limited by a maximum allowable dc resistance of 1300 Ω . This is to ensure that sufficient switching current is available to operate relays in the CO. Standard practice is to use 26 gauge cable for distances less than 15,800 feet corresponding to the 1300 Ω limit. Beyond this, heavier gauge cable such as 24, 22 or 19 gauge is used. To avoid excessive attenuation on these extra long loops, they are 'loaded' at appropriate locations with series inductors, thus reducing the loss at 3kHz considerably. This is illustrated in Fig. 2.3 which shows theoretical loss characteristics for 1300 Ω unloaded 26 gauge cable (B) and H88* loaded 24, 22 and 19 gauge cable (A). A third curve (C), shows the actual measured loss of a statistical survey taken by A.T. & T. (2.2)

*H88 refers to one of the standard Bell System loading coils.

As can be seen, it is higher than the theoretical loss of any of the cable types. This is due partly to non-ideal installation practices such as not cutting off bridge taps on the cable. These are originally there because the final route of a cable may not be known at the time of laying so it is tapped off in several directions. When it is used for a subscriber, the unused taps are supposed to be removed but this does not always occur. The result can be excess stray capacitance shunting the cable and increasing its insertion loss.

Fig. 2.3 Insertion loss of 1300Ω of cable for various gauges

Central Office Receive Signal Level

From Table 2A and Fig. 2.3 curve C, the total range of received signal amplitudes can be calculated. They are shown in Table 2B.

<u>Frequency Hz</u>	<u>Max. CO receive level, dBV</u>	<u>Min. CO receive level, dBV</u>
697	-5.8	-15.5
770	-5.3	-15.5
852	-5.1	-15.8
941	-4.8	-16.0
1209	-3.0	-15.7
1336	-2.9	-16.3
1477	-2.8	-16.8
1633	-2.7	-17.7

Table 2B - Maximum and minimum CO receive levels

2.2.2 Dialing to PBX's

PBX telephone systems normally have shorter loop lengths than CO systems and generally have fewer bridge taps so the received tone signal levels are generally higher than for CO's. However, to accommodate the largest PBX systems, it is desirable to cover a similar signal range as for CO's (see Table 2B).

2.2.3 End to End Signalling Levels

Fig. 2.4 Simplified system schematic for end to end signalling

Fig. 2.4 shows the arrangement occurring during end to end signalling. In this case there are three insertion losses to be considered, caused by the two telephone loops and the toll loss between the two end offices. The maximum total loss of the two loops can be taken as double the single loop loss shown in curve (C) Fig. 2.3.

Insertion loss between end offices is flat to within a dB over the tone signalling frequency range because of equalisation networks which are used. The maximum transmission plant design value of this loss is 17 dB. Statistical information on a 1969 survey of the loss between Bell System end offices at 1,000 Hz is shown in Table 2C. (From reference 2.3, p. 1329). The worst case loss of 17 dB is approximately 3.5σ away from the statistical mean, where σ is the standard deviation.

<u>Airline Distance (miles)</u>	<u>Loss (dB)</u>	<u>Standard Deviation (dB)</u>
0 - 180	6.5	2.0
180 - 725	7.3	2.3
725 - 2900	7.7	2.5

Table 2C - Measured inter-office insertion loss
at 1000 Hz (1969 Bell System survey)

Adding the worst case inter-office insertion loss of 17 dB to the two subscriber loop insertion losses enables the minimum end to end tone signalling receive level to be calculated using the output levels of Table 2A. The maximum receive level is the same as that for the CO case. These values are summarised in Table 2D.

<u>Frequency Hz</u>	<u>Maximum end to end receive level, dBV</u>	<u>Minimum end to end receive level, dBV</u>
697	-5.8	-40.5
770	-5.3	-41.0
852	-5.1	-41.8
941	-4.8	-42.5
1209	-3.0	-42.7
1336	-2.9	-44.0
1477	-2.8	-45.1
1633	-2.7	-46.0

Table 2D - Maximum and minimum end to end receive levels

2.3 Frequency Tolerance Considerations

2.3.1 Tone Generator Tolerance

The tone generator has a worst case frequency tolerance of $\pm 1.5\%$ about the nominal values. This is the only consideration for CO or PBX receivers but, with end to end signalling, carrier induced frequency offset can cause additional tolerance allowance to be necessary.

2.3.2 Carrier Induced Frequency Offset

The frequency of a signal may shift when transmitted over a carrier telephone channel. This will occur when modulating and demodulating carrier supply frequencies are not identical and has to be considered for end to end signalling applications. A statistical survey of toll connections in North America^(2.3) found that 90% of all connections had a frequency offset of less than 0.2 Hz and none had more than 3 Hz. The Western Electric 403 data set allows for up to 15 Hz carrier induced frequency offset which appears to be an excessively generous figure compared to the above values.

2.4 Signal to Noise Ratio

For normal dialing to CO's or PBX's signal to noise ratios are typically in the 40 - 60 dB region. However, with end to end signalling, the signal to noise ratio can drop much lower as the following study shows. The results of a 1966 Bell System survey of balanced noise levels with C message weighting are shown in Table 2E (2.3). The units of noise measurements are dBrnC. dBrnC are units of noise power measured after the noise has been passed through a 'C weighted filter.' 0 dBrnC then correspond to -90 dBm. The C message weighting curve is shown in Fig. 2.5.

These units have been used in telephony since the 1950's and were derived such that 0 dBrnC represents a negligible noise level. The C message weighting curve is shown in Fig. 2.5. The low frequency roll off makes a good approximation for the working conditions of tone receivers which normally have high pass pre-filters for removal of hum and the dial tone.

Table 2D gives the minimum receive levels for end to end signalling assuming simultaneous worst case values for dial output, loop losses and inter-office loss. According to this the minimum composite amplitude of a tone pair is about -41 dBV. The mean noise level on maximum length connections is -59 dBV. Thus, the minimum expected signal to noise ratio is 18 dB.

<u>Airline Distance (miles)</u>	<u>Noise dBrnC</u>	<u>Standard Deviation (dB)</u>	<u>Equivalent dBV</u>
0 - 180	21.6	6.4	-70
180 - 725	29.6	4.2	-62
725 - 2900	32.5	4.1	-59.1

Table 2E C message weighting noise according to 1966
survey of the North American telephone network

2.5 Dial Tone

A tone receiver for CO or PBX use must reject the dial tone since this is present for the first digit dialed. The standard North American dial tone used in tone dialing systems consists of two equal amplitude frequency components at 350 and 440 Hz. The total amplitude is 150 mV rms.

Fig. 2.5 C message filter weighting characteristic

2.6 Common Mode Noise

The main contribution to common mode noise is power line hum at 60 Hz and its associated harmonics. A 1964 survey of the Bell System^(2.2) found that the mean 3 kHz flat weighting noise for station sets was about 1V rms, with some sets reaching up to 15V rms. This emphasises the desirability of balanced inputs for tone receivers.

2.7 Echoes

With certain connections, especially trans-continental ones, a certain amount of echo can occur due to termination mismatch. According to reference 3.1, echoes may be expected with up to 40 msec. delay and an amplitude 14 dB below the signal level.

Chapter III

TONE RECEIVER DETECTION SYSTEM DESIGN

3.0 Introduction

Chapter II described the signal conditions which are encountered in the various applications of tone signalling. The objective of this project was to create a universal tone receiver which could be used for any application by making minor changes. To reduce the manufacturing cost a new detection system was conceived which eliminates the need for separating the two component frequencies and thus eliminates much of the filtering normally required. To gain perspective on this it is of interest to briefly summarise existing detection systems designed by BTL - (most non-BTL tone receivers are modelled on the BTL design).

These fall into two basic classes: Those with moderate dynamic range for dialing use with PBX, key telephones or CO's, and the 403 data sets which have wide dynamic range for end to end signalling.

3.1 BTL CO, PBX and Key Telephone Tone Receivers

The block diagram for BTL CO, PBX or key telephone systems can be summarised by Fig. 3.1. The different types have different dynamic ranges, band splitting filtering and timing arrangements, but use the same basic system. (1.4)

Fig. 3.1 Block schematic of BTL tone receiver

In this system the amplitude variation of the input signal, due to variations in dial output and loop insertion loss, is accommodated by separating the two tones into their groups and hard limiting. This results in two square wave of well defined amplitudes. These drive banks of second order bandpass filters with Q's of 15 which are followed by ac level detectors to determine which frequencies are present.

The high quality CO tone receivers use band elimination filters to separate the frequencies into the two groups rather than somewhat simpler bandpass or highpass/lowpass filters. If only two valid frequencies are present, this makes no difference; but, if extraneous noise is also present it is passed with reduced loss to the limiters. The idea behind this is that speech signals may have components present at two valid frequencies but they usually will have broadband power as well. This power is allowed to pass through the system and desensitise it to avoid a tone simulation. This technique is referred to as 'Limiter Guard Action.' It is basically a scheme to demand high signal to noise ratio before accepting that a valid signal is present. The technique can be extended

further by emphasising non-tone frequencies. The limitations of this are when the signal to noise ratio of valid signals is reduced excessively.

One drawback to this approach is that the receiver can accept tone pairs with up to 20 dB relative amplitude ratios since there is no positive ratio checking mechanism. The 20 dB limit on the ratio is due to imperfect band split filtering which allows the dominant frequency to feed through into the other channel and interfere with the smaller amplitude frequency. Since the relative amplitude ratio of valid signals should be 4 dB or less, tighter control on the acceptable ratio than 20 dB is desirable to minimise simulations.

Fig. 3.2 Block schematic of Bell System 403 tone receiver

3.2 Bell System 403 Data Set

The block diagram of the tone receiver part of the Bell System 403 data set is shown in Fig. 3.2. It is similar to that of Fig. 3.1 except that instead of the line amplifier having a fixed gain, an automatic gain control amplifier is used. (3.1) This compresses input amplitude variation

of 41 dB down to ± 1 dB so that a well controlled amplitude is fed to the band split filters and limiter amplifiers. This enables accommodation of the wide dynamic range of input signal that can occur with end to end connections (Table 2C). The AGC amplifier also avoids response to echoes by having a slow recovery time.

3.3 New Detection System Using Pivoting Threshold Control

A block diagram of the new tone detection system is shown in Fig. 3.3. The system operates as follows: The balanced input, consisting of a 20K Ω bridging transformer, removes common mode noise due to power line hum and other factors. (See section 2.6). The highpass pre-filter has a cut-off frequency of 670 Hz to remove dial tone and low frequency noise, and a 3 kHz emphasis circuit to help reduce voice simulations. (See section 7.3.) This is followed by a precision AGC circuit which

Fig. 3.3 Block schematic of the new tone receiver (MIL ME8800)

compresses input variations of up to 46 dB to output variation of only ± 0.14 dB. By way of comparison, this represents a factor of seven times improvement in control compared to the AGC circuit used in the 403 data set. (See section 3.2.) This precise control of the output level enables the ME8800 receiver to dispense with the band separating filters and limiter amplifier used in the BTL receiver designs. (See section 3.3.1.) The dual frequency signal is passed directly from the AGC circuit to the frequency selecting channel filters and detectors. The ac level detectors are divided into high frequency and low frequency groups as shown, so that the ratio of the amplitudes of the two frequencies present in a valid tone can be monitored and the allowable ratio controlled. This ratio should be within 3.1 dB for CO receivers, and 6 dB for end to end receivers. (Tables 2B, 2D.)

3.3.1 Receiver Guard Action

The amplitude of the input signal to the eight bandpass filters is maintained constant by the AGC amplifier at a level β_s . (See Fig. 3.3.) With valid input tones to the receiver, this signal will be composed of two sine waves, with one frequency present from each group, such that:-

$$\beta_l + \beta_h \cong \beta_s \quad 3.1$$

where β_l is the amplitude present at a valid low group frequency and β_h is the amplitude present at a valid high-group frequency. The nominal amplitudes of the high- and low-group frequencies are approximately equal so:-

$$\hat{\beta}_l = \hat{\beta}_h \cong \beta_s/2 \quad 3.2$$

where $\hat{\beta}_l$ and $\hat{\beta}_h$ represent the nominal signal amplitudes of the low-group and high-group frequencies present at the input to the bandpass

filters. The threshold levels of the detectors at the outputs of the filters are set such that input amplitudes of close to $\frac{\beta_s}{2}$ at valid frequencies are required to operate them. Limited variations around this nominal level are allowed by the pivoting trigger voltage circuit, (Fig. 3.3), in order to accommodate valid input tones with non-equal component amplitudes. The effect of the pivoting trigger voltage circuit is to demand that:- (See section 3.3.2.)

$$\beta_l + \beta_h \cong \beta_s \quad 3.3$$

i.e. that nearly all the energy present is composed of two valid frequencies only .

The approximation sign is necessary in equation 3.3 due to the combined effects of component tolerances and allowances for normal signal to noise conditions, and residual ripple in the detector circuitry of the receiver. However, the equation is accurate to within approximately 1 dB.

Consider now the effects of an extraneous 'noise' signal, with amplitude β_n , present concurrently with valid equal amplitude low- and high-group frequency components. Equations 3.4 to 3.6 then apply:-

$$\beta_l = \beta_h \quad 3.4$$

$$\beta_l + \beta_n + \beta_h = \beta_s \quad 3.5$$

$$\beta_l = \beta_h = (\beta_s - \beta_n)/2 \quad 3.6$$

By comparing 3.6 with 3.2 it can be seen that the effect of the extraneous signal is to reduce the amplitudes at valid frequencies fed to the band-pass filters. If the extraneous signal has a large amplitude, the valid

frequency amplitudes will be reduced sufficiently to prevent actuation of the receiver outputs. Thus, guard action is provided corresponding to that of the BTL receiver 'Limiter Guard Action.' To compare the relative effectiveness of the two systems, Fig. 3.4 was plotted. This shows the sensitivity reduction of the receiver vs. signal to noise ratio. The sensitivity reduction is defined as the drop in energy at valid frequencies presented to the bandpass filters caused by the extraneous noise signal. Thus an infinite signal to noise ratio gives 0 dB sensitivity reduction. Curve A was taken from reference 1.4. Curve B was plotted from equations 3.7 and 3.8. A single frequency noise signal is assumed for each of the curves.

$$\text{Signal to noise ratio (per frequency)} = (\beta_s - \beta_n) / 2\beta_n \quad 3.7$$

$$\text{Sensitivity reduction} = \beta_s / (\beta_s - \beta_n) \quad 3.8$$

Fig. 3.4 Receiver sensitivity reduction vs. signal to noise ratio for the limiter guard system (A), and the new system (B)

Comparing the two curves, it can be seen that if the receivers are adjusted such that a 2 dB drop in sensitivity is sufficient to disable the outputs, the Limiter Guard system will tolerate signal to noise ratios down to 1.7 dB, whereas the new system only tolerates signal to noise ratios down to 6 dB. Since the worst case expected signal to noise ratio (per frequency) is approximately 12 dB, (see section 2.4), it would appear that the new receiver has more desirable signal quality checking characteristics than the Limiter Guard system and thus has the possibility of giving fewer false outputs due to speech and other noise inputs.

3.3.2 Amplitude Ratio Control

The AGC circuit keeps the total input level to the bandpass filters constant at β_0 but, since the two component frequencies of a valid tone do not necessarily have equal amplitudes, the input level at each frequency will not necessarily be at the nominal value of $\beta_0/2$. This is allowed for in the new receiver by having variable group threshold levels in the trigger circuits which are controlled by the pivoting trigger voltage circuit (Fig. 3.3). The highest level present in the four detectors of each group is monitored and, if the level appears at Δ volts above the nominal expected value, the threshold voltage of the trigger circuits of the other group is correspondingly lowered by Δ volts. This approach allows for relative amplitude differences between the two component frequencies of an input tone but still constrains equation 3.3 to hold and thus maintain guard action against simulations. By limiting the maximum allowable value of Δ to $\Delta_{\max}$, the maximum allowable ratio of the amplitudes of the two component frequencies can be controlled as desired.

For a numerical example, consider the case where the maximum allowable amplitude ratio is ± 6 dB, and the nominal trigger voltage is 5 volts.

$$\text{Assume: } \beta_l = 2 \beta_h, \quad \beta_n = 0$$

From 3.2, 3.5;

$$\beta_l = \hat{\beta}_l + 2.5\text{dB}$$

$$\beta_h = \hat{\beta}_h - 3.5\text{dB}.$$

Thus, to ensure recognition of the high-group frequency, its threshold must be lowered 3.5 dB from the nominal, i.e. from 5V to 3.3V. A clamp preventing the thresholds from dropping below 3.3V will allow acceptance of tones with component frequency amplitudes of up to 6 dB difference but no more.

3.3.3 Echo Protection

The new system protects against responding to echoes in a similar manner to that of the 403 data set. The AGC amplifier has a slow recovery time so that echoes of reduced amplitude are ignored for up to 40 ms from the time the original signal ceases.

3.3.4 Output Format

The output of the basic tone receiver is undecoded, i.e. there are eight outputs corresponding to the eight tone frequencies. When an input tone is present two of the eight outputs are activated. The output switches are grounded in the active state, floating in the inactive state. It is straightforward to go from this format to decimal, binary or other formats by using standard logic circuitry.

Chapter IV

AGC AMPLIFIER CIRCUIT

4.0 Introduction

Chapter III described the new tone receiver algorithm which has considerably reduced filtering requirements compared to previous approaches but requires for its realisation an audio AGC amplifier with tight control and rapid attack time. The basic problem with designing such a circuit is to avoid instability caused by control voltage feedthrough into the feedback control loop. This chapter describes a novel approach to solving this problem, which is basically to control the phase of the control voltage feedthrough by swamping out random feedthrough by the addition of a small amount of degenerative feedthrough. A realisation of this concept as a silicon integrated circuit is then described along with details of the excellent performance and stability that are achieved.

4.1 Background

Feedback AGC circuits typically use an arrangement as shown in Fig. 4.1. The principle of operation is as follows:

Fig. 4.1 Feedback AGC system block diagram

A detector incorporating rectification and smoothing converts the ac output level into a dc control voltage V_c which can be used to control the gain of a variable gain element. When the dc voltage reaches a reference voltage V_r , the gain is reduced according to a function of the difference between V_c and V_r , i.e.

$$A = K.F (V_c - V_r) \quad 4.1$$

where K is a constant and F is some function. The design of practical AGC circuits is complicated by the fact that all variable gain elements suitable for IC structures have some control voltage feedthrough to the signal path. If the frequency range of the control voltage is very low compared to the signal band of interest, it is easy to attenuate it by the bandpass amplifier so that it is prevented from circulating around the control loop. This is the case with, for example, intermediate frequency (IF) AGC amplifiers which operate at frequencies like 455 kHz, 35 MHz, etc., and have control voltages operating in the 50 mS time constant region. With these systems, unbalanced variable gain elements are used with no problem.

However, in the case of a tone receiver system, the signal band of interest is down to 670 Hz and, since rapid attack is desired, the control voltage has a time constant of about 3 mS. With this situation, the control voltage feedthrough is not eliminated by the signal bandpass amplifier and circulates round the control loop. If the feedthrough is in the degenerative direction, the gain becomes stabilised at a rate depending on the time constants in the bandpass amplifier, i.e. the time to discharge the coupling capacitors by the dc change caused by the change

in control voltage. If the feedthrough is in the regenerative direction the system can become unstable and low frequency oscillation can occur.

Two approaches to the problem which can be taken are to:

- (a) use an unbalanced system with the phase of the control voltage feedthrough in the degenerative direction;
- (b) use a near-perfectly balanced system.

Approach (a) has the disadvantage that the large amount of control voltage feedthrough takes a relatively long time to be discharged, thus slowing the settling time of the system. Approach (b) is difficult to achieve with sufficient precision to allow its use with tightly controlled loops, i.e. when the output level is controlled to a fraction of a dB instead of several dB -- a high 'loop gain' configuration. Thus, approach (b) has the advantage of rapid response time but the disadvantage of potential instability due to feedthrough in the wrong phase. This can either cause poor SIC yield or unreliable operation. The circuit described here, which was designed for the tone receiver, uses a balanced voltage controlled amplifier as the variable gain element. This is subsequently unbalanced by adding a small amount of control voltage feedthrough in the degenerative direction. This is sufficient to ensure stability but is much smaller than if a completely unbalanced circuit were used. Thus the system has a rapid settling time as the amount of control voltage feedthrough to be discharged is small.

4.2 Description of Circuit Operation

Fig. 4.2 shows the circuit schematic. The operation is as follows. Differential input pair Q8, Q9 have degenerative resistors R1 and R2

in their emitters to enable large input signals to be handled of up to 2V rms.

The input voltage is converted into differential currents in the collectors of Q8, Q9. These currents are then split between Q2, Q3 and Q4, Q5 depending on their base voltages which are controlled by leads 1 and 3. Only the currents via Q3, Q4 are directed to the load so the more current that passes through Q2, Q5 the greater the attenuation of the signal. The outputs of Q3, Q4 are coupled via Q6, Q7 and Q18, Q19 to a single ended resistive load R11. Up to this point the circuitry is in a balanced configuration and, with perfect matching of components, the gain controlling voltage on lead 1 can be varied with respect to the bias voltage on lead 3 without varying the voltage across load R11. However, due to component mismatches, there is in practice, always some control voltage feedthrough and this will be in a somewhat random phase and magnitude depending on process parameter variations and other factors. This undesirable situation is overcome in this circuit by swamping out this small but random direction control voltage feedthrough by adding a larger feedthrough in the degenerative direction. Further, this is added such that it is smoothly applied over the whole gain range of the voltage controlled amplifier. The degenerative feedthrough is added by varying the current through Q20 and D1 by means of Q22, which is operated in the same manner as Q3, Q4, i.e. as the control voltage is increased the currents through Q3, Q4 and Q22 are reduced. The potential on the base of Q20 is nominally constant so the potential on the base of Q23 will vary at about 120 mV per decade change in current through Q22 due to the change in voltage drop this current change produces across Q20 and D1. Q23 then provides a low

Fig. 4.2 Schematic of AGC amplifier used in the tone receiver

output impedance for the voltage controlled amplifier at lead 10.

The maximum gain of the voltage controlled amplifier is about unity and it can be reduced 46 dB or more by applying an appropriate voltage on lead 1. The external amplifier with gain $-K$, then amplifies the audio signal which is detected by rectifier Q25, Q26 charging C store via Rch. The value of K controls the dynamic range of AGC that is available. When the ac voltage swing at lead 9, plus the dc bias on lead 11, is sufficient to charge C store to approximately 9.8 V, the gain of the voltage controlled amplifier is reduced by means of lead 1. This is done at the rate of about 20 dB for each 85 mV increase in the potential on lead 1. Thus, for a typical dynamic range of 40 dB, a total change of about 170 mV occurs in the potential on C store. This has to be produced by an increase in the peak value of the ac at lead 9. Since the nominal ac voltage at lead 9 is about 6 V peak, the change of 170 mV represents only a 3% change in the output for an input change of 40 dB. This represents unusually 'tight' AGC action.

Q10 to Q17 together with their emitter resistors provide appropriate biasing current sources. D3 has the function of precharging C store to just below the critical voltage of about 9.8 V. This is to avoid a long delay of charging C store from ground to 9.8 V via Rch when the system is hit with a tone burst after being idle. Rdch discharges C store and controls the recovery time of the system. This is normally set slow to avoid responding to echoes in telephone lines and is typically about 1/4 dB/mS. Attack settling time is determined partly by Rch and partly by the time taken to discharge the coupling capacitors by the dc voltage change occurring due to the control voltage feedthrough. For tone receiver

operation an attack settling time of about 5 ms can be achieved with this circuit. This is adequate since the minimum tone burst specified in the telephone system is 40 ms.

The advantage of the logarithmic unbalancing circuit over a linear one can be seen by considering some numerical values. With the circuit shown the amount of feedthrough applied via Q20 and D1 is +60 mV for each 10 dB of gain reduction and is smoothly applied over the whole gain range. If the unbalance were applied linearly, as for example by connecting the collector of Q22 to some portion of R11 instead of to the cathode of D1, the first 10 dB of gain reduction could be arranged to cause a 60 mV feedthrough, however, the second 10 dB reduction would only cause a feedthrough of about 20 mV, the third 6 mV and the fourth 2 mV. Clearly, the circuit would be nearly ineffective over some of the operating range and instability could occur under certain conditions.

To sum up, the careful control of the phase and magnitude of the control voltage feedthrough of the SIC voltage controlled amplifier enables a very tightly controlled audio AGC system with rapid attack and settling time to be realised without tendency for instability.

4.3 Theoretical ML1314K Circuit Characteristics

These are based on the assumption of a nominal power supply of 18 V and refer to the schematic of Fig. 4.2.

4.3.1 Maximum Input Level

The input stage is biased at a current of 0.5 mA. Degeneration resistors R1 and R2 of 6 K Ω each enable an input of voltage of 3 V peak to be accommodated before symmetrical clipping occurs.

4.3.2 Input Impedance

The single ended input impedance is mainly determined by RV1 since the input impedance to the bases of Q8, Q7 is approximately $\beta_n \cdot 12 \text{ K}\Omega = 1.2 \text{ M}\Omega$ minimum. (β_n refers to the gain of small geometry transistors on the SIC chip. A value of 100 is used in the above calculation.)

4.3.3 Maximum Gain of Voltage Controlled Amplifier

This occurs when lead 1 has a low voltage compared to lead 3 so that all the signal current is directed to Q6, Q7, which act as diode loads.

The dynamic impedance of an SIC diode is approximately 26Ω at 1 ma and is inversely proportional to current. The total biasing currents for Q6, Q7 are derived from Q10, Q13, Q14 and are $300 \mu\text{A}$ each under maximum gain conditions. Thus, the dynamic impedances of Q6 and Q7 are each 87Ω . The voltage gain of Q19 is approximately R_{L1} divided by the dynamic impedance of the base emitter diode of Q19. This calculates to be 180. The overall voltage gain from lead 13 to lead 10 is thus a maximum of:

$$A_v = \frac{87 \Omega}{12 \text{ K}\Omega} \cdot 180$$

$$= 1.3$$

4.2

The diode loads Q6, Q7 do not cause distortion since their non-linearity is exactly cancelled by the compensating non-linearity of the following stage, Q18, Q19.

4.3.4 Common Mode Range

The maximum common mode swing of the signal is determined on the negative side by the saturation of Q10 and on the positive side by the saturation of Q8 or Q9. The respective limits are: -1.7 V, +3.5 V

4.3.5 Current Drain

The currents drawn by each section are noted in Fig. 4.2. The total drain due the to AGC circuit is $6 \text{ mA} \pm 20\%$.

4.3.6 Control Voltage Law

When lead 1 becomes equal in voltage to lead 3, half the signal current is diverted from loads Q6, Q7. However, the gain is not reduced by 6 dB since a corresponding increase in the dynamic impedances of diode loads Q6, Q7 initially occurs due to $I_{e6,7}$ nearly halving. When only about 10% of the signal current is passing through Q3, Q4 (lead 1 = 9.87V) the bias current of Q6, Q7 begins to be dominated by current sinks Q13, Q14 and significant gain reduction then occurs at about 20 dB for each 80 mV increase in lead 1.

4.3.7 Pre-charge of C Store

The most desirable pre-charge for C store is just below 9.8 V so that the voltage controlled amplifier has maximum gain in the ideal condition but C store does not have to be charged very much to begin reducing the gain. This ensures a rapid attack time. In the SIC circuit, the pre-charge voltage is set at VD3 - 0.3 V below the reference voltage VA (the 0.3 V drop being due to the dc drop across R13). At its operating current of $1 \mu\text{A}$, VD3 has a room temperature drop of 0.5 V.

$$\text{Thus, } V_{\text{pre-charge offset}} = 0.2 \text{ V} \quad 4.3$$

4.3.8 AC Output Voltage

AGC is applied when the ac peak voltage at lead 9, superimposed on the dc bias of V_b , is sufficient to charge C store to V_a or above. (In fact to exactly V_a for 1 dB gain reduction and to $V_a + 170 \text{ mV}$ for 40 dB gain reduction.) Since there is a voltage drop due to the base-emitter diodes of Q25, Q26, and a charging drop due to R_{ch} , the output voltage is given by:

1 dB gain reduction:

$$V_{out}(-1) = V_{rch} + V_a + V_{be25} + V_{be26} - V_b \text{ a.c. volts peak} \quad 4.4$$

40 db gain reduction:

$$V_{out}(-40) = V_{out}(-1) + 0.17 \text{ a.c. volts peak} \quad 4.5$$

The effective operating current of Q26 is approximately $70 \mu\text{A}$ since it charges C store in short pulses and the discharge current is steady at about $1 \mu\text{A}$. This represents a steady state charging duty cycle of about 1.5% which is observed in practice. V_{rch} is thus 0.7 volts. The mean base-emitter voltage for small geometry transistors ($24 \times 24 \mu$ emitter) on the MIL QL12 process at 25°C is 0.65 V at $100 \mu\text{A}$, and varies at 60 mV/decade of current. The temperature coefficient is $-2.2 \text{ mV}/^\circ\text{C}$ at 1 mA , and increases $0.2 \text{ mV}/^\circ\text{C}$ for each decade reduction in current. The typical dc current gain is about 250. Using these figures, the calculated ac output voltages and temperature coefficients of voltage are as follows:

$$\begin{aligned}
 V_{out} (-1) &= 0.70 (+2\text{mV}/^{\circ}\text{C}) + 9.7(+0.4 \text{ mV}/^{\circ}\text{C}) + 0.500 (-3.0\text{mV}/^{\circ}\text{C}) \\
 &\quad + 0.630 (-2.5 \text{ mV}/^{\circ}\text{C}) - 5(-1 \text{ mV}/^{\circ}\text{C}) \\
 &= 6.53 (-4.1 \text{ mV}/^{\circ}\text{C})
 \end{aligned}
 \tag{4.6}$$

$$V_{out} (-40) = 6.70 (-4.1 \text{ mV}/^{\circ}\text{C}) \text{ peak ac volts} \tag{4.7}$$

Thus, the calculated temperature drift of the output voltage over the range 0 to $+60^{\circ}\text{C}$ amounts to a variation of $\pm 2\%$, which compares well with the measured results shown in Fig. 4.5, Section 4.5.3.

4.3.9 Attack Tone Burst Transient Response

The attack time for the AGC output to settle when a tone burst is applied to its input depends partly on the time to charge C store to its new value, and partly on the time taken to discharge the control voltage feedthrough (via CV4, RV7, RV2, Fig. 4.3). The worst case voltage change that C store must be charged through is the initial voltage offset of 0.2V required for reducing the amplifier gain by 46 dB, i.e. a total of 0.4V (Section 4.3.6).

The maximum slew rate of C store is when the detector output is limiting at approximately 0.5 mA and is 0.5V/msec. The worst case control voltage feedthrough at the output of the AGC SIC (lead 10) is +280 mV. Referred to the controlled output under maximum gain conditions of 180 times, this is an equivalent initial offset of 50 V. This is discharged via a single pole time constant of 0.13 mS involving CV4. Since the nominal controlled output voltage is 6.7 V peak, the offset must decay to about 0.7 V for the output to be within 10% of nominal. This represents about 5 time constants or 0.65 mS. Thus, the dominant factor in the

attack time is the charging of C store.

4.3.10 Recovery Time

The recovery time is the time taken for the system to increase its gain enough for the output to come back to the nominal level when the input suddenly drops in amplitude. As mentioned in section 3.3.3, this should not be too rapid to avoid responding to echoes occurring in some connections. The discharge current for C store is set at $1 \mu\text{a}$ by R_{dch} and gives a rate of IV/S . Since the gain changes by $1/4 \text{ dB per mV}$, the recovery rate is $1/4 \text{ dB/mS}$. This means that the system will not respond to echoes 10 dB down from the signal for a period of 40 mS from the end of the valid signal. This gives adequate protection.

4.4 External Amplifier

The amplifier shown as -K in Fig. 4.2 is realised with two 741 type operational amplifiers in series. (See Fig. 4.3.) Alternative realisations are possible, but this approach is economical since low quality 741's with high offset voltages can be used. It is important to avoid any tendency for ringing in the transient response of the amplifier since this would nullify the careful control of the phase of the control voltage feedthrough already described, and could cause instability.

It is desirable to have symmetrical clipping during transient overloads to the operational amplifiers so they are biased at approximately half the supply voltage which is derived from lead 3 of the SIC. The first operational amplifier provides a non-inverting voltage gain of 1 to 10 times for the two options of 26 or 46 dB dynamic range. The second operational amplifier provides a fixed inverting voltage gain of

40 times. Two of the coupling capacitors CV3 and CV5 involve large time constants while CV4 causes a roll off point at 1.3 kHz and is actually part of the pre-filter circuitry, so that the overall response does not roll off until about 670 Hz. (See section 7.3).

Fig. 4.3 Complete AGC circuit including external amplification

It is desirable to have as small a coupling time constant as possible in the AGC loop (consistent with having a relatively flat overall frequency response down to 670 Hz) so that the dc shift caused by an incoming tone burst, due to control voltage feedthrough, is discharged as rapidly as possible. This ensures a rapid settling time in the controlled output of the AGC circuit when responding to tone bursts. Thus, since a small

time constant pole is required for the pre-filter frequency response anyway, it is inserted within the control voltage loop to obtain a more rapid settling time than would otherwise be possible over the required frequency passband. For stability reasons it is desirable to make the low frequency roll off of the AGC loop approximately single pole so CV3 and CV5 are used with large time constants to make CV4 dominate the roll off.

The high level output from the AGC loop of 4.7 V rms is attenuated by a potential divider and buffered with an operational amplifier to provide a low level, low impedance output of about 135 mV rms suitable for driving the filter bank. A dc bias suitable for the filter bank of 5.0 V is derived from lead 11 of the AGC SIC.

4.5 Measured Performance of AGC Amplifier

4.5.1 Reproducibility

Table 4.A shows the results of measurements taken on 15 randomly selected ML1314K SIC, together with the calculated means and standard deviations. The mean rms ac output level of 4.7 V (6.65 V peak) corresponds well with that calculated in equation 4.9 (6.7 V peak).

4.5.2 Variations with Supply Voltage

Fig. 4.4 shows the variation with supply voltage of the two bias rails and the controlled ac output level. This is of interest both from the variation of the nominal 18 V supply voltage around its nominal, and for operation of 12 V for mobile applications. The bias rails derived in the ML1314K SIC are also used for biasing the filter block, and the pivoting amplifier which control the detector blocks.

Fig. 4.4 Effects of supply voltage variation on levels in AGC circuit

Unit #	Va	Vb	AC Output Volts RMS
1	9.74	5.04	4.70
2	9.70	4.96	4.74
3	9.72	5.05	4.67
4	9.71	5.00	4.71
5	9.74	5.04	4.70
6	9.71	4.99	4.72
7	9.66	4.94	4.72
8	9.70	5.01	4.69
9	9.69	4.96	4.73
10	9.72	5.02	4.70
11	9.67	4.92	4.75
12	9.72	5.04	4.68
13	9.70	4.98	4.72
14	9.72	5.02	4.70
15	9.77	5.04	4.73
Mean	9.71	5.00	4.71
Std. Dev.	.03	.04	.02

Vs = 18.04 Measurements at 25°C ambient temperature
Input signal: 0.5 V rms @ 1 kHz

Table 4A - Measurements on 15 randomly selected AGC microcircuits

4.5.3 Temperature Variation of Output Level

This is shown in Fig. 4.5 for an input of 1,000 Hz at 100 mV rms.

Fig. 4.5 Temperature variation of AGC amplifier output level

4.5.4 Distortion Levels with Continuous Sine Wave Input

Distortion due to the internal AGC amplifier and the external amplifier is less than 1%. However, due to the requirement for a rapid attack time and a reasonably rapid recovery time, the effective smoothing time constant is not large enough to eliminate ac feedback around the control loop. This results in frequency dependent harmonic distortion which is plotted in Fig. 4.6. The effects of changing the value of C store from its nominal value of 1 μ F are also shown.

Fig. 4.6 AGC amplifier output distortion vs. frequency

4.5.5 Tone Burst Response

Fig. 4.7 shows the output of the AGC amplifier responding to an input tone burst at 1000 Hz under various conditions. In Fig. 4.7a, a 40 mS input burst is shown, with the output settling in about 3 mS. In Fig. 4.7b, the input signal is periodically attenuated so that the recovery time of the AGC output signal can be seen.

1 kHz input burst
0.5 Volts per div.

Output or signal from
AGC amplifier
5 Volts per div.

Time base = 10 mS per div.

Fig. 4.7 a Tone burst response of AGC amplifier

Input:
1 kHz attenuated
periodically 20 dB
0.5 Volts per div.

Output of AGC amplifier
5 Volts per div.

Time base = 10 mS per div.

Fig. 4.7b Response of AGC amplifier to periodically attenuated input

Fig. 4.8 Regulation of AGC amplifier output voltage

4.5.6 Regulation

Fig. 4.8 shows the small variation in output level that occurs for input ranges of 26 dB and 46 dB corresponding to amplifier gains of 40 and 400 respectively. The measured variation of about 3% over 46 dB corresponds closely with the calculations of section 4.3.8.

4.6 Fabrication

The part of the circuit within the dotted lines in Fig. 4.2 is fabricated as an SIC in a 14 lead dual-in-line plastic package with lead numbers as shown. The external amplifier and a few of the other external resistors could have been included on the SIC but were not in this case for reasons of flexibility and time. Because of the need to get the complete tone receiver system working on schedule, the SIC design was simplified as much as possible within the basic requirements. This enabled the SIC layout to be done before the complete system operating parameters were finalised, thus cutting the total lead time involved and reducing the probability of errors. In fact, the first SIC layout was successful. A photograph of the die is shown in Fig. 4.9. The die size is 70 x 65 mil. The circuitry is non-critical and any standard operational amplifier type process is suitable such as the Microsystems International Limited QL3 or QL12 processes.

Fig. 4.9 MLL314K AGC Integrated Circuit Die (70 x 65 mil.)

Chapter V

THE EIGHT BANDPASS FILTER MODULE

5.0 Introduction

The new tone receiver system eliminates the band separation filtering required by BTL receivers. The highpass pre-filter required for elimination of dial tone and power line noise has non-critical performance and is realised using simple voltage follower discrete component Sallen and Key sections (section 7.3). The only critical filters in the tone receiver are the eight bandpass filters required for detection of the eight signal frequencies. It is desirable for these to have Q's of 15 with $\pm 5\%$ overall accuracy and centre frequencies of $\pm 1\%$ overall accuracy, including initial offset, temperature, and aging effects. These requirements call for careful filter design and component selection. An approach taken recently by Bell Northern Research Laboratories in the new receiver for the Northern Electric 'Pulse' electronic PBX uses nichrome thin film hybrid gyrators with discrete frequency determining components. (5.1, 5.2) Thus, for the eight bandpass filters, eight hybrid modules, sixteen discrete capacitors, and eight potentiometers are required.

For the ME8800 receiver, a single tantalum thin film hybrid module containing all the eight filters was designed which represents significant size and cost reductions over previous filter approaches of comparable performance. This was achieved by minimising the filter and amplifier complexity without sacrificing performance and by optimising the filter configuration with respect to tantalum thin film resistors and capacitor characteristics.

Also, a fully automatic trimming method was designed which enables precise setting of centre frequency and Q in only six seconds per section.

This scheme uses the standard laser trimming equipment available at MIL, together with a simple external circuit which causes the filter to oscillate and enables it to be trimmed using a period counter to monitor the centre frequency.

5.1 Choice of Filter Type

To minimise cost, the simplest possible filter structure was desired consistent with the required performance levels. The most compact filter configurations for hybrid microcircuit realisation are the Sallen and Key positive feedback class.^(5.3) This is because, even for high Q values, the component ratios can be small and thus layout area can be minimised. It is also possible to have independent adjustment of centre frequency and Q which simplifies trimming operations considerably, since it eliminates the need for iteration.

This class of filter has been avoided in the past, except for very low Q applications, because of its high Q sensitivity to both closed loop amplifier gain K and passive component ratios. The Q sensitivity refers to the fractional change in Q caused by a unit fractional change in the parameter of interest. Thus, the sensitivity of Q to amplifier gain K is:

$$S_K^Q = \partial Q / \partial K \cdot K / Q \quad 5.1$$

Because of their excellent ratio tracking properties (better than 5ppm/°C), the use of tantalum thin film resistor and capacitor substrates makes the Q sensitivity to passive component ratios unimportant for the Q = 15 region of interest. For this class of filters, typically $S_Z^Q \cong 3Q = 45$ for Q = 15, which is a high value. However, due to the good tracking mentioned,

total Q drift over 0 to 60°C caused by passive component variations, will be smaller than $\pm 0.5\%$.

The high sensitivity of Q to the amplifier gain K has also been considered to be a disadvantage of the positive feedback class of filters. However, Moschytz^(5.5) has recently shown that the 'Gain Sensitivity Product' of a filter is a more suitable parameter for judging Q stability than the Q sensitivity to gain alone. This follows similar reasoning to that of Saraga^(5.6) namely, that the higher the gain value required, the more unstable it will be for a given open loop gain. Moschytz shows that positive feedback structures can have lower GSP's than the more generally favoured negative feedback structures. This is the deciding argument in favour of the positive feedback Sallen and Key class of filters for this application since, as already mentioned, they also can have low component ratios and independent adjustment of f_0 and Q . The Moschytz GSP approach was extended to derive the optimum component ratios for minimising GSP for the six positive feedback filters shown in Fig 5.1. These all have the minimum component count of two frequency determining capacitors and resistors and none have true bandpass numerators. In this application the slight skewing of the response caused by a non-ideal numerator is unimportant and the extra component required for a bandpass numerator is not justified since it would double the resistor layout area. Two examples of true bandpass positive feedback filters are shown in Fig. 5.2. Since the additional resistor required (R_3 in each case), is effectively in parallel with R_2 , it not only adds layout area for itself but increases the values required for other components.

Fig.5.1 Six positive gain, minimum complexity filter sections

5.2 Gain Sensitivity Optimisation - Approximate Analysis

The voltage transfer functions of the six circuits analysed are shown in Fig. 5.1. The objective is to find the optimum ratios for minimising GSP for each case to evaluate their relative desirability. They all have the form (normalised to give $\omega_0 = 1$ Hz by putting $S1 \cdot S2 \cdot G1 \cdot G2 = 1$):

$$V2/V1 = N (p^2 + p/Q + 1)^{-1} \quad 5.2$$

In each case Q has the form:

$$Q = (x - Ky)^{-1} \quad 5.3$$

where x is the sum of the non K terms, and y is the sum of the terms in K.

Filter Type A

$$\frac{V2}{V1} = \frac{p \cdot K \cdot S1 \cdot G3}{p^2 + p(S1(G1 + G2 + G3) + S2 \cdot G1 - K \cdot S1 \cdot G2) + S1 \cdot S2 \cdot G1(G2 + G3)}$$

Filter Type B

$$\frac{V2}{V1} = \frac{p \cdot K \cdot S1 \cdot G3}{p^2 + p(S1(G1 + G2 + G3) + S2(G2 + G3) - K \cdot S1 \cdot G2) + S1 \cdot S2 \cdot G1(G2 + G3)}$$

Fig. 5.2 Two true bandpass filter sections

The sensitivity of Q to amplifier gain K is defined as:

$$S_K^Q = K/Q \cdot \partial Q / \partial K \quad 5.4$$

Combining 5.4, 5.3:-

$$S_K^Q = K \cdot y \cdot Q \quad 5.5.$$

Now an approximation can be made which is valid for high Q:-

$$K = y^{-1} (x - Q^{-1}) \approx x \cdot y^{-1} \quad 5.6$$

and it follows that:

$$S_K^Q \approx x \cdot Q \quad 5.7$$

and the Gain Sensitivity Product becomes:-

$$GSP = K \cdot S_K^Q \quad 5.8$$

$$GSP \approx x^2 \cdot y^{-1} \cdot Q \quad 5.9$$

This parameter can now be minimised with respect to the component ratios for the six filter cases:

5.2.1 Filters Types 1,2

These have the same demonimators, so for both,

$$x = S1 \cdot (G1 + G2) + S2 \cdot G1 \quad 5.10$$

$$= a \cdot b^{-1} + a \cdot b + a^{-1} b^{-1} \quad 5.11$$

$$y = S1 \cdot G2 \quad 5.12$$

$$= a \cdot b \quad 5.13$$

$$GSP = x^2 \cdot Q \cdot y^{-1} = (a^2 \cdot b^2 + a^2 + 1)^2 \cdot a^{-3} \cdot b^{-3} \cdot Q \quad 5.14$$

Setting $\partial GSP / \partial a = \partial GSP / \partial b = 0$ does not converge directly with this equation. However, inspection of the asymptotes reveals that a limit is reached when:

$$a \rightarrow 0, \quad b \rightarrow \infty, \quad a \cdot b = \text{constant}, \quad d \quad 5.15$$

$$\text{Then: } x^2 \cdot y = (d^2 + 1)^2 \cdot d^{-3} \quad 5.16$$

$$\text{Setting } \partial GSP / \partial d = 0 \text{ gives } d = \sqrt[3]{3} \quad 5.17$$

$$\text{and } \hat{GSP} = 16 / (3 \cdot \sqrt[3]{3}) \cdot Q = 3.08Q \quad (\text{IDEAL}) \quad 5.18$$

This is the best that can be achieved but requires a near infinite capacitor ratio, a^{-2} , and a near infinite resistor ratio, b^2 , with the resistor ratio being $d^2 = 3$ times the capacitor ratio. This is an impractical solution but an interesting one since equation 5.17 gives us a measure of how good the GSP can be so we can see how far practical solutions are from the ideal.

Constraints on the component ratios will now be imposed according to desirable hybrid microcircuit layout considerations.

One desirable feature is to have equal capacitor values to give good tracking and minimum total capacitance.

Assuming capacitors equal:

$$a = 1 \quad 5.19$$

$$x^2 y^{-2} = (b^2 + 2)^2 \cdot b^{-3} \quad 5.20$$

$$\partial \text{GSP} / \partial b = 0 \text{ gives } b = \sqrt{6} \quad 5.21$$

$$\text{and GSP} = 64 / (6\sqrt{6}) = 4.35Q \quad 5.22$$

Thus, the best GSP value, if capacitors are equal, is achieved with a resistor ratio 6:1.

If a further constraint is placed on the resistor ratio for the same reasoning used for the capacitors, but a value of $b^2 = 2$ is allowed instead of 1, we get:

$$\hat{\text{GSP}} = 4\sqrt{2}Q = 5.65Q \quad 5.23$$

(capacitors 1:1 , resistors 2:1)

5.2.2 Filter Types 3,4

Performing a similar analysis gives:

$$x = S1.G2 + S2 (G1 + G2) \quad 5.24$$

$$= a.b + a^{-1} \cdot b^{-1} + b.a^{-1} \quad 5.25$$

$$y = S1 \cdot G2 \quad 5.26$$

$$= a.b \quad 5.27$$

$$x^2 \cdot y^{-1} = (a^2 \cdot b^2 + b^2 + 1)^2 \cdot a^{-3} \cdot b^{-3} \quad 5.28$$

$$\text{This leads to } \hat{\hat{\text{GSP}}} = 3.08Q \quad (\text{IDEAL}) \quad 5.29$$

i.e. the same as 5.29 however, $a \rightarrow \infty$, $b \rightarrow 0$,

$$d = 1/\sqrt{3}$$

i.e. the capacitor ratio should be 3 times the resistor ratio.

If capacitors equal ($a = 1$) the optimum value for b is 1.5

giving $\hat{GSP} = 8.72Q$ 5.30

Note that this is worse than is achieved by filters 1,2 with practical values (Equation 5.23).

5.2.3 Filter Types 5, 6

Performing a similar analysis gives:

$$x = S1 \cdot (G1 + G2) + S2 \cdot G2 \quad 5.31$$

$$= a \cdot b + a \cdot b^{-1} + b \cdot a^{-1} \quad 5.32$$

$$y = S1 \cdot G2 \quad 5.33$$

$$= a \cdot b \quad 5.34$$

$$x^2 \cdot y^{-1} = (a^2 \cdot b^2 + a^2 + b^2)^2 \cdot a^{-3} \cdot b^{-3} \quad 5.35$$

In this case, equating the partial derivatives to zero yields a solution:

$$\partial GSP / \partial a = 0 \text{ gives } a^2 = 3b^2 \cdot (1 + b^2)^{-1} \quad 5.36$$

$$\partial GSP / \partial b = 0 \text{ gives } b^2 = 3a^2 \cdot (1 + a^2)^{-1} \quad 5.37$$

$$\text{hence } a = b = \sqrt{2} \quad 5.38$$

and $\hat{GSP} = 8Q$ and occurs with capacitor and resistor 5.39
ratios of 2:1.

Thus, the theoretically best possible solution with this format is not so good as the practical solution of 5.23.

5.2.4 Summary of Approximate GSP Optimisation

Table 5A summarises the results of the analysis for optimising GSP for the six filters obtained using the approximation of Equation 5.6.

5.3 GSP Optimisation - Exact Computed Results

The approximation used in Equation 5.6 leads to very simple solutions for GSP optimisation. To verify the validity of the approximation used, which assumes $1/Q \rightarrow 0$, the exact values for GSP were numerically solved for the three different types of denominator, and parametric curves obtained for various values of component ratios a^2 and b^2 . A Q of 15 was assumed for each case as this is the value required in the receiver application.

The exact equation for GSP is found from equations 5.3, 5.5 and is:

$$\text{GSP} = K^2 \cdot y \cdot Q \quad 5.40$$

$$\text{GSP} = y^{-1} \cdot (x - Q^{-1}) \cdot Q^2$$

$$\text{For filters 1, 2} \quad x = a \cdot b^{-1} + a \cdot b + a^{-1} \cdot b^{-1}, \quad y = a \cdot b \quad 5.41$$

$$3, 4 \quad x = b \cdot a^{-1} + a \cdot b + a^{-1} \cdot b^{-1}, \quad y = a \cdot b \quad 5.42$$

$$5, 6 \quad x = b \cdot a^{-1} + a \cdot b + a \cdot b^{-1}, \quad y = a \cdot b \quad 5.43$$

FILTER #	IDEAL VALUES FOR MINIMUM GSP	BEST PRACTICAL VALUES TO MINIMISE GSP (Capacitors Equal; $a^2 = 1$, Resistor ratio $\leq 2:1$, $b^2 \leq 2$, $b^{-2} \leq 2$)	
1, 2	$a \rightarrow 0, b \rightarrow \infty, a.b. = \sqrt{3}$ $K \approx 1.33$ $\hat{\hat{GSP}} = 3.08Q$	$a = 1$ $b = \sqrt{2}$ $K \approx 2$ $\hat{GSP} = 5.65Q$	
3, 4	$a \rightarrow \infty, b \rightarrow 0, a.b. = 1/\sqrt{3}$ $K \approx 1.33$ $\hat{\hat{GSP}} = 3.08Q$	$a = 1$ $b = \sqrt{1.5}$ $K \approx 2.66$ $\hat{GSP} = 8.72Q$	
5, 6	$a = \sqrt{2}$ $b = \sqrt{2}$ $K \approx 2$ $\hat{\hat{GSP}} = 8Q$	$a = 1$ $b = \sqrt{1.5}$ $K \approx 2.66$ $\hat{GSP} = 8.72Q$	

Table 5A Summary of Results of Approximate Analysis of GSP Optimisation

The results are summarised in Fig. 5.3, 5.4, 5.5. The Univac Fortran V program used for the analysis and the output data are listed in Appendix A, Table A1. The exact computed curves confirm the accuracy of the approximation used in the analysis in section 5.2. For example, in Fig. 5.3, for filters 1, 2 with equal capacitors ($a = 1$) the best value for b is 2.5 and gives a GSP of 62. The values predicted by the approximate analysis are 2.45 and 65 respectively (see equations 5.21, 5.22). Similarly, the overall optimum GSP tends to a value of around 43 and occurs with small a and large b and $a.b = 1.84$. The values predicted by approximate analysis are 46 and 1.73 respectively (see equations 5.17, 5.18). Figures 5.4, 5.5 for filters 3 to 6 also provide good correlation with the values predicted by the approximate analysis.

5.4 Chosen Filter Configuration

From the preceeding analysis it is evident that configurations 5 and 6, which have a minimum GSP of $8Q$, will have inferior Q stability performance, and should be avoided. Configurations 1 to 4 can all achieve GSP's approaching $3.08Q$ in the limit with unconstrained component ratios, but with the desirable constraints of equal capacitors and resistor ratios maximum 2:1, filters 1 and 2 are superior giving GSP's of $5.65Q$. From these two, filter 1 was selected because its numerator is closer to that of the ideal bandpass response.

Fig. 5.3 Filters 1,2 Exact GSP vs. component ratios (a^2 = capacitor ratio, b^2 = resistor ratio)

Fig. 5.4 Filters 3,4 Exact GSP vs. component ratios (a^2 = capacitor ratio, b^2 = resistor ratio)

Fig. 5.5 Filters 5,6; Exact GSP vs. component ratios. (a^2 = capacitor ratio, b^2 = resistor ratio)

697 Hz C = 5.710 nf
 770 Hz C = 5.168 nf
 852 Hz C = 4.673 nf
 941 Hz C = 4.228 nf

1209 Hz C = 3.290 nf
 1336 Hz C = 2.978 nf
 1477 Hz C = 2.693 nf
 1633 Hz C = 2.436 nf

Fig. 5.6 (a) Filter configuration selected for use in the new receiver

Fig. 5.6 (b) Normalised frequency response of bandpass filter section

5.4.1 Design Values of Selected Filter

Figure 5.6 (a) shows the selected filter configuration in which:

$$\text{resistor ratio } b^2 = 2$$

$$\text{capacitor ratio } a^2 = 1$$

The scaled values of impedance shown in Fig. 5.6 (a) were chosen for minimal layout area based on the nominal design parameter of $0.419 \mu F / in^2$ and $100 \Omega / \text{square}$ for the C and R substrates.

From 5.10, 5.11

$$x = 2\sqrt{2} \quad 5.44$$

$$y = \sqrt{2} \quad 5.45$$

The desired Q of 15 can be realised with a gain value K calculated from 5.6.

$$Q = (2\sqrt{2} - K\sqrt{2})^{-1} \quad 5.46$$

$$K = 2 - (\sqrt{2} Q)^{-1} = 1.9528 \text{ for a Q of 15} \quad 5.47$$

The sensitivity of the Q to the gain is given by 5.5

$$S_K^Q = 1.9528 \cdot \sqrt{2} \cdot 15 = \underline{41.5} \quad 5.48$$

Thus, for $\pm 3\%$ Q variation, the gain K must be held constant to within $\pm .07\%$. Assuming a temperature range of 0 to $60^\circ C$, this requires resistor tracking to within $40 \text{ ppm}/^\circ C$ and capacitor tracking within $80 \text{ ppm}/^\circ C$.

Both of these requirements are easily met with tantalum thin film resistors and capacitors.

The sensitivity of the Q to each of the passive components is summarised in Fig. 5.6 (a).

The normalised frequency response of the bandpass filter section is shown in Fig. 5.6 (b).

5.5 Filter Amplifier SIC - ML1302

Eight bandpass filters are required for the tone receiver. As described in section 5.3, a simple active filter section requiring a positive voltage gain of about two was selected and optimised for realising this application. The simplicity of this circuitry enables the passive components to be contained in a single low cost hybrid module. A simplified amplifier design was also devised to minimise unit amplifier cost and simplify assembly by incorporating the eight amplifiers required on a single die measuring 85 x 75 mil. Figure 5.7 shows the amplifier schematic which consists of 4 transistors connected in a feedback loop with the voltage gain defined by external thin film resistors Rx, Ry, so that

$$K = 1 + R_x/R_y \quad 5.49$$

For maximum voltage swing, the output is biased at one half the supply voltage of 18 V. Thus, lead 3 is biased at 9 V, lead 2 at 4.5 V, and lead 1 at 5.0 V. The input signal thus requires a 5.0 V dc bias with the ac signal superimposed on it.

The amplifier draws a total current of 1.5 mA. Ra and Rb set the operating currents of Q2, Q3 at 15 and 30 μ A. C1 increases the closed loop stability margin.

Fig. 5.7 Filter amplifier schematic

5.5.1 Computer Aided Analysis of the Filter Amplifier

The SLIC* circuit analysis program^(5.7) was used to analyse the performance of the amplifier and to verify its stability over the temperature range of operation. Fixed values of $R_x = R_y = 3k\Omega$ were used, representing a nominal voltage gain of 2. Figure 5.8 shows the computed gain error vs. temperature for the maximum range of interest -45°C to $+65^\circ\text{C}$. The error of approximately 0.2% represents an effective open loop voltage gain for the amplifier of about 1,000. Figure 5.9 shows the computed amplifier frequency response for magnitude and phase and Table 5B shows the computed pole -- zero locations for room temperature at the two temperature extremes. These results show the amplifier to have stable gain and adequate bandwidth for audio active filter use. The slightly underdamped roll off at 1.5 MHz has a maximum Q of 2 and has adequate stability margin. The input data used for the analysis is shown in Appendix C, Table C1. Parameters used for

* The cooperation of Dr. D.O. Pederson, Professor of Electrical Engineering University of California, Berkley, in the provision of the SLIC program to Microsystems International Limited, is gratefully acknowledged.

Fig. 5.8 Computed gain error of filter amplifier
vs. temperature: $R_x = R_y = 3 \text{ K}\Omega$

Fig. 5.9 Computed amplifier frequency response: $R_x = R_y = 3 \text{ K}\Omega$

POLES		ZEROS	
REAL	IMAG	REAL	IMAG
-3.09982+06	-1.09443+07	5.57859+07	0.00000
-3.09982+06	1.09443+07	-2.37667+08	2.15753+08
-3.06846+07	0.00000	-2.37667+08	2.15753+08
-4.21444+08	0.00000	2.96055+08	0.00000
-2.18067+09	0.00000	-4.93890+09	0.00000
-1.47608+09	0.00000	-4.70598+09	0.00000
-6.47397+09	0.00000	-2.03271+09	0.00000
-5.94826+09	0.00000	-1.17203+10	0.00000
-4.89001+09	0.00000	-1.37373+10	0.00000

-45°C

POLES		ZEROS	
REAL	IMAG	REAL	IMAG
-2.18475+06	-8.11978+06	3.72555+07	0.00000
-2.18475+06	8.11978+06	-1.53575+08	-1.52213+08
-2.17452+07	0.00000	-1.53575+08	1.52213+08
-2.52756+08	0.00000	1.92395+08	0.00000
-1.98878+09	0.00000	-4.65667+09	-2.64387+08
-1.43472+09	0.00000	-4.65667+09	2.64387+08
-5.15674+09	0.00000	-1.82329+09	0.00000
-6.26370+09	0.00000	-1.13977+10	-5.47883+08
-3.78767+09	0.00000	-1.13977+10	5.47883+08

30°C

POLES		ZEROS	
REAL	IMAG	REAL	IMAG
-1.92229+06	-6.91164+06	2.98832+07	0.00000
-1.92229+06	6.91164+06	-1.26551+08	-1.29957+08
-1.87873+07	0.00000	-1.26551+08	1.29957+08
-2.01887+08	0.00000	1.55917+08	0.00000
-1.90023+09	0.00000	-4.69915+09	-3.09873+08
-1.44903+09	0.00000	-4.69915+09	3.09873+08
-5.01190+09	0.00000	-1.77441+09	0.00000
-6.37797+09	0.00000	-1.10907+10	-5.11269+08
-3.39514+09	0.00000	-1.10907+10	5.11269+08

65°C

Table 5B - Amplifier transfer function vs. temperature

modelling the transistors are listed in Appendix C, Table C2, and are typical values for the MIL QL7 operational amplifier process.

5.6 Computer Simulation of Complete Filter Amplifier Units

To simulate the effects on the filter performance due to temperature variations of the amplifier, a SLIC program was written including both filter and amplifier components. The filter components were assigned zero temperature coefficient so that the effects of the amplifier alone could be evaluated. The configuration selected for the ME8802 filter module was analysed (type 1) and, for comparison, an optimised type 5 filter was also analysed using identical amplifier parameters. These have GSP of 5.65Q and 8Q respectively. Both filters were set up for a Q of 15 at 30°C. The resulting temperature drift of Q caused by amplifier temperature variations from -45°C to +65°C is shown in Fig. 5.10 for the two cases. The total Q drifts were 2.7% and 3.9% respectively. The ratio of the GSP's corresponds closely with the ratio of Q drift observed -- 1.42 vs. 1.45, thus showing that GSP is an accurate quality criterion for comparing different filter configurations. The frequency pulling effect of the amplifier was negligible in both cases. The SLIC programs used for this analysis are listed in Appendix C, Table C1. The nodal circuits used for the analysis are shown in Fig. C1, C2.

Fig. 5.10 Predicted Q drift due to amplifier for filters 1,5

5.7 Fabrication

5.7.1 Process Outline

The ME8802 bandpass filter module is fabricated using standard tantalum thin film production processes at MIL which were developed for a similar product, the ME8900 tone generator.^(5.8) A ceramic resistor substrate is sputtered with oxygen doped tantalum to give $115 \pm 30 \Omega$ per square, at a nominal temperature coefficient of resistance of $-200 \text{ ppm}/^\circ\text{C}$. After photo etching the resistors are laser trimmed to within 1% of their nominal value. The 32 accurate resistors required for the eight filters are all on the substrate which measures $1.2'' \times 0.94''$. The total resistance is $720\text{K}\Omega$. Two glass capacitor substrates are fabricated to provide the eight filter capacitors required for each group. The capacitors are formed by anodising undoped tantalum, sputtered to a sheet resistance of 8Ω per square. The 200V anodising potential results in a sheet capacitance of $0.419 \text{ microfarads per square inch}$. Gold upper electrodes are evaporated on and photo etched to give an overall capacitor accuracy of $\pm 5\%$. The

capacitors have a nominal temperature coefficient of $+200 \text{ ppm}/^{\circ}\text{C}$, giving a nominally zero time constant temperature coefficient when combined with the resistors. The low frequency group capacitor substrate is $0.33'' \times 0.7''$ and has a total capacitance of 19.8 nanofarads. The high frequency group capacitor substrate is $0.26'' \times 0.7''$ and has a total capacitance of 11.4 nanofarads.

The two capacitor substrates are glued on top of the resistor substrate with RTV adhesive and gold wire bonds used for electrical interconnection. The eight amplifier SIC die, measuring $85 \times 75 \text{ mil}$, is pulse bonded to the resistor substrate and the pads interconnected via aluminum wire bonds. At this point the complete unit is functionally trimmed to give accurate centre frequencies and Q's by laser adjustment of appropriate resistors and using the measurement techniques described in section 5.7.2. After trimming, external leads are attached by thermo-compression bonding and the unit is potted with epoxy in a rubber mould. Figure 5.11 shows the three passive substrates, Fig. 5.12 shows the SIC die and Fig. 5.13 shows the complete assembly. Figure 5.14 shows the complete module schematic.

5.7.2 Automatic Trimming Method for Frequency and Q Adjustment

Functional laser trimming of the completed filter assembly enables automatic adjustment of centre frequency and Q to be achieved. However, conventional techniques of monitoring filter centre frequencies tend to be slow and/or inaccurate (e.g. sweep generators and phase detectors). To eliminate this problem, a simple external circuit was designed which is probed into the filter circuit to make it oscillate at a frequency very

Fig. 5.11 Substrates for the filter module

Fig. 5.13 Complete filter module assembly (14B802)

Fig. 5.13 Complete filter module assembly (MH8802)

Fig. 5.14 Schematic of eight bandpass filter module (MH 8802)

close to its centre frequency (see Fig. 5.15). A period counter is then used to monitor the frequency while R_1 is adjusted upwards until the desired value is reached. After this, the external circuit is removed and the Q is trimmed by increasing the amplifier gain by increasing R_x until the gain at the centre frequency is the desired value. At this point a simple ac level detector stops the trimming. The desired centre frequency gain can be calculated from equation (a), Fig. 5.1:

$$V_2/V_1 = (K \cdot j \cdot \sqrt{2} + 1) Q/j = KQ (\sqrt{2} - j) \quad 5.50$$

$$= 50.6 @ -35.3^\circ \text{ phaseshift for } Q = 15 \quad 5.51$$

The principle of operation of the oscillation producing circuit is as follows. The amplifier gain K is equal to $1 + R_x/R_y$. As equation 5.46 shows, the Q value tends to infinity when $K = 2$. This means that oscillation will occur if K is slightly above 2. R_a in the external circuit is placed in parallel with R_y , thus increasing the gain enough to cause oscillation. As with all oscillators, the amplitude of oscillation

Fig. 5.15 External circuit for making filter oscillate

increases until non-linearity in the gain is reached which limits the amplitude by decreasing the effective loop gain to unity. A symmetrical non-linear resistive load in the form of a Silicon Carbide (SiC) varistor limits the amplitude. The varistor is capacitively coupled across R_x . Since the resonant frequency of this class of filter is independent of Q value, the undistorted oscillating frequency will be the exact centre frequency of the filter. However, undistorted oscillation requires that the loop gain be exactly unity over the full amplitude of oscillation. For this case the varistor would be inoperative and R_a would have to be precisely set to just the right value which would vary due to component tolerances. The value required would also vary during the frequency trimming cycle due to the change in Q caused by trimming only one resistor. This situation is thus impractical since it would require continuous manual adjustment of a potentiometer during trimming. For this reason it is desirable to set R_a to a value which will guarantee oscillation over the range of component tolerances. The varistor then limits the oscillation to a varying degree depending on the initial amount of excess loop gain. The non-linear limiting process causes a certain amount of frequency pulling. It is desirable that this be as low and as uniform as possible so that precise centre frequency trimming can be achieved. In this respect the SiC varistor was found to have superior characteristics than other limiting devices such as diodes or zeners in that it causes smaller and more uniform frequency pulling.

SiC varistors generally have the characteristic $I = \alpha V^n$ when n is typically 3. A suitable varistor for this application has $\alpha = 10^{-5} A/V^3$, $n = 3$, which draws 10 mA at 10 V, e.g. Carborundum type No. 328-BNR-44.

Figure 5.16 shows the oscillating frequency and amplitude vs. value of R_a . As can be seen, the frequency is very stable between the start of oscillation and the onset of clipping. The oscillating frequency has a nearly fixed offset of 0.57% below the true centre frequency of the filter. If this is allowed for in trimming the filter, the filter centre frequency accuracy can be 0.02%. To make the measurements, R_x was set at 2.87K and R_y at 3K. Since R_a is in parallel with R_y , the amplifier gain range between start of oscillation and onset of clipping is about 2.1 to 2.25 which is a broad enough range to allow for component tolerances within the filter. The oscillating amplitude should be centred on about 7 V peak to peak to give optimum tolerance range.

Fig. 5.16 Filter oscillation frequency offset and amplitude vs. R_a

5.7.3 Special Layout Considerations

Standard layout geometries are used for all the resistors except for R_x which is the amplifier gain setting resistor used to trim the Q . Because of the high sensitivity of the Q to the value of this resistor, it must be set with very fine resolution. This is difficult to do with normal trimming geometries due to the speed of the trimming laser beam and temporary local thermal effects. To get around this difficulty a coarse and fine trim geometry is used, as shown in Fig. 5.17. The first cut increases resistance at the rate of about $100\ \Omega$ per mil of laser cut, and is used until a Q of 8 is reached. The second cut increases resistance at the rate of about $2\ \Omega$ per mil of laser cut and is used to trim the Q to its final value of 15 with good resolution.

Fig. 5.17 Layout geometry for R_x for good Q trimming resolution.

5.8 Measured Performance

5.8.1 Temperature Effects of Amplifier Alone

A breadboard of the eight bandpass filters was constructed, with centre frequencies and Q appropriately adjusted. The SIC was then cycled over a temperature range without affecting the filter components so that the pulling effect due to the amplifier alone could be measured. Frequency shift was negligible (resolution of measurement $\pm 0.1\%$). The effect on the Q is shown in Fig. 5.18 along with the curve predicted in section 5.6, which gives reasonably good correlation. The curve shown is typical of several SIC's measured. Over the normal receiver range of operation of 0° to 60°C , the overall Q drift due to the amplifier alone is less than $\pm 1.5\%$.

Fig. 5.18 Measured Q drift with temperature due to amplifier alone
(Q = 15)

5.8.2 Overall Filter Drift vs. Temperature

A complete ME8802 filter module was temperature cycled and the centre frequency and Q measured. Figure 5.19 shows the overall centre frequency drift and Fig. 5.20 shows the overall Q drift. Over the normal range of operation the measured Q drift is about $\pm 2\%$. These performance values are quite acceptable for stable receiver operation. The curves show the average drift of several sections measured.

Fig. 5.19 Measured overall filter centre frequency drift vs. temperature

Fig. 5.20 Measured overall filter Q drift vs. temperature
(Q = 15)

Chapter VI

DETECTOR/TRIGGER UNITS - MH8802

6.0 Introduction

Sections 4 and 5 described how the AGC circuit and the bandpass filters were designed with particular regard to ensuring stability of the output level and Q versus temperature. This approach was also used in the design of the detector/trigger circuitry which has excellent threshold stability versus temperature. The combination of the three stabilised circuits ensures that the signal recognition bandwidth of the complete receiver remains nearly constant with temperature.

6.1 Circuit Requirements

Eight detector/trigger circuits are required for the receiver, one for each bandpass filter output. The basic function of each circuit is to provide a logic voltage output when the ac output of the bandpass filter driving it is above a threshold voltage for 40 ms or more. Secondary functions required for operating the pivoting threshold control system (see section 3.3.2) are an externally controlled threshold voltage, and an output suitable for indicating the highest level present in the frequency group.

Fig. 6.1 Detector/trigger schematic

6.2 Description of Circuit Operation

6.2.1 Detector Circuit

Figure 6.1 is a schematic of one of the detector/trigger circuits. Input capacitor C1 couples the output of one of the filters to node 1 which is biased at a quiescent voltage of V_d (where V_d = the approximate voltage drop of a SIC base-emitter junction ≈ 0.65 V). This is derived from node 2 (biased at $2 V_d$) by D1, R1. D1 allows node 1 to swing positively but clamps it from going below V_d . With an ac input signal, a dc bias shift equal to V_p (where V_p = the peak ac voltage) builds up at node 1 within the first cycle. This is due to the unequal charging and discharging paths for C1 at node 1. As well as the dc bias shift, the full ac signal also appears at node 1. This ac component is removed by lowpass filter R2, C2 to provide a dc voltage equal to $V_p + V_d$ at node 3. (The residual ac ripple = 0.1 V peak.) This is buffered by voltage follower Q2 to give a low impedance dc detector output, $V_o = V_p$. A second emitter on Q2^{6.1} is joined to similar points in the three other detectors of the group to provide a voltage representing the group signal level. (This is used for operating the receiver threshold voltage pivoting circuitry) (See section 7.4). Due to diode OR type action, this voltage follows the highest detector output of the group without being loaded by the others.

6.2.2 Trigger Circuit

The trigger circuit involves transistors Q3 to Q8 in Fig. 6.1. The quiescent state is with Q6, Q7 and Q8 ON, and Q3, Q4, Q5 OFF. In this condition the base of Q6 is within a few mV of V_t , since only the base currents of Q6 and Q7 are flowing through R5. Thus, node 4 is held at

$V_t - V_d$. The value of V_t is controlled by the pivoting circuitry in the receiver, but has a nominal value of 5 V. (See section 7.4.) Q3, Q4, Q5 are series connected so are all OFF until Q3 turns ON. Q3 draws no current until the base of Q3 reaches the same potential as the base of Q6. This requires the detector output V_p to be near V_t . When node 3 approaches this voltage, Q3 begins to share with Q6 the sink current flowing in R3. The collector current flowing in Q3 causes an equal collector current in Q4 (which is a unity gain current reflector, a well known SIC lateral PNP device,^{6.1}). When this reaches 100 μA , it produces 0.5 V at the base of Q5, thus beginning to turn it ON. At this point regenerative feedback occurs and the trigger snaps since, as Q5 turn ON, it reduces the voltage on the base of Q6, thus allowing Q3 to draw more of the sink current of R3 and turning Q5 ON harder, etc. Thus, Q5 rapidly saturates which results in a turn OFF threshold voltage reduced by resistor divider R5, R6 and giving 30% hysteresis. This makes the trigger threshold very sharp and prevents any tendency for the output voltage to toggle up and down in response to the residual ac ripple at node 3.

The trigger part of the circuit bears some resemblance to a configuration independently derived by M. Callahan.^{6.2} However, his circuitry has severe temperature problems due to five cascaded transistors with no provision for leakage degeneration.

6.3 Calculated Performance Characteristics

6.3.1 Temperature Effects on the AC Input Voltage at which Triggering Occurs

With the simplifying approximation of V_d a constant, and not a function of temperature, Section 6.2 shows that triggering occurs with an $\approx$ peak input voltage $V_p = V_t$. Allowing for temperature dependence of V_d ,

$$V_p \approx V_t$$

6.1

If equation 6.1 were exact, the temperature coefficient of the ac triggering voltage would be zero.

A more accurate calculation of temperature coefficient can be made by taking into account the non-equal diode currents which cause offsets so that equation 6.1 is not exact.

The temperature coefficient of an SiC diode is typically $-2.5 \text{ mV}/^\circ\text{C}$ at $100 \text{ } \mu\text{A}$ bias current and changes by $0.2 \text{ mV}/^\circ\text{C}$ for each decade increase in current. Referring to Fig. 5.1, if we assume $V_{\text{ref}} = 5 \text{ V}$, the bias current of D33, D34 is about $350 \text{ } \mu\text{A}$ giving a combined temperature coefficient of $-4.8 \text{ mV}/^\circ\text{C}$. Assuming a 5 V peak input signal, the effective average bias current of Q1 is about $50 \text{ } \mu\text{A}$ so that the voltage across R1 has a temperature coefficient of $-2.2 \text{ mV}/^\circ\text{C}$. Triggering occurs when Q2 is biased at $0.5 \text{ } \mu\text{A}$, Q3 at $100 \text{ } \mu\text{A}$, Q6 at $300 \text{ } \mu\text{A}$, (assuming a room temperature, NPN transistor current gain, $\beta_n = 200$). This means the emitter of Q3 has a temperature coefficient of $+3.3 \text{ mV}/^\circ\text{C}$, the emitter of Q6 a temperature coefficient of $+2.4 \text{ mV}/^\circ\text{C}$ (assuming temperature coefficient of $V_t = 0$). Thus, there is an overall temperature coefficient of triggering voltage of $-0.9 \text{ mV}/^\circ\text{C}$, i.e. signals can be 54 mV smaller at 60°C than those at 0°C and still have the same effect. Since the ac peak value of the signal being detected is of the order of 5 V , the calculated amount of temperature drift to be expected is of the order of 1% . This corresponds reasonably well with the measured performance of Fig. 6.7.

6.3.2 Time Delay Before Triggering

R2C2 provide a delay time constant, $\zeta = 10$ mS. Thus, the value of V_o in response to a step input dc bias change at node 1 is given by:

$$V_o = V_p (1 - e^{-\frac{t}{\zeta}}) \quad 6.2$$

With nominal tone dialing input frequencies, the ac input voltage to the detectors is $V_p = 5.9$ V peak (see section 8.11). Triggering occurs when V_o reaches 5 V. Thus from equation 6.2:

$$e^{-\frac{t}{\zeta}} = 0.2 \quad \therefore t = 2\zeta = 20 \text{ mS} \quad 6.3$$

This time delay of 20 mS increases if the input frequencies are off from their nominal values, until it reaches 40 mS at the recognition band edges.

6.4 Fabrication

The eight detector/trigger circuits required are realised in two identical hybrid modules containing four each. The four low-group filter outputs are coupled to one module and four high-group filter outputs to the other module. The signal level, and V_t terminals of each module, are cross-coupled via the threshold control circuitry described in section 7.4 to control the acceptable ratio of frequency amplitudes. Fig. 6.2 shows the complete schematic of the quad module including the partitioning of the circuit for fabrication. The bulk of the circuitry is contained on an SIC chip measuring 64 mil x 60 mil, fabricated on a standard operational amplifier process, e.g. MIL QL3 or QL12 processes. D1 of Fig. 6.1 is realised as a lateral PNP transistor with collector base short (Q1 in Fig. 6.2) in order to provide a high breakdown diode without parasitic action. (NPN emitter base diodes break down at 7 to 8 volts and NPN collector-base diodes have high parasitic PNP gain to the substrate.)

Q4 is a split collector lateral PNP transistor with equal collector areas. This provides a unity gain current reflector action, coupling the output current of Q3 to the base of Q5. Standard IC layout geometries are used for the remainder of the components except for the output transistors, e.g. Q8, which have medium geometry with a saturation resistance of about 150Ω .

The eight 5 nanofarad capacitors required for coupling and smoothing are fabricated on a tantalum thin film glass substrate measuring 383 x 700 mil. The eight $2\text{ M}\Omega$ resistors required are fabricated on a thick film substrate measuring 1200 x 940 mil which also serves as a carrier for the capacitor module and the SIC chip. These are interconnected to the resistor substrate via wire bonds. External leads are soldered to the substrate and the module potted in epoxy. The only trimming required for this unit is to trim the value of the $2\text{ M}\Omega$ resistors which can either be by laser or conventional sandblast techniques. Figure 6.3 shows the resistor and capacitor substrates; Fig. 6.4 shows the SIC chip, and Fig. 6.5 shows the complete assembly.

Fig. 6.2 Quad detector/trigger module schematic (MH8803)

Fig. 6.3 Substrates for the detector/trigger module (M18803)

Fig. 6.4 Quad detector/trigger IC chip (ML1313)

Fig. 6.4 quad detector/triplet chip (CMA 112)

Fig. 6.5 Complete MH8803 detector/trigger module assembly

Fig. 6.5 Complete MH8803 detector/trigger module assembly

6.5 Measured Performance

6.5.1 Linearity

Figure 6.6 shows the peak ac voltage at 1 kHz required to trigger and release the detector/trigger circuit vs. the dc voltage applied to V_t . The curves show the excellent linearity that is achieved.

Fig. 6.6 Peak ac triggering and release voltages of MH8803 vs. V_t

6.5.2 Temperature drift

Figure 6.7 shows the drift of the triggering and release voltages of the circuit vs. temperature with $V_t = 5$ V.

Fig. 6.7 Peak ac triggering & release voltages of MH8803 vs. $^{\circ}\text{C}$ ($V_t = 5$ V)

Chapter VII

SUPPORT CIRCUITRY FOR COMPLETE RECEIVER

7.0 Introduction

Figure 7.1 shows the schematic of a complete ME8800 tone receiver. The basic block diagram and the mode of operation of the system have been discussed in section 3.3. The ME8801 AGC circuit, the ME8802 filter module, and the ME8803 detector module have been described in Chapters 4, 5 and 6. This chapter describes the rest of the circuitry required to construct the complete working receiver around these components.

7.1 Power Supply

The basic tone receiver is designed to operate off standard telephone system batteries which have voltage ranges of 42 V to 55 V and, depending on the type, 21 V to 27 V. To accommodate this wide variation, a simple series regulator is included which supplies the internal circuitry with an 18 V supply rail. This is comprised of QS1, DS2 and RS1. The maximum power dissipated by QS1 occurs at 55 V and is 1.5 watts, so a medium power device is used. DS1 prevents damage if the power supply is connected incorrectly. CS3 prevents long inductive supply leads from causing stability problems. CS1 and CS2 are decoupling capacitors to provide a low ac impedance on the 18 V supply rail.

The maximum power dissipation of RS1 is 0.33 watts and that of DS1 is 0.2 watts. A voltage tolerance of $\pm 3\%$ is suitable for DS1.

7.2 Input Stage

A bridging input transformer of $20K\Omega$ input impedance is used to couple to the telephone line. This is desirable because of the high common mode noise levels present with telephone lines of up to 15 V rms (section 2.6).

Fig. 7.1 Complete tone receiver schematic

The high dynamic range and the degree of balance required make an input transformer the best way of coupling to the line. A step-down turns ratio of 4.5 : 1 is used. Diodes D01, D02 clip the signal to approximately 0.5 V rms corresponding to an input amplitude of 2.25 V rms. Due to the presence of R02, the line is not loaded by this clipping, an important consideration to avoid attenuation of rotary dial 50 V pulses.

This clipping level allows an ample margin on the maximum composite tone amplitude 1.8 V rms. The purpose of the clipping is to prevent input transients caused by lightning and inductive surges, etc., from overloading any of the tone receiver circuitry. R01 provides a suitable loading impedance appropriate for ensuring the input impedance is greater than 20 K Ω . The nominal input impedance is 30 K Ω . A blocking capacitor CT1 is used on the input to allow direct connection to the telephone line.

A buffer amplifier follows the input transformer with the gain set by Rf2/Rf1. This provides a low source impedance for driving the prefilter and, together with the gain of the AGC circuit, determines the input voltage operating range of the receiver.

7.3 Prefilter

The prefilter has three main functions:

- (a) removal of low frequency noise (including dial tone)
- (b) removal of high frequency noise -- above voice band.
- (c) providing pre-emphasis of 3 kHz to reduce speech simulation of tones.

The filter consists of the second order highpass sections involving operational amplifiers 2 and 3, followed by a second order underdamped lowpass section involving operational amplifier 4. These are all of the

voltage follower Sallen & Key type which is the simplest possible realisation and has adequate performance in this application. These active sections are followed by two passive poles formed by CV1, RV1 and CV4, RV7/RV8. The overall response was simulated using the KRON analysis program* which gives the transfer function shown in Fig. 7.2. The nodal circuit diagram and input and output data of the KRON simulation are shown in Appendix A. This is an insignificant value compared to the minimum input tone burst time of 40 mS.

7.3.1 Dial Tone Rejection

The low frequency roll off of the filter reduces the total effective dial tone amplitude to 5 mV rms from its nominal value of 75 mV rms at each of 350 Hz and 440 Hz. The minimum CO receive signal level is a total of 260 mV rms (Table 2B), so the minimum signal to noise ratio when dial tone is present is approximately 34 dB which is ample for proper reception of signals.

7.3.2 3 kHz Pre-emphasis

The 12 dB peak in the filter at 3 kHz has the purpose of pre-emphasising the 'third formant' sounds present in speech. Detailed treatment of the advantages of using this type of pre-emphasis is given in references 7.1, 7.2, 7.3. The principle is that with valid tone signals present, there should be no energy present in the 3 kHz region so the filter pre-emphasis has no effect except to reduce the signal to noise ratio slightly. With speech signals there is nearly always some significant energy present in the 3 kHz band and pre-emphasis helps this to dominate the signal and thus improve the guard action described in section 3.3.1.

* Thanks are due to Mr. J. Pinel, of Bell Northern Research, Ottawa, for providing Microsystems International with the KRON program.

Fig. 7.2 Frequency response of prefilter

A greater amount of pre-emphasis could be used to further improve the guard action but the signal to noise ratio of valid signals begins to deteriorate on long loops if this is carried too far. Also, the harmonic distortion of valid signals can be increased excessively. The 12 dB pre-emphasis of this filter provides about two times fewer simulations than a flat response.

7.3.3 Response at Signal Frequencies

The filter response has a gain of 3 dB at 1633 Hz which is $2\frac{1}{2}$ dB above that at 697 Hz. This is to allow an additional pre-emphasis of high frequencies, to allow for attenuation slope in the telephone loop, over and above that provided in the tone dial (Table 2A). This is desirable due to the increased attenuation slope that can be encountered on end to end connections.

7.4 Trigger Threshold Voltage Control Circuitry

Section 3.3 described the principle of pivoting threshold control which is used in the new receiver to allow for relative amplitude variation between the individual frequencies of an input tone. Referring to Fig. 7.1, this concept is realised using the circuitry involved with operational amplifiers 8 and 9. The basic object of this circuitry is to reduce the threshold voltage of one group when a large signal is detected in the other group.

Operational amplifiers 8 and 9 are connected in a unity gain inverting configuration with a bias of 5 V on their positive inputs. Thus, V_{t1} is controlled by V_{oH} and V_{t2} is controlled by V_{oH} , according to equations 7.1, 7.2.

$$Vt_1 = 5 - (Vo_h - 5) \quad 7.1$$

$$Vt_h = 5 - (Vo_l - 5) \quad 7.2$$

Thus, if equal signals are present in the two detector groups of 5.5 V, equal threshold voltages result:

$$Vo_h = Vo_l = 5.5 \text{ V} \quad 7.3$$

$$Vt_h = Vt_l = 4.5 \text{ V} \quad 7.4$$

However, if non equal signals are present, e.g.

$$Vo_l = 6.5 \text{ V} \quad 7.5$$

$$Vo_h = 4.5 \text{ V} \quad 7.6$$

the threshold voltages are set to non equal values:-

$$Vt_h = 3.5 \text{ V} \quad 7.7$$

$$Vt_l = 5.5 \text{ V} \quad 7.8$$

This allows for a small amplitude high-group signal to trigger its output provided a large amplitude low-group signal is present. The converse also applies, since the circuitry is symmetrical.

Transistors Q1, Q2 have the function of limiting the lower excursion of the threshold voltages to 3.3V. This is to avoid excessively large amplitude ratios from being accepted, and corresponds to an acceptable amplitude ratio limit of about 7 dB.

Transistors Q3, Q4 ensure that only one trigger in each group will operate, since they do not allow the threshold voltage of a detector module to drop more than one diode drop below the signal level present in that module.

Chapter VIII

COMPLETE TONE RECEIVER

8.0 Introduction

Chapter 7 described the support circuitry which, together with the microcircuit components, are used in the complete receiver. This chapter discusses the performance of the complete receiver, its fabrication and its competitive position in relation to other tone receivers.

8.1 Overall Performance Characteristics

A schematic of the complete tone receiver is shown in Fig. 7.1. The standard configuration shown includes an input transformer and pre-filter, has about 46 dB dynamic range, and undecoded open collector outputs. The maximum acceptable amplitude ratio is 9 dB. The input transformer can be removed for applications which are single ended. The gain of the input buffer, operational amplifier 1, can be adjusted to suit the maximum expected input level. The pre-filter is desirable for most applications as it removes dial tone, power line hum and other low frequency noise and also has a 3 kHz peaking section which reduces voice simulations. (See section 7.3). If only about 26 dB dynamic range is required (applications other than end to end signalling), operational amplifier 5 and its associated circuitry can be removed and bypassed directly to CV4. Variations of dynamic range can also be made by changing RV5 or RV9. If it is desired to reduce the acceptable input amplitude ratio, Rt9 can be reduced. The acceptance bandwidth is controlled by the input ac level to the bandpass filters. This is set to 150 mV rms by adjustment of potentiometer RV12 which gives an acceptance bandwidth of $\pm 2\%$ about nominal. The overall response time to tone bursts is 20 mS with nominal input frequencies,

and 40 mS worst case with input frequencies at the extremes of their $\pm 1.5\%$ tolerance. Figure 8.1 shows response waveforms to an input 40 mS tone-burst consisting of 697 Hz and 1633 Hz frequency components. The upper trace shows the input tone burst. Beneath it are the AGC amplifier output, the 1633 Hz filter output, and the 1633 Hz detector-trigger output waveforms. The overall response time to operate the logic output is 20 mS. (This time increases to 40 mS if the input frequencies are at their tolerance limits of $\pm 1.5\%$.) The detector outputs were terminated in loads of 10 k Ω to a +5 V supply.

Time base = 10 mS per division.

Fig. 8.1 Receiver waveforms in response to
40 mS input tone burst

and 40 mS worst case with input frequencies at the extremes of their $\pm 1.5\%$ tolerance. Figure 8.1 shows response waveforms to an input 40 mS tone-burst consisting of 697 Hz and 1633 Hz frequency components. The upper trace shows the input tone burst. Beneath it are the AGC amplifier output, the 1633 Hz filter output, and the 1633 Hz detector-trigger output waveforms. The overall response time to operate the logic output is 20 mS. (This time increases to 40 mS if the input frequencies are at their tolerance limits of $\pm 1.5\%$.) The detector outputs were terminated in loads of 10 k Ω to a +5 V supply.

Time base = 10 mS per division.

Fig. 8.1 Receiver waveforms in response to
40 mS input tone burst

8.1.1 Signal Levels

The secondary of the input transformer is limited to an amplitude of 0.5 V rms by diodes Do1, Do2. This corresponds to an input amplitude of 2.25 V rms which is acceptably high. Limiting is effected to prevent large input transients from overloading any of the receiver stages. The input transformer has a step down turns ratio of 4.5: 1 and this is followed by a buffer, operational amplifier 1, with a voltage gain of 3. The prefilter has a voltage gain of approximately 1.2, so that the overall voltage gain from the line to the AGC amplifier input is 0.8. The maximum signal handling ability of the AGC amplifier is about 2 V rms so the maximum expected input tone amplitude can be handled comfortably. The maximum voltage gain through the AGC amplifier is 1.3 (see equation 4.2) so the overall voltage gain from the line to the output of the AGC amplifier is a maximum of close to unity. The minimum composite signal for operating the filter-detector combination is about 4 V rms. Thus, the minimum composite waveform amplitude that will operate the receiver is $4/\Psi$ V rms or $2/\Psi$ V rms per frequency, where Ψ is the combined voltage gain of operational amplifiers 5, 6. The level at the input to the bandpass filters is adjusted to be 150 mV rms, or 75 mV rms per frequency, assuming two equal amplitude frequencies are present.

The voltage gain of the bandpass filters is 50.6 at their centre frequencies, (see equation 5.51). Their voltage gain at the frequencies of the other group is approximately 5. Thus the ac filter output voltage to the detectors is:

$$\begin{aligned} 3.8 + .38 &= 4.18 \text{ V rms} \\ &= 5.9 \text{ V peak} \end{aligned}$$

Since about 4.9 V nominal (allowing for .1 V ripple) is required to trigger the detectors, this allows margin for reduced gain of the filters at the band edges. By varying the amplitude of the input signal to the bandpass filter module by means of potentiometer RV12 the recognition bandwidth of the receiver can be adjusted. The normal adjustment is for $\pm 2\%$ bandwidth.

8.1.2 Overall Signal Recognition Bandwidth Vs. Temperature

Figure 8.2 shows the variation of receiver recognition bandwidth versus temperature. This includes the composite effects of temperature drift in the output amplitude of the AGC circuit, the Q of the filters and the threshold of the detector/triggers which are shown in Figs. 4.5, 5.18 and 6.7 respectively.

Fig. 8.2 Overall receiver recognition bandwidth vs. temperature

8.1.3 Voice Immunity Performance

Voice immunity is difficult to evaluate quantitatively due to the statistical nature of voice sounds. However, BTL has prepared a test tape specifically for testing the voice immunity of tone receivers and this provides a convenient standard to measure the comparative performance of different receivers. The Bell Northern Research telecommunications group uses this tape in their receiver evaluations. (See Table 8A.) The tape consists of approximately half an hour of recorded telephone trunk conversation which was edited from an original six hour recording and contains all the passages likely to cause simulations. A simulation is defined as an output pulse from the receiver when no valid input signal is present. The total number of potential simulations on the tape is about 200. To test a receiver, the tape recording is played back into the receiver at a calibrated input level. A counter connected to the outputs of the receiver records any hits that occur. Ideally, the number of hits caused by voice should be zero when the tape is finished. Typically, the number varies from two to one hundred depending on the receiver.

The dynamic range of a receiver is the range of input signal amplitudes that it will accept. Normally the upper limit is set at about 1 V rms per frequency, so if the minimum acceptable amplitude is 0.1 V rms/frequency, the receiver has a dynamic range of 20 dB. Due to the random nature of speech signals, the wider the dynamic range the more simulations it will pick up.

The new receiver was tested with various settings of dynamic range and a bandwidth of $\pm 2\%$ and the resulting number of simulations is shown in Fig. 8.3.

Fig. 8.3 Simulations on ME8800 receiver vs. dynamic range

8.1.4 Error Rate Vs. S/N Ratio

Errors in a receiver are defined as occasions when the output of the receiver does not give the correct response to a valid input signal. They are caused by noise present together with the input signal. At signal to noise ratios in excess of 20 dB, the error rate of the new receiver is close to zero. But, it climbs rapidly as the noise level increases. Fig. 8.4 shows the measured error rate of the receiver vs. signal to noise ratio. This was measured by repetitively pulsing a valid input tone at the input to the receiver. To this signal white noise was added at the appropriate levels. The values for the signal to noise ratios were computed from true rms voltage measurements on the noise and on the uninterrupted tone. The input signal voltage was set at 0 dBV for the

measurements. Digit 1 (697 Hz and 1209 Hz) was used for the input tone, in 45 ms tone bursts.

Fig. 8.4 Error rate of ME8800 receiver vs. S/N ratio

8.2 Fabrication of Complete Receiver

The complete receiver is fabricated as a PCB assembly of two MH8803 detector modules, one MH8802 filter module and a ML1314K AGC SIC, together with four 1458 operational amplifiers (dual 741 type), a 3045 transistor array, 15 capacitors, 34 resistors, a 2-watt transistor, and two diodes. The circuitry occupies about 25 square inches of PCB without crowding and using standard sized capacitors. The components required are low enough in profile for the PCB to be mounted in racks on 1/2" centres. The overall volume requirements of the receiver are much less than for

existing circuitry. This is a major advantage with modern applications in electronic PBX's and end to end terminal equipment. The small size is due partly to simplification of the overall system and circuitry and partly due to the fabrication of the filter and detector modules as hybrid microcircuits. Further miniaturisation is possible since the bulk of the discrete component circuitry could be contained in another module the size of the filter block which would reduce the space requirements by a further factor of two. Due to limitations of the MIL production capacity, this was not attempted for the initial design but may be introduced at a later date. After assembly the only adjustment that is made is to set the input level to the bandpass filters at 150mV rms by means of potentiometer Rv12. Figure 8.5 shows a completed receiver board; Appendix E contains a MIL data sheet on the receiver.

8.3 Comparison with Other Receivers

It is difficult to provide an exact picture of the competitive position of the new receiver compared to all the presently available designs because of the lack of published information. However, some broad outlines can be discussed concerning the present situation and likely future trends.

The older equipment used presently by most telephone companies follows the original BTL design^(1.4) in various qualities of realisation. Fabrication is by conventional LC filter techniques using ferrite pot core inductors and polystyrene capacitors, together with discrete transistor circuitry on multiple PCB's which are assembled in a rack. The unit price varies from about \$300 for a simple low performance key telephone receiver, to about

Fig. 8.5 Complete ME8800 tone receiver PCB

\$1,000 for the best CO receivers. The simpler units have only about 12 dB dynamic range and crude band separation filtering, while the more complex units have about 20 dB dynamic range and use band elimination filtering. A new receiver recently designed by Bell Northern Research Laboratories for use in the new Northern Electric 'PULSE' electronic PBX, duplicates the receiver system of the older designs but replaces the inductors with hybrid gyrator-capacitor combinations. ^(5.1) A number of small companies in Europe and the United States have recently started marketing receivers on PCB's using techniques which range from miniaturised LC approaches to phase lock loops. At least two development groups are known to be working on combination analog-digital approaches using LSI MOS circuitry and crystal oscillators, although none have yet reached the manufacturing stage.

The present competitive position of the ME8800 tone receiver is very good. It has the advantages of small size (occupying less than half the volume of any previously available receiver), excellent performance, wide dynamic range and simple interfacing. The selling price of about \$180 is competitive and provides MIL with a good profit margin, which should improve further as volume builds up. The strongest future competition is likely to come from a combination of linear circuitry with specially designed MOS integrated circuits which could be compact and have relatively low cost. However, manufacturing costs of these approaches are projected to be somewhat higher than the ME8800 and, at present, the available designs are only capable of about 30 dB dynamic range which is too small for end to end use.

Some idea of the relative performance of receivers discussed above can be obtained from Table 8A which shows data from tests made at Bell Northern Research Laboratories. The number of simulations occurring with

Receiver Type	Min. Sig. (dBV)	Min. S/N Ratio (dB)	Number of Simulations on Bell Labs Voice Test Tape
Northern Electric Co. (A3)	-21	18	5
Northern Electric Key Telephone (247KTU)	-12	7	3
A.T. & T. Data Set (403D)	-45	20	32
Northern Electric 'PULSE' SGI	-28	22	30
BNR Labs Analog Digital	-29	19	17
'Tel Tone' (independent U.S. company)	-17	5	80
Microsystems International Ltd. ME8800	-25	17	14

Table 8A Receiver comparison data

the new receiver is among the lowest of those tested. The low value achieved with the 247 KTU receiver is due to its very small dynamic range. The value achieved with the A 3 receiver is also low but this unit is a rack of equipment, sells for \$700 and has a somewhat small dynamic range.

To summarise, the new receiver has the smallest size and lowest price available at present. It offers excellent versatility, with a dynamic range of from 10 to 46 dB . It also has a simulation rate that is among the best available at any price or size.

Chapter IX

CONCLUSIONS

This successful and rapid realisation of an advanced technology product occurred due to cooperative interaction between Carleton University and Microsystems International Limited. The project was conceived and the theoretical work done at Carleton University, with support from MIL. Subsequently, the full cooperation of MIL in provision of facilities enabled the three custom SIC's and the total of five custom resistor and capacitor substrates to be successfully designed and fabricated. The result is a competitive product which is expected to provide significant profitable return for MIL over the next few years.

Apart from the desirable result of a profitable end product, this project resulted in innovation in five areas which can be summarised as follows.

- a receiver system configuration which cuts the critical filtering requirements by half*
- an audio AGC circuit with about five times tighter control than previous designs of comparable speed*
- a simplified approach to optimum design of positive gain single amplifier filters, enabling excellent visibility of performance/layout area trade-offs.
- a rapid and accurate method of monitoring filter centre frequencies to enable automatic trimming of each section in less than six seconds*
- a precisely controlled ac level detector*

* Four patent applications (assigned to Microsystems International Ltd.) were filed on the items asterisked above.

These innovations may prove to be useful in areas other than tone receivers. The filter optimisation approach and the methods of fabrication and trimming are applicable to a wide variety of audio filter applications. The AGC circuit approach offers tighter control than previously available and could be useful for other functions. The ac level detectors are also suitable for general purpose use, providing excellent linearity and accuracy over temperature.

Chapter X

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Appendix A

Program Used for GSP Computation

1*		DIMENSION A(10),B(10)
2*		DATA B/.1,.2,.4,.8,1.,1.4,2.,4.,6.,10./
3*		DATA A/.1,.2,.4,.8,1.,1.4,2.,4.,6.,10./
4*		Q=15.
5*		WRITE(6,22)
6*	22	FORMAT(5X,'ROOT C RATIO,A',5X,'ROOT R RATIO ,B',10X,
7*		C 'GSP FILTERS 1,2')
8*		DO 10 J=1,10
9*		DO 10 I=1,10
10*		Y=A(J)*B(I)
11*		X=A(J)/B(I)+Y+1/Y
12*		GSP=Q/Y*(X-1/Q)**2
13*		WRITE(6,11) A(J),B(I),GSP
14*	10	CONTINUE
15*	11	FORMAT(10X,F7.2,12X,F7.2,15X,F10.1)
16*		STOP
17*		END

END OF COMPILATION: NO DIAGNOSTICS.

Table A1 - Univac Fortran V program for exact GSP computation

Appendix B

Data Used in Amplifier Stability Analysis Program

INPUT DATA--

```

EECS  B.P.FILTER ANALYSIS
GAIN PZ VOUT VIN 5 0 2 0 10 1K 10E6
TEMP 400 330 300 270 225
VB 1 0 5
VS 10 0 18
RS 1 2 600
R1 2 3 56.6K
R3 8 7 20K
R4 7 5 10K
R5 5 6 3K
R6 6 0 3K
XN NP 300 6N 150 / 7M / / 450 // 200 //// 300MEGHZ 2M 5 2P .6 .6P -10 2P
XP PNP 20 / / /-2M / / 50 // 300 //// 3.MEG .1M -5 1P -.6 2P / 2P
Q1 XN 9 3 6
Q2 XP 8 9 10
Q3 XN 10 8 7
Q4 XN 10 7 5
END

```

Table B1 - Amplifier stability analysis data

TRANSISTOR MODELS--

PARAMETER	VALUE	
NAME	XN	XP
TYPE	NPN	PNP
BF/BE _{MAX}	3.000+02	2.000+01
/IC _{MAX}	6.000-03	0.000
/BR _{LOW}	1.500+02	0.000
BR/BE _{MAX}	1.000+00	1.000+00
TC1	7.000-03	-2.000-03
TC2	0.000	0.000
IS	1.000-14	1.000-14
RB	4.500+02	5.000+01
TC1	0.000	0.000
TC2	0.000	0.000
RC	2.000+02	3.000+02
TC1	0.000	0.000
TC2	0.000	0.000
PO	5.000+04	5.000+04
IC	1.000-03	1.000-03
FT	3.000+03	3.000+06
IC	2.000-03	1.000-04
VCE	5.000+00	-5.000+00
CJE	2.000-12	1.000-12
VBE	6.000-01	-6.000-01
CJC	6.000-13	2.000-12
VBC	-1.000+01	0.000
CS	2.000-12	2.000-12

Table B2 - Summary of data used for NPN
and PNP Transistor Models

Appendix C

Data Used in Q Stability Analysis

Fig. C1 - Type 5 circuit for computer analysis of Q stability

Fig. C2 - Type 1 circuit for computer analysis of Q stability

INPUT DATA--

```

SECS  B.P.FILTER ANALYSIS
GAIN PZ VOUT VIN 5 0 2 0
TEMP 300 270 285 315 330
VB 1 0 5
VS 10 0 18
RS 1 2 600
R1 2 3 56.6K
R2 5 4 28.3K
R3 8 7 20K
R4 7 5 10K
R5 5 6 2.37K
R6 6 0 3K
C1 3 4 2436P
C2 4 0 2436P
XN NPN 300 6M 150 / 7M / / 450 // 200 //// 300MEGHZ 2M 5 2P .6 .6P -10 2P
XP PNP 20 / / / -2M / / 50 // 300 //// 3.MEG .1M -5 1P -.6 2P / 2P
O1 XN 9 3 6
O2 XP 8 9 10
G3 XN 10 8 7
O4 XN 10 7 5
END

```

Table C1 (a) - Input data for SLIC analysis of effect of amplifier variations
on Q and f_0

INPUT DATA--

```

EECS  B.P.FILTER ANALYSIS
GAIN PZ VOUT VIN 5 0 2 0
TEMP 300 270 285 315 330
VB 1 0 5
VS 10 0 18
RS 1 2 600
R1 2 3 56.6K
R2 3 4 28.3K
R3 8 7 20K
R4 7 5 10K
R5 5 6 2.9115K
R6 6 0 3K
C1 4 5 3445.6P
C2 3 0 1722.8P
XN 1PN 300 6M 150 / 7M / / 450 // 200 //// 300MEGHZ 2M 5 2P .6 .6P -10 2P
XP 1PN 20 / / -2M / / 50 // 300 //// 3.MEG .1M -5 1P -.6 2P / 2P
Q1. XN 9 3 6
Q2 XP 8 9 10
Q3 XN 10 8 7
Q4 XN 10 7 5
END /

```

Table C1 (b) Input data for SLIC analysis of effect of amplifier variations on Q and fo

Appendix D

Data Pertaining to Analysis of Prefilter Frequency Response

Fig. D1 - Nodal circuit diagram for ME8800 prefilter used for KRON simulation

KRON'S METHOD

8800 PREFILTER

NETWORK INPUT DATA

20 PASSIVE ELEMENTS

4 DEPENDENT SOURCES

13 NODES

NODE 1 + INPUT TERMINAL

NODE 0 - INPUT TERMINAL

NODE 16 + OUTPUT TERMINAL

NODE 0 - OUTPUT TERMINAL

1	2	C		.1000000-07
2	4	C		.1000000-07
2	5	R		.8000000+04
0	4	R		.4300000+05
0	5	R		.1000000+01
5	6	C		.1000000-07
6	8	C		.1000000-07
6	9	R		.3400000+04
0	8	R		.1780000+06
9	10	R		.5100000+05
10	12	R		.5100000+05
9	0	R		.1000000+01
10	13	C		.1000000-07
12	0	C		.1000000-09
13	0	R		.1000000+01
13	14	C		.1000000-07
14	0	R		.2430000+05
15	0	R		.1000000+01
15	16	C		.1000000-07
16	0	R		.1330000+05
0	5	4	0 G	.1000000+01
0	9	8	0 G	.1000000+01
0	13	12	0 G	.1000000+01
0	15	14	0 G	.1000000+01

40 POINTS PER DECADE MIN FREQ

.3000+03 MAX FREQ

.4000+04

Table D1 - Input data for KRON simulation of ME8800 prefilter response

FREQ	FUNCTION	V2V1	
CPS	GAIN	ARG	TD
	DB	DEG	SEC
.3000+03	-.4878+02	.1119+03	.6755-03
.3182+03	-.4546+02	.1075+03	.6889-03
.3376+03	-.4210+02	.1026+03	.7058-03
.3581+03	-.3869+02	.9731+02	.7275-03
.3799+03	-.3522+02	.9150+02	.7558-03
.4030+03	-.3167+02	.8507+02	.7933-03
.4275+03	-.2804+02	.7785+02	.8444-03
.4535+03	-.2430+02	.6963+02	.9157-03
.4811+03	-.2042+02	.6005+02	.1019-02
.5104+03	-.1638+02	.4855+02	.1174-02
.5414+03	-.1219+02	.3418+02	.1414-02
.5743+03	-.7903+01	.1537+02	.1778-02
.6093+03	-.3850+01	.3501+03	.2229-02
.6463+03	-.7721+00	.3185+03	.2430-02
.6856+03	.7727+00	.2860+03	.2088-02
.7273+03	.1273+01	.2586+03	.1590-02
.7716+03	.1435+01	.2363+03	.1234-02
.8185+03	.1552+01	.2175+03	.1010-02
.8683+03	.1676+01	.2009+03	.8570-03
.9211+03	.1792+01	.1857+03	.7395-03
.9771+03	.1887+01	.1719+03	.6407-03
.1037+04	.1958+01	.1591+03	.5545-03
.1100+04	.2014+01	.1474+03	.4791-03
.1166+04	.2069+01	.1367+03	.4141-03
.1237+04	.2136+01	.1268+03	.3587-03
.1313+04	.2229+01	.1178+03	.3122-03
.1392+04	.2358+01	.1094+03	.2733-03
.1477+04	.2532+01	.1016+03	.2410-03
.1567+04	.2759+01	.9422+02	.2144-03
.1662+04	.3049+01	.8726+02	.1926-03
.1763+04	.3411+01	.8058+02	.1752-03
.1871+04	.3859+01	.7409+02	.1617-03
.1984+04	.4410+01	.6767+02	.1524-03
.2105+04	.5087+01	.6118+02	.1475-03
.2233+04	.5923+01	.5438+02	.1485-03
.2369+04	.6962+01	.4693+02	.1580-03
.2513+04	.8266+01	.3819+02	.1823-03
.2666+04	.9904+01	.2688+02	.2346-03
.2828+04	.1187+02	.1038+02	.3415-03
.3000+04	.1366+02	.3441+03	.5072-03
.3182+04	.1360+02	.3086+03	.5233-03
.3376+04	.1117+02	.2707+03	.5299-03
.3581+04	.8118+01	.2602+03	.1080-03
.3799+04	.5324+01	.2485+03	.1172-03
.4030+04	.2877+01	.2405+03	.8010-04

Table D2 - Output data from KRON simulation of ML8800
prefilter response

Appendix E

Microsystems International Limited Preliminary
Data Sheet on the New Receiver

APRIL 1973.

Telecommunications Group
Tone Receiver
For TOUCH-TONE^R Compatible Systems

ME8800J
ME8810J

PRELIMINARY DATA SHEET

FEATURES

Small size - printed circuit board 5 3/16" x 5 1/64" with plug-in end connection.

Good immunity to false dialing and voice frequencies.

Balanced input with built-in high pass filter.

Open-collector output switching 2 out of 8.

Sensitivity 5mV/frequency (range 46db), for ME8810, and Central Office applications 50mV/frequency range 26db for ME8800.

DESCRIPTION

The Microsystems International ME8800J and ME8810J are complete receivers for TOUCH TONE^R compatible systems, suitable for Central Office and end-to-end signalling applications. The receiver consists of three dual-in-line hybrid modules, packaged integrated circuits, and discrete components on a p.c. board. A power supply of 20 to 55 volts is required with current drain normally less than 50mA. The receiver is protected against polarity reversal. Open-collector, undecoded outputs are provided (normally grounded) and these can sink up to 3mA. The System Block Schematic shows the various parts involved in the receiver. ME8801 is a signal conditioning I.C. consisting of balanced input and a fast attack a.g.c. system, with adjustable range and hold time. ME8802 provides eight bandpass filters with a Q factor of 15. The response time of these filters is typically 15mS. Units ME8803 provide four envelope detectors and triggers, two units used. The receiver response time can be adjusted during manufacture by custom trimming the detector stage, but this is normally set at 40mS.

The receiver provides a signal quality checking function, providing voice protection with a pivoting trigger voltage circuit. This demands that only two frequencies be present and also that the amplitudes be within a certain ratio before the receiver outputs are triggered. Cross coupling between ME8803 units enables this ratio checking function, which can be adjusted but is normally set at 4db for ME8800J and 10db for ME8810J.

For central office applications dial tone rejection is obtained with the front end pre-filter.

TOUCH-TONE is a trademark of A.T. & T.

MICROSYSTEMS INTERNATIONAL LIMITED,
Ottawa, Ontario.

ABSOLUTE MAXIMUM RATINGS

Power Supply	55 Volts
Input Signal Level	1.0 Vrms/frequency
Output Sinking Current	5mA
Operating Temperature Range	0 to +60°C
Storage Temperature Range	-40 to +70°C

tone RECEIVER SPECIFICATION

Temp. Range: 0 - 60°C

Frequencies: Low 697, 770, 852, 941 Hz
 High 1209, 1336, 1477, 1633 Hz BANDWIDTH $\pm$ 2.5%

A.G.C. Range: 26db (ME8800), 46db (ME8810)

Ratio Check: Amplitude ratio of high group to low group can be checked for between 4db to 10db (adjustable)

Minimum tone burst time 40 msec (Standard Unit)

Minimum inter burst time 40 msec

Power supply: 22V to 55V (the circuit is protected against wrong polarity)

Current Drain at 24V 40ma
 at 55V 50ma

Output: Two out of eight undecoded, open collector NPN transistors with the emitters on negative rail. Max. current sinking is 3ma, Max. Voltage rating 35V. Transistors are 'on' normally, 'off' when tones present. Minimum output pulse is 5 msec.

Voice Immunity: Near perfect immunity against voice talk off is provided by guard action and frequency pre-emphasis.

Dial Tone Rejection is provided with a high pass filter.

A balanced input is provided for direct connection to telephone lines.

Input impedance is 20K Ω .

The complete unit is provided on printed circuit board 5 3/16" x 5 1/64" with plug-in end connector. Versions including decode, steering output, etc. can be provided on a custom basis.

CIRCUIT DIAGRAM FOR COMPLETE TONE RECEIVER ME8800J (8810J)

TYPICAL PERFORMANCE CHARACTERISTICS

SYSTEM BLOCK SCHEMATIC

PRINTED CIRCUIT BOARD DETAILS

CONNECTOR B COLUMN 2 CONTACT

FUNCTION

A	GND	
B	TIP	INPUT
C	RING	
D	SUPPLY VOLTAGE	
E	10V	GENERATED
F	18V	
G	5V	INTERNALLY
H	AGC OUTPUT SIGNAL	
J	N.C.	
K	1633 Hz	OUTPUTS
L	1477 Hz	
M	1336 Hz	
N	1209 Hz	
P	697 Hz	
R	770 Hz	
S	852 Hz	
T	941 Hz	
U	GND	
V		

CONNECTOR TYPE:

DIGITAL EQUIPMENT
CORPORATION CONNECTOR
BLOCK - H802.