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Contents

Orbiting at an altitude of 900 km, the infrared astronomical satellite (IRAS) scans the 8 μm to 120 μm spectrum for cosmic infrared sources not observable by earthbound instruments. Launched early this year, IRAS is to spend its first six months mapping infrared sources and the next six months measuring them. IRAS is a joint American, British, Dutch venture in which Philips has played an active part.

Bipolar IC merges telephone dialling and speech

F. VAN DONGEN, J. J. A. GEBOERS and H. J. M. OTTEN

The manner in which integrated circuits are incorporated in telephone subscriber sets depends on:

- the functions which must be performed electronically
- the required compatibility with a conventional subscriber set
- the desired degree of design flexibility
- the cost of the subscriber set.

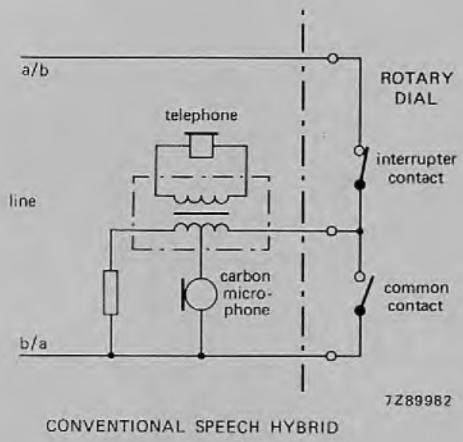
Figure 1 shows four design approaches to a basic telephone with DTMF dialling. In Fig.1(b), the rotary dial in a conventional subscriber set Fig.1(a) has been replaced by a three-wire DTMF dialling unit insert. The rest of the set, including the carbon microphone and transformer hybrid, remains unchanged (Ref. 1).

When less compatibility is required but it is still essential to limit the speech/transmission circuit connections to two wires, the arrangement shown in Fig.1(c) may be used. Switching over from dialling to speech and vice-versa is performed by discrete transistors controlled by the mute output from a bipolar integrated DTMF generator. In this arrangement, a conventional transformer hybrid may also be used for the speech function.

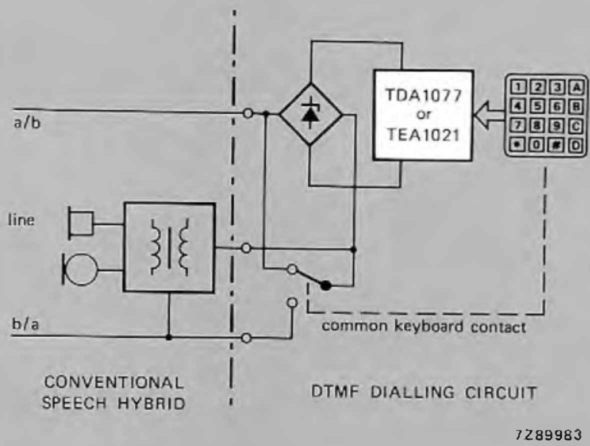
The duplication of line interface functions which is necessary in the Fig.1(c) approach can be avoided by using an electronic speech circuit with a mute input, a DTMF signal input and a supply output as shown in Fig.1(d). Electronic switching from dialling to speech and vice-versa is then performed inside the speech/transmission circuit. The voltage regulator and output stage of this IC are used for transmission of both speech signals and DTMF dialling tones, depending on the operating mode (Ref.2). The dialling function is performed by a separately integrated small-signal generator (usually CMOS). This two-chip approach offers considerable design flexibility to the subscriber set manufacturer. The dialling unit (IC plus keyboard) can be replaced by a pulse dialling circuit or dual-standard dialling circuitry, with or without additional features and connected to the speech/transmission section by only a few wires.

Experimental printed-circuit board with the TEA1046, its peripheral components, keyboard and handset. The basic functions of the IC are dialling tone generation, speech amplification, four to two-wire conversion, dialling/speech switching and line adaptation

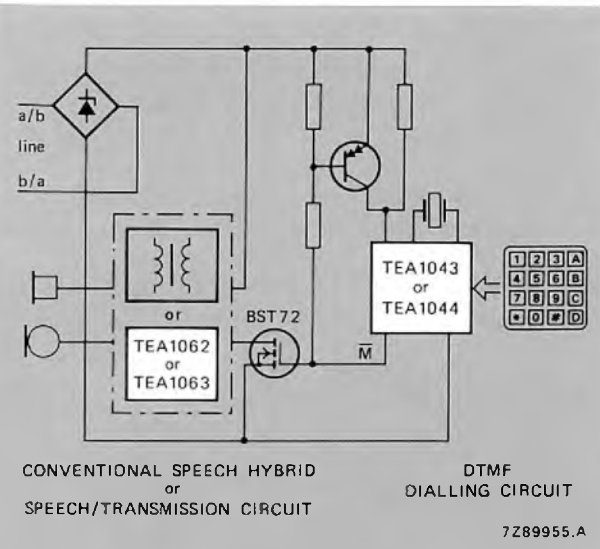
As shown in Fig.1(e), the speech/transmission and dialling functions can be combined on the same chip if the DTMF generator is bipolar. This is the principle of the integrated DTMF dialling/speech transmission circuit TEA1046 described in this article. The philosophy behind this approach is to sacrifice some design flexibility to construct an inexpensive basic DTMF subscriber set.



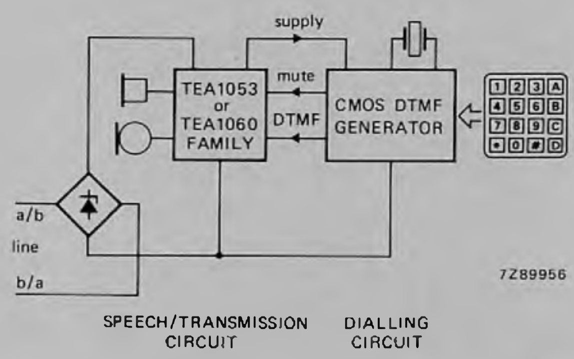
(a) conventional rotary-dial subscriber set



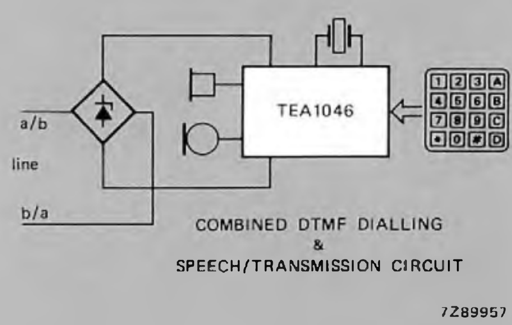
(b) rotary dial replaced by a DTMF insert



(c) DTMF subscriber set with conventional or electronic speech part



(d) DTMF subscriber set with two ICs and a common line interface



(e) subscriber set with speech and DTMF dialling circuits on a single chip

Fig.1 Architecture for basic electronic subscriber sets with DTMF dialling

BRIEF DESCRIPTION

As shown in Fig.2, the timebase from which the two groups of four dialling tones are generated is a crystal-controlled oscillator, the output from which is divided by 9 or 12 (depending on the crystal frequency) to obtain a 398 kHz clock. The clock pulses are then passed through two dividers with programmable division ratios controlled by the logic state of the tone selection inputs at pins 15 to 17 and 19 to 23. The tone selection inputs can be directly driven by a 4 x 4 matrix keyboard or indirectly via a microcomputer if extended dialling features such as redial and repertory dialling are required. The tone-frequency squarewave from each programmable divider is converted into stepped sinus-

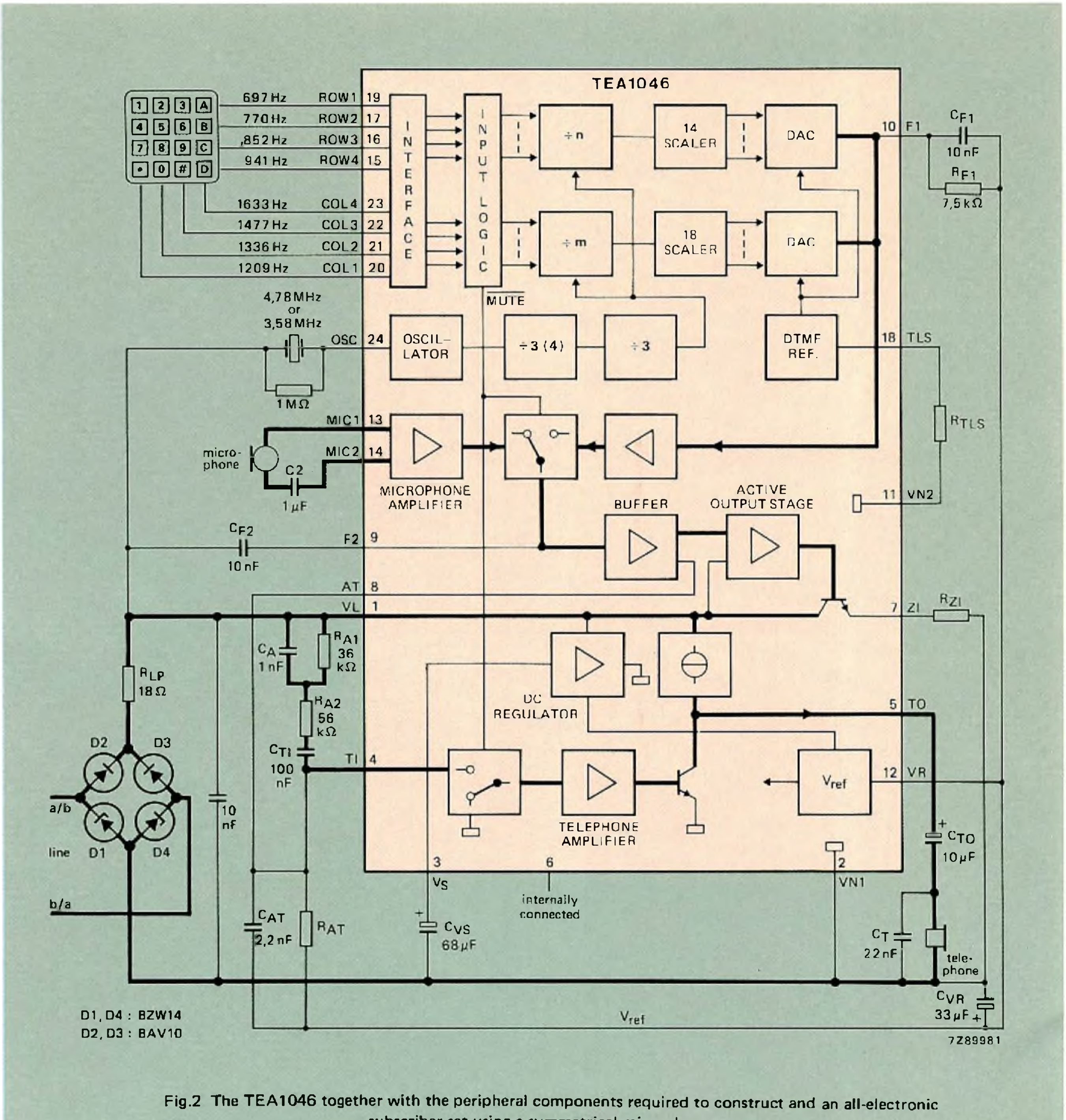


Fig.2 The TEA1046 together with the peripheral components required to construct and an all-electronic subscriber set using a symmetrical microphone

oidal form by a DAC comprising a scaler and an array of weighted current sources. Since a 14-scaler is used for the lower-frequency tone group, the lower-frequency synthesised tones consist of 14 steps per cycle. Similarly, the use of an 18-scaler for the higher-frequency tone group results in 18 steps per cycle for the higher-frequency synthesised tones. The amplitude of the tone currents at pin 10 is accurately maintained because the weighted currents in the DAC are referred to a temperature compensated level which is also independent of the line current. The sinusoidal current steps at pin 10 are converted into a voltage by an external RC network which forms the first stage of a two-section low-pass filter for attenuating the higher harmonics of the synthesised sinewaves. The second capacitor of the low-pass filter is connected to the output of the speech/dialling changeover switch at pin 9. From this point, the tones or speech signals are passed to the line at pin 1 via a buffer and an active output stage, the dynamic resistance of which can be set by an external resistor at pin 7.

The microphone amplifier at pins 13 and 14 is basically a voltage-controlled current source with symmetrical differential inputs. When the tone selection inputs are inactive, an electronic changeover switch, activated by a mute signal from the keyboard input logic circuit, passes the output current from the microphone amplifier to the second section of the low-pass filter at pin 9. From here, the amplified microphone signal is applied to the line via the same path as previously described for the dialling tones.

The telephone amplifier provides sufficient output at pin 5 to drive a dynamic earpiece in the handset. The input signal for the amplifier is extracted from the composite signal on the telephone line via an RC network connected between the line and pin 4. An antiphase sidetone-cancelling signal from pin 8 prevents the microphone signal from being amplified by the telephone amplifier. During dialling, the input to the telephone amplifier is disconnected by an electronic switch activated by a mute signal from the keyboard input logic circuit.

If the line current into pin 1 exceeds the requirements of the IC (about 10 mA), a shunt regulator bypasses the excess current and fixes the average voltage applied to the IC.

Register recall (flash) by a timed loop break can be obtained by using an external network. This feature will be described in a future publication.

DETAILED DESCRIPTION

Tone selection inputs

As shown in Fig.2, the TEA1046 has eight buffered tone selection inputs divided into two groups of four. Pins 15, 16, 17 and 19 are active LOW 'row' inputs for selecting the four lower-frequency tones. Pins 20 to 23 are active HIGH 'column' inputs for selecting the four higher frequency tones. Two types of keyboard can be used to select the two-tone combinations required for DTMF dialling:

- A keyboard in which a single contact connects one of the row inputs to one of the column inputs to select each required two-tone combination
- A keyboard in which one contact per pushbutton connects one of the row inputs to a LOW level and a second contact per pushbutton connects one of the column inputs to a HIGH level to select each required two-tone combination.

Since the maximum 'on' resistance for each contact is 10 kΩ and the minimum 'off' resistance is 250 kΩ, the IC is suitable for use with keyboards in which the contacts are operated by a rubber membrane.

The appropriate two-tone combination is generated immediately after a pushbutton on the keyboard is pressed. A tone, or tone combination is then produced for two full cycles of the lower-frequency tone before the keyboard entry is scanned again. The tones continue as long as the pushbutton remains pressed. As soon as the pushbutton is released for more than two cycles of the lower-frequency tone, tone generation ceases. Simultaneous depression of more than one pushbutton has no undesirable effect because all other keyboard inputs are inhibited as soon as one input is activated. Thus, only the tone combination corresponding with the first pushbutton to be pressed is generated. If the first pushbutton is released after a second has been pressed, the tone combination corresponding to the second pushbutton is generated two periods of the lower frequency tone after release of the first pushbutton (two key rollover). This procedure, which was chosen after extensive investigations of keyboard behaviour, prevents the generation of incorrect tones when the user unintentionally presses two adjacent pushbuttons. For testing purposes, the tones can be generated singly by making the appropriate 'row' input LOW, or the appropriate 'column' input HIGH.

A keyboard will only operate in the manner described as long as at least one of the 'column' inputs (pin 20, 21 or 23) remains LOW (inactive). If all these three 'column' inputs are held HIGH, pin 22 acts as an enable input for the tone generator and the four 'row' inputs can be driven with a 4-bit binary code derived from a 4-bit or 8-bit microcomputer. When using a 4 x 4 keyboard, pins 20, 21 and 23 can of course be made HIGH at the same time by simultaneously pressing pushbuttons in the first, second and fourth column. This type of incorrect operation is, however, considered so unlikely that malfunctioning can be accepted if it does occur. It cannot occur with the more commonly used 3 x 4 keyboard.

Table 1 is the truth table for the dialling tone selection inputs during microcomputer operation.

TABLE 1
Dialling tone control by microcomputer

tone selection inputs ROW 1	2	3	4	COL (1, 2, 4)	3	generated tones (Hz)	symbol	mute	remarks
H	H	H	H	L	L	–	–	off	speech mode
X	X	X	X	H	L	–	–	on	separate mute
H	H	H	H	H	H	697 + 1209	1	on	
H	H	H	L	H	H	697 + 1336	2	on	
H	H	L	H	H	H	697 + 1477	3	on	
H	H	L	L	H	H	697 + 1633	A	on	
H	L	H	H	H	H	770 + 1209	4	on	
H	L	H	L	H	H	770 + 1336	5	on	
H	L	L	H	H	H	770 + 1477	6	on	
H	L	L	L	H	H	770 + 1633	B	on	
L	H	H	H	H	H	852 + 1209	7	on	
L	H	H	L	H	H	852 + 1336	8	on	
L	H	L	H	H	H	852 + 1477	9	on	
L	H	L	L	H	H	852 + 1633	C	on	
L	L	H	H	H	H	941 + 1209	★	on	
L	L	H	L	H	H	941 + 1336	0	on	
L	L	L	H	H	H	941 + 1477	#	on	
L	L	L	L	H	H	941 + 1633	D	on	

TABLE 2
Generation of tones

required frequency (Hz)	divisors of prog. divider	overall divisor	dividing error (%) (4.78 MHz crystal)	dividing error (%) (3.58 MHz crystal)
697	41 (×12) + 40 (×2) =	572	+0.03	–0.17
770	37 (×14) =	518	–0.01	–0.22
852	33 (×12) + 36 (×2) =	468	–0.02	–0.23
941	30 (×12) + 32 (×2) =	424	–0.04	–0.25
1209	18 (×16) + 21 (×2) =	330	–0.04	–0.25
1336	16 (×16) + 21 (×2) =	298	+0.16	–0.05
1447	15 (×18) =	270	–0.01	–0.22
1633	13 (×16) + 18 (×2) =	244	+0.08	–0.13

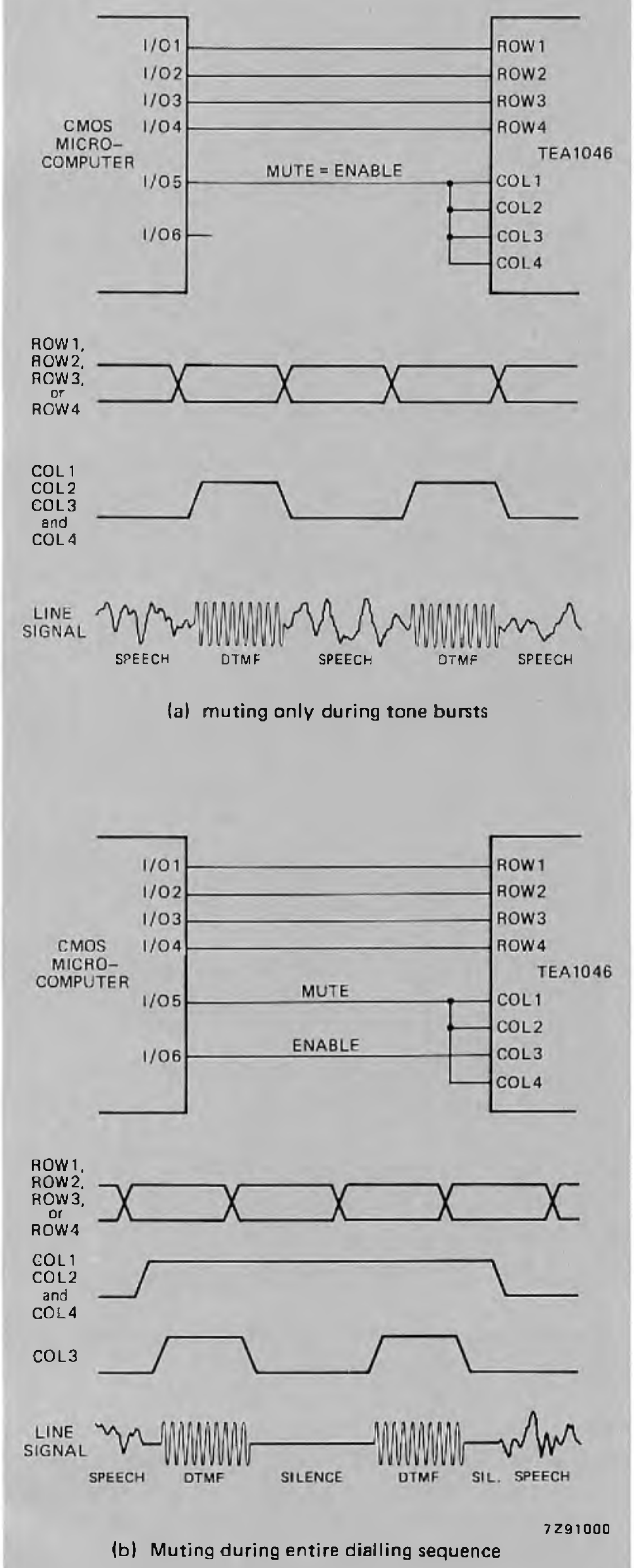


Fig.3 Muting in the microcomputer mode

Figure 3(a) shows the waveforms when pins 20 to 23 are all connected together and held HIGH by the microcomputer during dialling. The signal on the line is the same as when a keyboard is being used.

Figure 3(b) shows how the speech can be muted throughout a dialling sequence. Tone selection input pin 22 acts as the enable input for the tone generator whilst the speech signal on the line is muted by setting pins 20, 21 and 23 HIGH. In this way, a minimum number of microcomputer I/Os are used to ensure that any speech on the line remains muted throughout a dialling sequence instead of only during the tone bursts.

Oscillator and prescaler

Figure 4 is a simplified diagram of the oscillator and prescaler. To ensure sufficient accuracy of the DTMF frequencies (Ref. 1), a crystal oscillator is used in conjunction with either a 4.78272 MHz crystal as used with our range of bipolar DTMF diallers, or a 3.579545 MHz crystal as used for an NTSC chrominance subcarrier oscillator in tv. The oscillator is a single-pin Pierce parallel type which requires a 1 MΩ external resistor in parallel with the crystal for correct biasing.

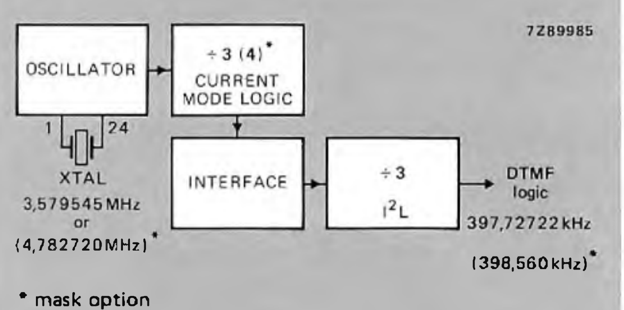


Fig.4 Oscillator and prescaler

A stable 398 kHz clock for the DTMF generator is obtained either by dividing 4.78272 MHz by 12 to obtain 398.56 kHz, or by dividing a 3.579545 by 9 to obtain 397.727 kHz. To this end, the oscillator is followed by a mask-programmable divider by 3 or 4, consisting of four latches with series gating. Current-mode logic is used for this first divider because, at these frequencies, I^2L is too slow and consumes too much power. The second divider has a fixed ratio of 3 and uses I^2L. When the 3.579545 MHz crystal is used, all the tone frequencies are reduced by 0.21%. Even with this excess inaccuracy, the CEPT recommendation of ±1.5% over the entire current and temperature range is easily met when the errors shown in Table 2 are added to the errors due to spread and ageing of the crystal.

Sinewave synthesis

The outputs from the divider for the higher and lower-frequency tone groups are used to synthesise stepped sinewave approximations as shown in Fig.5. Each half-cycle of the tone waveform comprises seven discrete current amplitudes for a tone from the lower-frequency group, and nine discrete current amplitudes for a tone from the higher-frequency group. As shown in Fig.6, each amplitude increment is generated by progressively switching-on weighted current sources for the duration of each step of the sinewave. The frequency of the tone is varied by changing the duration of the steps. To switch the current sources, squarewaves are generated with a period of 1/14th or 1/18th of the required tone frequency period by using scalers following the programmable dividers. Ideally, these scalers should have a length of 1/14th or 1/18th of the divisors given in Table 2. In most cases however, this would require scaling by non-integer divisors which is impractical. To minimize frequency errors, some of the steps of the synthesised waveform are therefore made a few clock periods longer or shorter than the others. For example, the 852 Hz tone requires the clock frequency to be divided by 468 and the duration of each level increment should be 468/14 = 33.43 clock periods. As shown in Fig.5, the problem is solved by making the duration of 12 of the level increments equal to 33 clock periods, and making the duration of the highest and lowest level increment equal to 36 clock periods. The d.c. content of the composite output current is compensated by two d.c. current sources, each having half the amplitude of the sum of the weighted current sources.

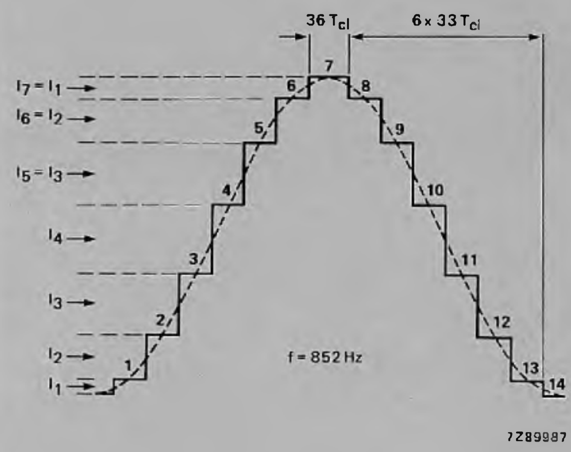


Fig.5 Waveform of synthesised sinewave from DTMF DAC

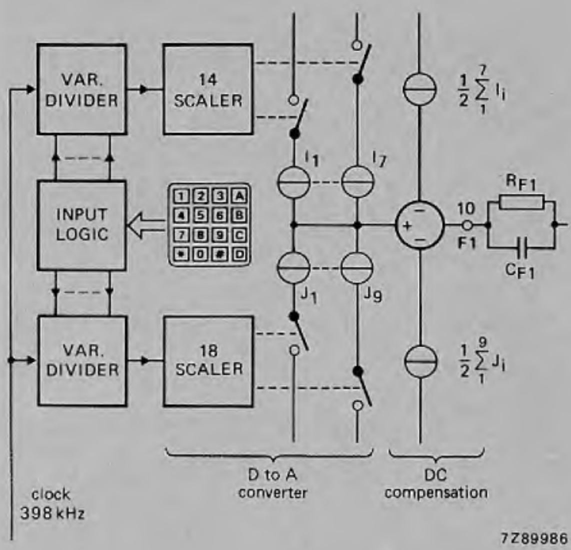


Fig.6 Block diagram of the DTMF generator

DTMF reference circuit

Since the total spread of the level of the dialling tones on the line must not exceed ±2 dB when all variables are taken into account, it is necessary to have a temperature-compensated reference level which is also independent of the line current. In the reference circuit, an unswitched direct current I_0 is derived from the same supply rail that supplies the current sources used for generating the group of lower-frequency tones. This current flows into an external resistor R_{TLS} which is connected between pin 18 and the common return. The value of the resistor can be varied to set the level of the tones. The voltage developed across R_{TLS} is internally fixed at:

$$I_0 R_{TLS} = 0.11 \frac{kT}{q} \ln \left(\frac{I_d}{I_0} + 1 \right) + 2 \frac{kT}{q} \ln 3$$

The first term is a fraction of a junction voltage across an internal reference diode and, as shown in Fig.7, it has a negative temperature coefficient. The second term is an offset voltage obtained from two internal junctions with unequal emitter areas and has a positive temperature coefficient. The sum of the two terms is a temperature-compensated 131 mV reference voltage for the tone DACs which also appears at pin 18. External resistor R_{TLS} at pin 18 can be varied between 35400 Ω and 62900 Ω to set the current sources of the DACs so that the amplitude of the dialling tones on the line meets the requirements of various telephone authorities. A 2 dB pre-emphasis of the higher-frequency tones is determined internally by weighting of the current sources. The amplitude of the tones at pin 1 terminated with 600 Ω is:

lower-frequency tones: $V_{L\,rms} = 13900/R_{TLS}$

higher-frequency tones: $V_{H\,rms} = 1.26\,V_L$.

For convenience, the tone levels are plotted as functions of R_{TLS} in Fig.8.

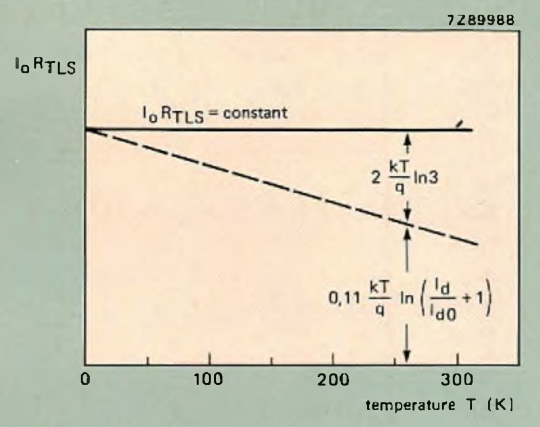


Fig.7 Temperature-compensated reference voltage

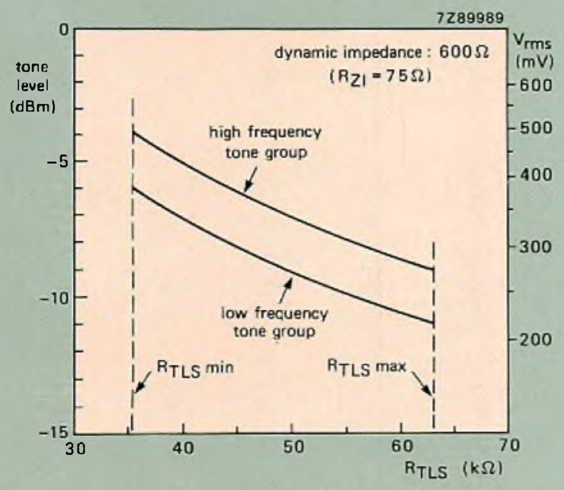


Fig.8 Tone level as functions of R_{TLS}

Filtering

To accord with the CEPT203 recommendations regarding distortion and spurious content of the DTMF tones (Ref.1), the higher order harmonics of the approximated sinewaves are attenuated by a passive second-order low-pass RC filter. A filter with two real poles is used because:

- Such a filter requires only two connections to the IC. Active filters such as those according to Sallen and Key require an additional pin.
- The time-constants can be so chosen that one of the poles can also be used as a low-pass filter in the microphone channel. This prevents instability due to crosstalk between the microphone and telephone wires in the handset cord.

The first section of the filter is formed by the RC network at the outputs of the tone DACs at pin 10. It has a –3 dB corner frequency of 2.2 kHz. The second section consists of an internal resistor and a capacitor at pin 9. It has a corner frequency of 4 kHz and filters both the DTMF tones and the microphone signal. Figure 9 shows Bode diagrams for both sections of the filter.

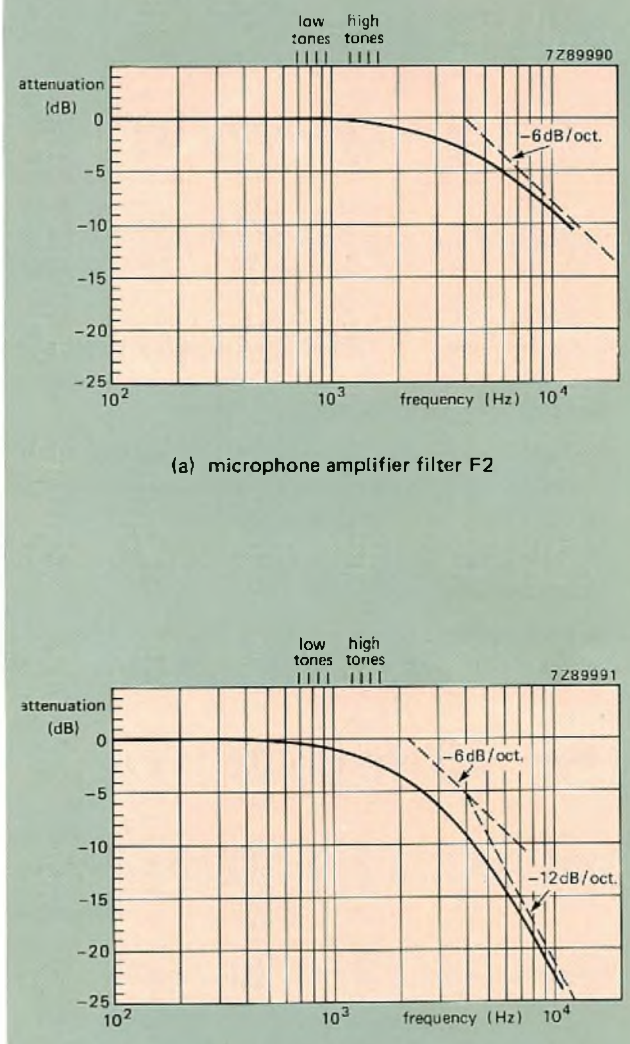


Fig.9 Bode diagram for the filters

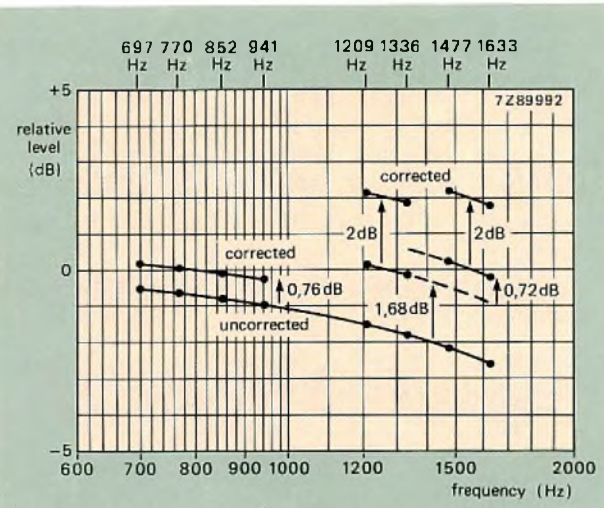


Fig.10 DTMF filter attenuation and correction

As shown in Fig.10, a disadvantage of this filter is that the attenuation in the passband over the fundamental frequency range of the tones varies from 0.55 dB at 697 Hz to 2,58 dB at 1633 Hz. This is unacceptable and is compensated internally in the following three steps:

- The amplitude of the four lower-frequency tones is increased by 0.76 dB, this being the average attenuation of the filter between 697 Hz and 941 Hz.
- The amplitude of the four higher-frequency tones is increased by 1.68 dB, this being the average attenuation of the filter between 1209 Hz and 1336 Hz.
- The amplitude of the two highest frequency tones is further increased by 0.72 dB.

After these corrections have been applied, the filter causes a theoretical deviation of only ±0.2 dB from the levels required for the two tone groups.

Microphone preamplifier

The microphone preamplifier provides the first stage of gain between the microphone inputs (pins 13 and 14) and the buffer input at pin 9. Requirements for the circuit are:

- It must have a low-resistance (few kΩ), symmetrical input with a high common-mode rejection ratio for use with magnetic or dynamic microphone capsules
- It must have a higher-resistance (few tens of kΩ), asymmetrical input for an electret microphone capsule with built-in FET impedance converter (source follower)
- To compensate for various microphone sensitivities, the overall gain between the microphone input and the line must be externally adjustable within the range 35 – 42 dB (electret microphone) or 35 – >50 dB (magnetic or dynamic microphone). The gain adjustment must not influence the input impedance. This is especially important for the electret microphone.

The circuit in Fig.11 meets all these requirements. It is basically a voltage to current converter with capacitively-coupled differential inputs and a d.c. biased output transistor. When the internal mute signal occurs during dialling, an electronic switch transfers the bias current to a similar voltage to current converter so that an almost click-free transition is made from the speech mode to the dialling mode and vice-versa.

For reasons of symmetry, $R_6 = R_7$ and, for a high common-mode rejection ratio, $R_3 + R_5 = R_4$. Since output filter resistor R_{F2} is equal to R_5, the voltage gain of the preamplifier when $f \ll 1/2\pi R_{F2}C_{F2}$ is:

$$A_V = \frac{V_u}{V_m} = \frac{R_3 + R_4 + R_5}{R_6 + R_7}$$

To ensure stability, the maximum overall gain between the microphone inputs and the line is 50 dB. The gain is almost independent of temperature and line current.

Fig.11 Microphone preamplifier and muting

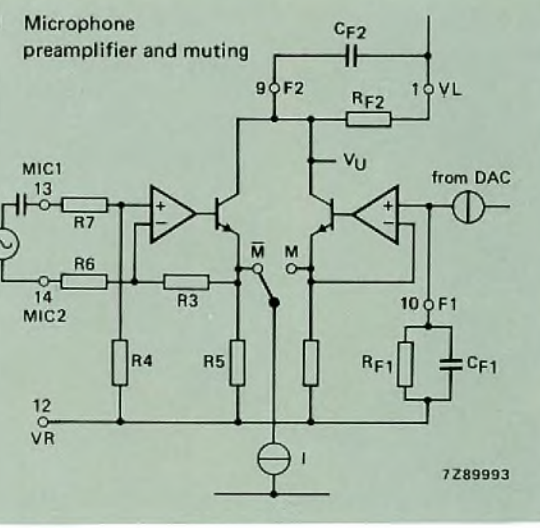


Figure 12 shows how different types of microphone can be connected to the microphone inputs at pins 13 and 14. A dynamic or magnetic microphone capsule is connected as shown in Fig.12(a). The symmetrical input resistance R_{i1} is 4 kΩ. By varying R_{MS} between 18 kΩ and 0, the overall voltage gain between the microphone terminals and the line can be adjusted within the range 35 dB to 50 dB to compensate for various microphone sensitivities. A microphone termination resistance of less than R_{MS} + 4 kΩ can be obtained by using a parallel resistor R_{MP}.

An electret microphone capsule with built-in source follower is connected as shown in Fig.12(b). The asymmetrical input resistance R_{i2} is 22 kΩ, which is sufficiently high compared with the output resistance of the FET (about 1 kΩ). By varying R_{MA} between 19 kΩ and 3.9 kΩ, the overall voltage gain between the microphone input and the line can be adjusted within the required range of 35 dB and 42 dB without affecting the input resistance.

Active output stage

As shown in Fig.2, the dialling tones or microphone signal from the electronic changeover switch are externally filtered at pin 9 and then applied to a current converter with two outputs. One output is the anti-sidetone output at pin 8 and the other drives the active output stage shown in Fig.13. This circuit determines the impedance presented to the lines by the IC and also amplifies the microphone signal or dialling tones to a level suitable for transmission.

Impedance setting function. The active output stage behaves as a voltage source with a dynamic resistance (R_a). An important advantage of this voltage source configuration is that power is not dissipated in the dynamic resistance as it would be by a lumped resistor of the same value in parallel with a current source. The dynamic resistance of the circuit is adjustable by external resistor R_{ZI} according to the expression:

$$R_a = 8.93\,R_{ZI}$$

in which the constant 8.93 is determined by internal resistors and R_{ZI} is the value of the external resistor connected to pin 7. The total impedance (R_i) presented to the lines, consisting of R_a in parallel with the 7 kΩ equivalent resistance of the remainder of the IC (R_p), can be set within the range 600 to 900 Ω by varying R_{ZI} between 75 Ω and 120 Ω. Apart from facilitating adjustment of the circuit impedance, resistor R_{ZI} has to be external because the spread of R_i and gain with an internally diffused resistor would be unacceptable.

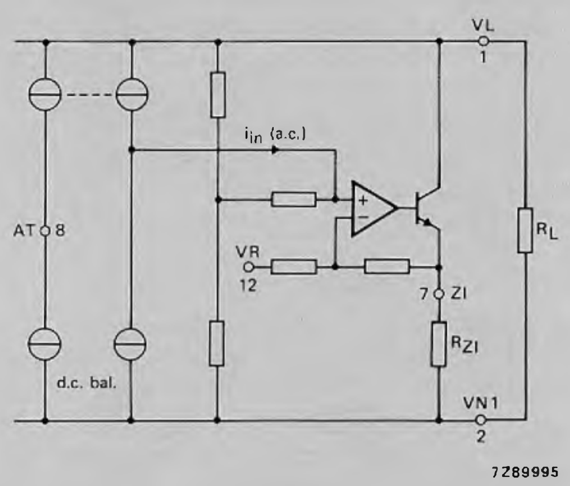
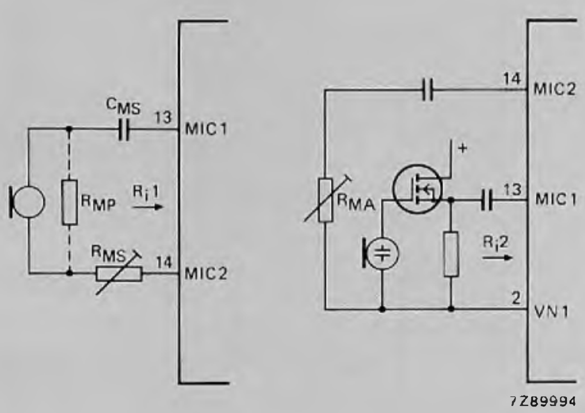


Fig.13 Active output stage



(a) (b)

Fig.12 Connection methods and gain setting
(a) dynamic or magnetic microphone

$$\frac{A(R_{MS} > 0)}{A(R_{MS} = 0)} = \frac{4}{4 + R_{MS}\,(k\Omega)}$$

(b) electret microphone with built-in source follower

$$R_{i2} = R_4 + R_7 = 22\ k\Omega$$

$$\frac{A(R_{MA} > 0)}{A(R_M = 0)} = \frac{22 + R_{MA}\,(k\Omega)}{22 + 11R_{MA}\,(k\Omega)}$$

Amplification function. The amplitude of the signal applied to the line is:

$$V_L = \frac{i_{in}\,R_t\,R_L}{R_L + R_i} = \frac{V_o\,R_L}{R_L + R_i}$$

in which R_t is the transimpedance (V_o/i_{in}) of the output stage which is fixed at 145 kΩ by internal resistors. An advantage of this active output stage is that a d.c. bias of only 3.65 mA is required to allow an undistorted sinusoidal signal with a level of +6 dBm (R_L = 600 Ω) to be transmitted. With a passive output stage, twice as much d.c. bias would be required to transmit the same signal level.

Earpiece amplifier

As shown in Fig.2, the earpiece amplifier receives its input from the line via the anti-sidetone network between pins 1 and 4 and an electronic muting switch. Its output is applied to the earpiece at pin 5. The amplifier is non-inverting type with a class-A output stage and a fixed gain of 20 dB defined by internally applied feedback. During dialling, the mute signal generated by the keyboard logic operates the electronic muting switch and reduces the gain of the circuit by at least 60 dB. Internal symmetrical soft clipping at the output limits transients which could cause irritating clicks in the earpiece.

The peak value of undistorted output voltage obtainable from the amplifier is determined by the d.c. bias voltage at the output. This is the reference voltage (about 1 V). The peak value of undistorted current available from the amplifier is fixed by the direct current bias of about 3.5 mA. The efficiency is maximum with a 350 Ω earpiece. If a lower impedance earpiece is used, the bias current limits the available power. If a higher impedance earpiece is used, the bias voltage is the limiting factor.

Anti-sidetone circuit

An alternating current, which is equal to the input current to the active output stage, is available as an anti-sidetone output at pin 8. This signal is used to cancel the signal from the microphone which would otherwise be reproduced as sidetone in the earpiece. The operating principle of the system is illustrated in Fig.14.

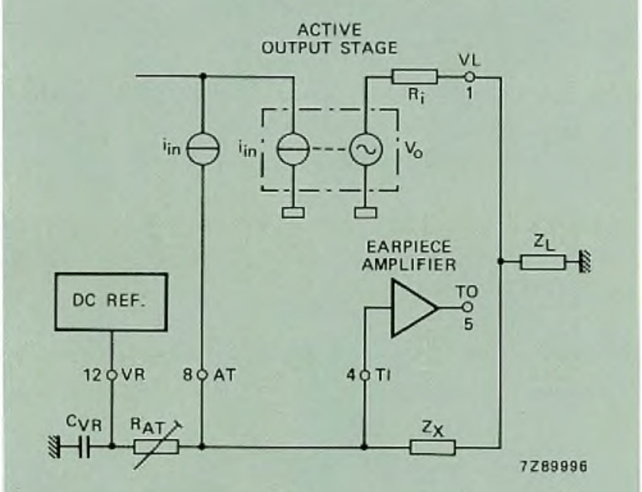


Fig.14 Anti-sidetone principle

When the IC is connected to a line pair with a complex impedance Z_L, the microphone signal which appears at the earpiece amplifier input (pin 4), is cancelled when:

$$Z_x = k \frac{R_i Z_L}{R_i + Z_L}$$

where

$$k = R_t \frac{R_a + R_p}{R_a R_p} = \frac{R_t}{R_i} = 237$$

in which R_t is the transimpedance of the active output stage (145 kΩ), and Ri = Ra//Rp which is the dynamic resistance of the total circuit (670 Ω//7 kΩ = 611 Ω).

Since Z_L varies considerably with the type and length of the lines, an optimum value of Z_X can only be chosen for one type and length of line. For the circuit in Fig.2, a twisted-pair copper cable with a diameter of 0.5 mm and a length of up to 10 km has been chosen. The impedance Z_X for a line of average length (5 km) is approximated by the network comprising R_{A1}, R_{A2} and C_A; it results in sidetone attenuation of more than 10 dB over the entire cable length.

External resistor R_{AT} sets the signal level at the input of the earpiece amplifier to suit the sensitivity of the earpiece. As shown by the formula for Z_X, the value of R_{AT} does not influence the sidetone cancellation.

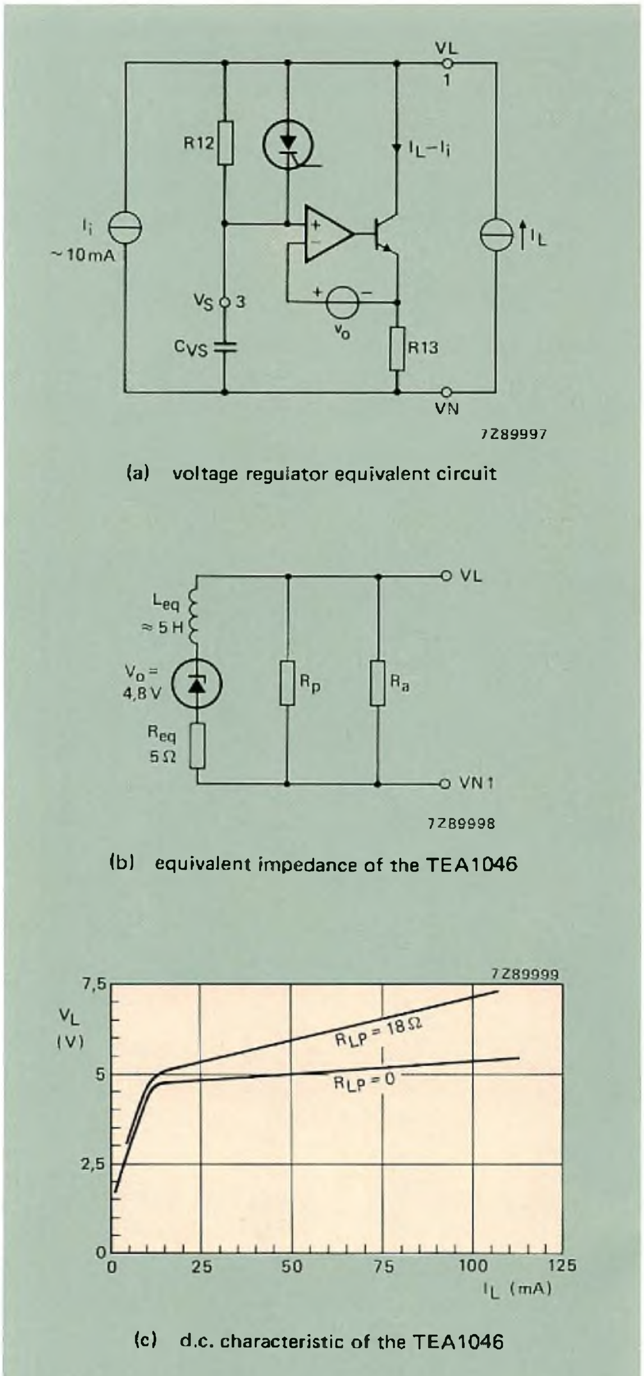


Fig.15 Voltage regulator

Voltage regulator

The basic circuit of the line voltage regulator is given in Fig.15(a). If the line current exceeds the current required to operate all the integrated functions ($I_i \approx 10\,mA$), the shunt voltage regulator sinks the excess current and sets the average voltage applied to the IC.

The voltage regulator is basically a reactance circuit. The RC network $R_{12}C_{VS}$ smooths line voltage V_L so that the regulator does not respond to a.c. signals. A direct current $I_L - I_i$ flows if an internally defined reference voltage V_o is exceeded. The equivalent impedance of the TEA1046 is shown in Fig.15(b).

Static behaviour of the circuit is expressed by:

$$V_L \sim V_o + (I_L - I_i)R_{13}$$

in which V_o is about 4.8 V at 25 °C and R_{13} is the slope of the static I_L/V_L characteristic shown in Fig.15(c).

The dynamic impedance of the regulator is equivalent to a resistor in series with a simulated inductor:

$$Z_R(\omega) = R_{eq} + j\omega L_{eq}$$

in which R_{eq} is about 5 Ω and $L_{eq} \sim R_{13}C_{VS}$ and is about 5 H.

In the telephone audio frequency range (300 Hz to 3400 Hz), this impedance is very much greater than the dynamic resistance of the active output stage.

An internal thyristor switch short-circuits R_{12} during switch-on. This minimises the start-up time and limits the voltage overshoot applied to the line.

Polarity guard and line transient suppression

Current surges on the telephone lines can cause voltage transients that may damage or destroy the IC. The circuit of Fig.2 therefore incorporates a bridge rectifier which not only acts as a polarity guard but also incorporates two zener diodes. During normal conditions, only one of the zener diodes is conducting. If the voltage across the circuit exceeds the zener voltage, the other zener diode conducts to protect the IC. The 18 Ω resistor R_{LP} in series with the bridge rectifier limits the current that can be drawn by the IC.

ACKNOWLEDGEMENTS

The authors wish to thank D.J.G. Janssen and R.J. van de Plassche for their helpful discussions and contributions to the design of the TEA1046.

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Circuits for modular design of consumer and industrial products – the CLIPS system

E. T. KEVE

Many design projects meant for 8-bit implementation, even for entirely different applications, look remarkably similar. Consider four different and seemingly unrelated application areas: telecommunications, industrial control, instrumentation, and home video/audio. How are they alike?

Well, let's further consider some typical 8-bit applications within these areas:

- a telephone set
- a process control monitor
- a digital voltmeter
- a tv controller.

None of these applications will require a high data throughput, but all will probably be pushed for space; serial data communication is therefore a must. The use of serial links also results in an overall cost saving compared to parallel data connections. Fewer pins means smaller packages on smaller boards and cheaper encapsulations; less wire means cheaper interconnections and simpler layouts.

In addition, each application will usually require some form of intelligent control, an 8-bit single-chip microcomputer being the best candidate. Finally, all will include a number of identical peripheral circuits: additional RAM, I/O expanders, LCD drivers etc. In fact, the only difference between them is to be found in the dedicated peripheral circuits which perform the specific application-oriented tasks, and the software resident in the microcontroller. The telephone set, for example, will require a ringer circuit, transmission circuit and, possibly, a DTMF dialler; the voltmeter, (or process controller) will need D/A and A/D converters; the tv will have to include special tuning circuits and a remote control facility.

In all these application areas, the combination of general-purpose and dedicated circuits can be found. Introducing standard products and design techniques for applications as diverse as these, leads to wide-ranging benefits:

- The designer is able to adopt a 'building block' approach, going straight from block diagrams to circuit implementation knowing that compatibility and interfacing problems are already solved.

  Such an approach leads to a straightforward, structured hardware design. Each device on a circuit board can immediately be recognised as fulfilling a particular function. Upgrading is then done simply by adding extra devices to perform extra functions. One-to-one recognition of functions-to-ICs also means easier diagnostics and repair, as functional problems can be immediately linked to particular devices.

  Structured hardware also leads naturally to modular software. If the same types of microcomputer are used continuously, a library of reusable software modules can be built-up to cover control needs for a variety of peripherals in many different application areas.

- The manufacturer could build up a family of products by simply adding or subtracting dedicated peripheral circuits.

- The component manufacturer would be able to produce large volumes of the same group of ICs for a wide variety of applications, thus keeping costs low: the dedicated devices would be added to the component range to match the demand from each product area.

THE CLIPS CONCEPT

To realise this 'building block' approach, a complete range of products is needed. To be at all feasible this range must include:

- low-cost industry-standard microcomputers
- general-purpose peripheral circuits
- application-specific peripheral circuits
- a versatile interconnecting bus.

Philips is introducing such a range under the collective designation CLIPS (Consumer and Industrial Peripheral Set). The aim of CLIPS is to create a design environment in which devices can be 'clipped' on to a bus. Therefore, numerous ICs with the necessary serial I/O are being released to interface with the Inter-IC (I^2C) bus (Ref.). Incorporating the hardware capable of supporting the I^2C bus directly on-chip enables the designer to concentrate on the functional aspects of a system without having to worry about the communication link between the different system parts.

To cover a broad range of applications, including battery-powered operation, most CLIPS circuits will be CMOS devices; because of its high noise immunity CMOS is ideal for noisy industrial and automotive applications. The specification of the I^2C bus, however, permits devices using any type of technology (CMOS, NMOS or I^2L) to be connected together in one CLIPS system. Besides standard DIL, all CLIPS circuits are also available in SO (Small Outline) packages.

CLIPS circuits are also adaptable to further integration. A typical system can be considered as various general purpose devices – microcomputers, LCD drivers, I/O expanders, extra memory (Fig.1), with specific application-oriented devices tagged on. If a particular CLIPS system has been successfully used in large numbers, the specific combination of devices can be integrated with the existing general-purpose devices. Separate microcomputer and A/D converter ICs, for example, can easily become one unit, with circuit functions and system performance remaining the same.

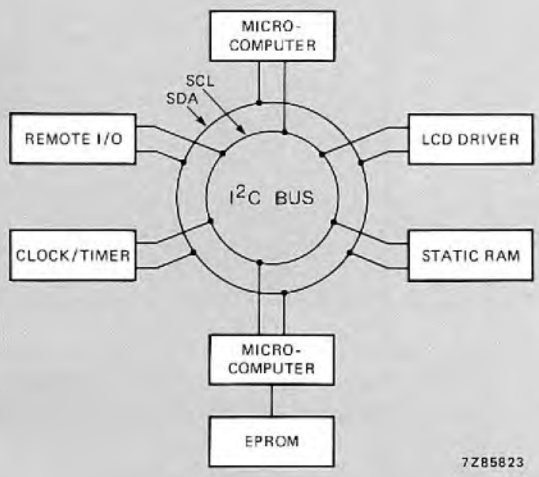

Fig.1 A typical CLIPS configuration showing the general purpose circuits present in most 8-bit designs

Standardisation within CLIPS systems leads to more economical use of any software developed. Modular application software can be used again and again in a variety of applications, carrying out the general-purpose functions within a CLIPS system. Subroutines associated with implementing a particular CLIPS IC will also be included in the support material for that circuit as an aid to designers.

CLIPS APPLICATIONS

The range of CLIPS applications is as wide as the microcomputer field; the following examples merely give an indication of the range and possibilities of CLIPS circuits.

Video entertainment and information centre

Video games, teletext and home computers are transforming tv sets from mere receivers of broadcast programmes to complete entertainment and information terminals. As extra features are introduced the tv receiver must be able both to accept them and integrate them into a user-friendly home appliance.

Let us consider a top-of-the-range tv set including at least a remote control, a games interface, and viewdata capability. The Fig.2 block diagram shows how these, as well as other functions, might be served by a CLIPS system. Using CLIPS, it is possible to relate each block to an equivalent IC.

A point worth noting is the use of the remote control receiver within the system. Since the microcontroller associated with this device has to be active all the time, it is best implemented in CMOS. The I^2C bus supports devices in any of the major manufacturing technologies and therefore allows each IC in the CLIPS system to be realised in the technology best suited to its function.

The D^2B interface (see Ref.) enables the tv to be linked to other 'intelligent' appliances, such as a VCR or video camera, which themselves benefit from CLIPS implementation.

Subscriber telephone set

The CLIPS implementation of a subscriber telephone set includes all the functional units usually found in top-of-the-range models (Fig.3). The RAM is additional to that found

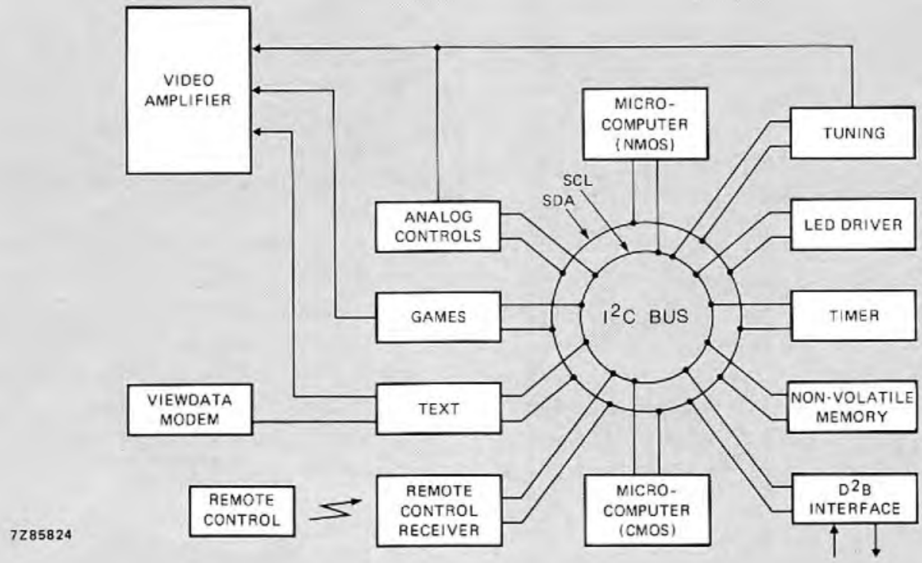

Fig.2 The many functions required for a modern tv can be easily implemented with CLIPS circuits. The specification of the interconnecting I^2C bus also allows CMOS, NMOS and I^2L devices to reside on the same bus

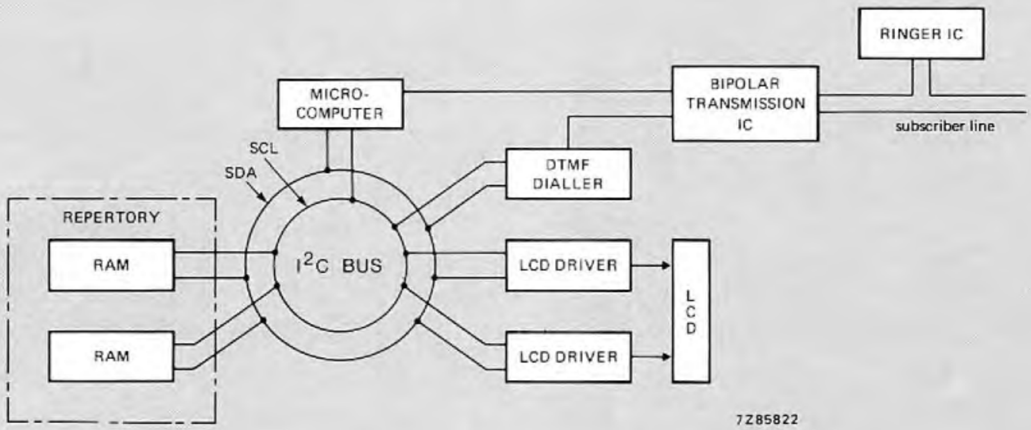

Fig.3 Telephone set is also well suited to CLIPS implementation. The additional RAM can be used to store telephone numbers for repertory dialling

on the microcomputer and can be used for such functions as repeated-call or repertory dialling. The DTMF dialler can be added whenever the local exchange converts to a tone-dialling system. The LCD and its drivers are also optional.

By providing the general-purpose devices within a telephone set, the CLIPS system gives the designer a great deal of freedom. Although the application demands that the interfacing ringer and bipolar transmission ICs are not connected directly to the I^2C bus, the other devices which are can be configured to match a wide range of specifications and demands from both telephone authorities and subscribers. We have already built-up a library of telephony-related software modules to help designers meet the various requirements they are likely to encounter.

The CLIPS concept is ideal to respond to the fragmented nature of present telephony applications. Each authority has its own requirements; each consumer has his own preference. To design a 'cover-all' telephone set is almost impossible. CLIPS offers a framework within which a designer can work effectively.

Industrial applications

In industrial environments, CLIPS systems (Fig.4) can perform a variety of functions. For intelligent machine control, level sensing, pressure monitoring – the implementation of CLIPS systems can provide distributed processing and microcomputer control on-site, where it is needed.

For analog applications, a whole range of sensors can be used. These can then be connected to an A/D converter to provide data for devices connected to the I²C bus. Keyboard encoding circuitry provides user-access to the system, with user-output sent via the display driver. For control purposes, evaluated data can be fed back to actuators such as temperature controls or electromechanical switches via another D/A converter.

For digital applications, data can be sent or received via remote I/O circuits. Due to the nature of the I²C bus, remote I/Os can be situated anywhere within the control environment; other CLIPS circuits can likewise be placed exactly where they will best perform their allotted task. Within a cabinet, wiring can therefore be drastically reduced. Instead of trailing parallel wires between devices, remote I/Os can be used to convert the parallel device output into serial bus input almost immediately – saving space, saving wire, and making life much easier for the service engineer. The remote I/O, for example, would be connected near simple on/off devices: to drive valves or relays, activate emergency sirens, or read keys and drive LEDs.

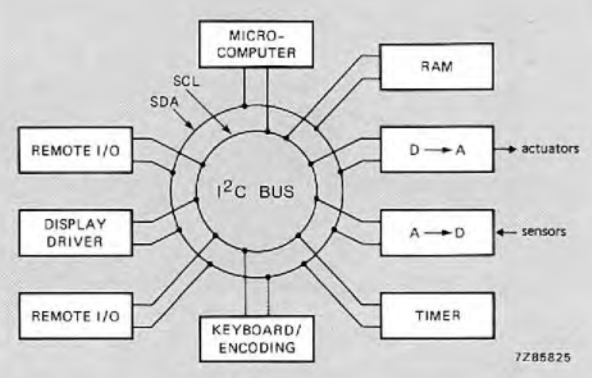

Fig.4 In industrial control applications, a CLIPS system can interface with many types of sensor and actuator, both digital and analog. Connection to a higher level of control is also possible via the controlling microcomputer

When used in an industrial environment, CLIPS systems can also be connected to higher levels of control, using the parallel ports of the microcontroller for example. A CLIPS system can therefore cover the complete range of industrial applications from hand-held, battery-powered measuring instruments to the first stage of a large scale process control system.

For low-volume Industrial designs, piggy-back versions of both the NMOS MAB8400 and CMOS PCF8500 single-chip microcomputers can be used. By connecting industry-standard EPROMs to these devices, designers can still build systems using CLIPS ICs.

CLIPS ICs – THE BUILDING BLOCKS

The previous examples highlight the CLIPS philosophy – general-purpose devices complemented by application-specific peripherals. By examining some CLIPS ICs in more detail, the interfacing simplicity of CLIPS circuits will also become apparent.

Static CMOS RAM (1024 bit)

The PCD8571 static CMOS RAM incorporates an interface for direct connection to the I²C bus. Its low data retention voltage (min 1.0 V) and low standby current (typ. 50 nA) make this memory suitable for numerous CLIPS applications.

In addition to acting as a general-purpose RAM expansion for CLIPS microcomputers, the PCD8571 can also find service as a more dedicated device. In the telephony application (Fig.3), for example, it can provide the extra RAM to store telephone numbers for repertory dialling. In radio or TV, it can be used to store channel presets.

Three hardware address pins are included on the PCD8571 to extend the specific device address. Up to 8 of these RAMs can be connected to the I²C bus, allowing the amount of memory used and its eventual purpose to rest with the designer.

Auto-increment for memory word addresses also facilitates efficient write/read of blocks of data; for example, channel preset information in a tv application would usually require 2 consecutive data bytes.

When connected to the I²C bus, the PCD8571 acts as a slave transceiver. Figure 5 takes us through the READ and WRITE modes associated with transfers to and from a master device. To ensure that a transfer occurs, data is sent in groups of 8-bits accompanied by an acknowledge bit, with failure to acknowledge a byte transfer resulting in the transfer being terminated by the master device.

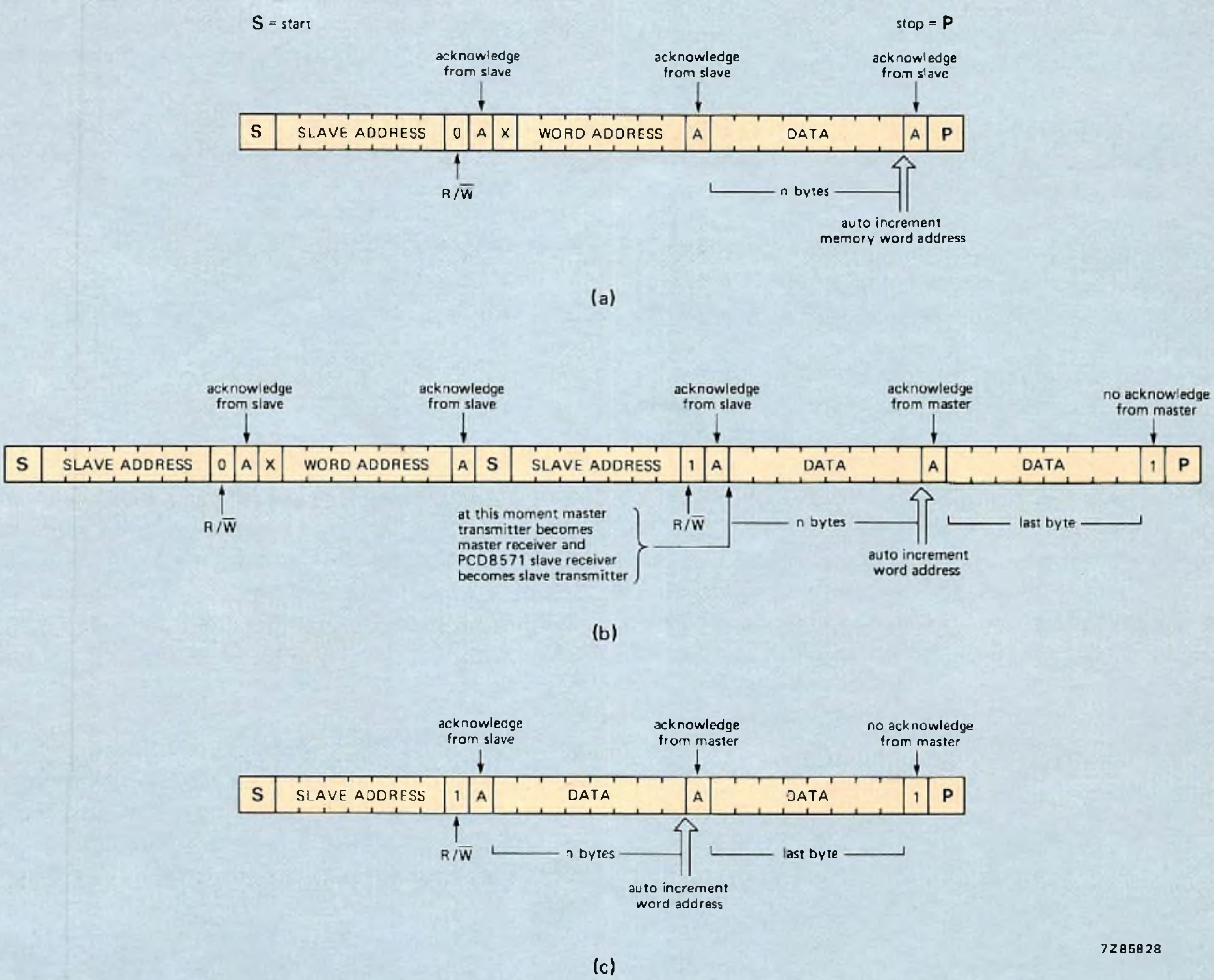

Fig.5 The I²C bus configuration for different READ and WRITE cycles of the PCD8571 static CMOS RAM. (a) WRITE mode. The master transmitter transmits to the PCD8571 which is at that moment a slave receiver. (b) WRITE word address, READ data. After setting the word address, the master reads the slave which then becomes a slave transmitter. (c) READ mode. Again the master reads the slave, but this time immediately after the first byte

LCD driver with serial I/O

The PCF8577 is a CMOS silicon-gate LCD driver for driving either 64 segments duplex or up to 32 segments directly. For systems with several PCF8577s, hardware sub-addressing and automatic address incrementing keeps wiring and bus traffic to a minimum.

CLIPS circuits such as LCD drivers require control information from the microcomputer in the form of pointers or control bytes before they can carry out a requested function. This information is provided within the I²C bus addressing procedure.

The first byte of the slave address is used to address all LCD drivers connected to the I²C bus. All PCF8577 devices connected to the bus will therefore respond to this address. Each device then loads the second byte into its control register (Fig.6).

The six least significant bits of this byte constitute the segment byte vector (SBV); they select the device and the segment byte register which is to be loaded next. The upper three bits of the SBV are compared with the preset hardware address of the device. If the address is the same, the device is enabled for loading.

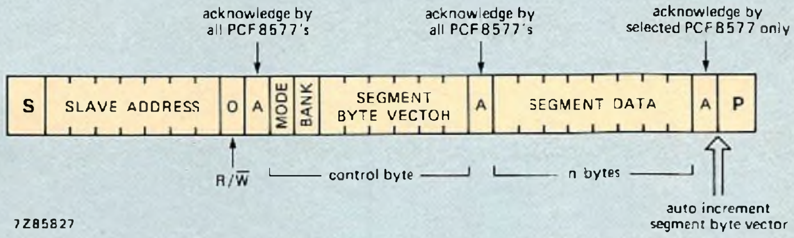

Fig.6 The bus protocol for the PCF8577 LCD driver. After acknowledging the slave address, the second control byte indicates which device is selected and the display mode of that device

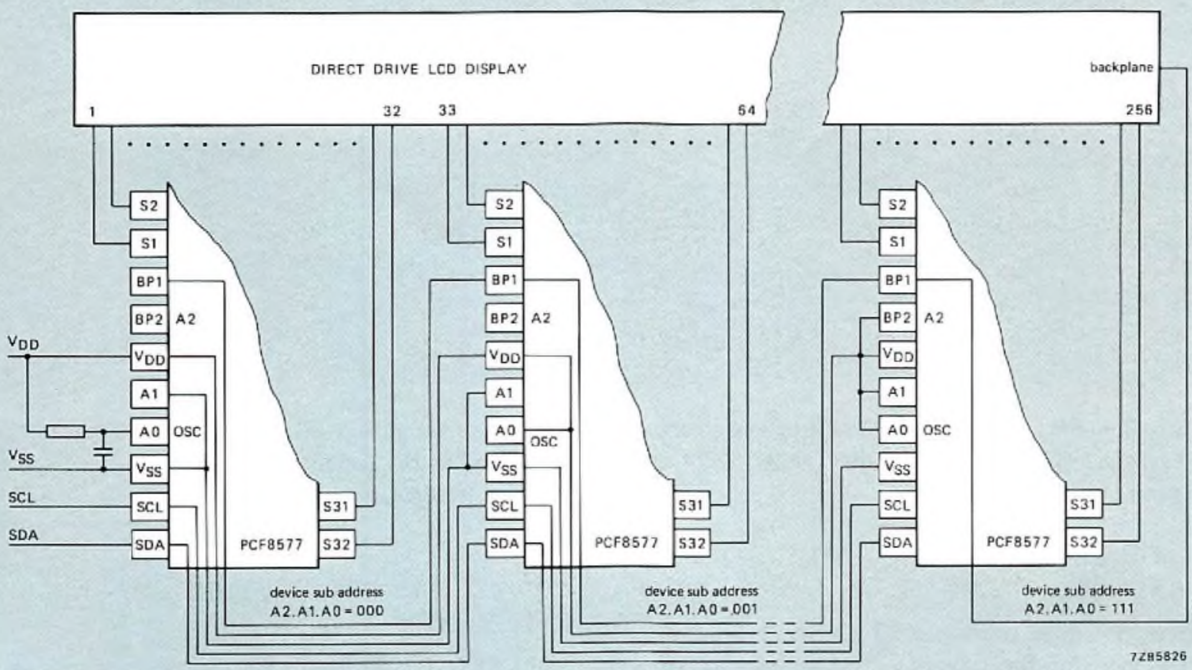

Fig.7 In direct drive applications, expansion to a 256 segment display is possible using 8 PCF8577 units

After each segment byte is loaded, the SBV is incremented automatically. In this way, sequential loading occurs when more than one segment byte is received in a data transfer. The control register also includes two display control bits, MODE and BANK. The MODE bit selects either direct drive or duplex; the BANK bit allows switching between two alternative control stores when the device is in the direct drive mode.

The programmable addressing of the PCF8577 allows the number of LCD drivers connected to a CLIPS system to be increased to a maximum of 8 directly driven units (4 duplex), giving 256 display segments (Fig.7). Therefore, the type of display output can be tailored to suit each member of a product range.

Remote 8-bit I/O

As shown in the industrial control example, remote I/Os within a CLIPS system can act as an interface between the I^2C bus and a digital source. The PCF8574 is an 8-bit I/O circuit. It consists of an 8-bit quasi-bidirectional port and an I^2C interface. This device can also interface microcomputers without a serial interface to the I^2C bus (as a slave function only). The PCF8574 has low current consumption and includes latched outputs with high current drive capability for directly driving LEDs.

The PCF8574 differs from the other CLIPS circuits already described in that it possesses an interrupt line (INT). This line is connected to the interrupt logic of the microcomputer on the I^2C bus. By sending an interrupt signal on this line, the remote I/O can inform the microcomputer if

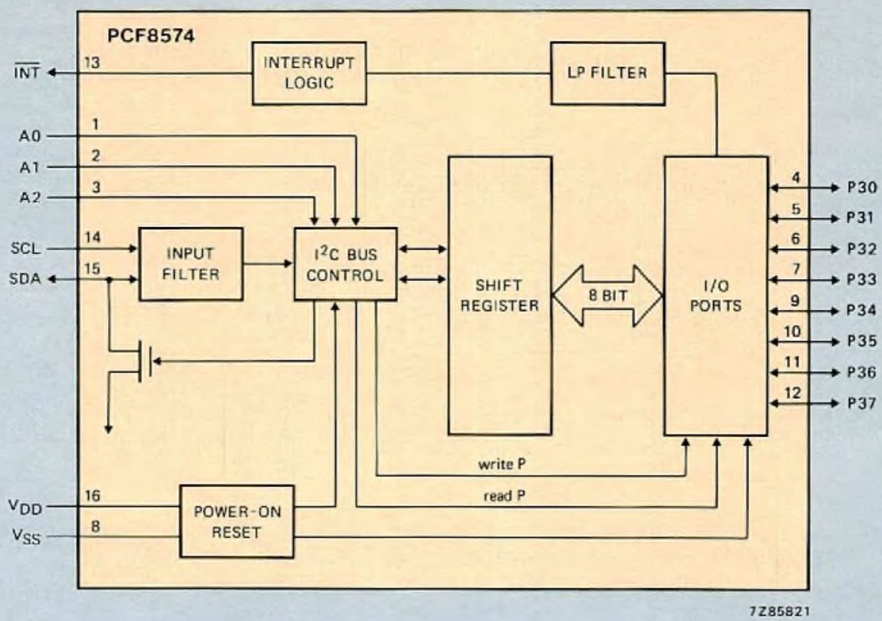

Fig.8 Block diagram of the PCF8574. Note the Interrupt Latch which informs the microcomputer of data on the I/O ports. The microcomputer can then access the expander in the READ mode, via the I²C bus

there is incoming data on its ports without having to communicate via the I²C bus. This means that the PCF8574 can remain a simple slave device (Fig.8).

CLIPS – GENERAL PURPOSE AND DEDICATED ICs

The devices described are some of the general-purpose CLIPS ICs; more of these types of circuit are available. This selection is meant to serve only as an example of the types of function, addressing schemes and read/write actions of devices within a CLIPS system. The dedicated blocks mentioned in the various application examples are also, or soon will become, readily available.

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Microwave integrated circuits — design and realisation

W. GOEDBLOED and S. D. VRIESENDORP

In recent years microwave technology has changed considerably. This has been due primarily to the introduction of miniaturisation techniques which, besides reducing costs significantly, have eased the problems of interfacing between transmission line systems and modern semiconductor devices.

The most popular miniaturisation technique is based on the principle of *microstrip*, which is an adaptation of the open transmission line. With suitable choice of substrate material, microstrip can provide significant size reduction compared with coaxial line and waveguide technology. So microwave integrated circuitry is now a real alternative to the larger, more costly systems.

Miniaturisation, however, brings problems of its own. In finding solutions for them, much expertise has been built up, both from an electrical and from a technological standpoint.

System designers who wish to avail themselves of the advantages offered by microwave integrated circuits (MICs) should become aware of these problems. The aim of this article, therefore, is to present briefly the current state of the art, as well as to indicate what can and cannot be done in MIC technology. This will be done using an example: a currently available device that employs modern MIC techniques. The example, shown in Fig.1, is a balanced downconverter. This example has been chosen principally because its action is well understood, and because it is highly representative of systems currently available in MIC technology. Our discussion will be confined to thin-film technology, principally because of the better line definition it provides and the advantages this offers at higher frequencies.

The system shown in Fig.1 is designed to operate as an r.f./i.f. converter in the 5 GHz region and has a relative bandwidth of about 10%. The r.f. aerial input and local oscillator signals are fed into a 3 dB branch-coupler that divides the power equally between two Schottky-barrier mixing diodes. After conversion, the i.f. signal carrying the information (about 70 MHz) passes through a filter (to remove r.f. components) and is subsequently amplified. An isolator located in the r.f. input port of the branch coupler improves isolation between the r.f. and local oscillator signals. A matching section is provided between the branch coupler and the diodes which themselves must be matched to the source impedance of the i.f. amplifier for maximum delivery of i.f. power.

The following sections describe the design and realisation of the different functions contained within the system.

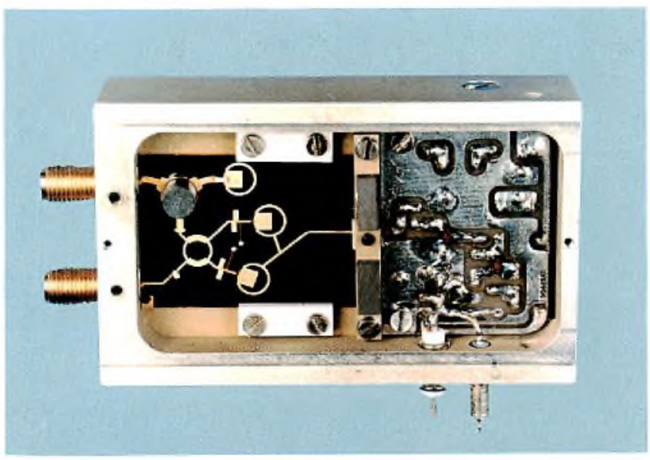

Fig.1 Balanced downconverter designed to operate in the 5 GHz region and having a relative bandwidth of 10%. The system is based on a ferrite substrate allowing an isolator to be integrated directly into the r.f. input of the branch-coupler

This will in fact be done in two stages: firstly, design and production of the circuit pattern and distributed elements (i.e. the microstrip line), and secondly, consideration of add-on components, i.e. functions that are added to the system after the circuit patterns have been formed.

FUNCTIONS IN MICROSTRIP – DESIGN AND MANUFACTURE

Microstrip line

Figure 2 is an enlarged view of the microstrip line showing the approximate field pattern.

Microstrip is in fact a derivative of the open transmission line, one conductor being a narrow strip and the other a ground plane. A dielectric substrate fills the space between them. The higher the dielectric constant (ϵ_r) of the substrate the smaller the microstrip can be, since the effective wavelength decreases with increasing dielectric constant.

The most widely used conductor material is gold (better than 99.99% pure). Gold provides the following advantages:

- chemical inertness, eliminating the need to passivate circuits
- ease of bonding of miniature components and semiconductor devices to the circuit.

The manufacture of microstrip demands a strictly controlled process that does not allow of any intermediate adjustment of circuit parameters. It is thus essential to obtain an accurate specification of circuit dimensions at an early stage.

Governing factors of microstrip behaviour

Discontinuous dielectric: dispersion effects

In a microstrip line the field pattern traverses both the substrate and the surrounding air, and so never exhibits a pure TEM (transverse electromagnetic) field configuration. The air/dielectric combination can only support a hybrid field with components in the propagation direction. This leads to dispersion effects which become more important as frequency increases. The effects of dispersion have in fact been observed at frequencies as low as 1 GHz.

While of relatively minor importance for narrow band systems (i.e. up to 10% relative bandwidth), dispersion introduces severe problems to the design of systems covering more than one octave. In such systems, the non-linear frequency dependence of parameters such as impedance and phase velocity may mean that a function optimised at one frequency fails completely when the function is scaled to another. It is thus essential to adopt a fundamental approach in solving the electromagnetic problem, one that takes account of the axial field components and the transverse currents in strip and ground plane. Such an approach demands the use of sophisticated computer-modelling techniques. However, this in itself is not sufficient, and the design of a function must also include a programme of optimisation carried out on an actual circuit.

Substrate: material properties and thickness

To confine the electric field between the conductor and ground plane, substrates of high dielectric constant (ϵ_r between 8 and 16) and low loss (tan $\delta < 10^{-4}$) must be used, e.g. alumina, ferrite and sapphire. Quartz fulfils only the low-loss requirements but nevertheless it is often used at high frequencies (above about 10 GHz), and where a well-defined dielectric constant is of major importance.

An important factor governing the behaviour of microstrip is the thickness of the substrate. For a given frequency the thickness must be significantly less than half a wavelength (i.e. about an eight of a wavelength), otherwise unwanted high-order modes may be induced in the substrate.

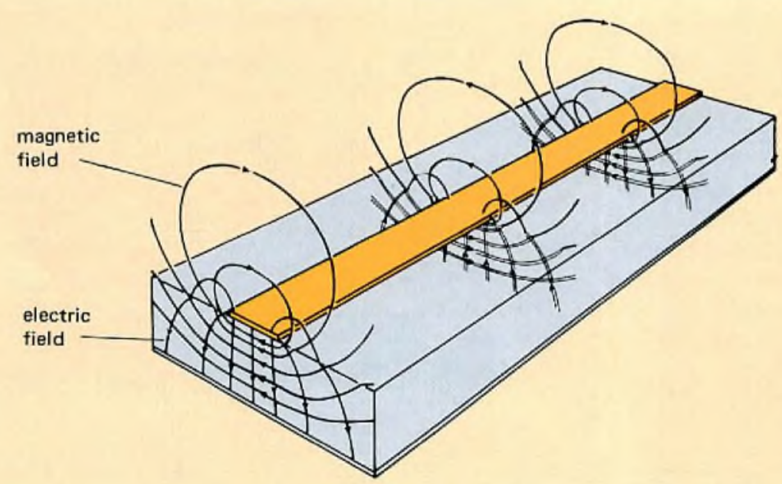

Fig.2 Electromagnetic field pattern in a microstrip line. In this figure, both the electric field and the magnetic field are transverse; this is an approximation that neglects dispersion and is valid only at the lower microwave frequencies. The magnetic and electric fields traverse both the air and the dielectric

Since the wavelength within the dielectric decreases with increasing dielectric constant, substrates of high-dielectric material (e.g. ferrite) may be so thin at higher frequencies (>30 GHz) that their mechanical strength is inadequate. The only alternative is then to use a thicker substrate of lower-dielectric material.

A standard thickness for alumina substrates ($\epsilon_r \approx 10$) is 635 μm, which means that they can be used at frequencies up to about 14 GHz.

Since the dielectric constant of ferrite materials (between 14 and 16) is significantly greater than that of alumina, ferrite substrates must be thinner for the same band of frequencies. Ferrites have a distinct advantage in one major respect: their magnetic properties can be used to produce non-reciprocal devices (such as circulators, isolators etc.). However, these same magnetic properties place a lower limit on the usable frequency, so a range of ferrite materials must be available to cover different bands of the microwave spectrum (see Table 1).

Each ferrite material in Table 1 has been produced to give optimum performance within a specified frequency band. At frequencies below the lower limit (e.g. about 7 GHz in material I) losses are severe, owing to the electron spins within the material becoming misaligned. Note: this can be an advantage in functions with a tendency towards low-frequency parasitic oscillation (e.g. X-band amplifiers), since damping below the lower limit can be 60 dB or more.

TABLE 1
Dielectric materials suitable for use as substrates in microstrip

material	ϵ_r	tan δ ($\times 10^{-4}$) measured at 25 °C 10 GHz	lowest operating frequency (GHz)
sintered alumina (99% pure)	9.7	1	–
ferrite I	15.5	2	7
II	14.8	2	3.4
III	14.4	2	2.2
IV	15.4	2	5
V	14.0	2	1.5
fused quartz	3.8	1	–
sapphire C-axis	11.5	0.3	–
A-axis	9.3	0.9	–
filled PTFE material			
Epsilam 10	10.3	18	–
duroid 5870	2.3	12	–
duroid 6010	10.5		–
polyguide	2.3	5	–

Another factor governing the selection of substrate material is dielectric loss. Ferrite materials induce marginally greater losses than alumina and other non-magnetic materials. These losses are, however, of little significance compared with conductor losses, and ferrite substrates are often favoured because of their ability to support non-reciprocal functions.

Losses can be reduced significantly by using material of high purity and small grain size. This demands, among other things, strict monitoring of raw materials and a well controlled fabrication process.

Conductor strip: properties and geometry

The characteristic impedance Z_0 of a microstrip line depends upon the electrical properties of the dielectric and the geometry of the line, in particular the ratio of strip width w to strip/ground-plane spacing h. Figure 3(a) shows the relationship between Z_0 and w/h for various substrate materials.

Figure 3(b) shows that the wavelength λ_g along a microstrip line is also a function of w/h. The horizontal scale indicates the practical limits imposed on w/h, since a strip that is too wide gives rise to undesired propagation modes of high order, and one that is too narrow is difficult to manufacture accurately.

Because of the extremely high current density within the strip at the dielectric/strip interface, conductor losses account for more than 80% of the total losses in a microstrip line. Significantly less than the conductor losses, although by no means negligible, are the radiation losses caused by currents induced in the strip at the air/strip interface. These currents, which are caused by the electromagnetic field within the substrate, decrease with increasing strip thickness, so radiation losses can be kept low by using a relatively thick conductor layer. For most applications a thickness of at least five skin depths is in fact necessary for reducing radiation losses to tolerable levels. At present, strip thickness is standardised at 10 μm, which at higher microwave frequencies (i.e. X-band and above) is significantly more than the five skin depths required.

Skin depth varies inversely with the square root of the frequency. This means that at high frequency it can become comparable with surface irregularities of the substrate, so the current path at the strip/dielectric interface is longer and losses increase. The surface finish of the substrate is thus an important factor governing conductor losses at high frequency, and a smooth, highly polished surface is preferred. If the surface is too smooth, however, it may impair the adhesion between substrate and conductor.

Much research has therefore been devoted to obtaining good adhesive layers that do not themselves introduce greater losses than would otherwise occur with rougher substrate surfaces.

Discontinuities in the microstrip line (e.g. stubs) can also cause radiation losses, as well as other losses produced by coupling between lines in close proximity. In circuits with many discontinuities, these losses can be considerable.

The above considerations are all reflected in the Q-factor of the microstrip line. High-Q systems, i.e. those exhibiting sharp resonances, are also low-loss systems. Q-factors of 250 are typical of the 50 Ω microstrip line. This is considerably less than the 2000 or more attainable with waveguides, but in most applications the space saved by microstrip more than offsets the increased losses.

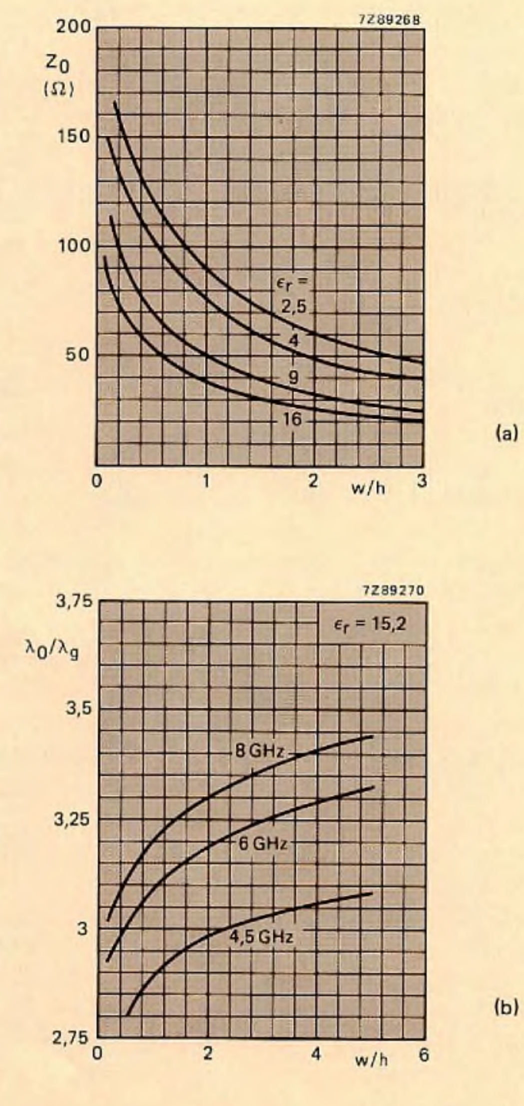

Fig.3 (a) The variation of characteristic impedance (Z_0) with w/h for various dielectric materials; (b) The wavelength λ_g along a microstrip line is strongly dependent upon w/h. This is illustrated by plotting λ_0/λ_g against w/h where λ_0 is the free-space wavelength

3 dB branch coupler

The 3 dB branch-coupler (Fig.4) performs two basic functions: isolation between the two input branches (1 and 4), and an equal (or almost equal) power split between the two output branches (2 and 3). The device exhibits two-fold symmetry about lines a and b. For optimum performance this two-fold symmetry must be preserved during the design phase: changes made at any one port must be duplicated at the other three.

Figures 5(a) and (b) show, respectively, the isolation between the input ports and the power split at the output produced by a branch coupler with optimised performance.

The branch-coupler is equivalent to the multi-hole coupler of waveguide technology and its behaviour is fundamentally the same. Impedance matching is thus essential for optimum performance. If the microstrip line has a characteristic impedance of Z_0 (50 Ω), then the characteristic impedances of the two different line sections of the coupler must be Z_0 and $Z_0/\sqrt{2}$ (nominally). Any mismatch will produce reflections leading to a drop in isolation (from about 30 dB to 10 dB or even less) and an incorrect power split.

Adequate isolation can only be maintained over a 12% to 13% relative bandwidth (giving 20 dB isolation at the

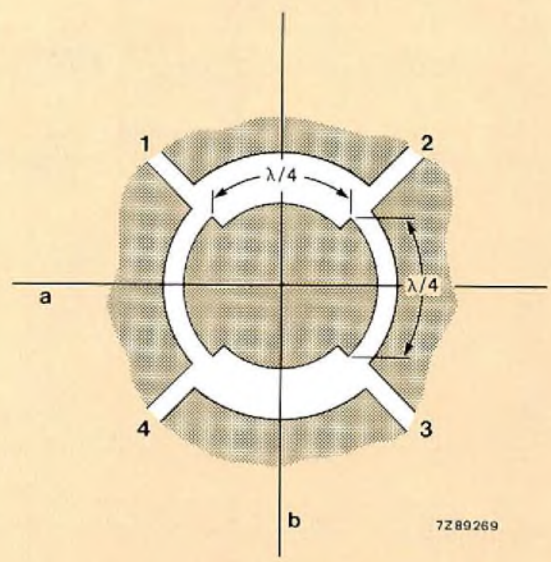

Fig.4 3 dB branch-coupler for isolation between the two input branches (1 and 4), and equal power split between the two output branches (2 and 3) band edges), but this presents no problem in the field of telecommunications where bandwidths of 10% or less are usual.

Reflections produced at the ports of the branch coupler by impedance mismatch will set up standing waves within the system. At the development stage therefore, a knowledge of the standing wave pattern at each of the four ports is essential, and the work of optimising performance whilst at the same time preserving the necessary symmetry, is to a great extent a process of successive approximation based on the measurement of VSWR.

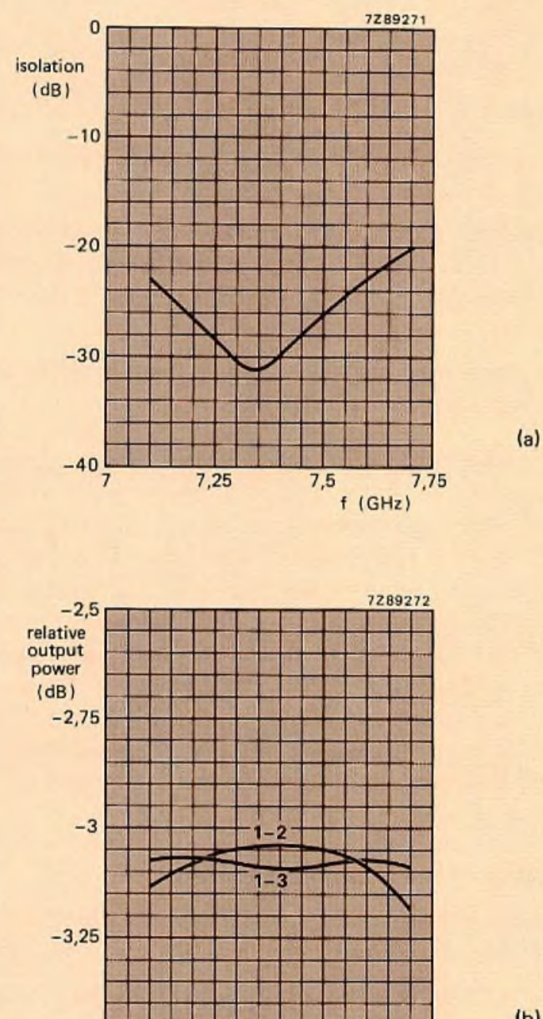

Fig.5 (a) The isolation between the input ports of an optimised branch-coupler as a function of frequency; (b) Power split of an optimised branch-coupler as a function of frequency. The difference in the two power curves arises because the path lengths between either input port and the two output ports are unequal (in the ratio of 2:1)

Technology

Substrate manufacture

Substrates are manufactured from the base materials, either by moulding a slurry to the approximate shape of the substrate before firing, or by machining from fired ceramic bars or from single crystals.

Alumina substrates are usually manufactured using the former method. Three grades of alumina substrate, based on surface quality, are currently available:

- As-fired substrates. Formed from wafers of Al_2O_3 slurry moulded and fired to the required thickness.
- Ground substrates. Formed in the same way as the 'as-fired' substrates, but then ground to a finer surface finish. They are thus slightly more expensive. Usually, only one surface of the substrate is of ground quality.
- Polished substrates. Generally used where optimum performance and reliability are critical factors and where cost is secondary.

The three grades are also available in glazed form. Glazing significantly reduces the number of surface defects.

For ferrite substrates, the bulk material is supplied in the form of rectangular bars. After machining for accurate squareness, the bars are sawn and lapped to the required thickness. The lapping produces a reasonably high surface quality which is subsequently polished to a very fine grade. The machining and sawing of the ferrite bars makes it possible to produce substrates to a very high degree of accuracy.

Substrates of fused quartz and, to a lesser extent, of crystalline quartz are becoming fairly common and present no major problems in manufacture. However, the cutting of crystalline quartz is complicated by its brittleness and by the fact that it must be accurately cut along specific directions to ensure a well defined dielectric constant.

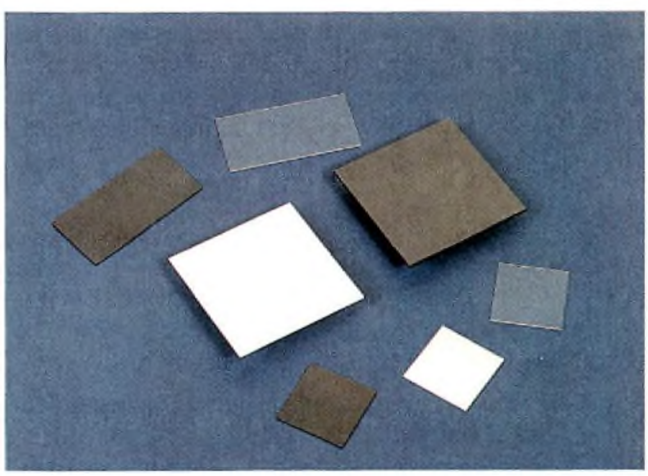

Fig.6 Substrates for microstrip. Centre: glazed alumina 5 cm X 5 cm; clockwise from the top: fused quartz 5 cm X 2.5 cm, polished ferrite 5 cm X 5 cm, fused quartz 2.5 cm X 2.5 cm, alumina polished on one side 2.5 cm X 2.5 cm, ferrite 2.5 cm X 2,5 cm, lapped ferrite 2.5 cm X 3.75 cm and polished ferrite 5 cm X 2.5 cm

TABLE 2
Mechanical properties of common substrates

substrate material	thickness tolerance (%)	surface roughness (μm)	thermal conductivity (W/°C)	thermal expansion coefficient ($\times 10^{-6}$/°C)	maximum standard size (cm^2)	density (g/cm^3)
alumina						
glazed	15	0.02	0.25	7.3		3.8
as-fired	10	0.5–1.3	0.25	7.3	9 × 11	3.9
ground	10	0.2–0.3	0.25	7.3		3.9
ferrite	1	polished	0.063	10	5 × 5	≈5
fused quartz	1	polished	0.014	0.55	2.5 × 5	2.2

Beryllia and sapphire are generally unpopular as substrate material in view of their inherent disadvantages compared with alumina and ferrite. Beryllia substrates in particular are little used nowadays owing to the health hazard involved in their manufacture; and sapphire crystals, besides being expensive and limited in size, are unsatisfactory because their dielectric properties are highly sensitive to crystal direction.

Table 2 compares mechanical aspects of commonly used substrates.

Grinding and polishing can strain the substrate. The risk is lessened by working both sides at once. The difficulty of polishing large surfaces with a high degree of accuracy imposes a limit on substrate size (particularly in the case of alumina which is very difficult to polish). Polished substrates are consequently limited to a size of 5 cm x 5 cm; the most popular size is 2.5 cm x 2.5 cm. In the case of ferrite substrates, the dimensions of the bars from which they are produced imposes a secondary limitation on size. Immediately after polishing, substrates are ultrasonically cleaned in hot water to remove waste. Their specifications are then checked and the surface is inspected microscopically to determine polishing grade.

Mask realisation

The microstrip circuit pattern is formed on the substrate by a photo-lithographic process employing a mask that is an accurate facsimile of the final conductor pattern. The steps involved in producing the mask are as follows:

- A large scale negative of the conductor pattern is produced on studnite film. The drawing may be between 10 and 100 times the true size of the conductor pattern, depending upon the line accuracy required. In the initial design phase the studnite pattern is produced manually to allow a fast turn-around. As soon as the design is satisfactory, it is frozen and all relevant data put onto punched tape. This then controls the machine that cuts the studnite pattern for the production mask.
- The studnite pattern is photographically reduced to obtain the master pattern. This is a life-size positive of the conductor pattern on glass. Once the design is frozen, the master is reproduced in chromium on glass.
- From the master, the final negative mask is produced, either on glass or on cellulose film.

The above procedure applies when a positive photolacquer is used on the substrate. Although this is usually the case, there are some instances where a negative lacquer is used. The final mask must then be a positive of the circuit pattern, this being obtained from the master either by photographic reversal, or by a double photographic process, i.e. with the production of an intermediate negative mask. The intermediate step could be omitted by producing a positive studnite pattern. However, this would entail the removal of large quantities of material from the studnite film, which would lead in turn to reduced definition in the final mask.

Formation of the conductor pattern

The process of forming the gold conductor pattern on the substrate consists of four stages:

- Complete metallisation of both surfaces of the substate with a layer of gold to a thickness of about 200 to 300 nm (including a 10 to 20 nm adhesion layer).
- Application of an ultraviolet-sensitive photo-lacquer to one of the metallised surfaces, either with a roller coater (preferable for batch production) or by spinning. Lacquered substrates can be stored until needed.
- Exposure of the photo-lacquer through the mask and development to print a negative circuit pattern on the substrate.

Fig.7 In the initial design phase of a system or function, the studnite pattern is produced manually to allow a fast turn-around. The apparatus in the foreground can do this job very accurately. It is, however, limited to relatively simple shapes; more complicated shapes must be cut by hand as shown in the background

- Growth of the conductor pattern (to a thickness of $10\,\mu$m) by electroplating. After this the photo-lacquer is removed, and where no conductor pattern has been formed, the initial gold layer and the adhesion layer are etched away to leave the substrate surface exposed.

Preparation of substrates for metallisation

After incoming inspection all substates undergo intensive cleaning to prepare them for metallisation. This includes compressed-air cleaning to remove dust, and ultrasonic cleaning in boiling freon, first in the liquid phase, and then in the vapour phase. After this the substrates are oven dried at 150 °C for 1½ hours.

Metallisation

Two metallisation techniques are currently in use: vacuum evaporation and r.f. sputtering. The choice depends principally upon the adhesion layer employed.

Medium to low-power systems often use aluminium or chromium as an adhesive. In systems with integrated resistors, nichrome (nickel-chromium) is often preferred, the resistors being produced by selective etching of the gold conductor layer to expose the nichrome. Aluminium, chromium and nichrome are usually deposited by vacuum evaporation.

In high-power systems employing add-on resistors, titanium may be used as an adhesive, either by itself or in combination with platinum to produce a triple layer of titanium-platinum-titanium. The layers are deposited by r.f. sputtering. This technique has the advantage that it can be readily adapted for different materials, and so makes the formation of multiple layers easier.

Vacuum evaporation. Many systems are available for vacuum evaporation. The one shown in Fig.8 works at a pressure of about 10^{-8} torr (= 1.33×10^{-6} Pascal) attained with a combination of dry-mechanical, adsorption and ion-getter pumps (occasionally augmented by a titanium sublimation pump). During the pumping stage, the substrates are degassed by heating to 300 °C. The whole evaporation system is oil-free and accordingly can attain pressures as low as 10^{-9} torr without difficulty.

To speed the pumping process, the vacuum apparatus is split into a working chamber which can open to the atmosphere for insertion of substrates, and a pumping chamber which can be sealed from the working chamber and which is always kept at low pressure. The substrates are mounted on a carousel which, besides rotating continuously to provide uniform deposition, can also tilt the substrates for two-sided metallisation. Eccentric location of the source relative to the rotational axis of the carousel together with the use of a shuttering system, guarantees minimum spread in layer thickness. A probe in the working chamber monitors the evaporation process.

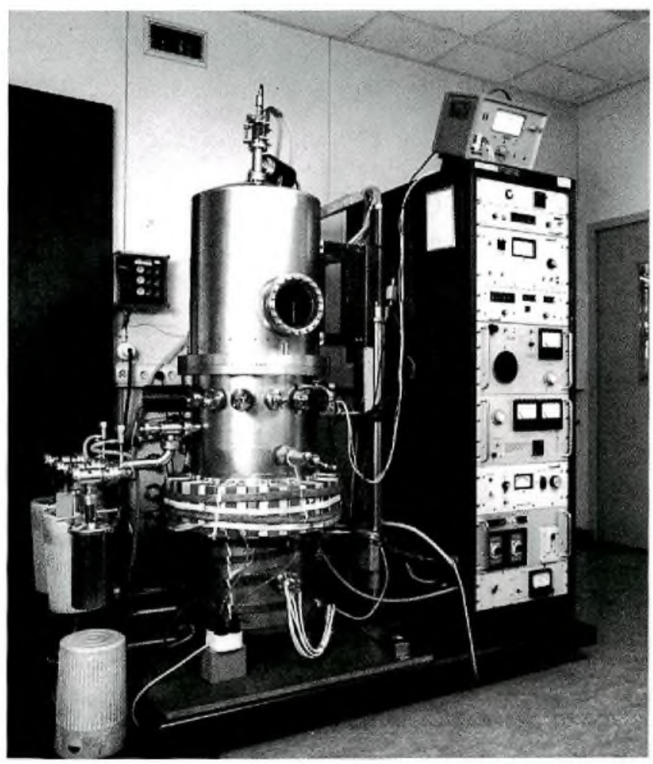

Fig.8 Vacuum evaporation apparatus. This apparatus can be pumped down to about 10^{-10} torr, although in operation it is usually kept at about 10^{-8} torr

R.F. sputtering. The apparatus shown in Fig.9 is for r.f. sputtering in argon at a controlled pressure of 10^{-2} to 10^{-3} torr. Individual targets of gold, titanium, platinum, etc. are located in the sputtering chamber and substrates are inserted via a 20 cm Intervac side loader (scavenged by an adsorption pump) which seals the chamber from the ambient air. When the side loader reaches a pressure of about 10^{-5} torr it is opened to the sputtering chamber which is maintained at about 10^{-7} torr by a turbo-molecular pump. Following a momentary rise, the pressure stabilises again at 10^{-7} torr, after which the chamber is filled with high-purity argon.

Before metallisation begins, contaminants are removed from the targets by pre-sputtering and from the substrates by back-sputtering. Each layer is then sputtered in turn, ending with a layer of gold. Only one surface can be metallised in each run, so two-sided metallisation requires two separate runs.

Fig.9 R.F. sputtering apparatus. The apparatus contains individual targets of gold, titanium, platinum, etc. The sputtering operation is carried out under argon at a controlled pressure of 10^{-2} to 10^{-3} torr

The sputter yield is governed by the sputter time, the r.f. power and the argon pressure; maximum yield can only be obtained through painstaking optimisation of these parameters.

Application of photo-lacquer

The advantage of the roller coater (Fig.10) for applying photo-lacquer lies in its ability to process a large number of substrates in a single run, and to produce uniform layers of any desired thickness (within reason) by using rollers of the appropriate size. It does, however, require a considerable amount of preparation before use. In particular, the roller must be saturated with lacquer to guarantee uniformity in production, and this can only be assured by a long run-up. For this reason, spinning is preferred for small batches (i.e. 10 substrates or less). In this process a small quantity of lacquer is applied to each substrate in turn which is then spun to spread the lacquer evenly over its surface. To facilitate spreading, the lacquer must be of rather low viscosity. This has the disadvantage of producing rather thin layers which, during electroplating, can lead to mushrooming of the conductor pattern.

The lacquer is dried by baking at 70 ℃ for about 10 minutes, after which the substrates are stored in a conditioned environment.

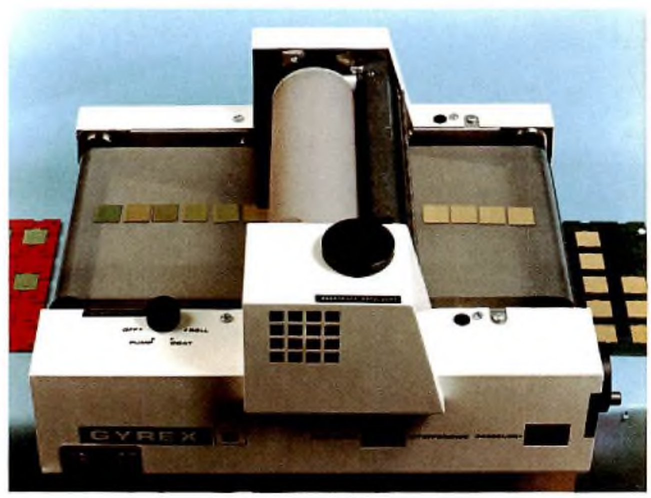

Fig.10 Roller coater used for covering substrates with photo-lacquer

Photo-lithographic process

The photo-lithographic process, which comprises alignment of the mask on the substrate, exposure of the photo-lacquer, developing, fixing, rinsing, drying, electroplating and etching, is carried out in an uninterrupted sequence.

The mask is aligned in a special jig that holds it in place by vacuum. After exposure to ultraviolet light for about 20 seconds, the mask is removed and the lacquer developed for 20 seconds, fixed, rinsed and spin dried.

After drying, the substrates that have been spin-coated with photo-lacquer are given an extra bake at 160 ℃ for about 10 minutes to strengthen the lacquer so that it can withstand the corrosive action of the electroplating bath.

Both surfaces of the substrate are electroplated simultaneously, the conductor pattern and ground plane being grown to a thickness of 10 μm.

The photo-lacquer is then chemically stripped from the substrates, and the thin gold and adhesive layers are etched away, either chemically or by sputter etching.

ADD-ON COMPONENTS

Add-on components are elements or functions added to the substrate after the conductor pattern has been formed. In the downconverter used as our present example these comprise the isolator located in the r.f. input of the branch-coupler, and the mixing diodes together with the associated r.f. filter and matching sections. Strictly speaking, the isolator is not an add-on component since its r.f. circuitry is part of the conductor pattern initially formed on the substrate. Nevertheless, it is included here because its operation requires an external magnetic field which is supplied by a pair of permanent magnets bonded to the device above and below the substrate.

Isolator

The 3 dB branch-coupler can provide isolation of some 20 dB over a 10% relative bandwidth, conditional upon perfect impedance matching at each of the four ports. This is often difficult to achieve in practice, and in the balanced mixer the isolation between the local oscillator and r.f. inputs can fall as low as 15 dB owing to mismatch at the output ports. To maintain the isolation at the 30 dB or more demanded in telecommunications, an isolator is therefore inserted in the r.f. input of the branch-coupler.

The isolator used in the present example is derived from the circulator shown in Fig.11. In this device, three ports are coupled to a disk that rests on a ferrite substrate magnetised normal to its surface by a pair of permanent magnets. The biasing field produced by these magnets orients the electron spins in the ferrite to produce a gyromagnetic effect, with the electrons precessing about the magnetic field direction (Larmor precession).

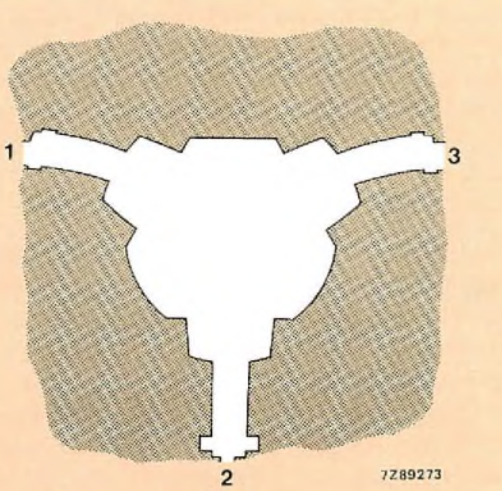

Fig.11 The circulator directs energy from port 1 to port 2, from port 2 to port 3 and from port 3 to port 1. If one of the ports, port 3 say, is terminated with a matched load, the device acts as an isolator allowing only unidirectional energy flow from port 1 to port 2

An r.f. field applied in a direction perpendicular to the biasing field will interact with the precessing electrons to produce a standing-wave pattern within the disk (which acts as a low-Q resonator). This pattern is governed by the magnitude of the biasing field so that, with suitable choice of permanent magnets, a nodal point can be made to occur at one of the ports, which is thus effectively decoupled from the input. For example, in Fig.11 port 3 is decoupled from port 1 so that all energy entering via port 1 exits via port 2. The device exhibits three-fold symmetry, which means that a similar energy flow will occur between ports 2 and 3 and between ports 3 and 1.

To convert the circulator into an isolator, it is only necessary to terminate one of the ports with a matching load that is able to absorb all the energy directed to it. This can be done using discrete 50 Ω resistors secured to the conductor pattern by an expoxy adhesive. Although these resistors have a somewhat limited power handling capacity (100 mW), for about 80% of all current applications they are more than adequate.

Major considerations in the design of circulators and isolators are:

- Symmetrical location of the permanent magnets relative to the substrate. The magnets must also be of equal strength so that, as far as possible, the magnetic field passing through the disk is everywhere normal to the substrate surface.
- Three-fold symmetry about an axis normal to the substrate surface.

As with the branch-coupler, optimisation of design can be reduced to an empirical process based upon measurement of VSWR at each port. An optimised device with a suitably matched load at the third port can provide isolation in the reverse direction of some 30 dB at the cost of an insertion loss of about 0.3 dB in the forward direction. The example used here is based wholly upon a ferrite substrate (material IV in Table 1) so that the isolator can be incorporated as an integral unit at very little extra cost. However, a family of isolators and circulators is available (ranging from 2 to 12 GHz with 10% to 35% relative bandwidth), each unit being supplied as a complete package that can be inserted into systems based on non-ferrite substrates.

Mixing diodes

The mixing action is performed by a matched pair of beam-lead Schottky-barrier diodes (Fig.12) pumped in unison by the local oscillator. The quarter-wavelength phase difference between outputs 2 and 3 of the branch-coupler is accounted for by reversing the connections of one of the diodes and increasing its distance from the branch-coupler by a quarter wavelength.

Beam-lead devices possess the advantage of highly predictable behaviour owing to the simultaneous manufacture of device and connections, and the straight forward gold-on-gold bonding to the microstrip circuitry.

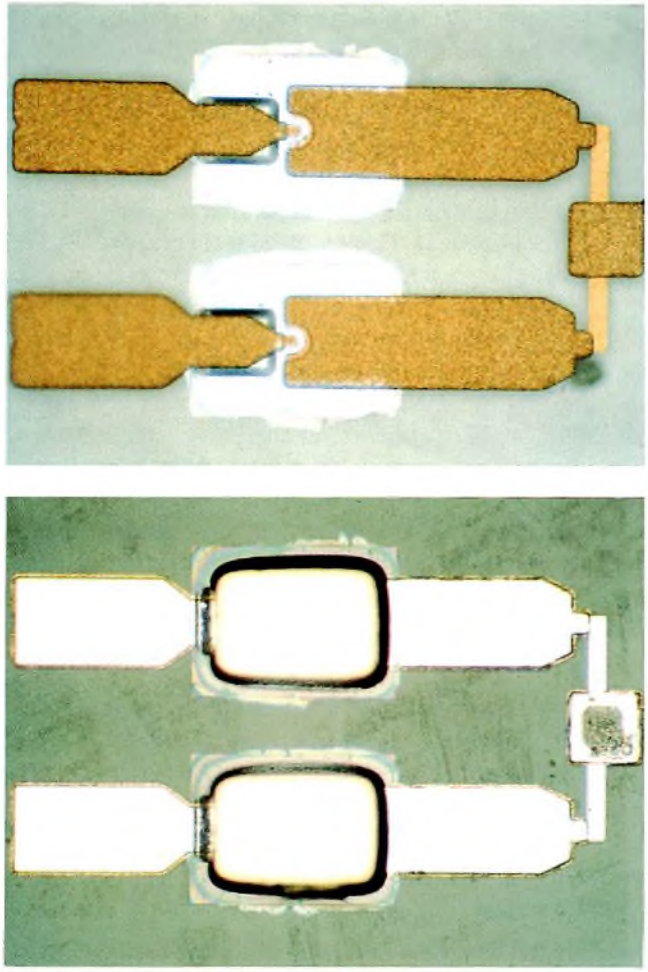

Fig.12 A pair of beam-lead Schottky barrier diodes formed next to each other on the slice, viewed from above (top photograph) and from below (bottom photograph). After separation from the slice, the pair is kept intact. Using two diodes that form an adjacent pair assures minimum spread in parameters

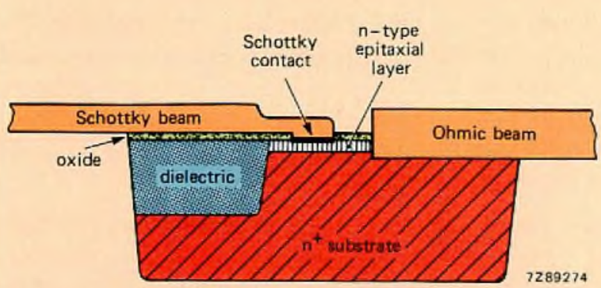

Fig.13 Cross-section of a typical beam-lead Schottky-barrier diode. Conversion losses can be reduced by reducing the parasitic capacitance, e.g. by increasing the Schottky-beam/substrate distance. In the example shown this is done by forming a cavity in the substrate below the Schottky beam and filling it with a dielectric

A cross-section of a typical beam-lead diode is shown in Fig.13. It comprises a doped n^+ silicon substrate supporting an n-type silicon epitaxial layer. The Schottky barrier is formed between the epitaxial layer and a contact electrode of palladium, nickel or titanium.

Factors governing the performance of a diode are:

- Contact electrode. A palladium or nickel contact gives rise to a lower reverse-leakage current and consequently to lower noise than a titanium electrode. However, diodes of palladium/silicon (and nickel/silicon) have a greater barrier height than those of titanium/silicon, and therefore require more power to drive them into their non-linear region (the region essential for the mixing action).
- Diode resistance and capacitance. The conversion losses produced by the diode are directly proportional to its series resistance R_S, and to the capacitance C which comprises the junction capacitance C_j and the parasitic capacitance C_p. Both R_S and C_j vary with the thickness of the epitaxial layer, R_S directly and C_j inversely. The choice of layer thickness must therefore provide a compromise in the values of R_S and C_j to give minimum conversion losses. An overall reduction in conversion losses is obtained by reducing C_p, which can be done by increasing the Schottky-beam/substrate distance, e.g. by forming a cavity in the substrate and filling it with an insulating material.
- Substrate material. The diodes used in our example are silicon. However, other materials such as gallium arsenide or germanium may also be used. Gallium arsenide in particular possesses the advantage of a higher mobility and a consequent lower R_S than silicon diodes. However, the introduction of an insulating region to reduce C_p is extremely difficult in gallium arsenide diodes since they are formed at a temperature of about 800 to 900 °C (compared with 1000 to 1100 °C for silicon), which is roughly the temperature at which the insulating region is formed. Its introduction would thus effectively destroy the epitaxial layer. Materials with lower melting points (such as polyamide) could be used but they introduce additional problems.
- Impedance matching at the i.f. side. Located behind the diodes is the i.f. amplifier whose source impedance must ideally be matched to the impedance of the diodes. The 40% relative bandwidth at the i.f. side, means that this requirement can be met only by using diodes of the appropriate impedance. Since the diodes operate in parallel, this leads in effect to the condition that the junction resistance R_j of each diode must be approximately equal to twice the source impedance of the amplifier. Furthermore, given limited local oscillator power, the specifications of the diodes must be such that they can be pumped sufficiently hard to provide the required impedance whilst giving suffient current to minimise diode noise.

R.F. filter

The local oscillator, in general, has to deliver power of a few milliwats per diode, and this power must be prevented from passing to the i.f. amplifier. For this reason an r.f. filter is inserted between each diode and the amplifier. This filter (Fig.14), which is in the form of a quarter-wave stub, effectively short circuits the r.f. signal, providing r.f. attenuation of more than 30 dB (the maximum to be expected in microstrip) over a 10% to 15% bandwidth, and at the same time giving negligible insertion loss in the r.f. region (less than 0.2 dB). A ring of about three-quarters of a wavelength circumference surrounds the stub to limit radiation. The exact dimensions are obtained empirically, and the design has been optimised for minimum radiation losses.

Matching section

The diode/r.f. filter combination is matched to the source impedance of the i.f. amplifier for maximum delivery of i.f. power, and this may result in mismatch on the r.f. side. For efficient delivery of r.f. power to the diodes, a matching section must therefore be inserted between the diodes and the 50 Ω line of the branch-coupler. This can be realised in a distributed transformer section (comprising a series arrangement of lines of differing widths), or alternatively in a series of short stubs on the 50 Ω line between the diodes and the branch-coupler. Also, since the diodes are in series with the microstrip line, a d.c. return, in the form of a shorted stub, must be provided on the r.f. side. In the case of stub matching, the d.c. return can be incorporated in one of the stubs already present. Other advantages of stub matching over transformer matching are lower insertion losses and better diode protection against power surges. Transformer matching can, however, provide a greater bandwidth than stub matching which is generally limited to a bandwidth of about 15%.

Other add-on components

The isolator and beam-lead diodes are add-on components specific to the example being considered. However, many other add-on components are available in MIC-compatible technology. These include passive components such as chip and beam-lead resistors and capacitors, as well as active components like FETs, Gunn diodes and IMPATT diodes.

The choice of packaging is not always straightforward since it depends very much upon the application and upon economic considerations. For example, beam-lead devices are usually very small, which means that their power handling capacity is limited. In special applications, however, they are often indispensable. This is true in particular at frequencies above about 18 GHz where the parasitics produced by larger devices often become troublesome. Moreover, in small series the use of beam-lead resistors may be less expensive than integrated resistors which often have to be trimmed to the correct value.

Blocking capacitors, which are generally of high capacitance, must be in chip form since the small size of beam-lead capacitors limits them to a maximum of about 20 pF.

In the absence of beam-leads, gold wire is often used to connect components to the circuit pattern. To this end the components can be provided with ready made bonding pads. Alternatively, they may be bonded to the circuit pattern by means of an epoxy adhesive.

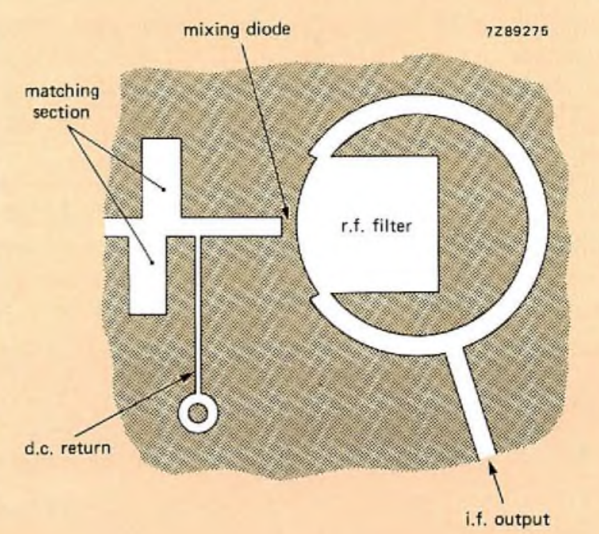

Fig.14 The r.f. filter, in the form of a quarter-wave stub, prevents the local oscillator energy from passing to the i.f. amplifier. To limit radiation, the stub is surrounded by a ring with a circumference of about three-quarters of a wavelength. The figure also shows the matching section made up of a series of short stubs on the 50 Ω line, and a d.c. return in the form of a shorted stub

Fig.15 The ceramic disk resonator in a balanced downconverter based on alumina. The downconverter shown here operates in the 11.4 to 11.7 GHz band

An add-on component recently introduced is the ceramic disk resonator (Fig.15). This is made of a ceramic such as barium titanate and functions as a dielectric resonator in oscillator circuits. It has the advantage that its characteristics are insensitive to temperature changes, an important factor in producing stable oscillators. The disk is bonded to the substrate close to the oscillator circuitry to which it is coupled electromagnetically.

Bonding techniques

There are many ways of bonding components to the conductor pattern. These include gluing with conductive or non-conductive epoxy resins, welding and thermosonic bonding (a combination of thermal compression and ultrasonic bonding).

Lead or tin soldering (either of components to the circuit or of circuit to baseplate) can damage the gold layer. However, there are special solders (e.g. gold/indium, gold/tin or gold/germanium) that bond to the gold pattern with minimum damage.

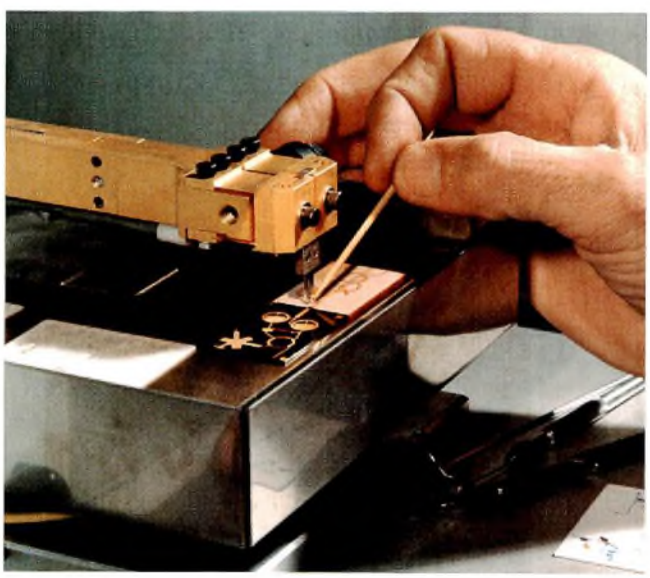

Fig.16 A ferrite and an alumina substrate are bonded together. To reduce the effects of heat dissipation, the substrates are preheated before the bonding tool is applied

In MIC technology the most common bond is gold-on-gold, and this is usually effected thermosonically or by welding. Welding is used wherever thick layers have to be joined, since in such cases the thermosonic technique is not so effective. In both thermosonic bonding and in welding, and in fact wherever heat has to be applied to the bond, the effects of dissipation can be offset by pre-heating the whole substrate before applying the bonding tool.

PACKAGING

There are two major requirements in packaging MICs:

- reliable connections between the microwave circuitry and other systems or subsystems.
- suitable encapsulation to provide the controlled environment essential for reliable operation.

These requirements are strongly interdependent, but in discussing them here we shall as far as possible treat them separately.

Electrical connections

The completed circuit, including the add-on components mounted on the substrate, must first be secured to a base plate which provides the earth connections to the ground plane. To assure good electrical contact, the base plate is first machined and ground to a surface finish comparable with that of the substrate, after which it may be covered with gold.

The substrate is either clamped or bonded (i.e. glued or soldered) to the base plate, depending upon factors such as the surface finish of the substrate and the environment in which the circuit will operate. For example, the fine surface finish of ferrite substrates is well suited to clamping, whereas alumina substrates (which are usually polished only on one side) have to be bonded to the base plate.

Thermal stresses within the substrate, which if extreme can result in fracture, have to be minimised. With clamped substrates this is not difficult, provided the clamping points are selected to allow relative movement between substrate and base plate. In the case of bonded substrates, it is often necessary to insert a layer of ductile material between ground plane and base plate. Ideally of course substrate and base plate should have equal expansion coefficients, but this is not always easy to attain in practice.

The problem of thermal stress is also important in considering the other electrical connections. These are often in the form of ribbons bonded to the microstrip, allowing the substrate to expand unrestrictedly. Alternatively they may be in the form of gold pressure contacts.

Encapsulation

The MIC must be provided with a controlled environment, and this is usually accomplished by placing it within a sealed box. The box is normally filled with inert gas at ambient pressure, or with dry air. This reduces the chance of moisture diffusing into the system, as might occur if the box were evacuated.

In some applications, however, where the environment is favourable (i.e. air-conditioned, low humidity) it is better to have an open box to allow the free escape of any moisture that might condense on the circuit.

A system may comprise two or more interconnected substrates all contained within the same box. Ideally, between the substrates pure TEM modes should propagate. This is best ensured by using purpose-built connectors of square cross-section containing a gold-ribbon centre conductor supported by dielectric material.

Design of the box is highly dependent upon application, and for most systems boxes have to be custom built. For example, a given application may need efficient heat-sinking which would require a box (preferably of aluminium or copper) with large surface area. Another application may require that the box be able to accommodate coaxial sockets or waveguides for connection to external systems. In fact, customer requirements have a greater influence upon box design than upon any other area of MIC technology.

QUALITY

Quality is not just a matter of test and inspections. It is determined by the design of the product, by the materials and processes used to manufacture it, and by the discipline and expertise of the factory staff. The function of in-line tests and inspections is simply to control the manufacturing process and to detect adverse process trends as early as possible; they effectively establish a series of go/no-go conditions in the manufacturing process.

At an early stage in design, a choice of materials and processes is made by the development engineers. This choice is based on data obtained from research into failure mechanisms as well as life test and field experience data gathered in co-operation with the Quality Laboratory. This data enables the quality of successive generations of the product to be maintained and possibly improved.

Failure mechanisms and their investigation

The initial stage of selecting the technology and processing is crucial. Before manufacturing can begin, all likely failure mechanisms must be thoroughly investigated. The technology and processing must then be selected to eliminate these as far as possible. The potential causes of failure listed below illustrate the problems that can arise in MIC manufacture:

- diffusion of surface contaminants into the adhesive layer. This usually reduces adhesion and can ultimately detach the conductor pattern from the substrate;
- interdiffusion of material from the adhesive layer into the conductor. In time this may alter the conduction properties so that the product fails to meet its electrical specifications;
- diffusion of material from the adhesive layer into the substrate. This can reduce adhesion and may alter the dielectric properties of the substrate so that ultimately the product fails to meet its electrical specifications;
- blemishes in the conductor pattern such as surface scratches caused by careless handling. Such blemishes can lead to crystal defects which often show as extrusions on the surface (Fig.17).

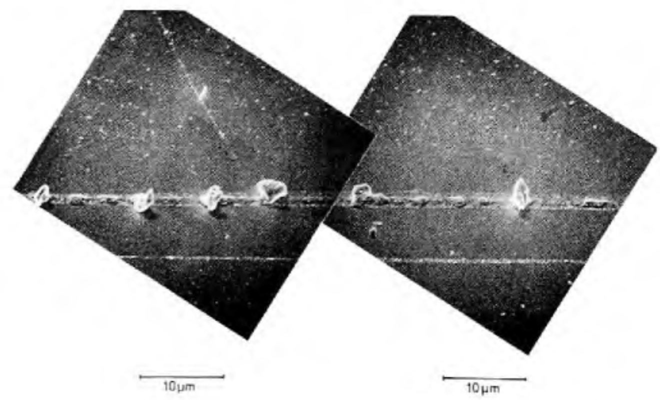

Fig.17 The conductor pattern can be scratched by careless handling. Such scratches can cause crystal defects which often show up as extrusions on the surface

The major problem with MICs, as with integrated circuits in general, is that often failure mechanisms are not immediately evident, so that zero-hour measurements may well give a false impression. Aluminium, for example, appears to be an excellent adhesive, but in combination with gold conductor layers its adhesive strength can diminish very rapidly. Other materials, which do not provide such strong adhesion at first may well be more suitable than aluminium in the long term.

We have in fact conducted an extensive investigation into the behaviour of adhesive layers with the object of finding a reliable recipe. From an investigation of many different adhesive layers, we find that the layers can be characterised by just four isochronal curves (Fig.18).

In curves 1) and 2) a single effect seems to dominate. The steady decrease of resistivity exhibited by 1) indicates a crystal recovery process, and the slow increase of 2) suggests a slow chemical reaction.

The reactions governing curves 3) and 4) are rather more complex, the sudden changes indicating that more than one process is involved.

Figure 19 illustrates the effect of pre-treatment on various metallisations. The figure shows isochronal ageing behaviour of Ti-Au-Au and Ti-Pt-Ti-Au-Au metallisations. The first layer (Ti) is the adhesive, which is sputtered onto a polished ferrite substrate.

Region (a) indicates the spread of characteristics of a sample batch with Ti-Au-Au metallisation, measured immediately after their manufacture. Note the significant deviation from the initial value. Curves (b) and (d), produced by samples heated respectively for 4 hours at 390 K and 100% relative humidity and 140 hours at 480 K, illustrate how pre-treatment can reduce this deviation. Curves (c) and (e) are produced by samples with Ti-Pt-Ti-Au-Au metallisation, pre-treated in the same way as the (b) and (d) samples respectively. The curves illustrate how inserting a Pt layer within the Ti layer can affect the characteristics. The Pt layer acts as a barrier to diffusion, reducing the rate of metal interdiffusion as well as the rate at which oxygen and other contaminants diffuse from the substrate surface.

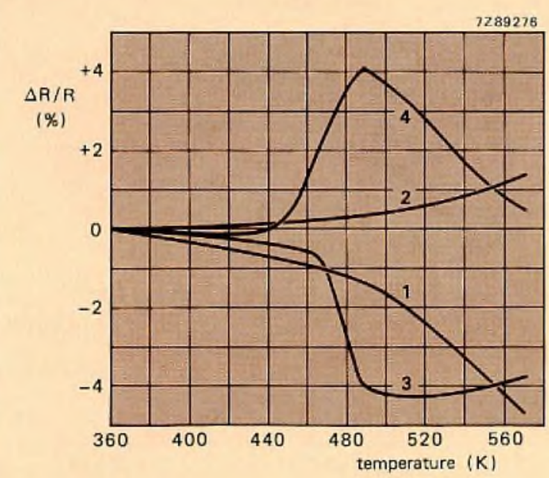

Fig.18 Adhesive layers can all be characterised by one of four isochronal curves: (1), (2), (3) and (4) above. The curves show the variation of resistance (ΔR/R) as a function of temperature. The elevated temperatures effectively accelerate the ageing process, and so provide an indication of how the conductor pattern would behave over a long period

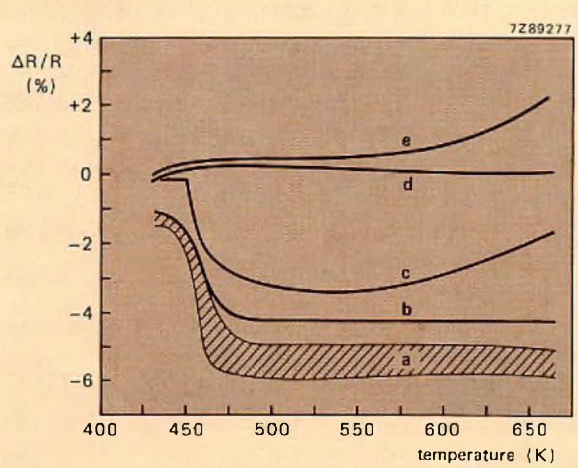

Fig.19 Isochronal ageing behaviour of Ti-Au-Au and Ti-Pt-Ti-Au-Au metallisations

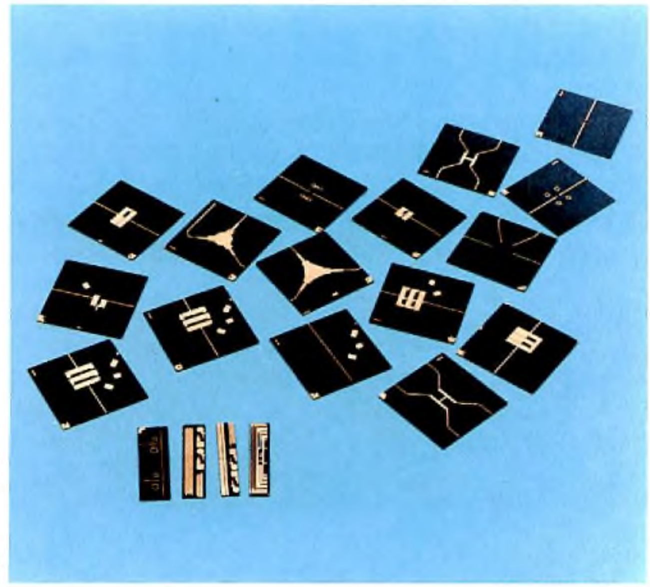

Fig.20 To allow statistical analysis of the processing as if the circuits were being mass produced, a test pattern (foreground) representative of the processing of all circuit pattern is manufactured with each batch

Quality control in manufacture

For professional applications MICs are usually manufactured in batches of 100 or less. To allow statistical analysis of the processing as if the products were being mass produced, a test pattern (Fig.20) representative of the processing of all circuits is manufactured with each batch. The test pattern is in fact a very highly defined product, one in which the correlation between behaviour and the quality of the processing is known with a high degree of certainty. In this sense it functions as part of the in-line inspection. It is also a product that can be destructively tested without destroying any members of the batch.

The test pattern is produced from the mask shown in Fig.21. The pattern is designed to make a wide range of tests available; some of these tests establish process reliability, and some are used for process control. The test pattern is made up as follows:

A Strips of differing widths. The wide strips are used to investigate electromigration phenomena for several different a.c. and d.c. current densities and gradients: the narrow strips are used for d.c. resistivity measurements carried out before, during, and after heat treatment.

B Square and round spots for bonding, soldering and adhesion tests, for investigating diffusion phenomena, and for microprobe and structure analysis.

C Strips used for peel strength measurements. The finger structures in the middle are used to measure pattern accuracy and to determine breakdown voltages under different ambient conditions.

D Spots and strips used to determine pattern accuracy (Fig.22).

E Fine line pattern used to measure power capability and d.c. resistivity of the conductor layers during batch processing.

F Resonant circuit used for in-line measurement of Q-factor at radio frequencies (Fig.23) to establish quickly whether or not the batch is acceptable

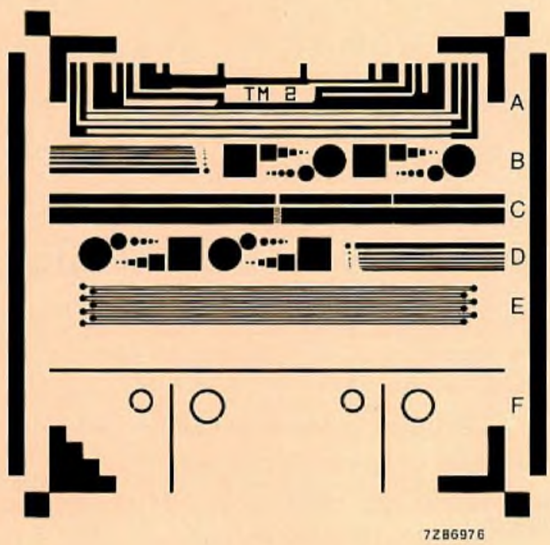

Fig.21 The test pattern is produced from the mask shown here. It functions as part of the in-line inspection, and it can also be destructively tested without destroying the actual product

The quality story does not end with the production of an acceptable batch of circuits. Each completed MIC subsystem, with circuitry and add-on components installed, must satisfy stringent quality and specification requirements. These requirements must be agreed upon by customer, development department and Quality Laboratory before any development work is started. After the final tests, only those subsystems that completely satisfy the requirements are released.

Reliability in MICs in essentially a matter of using established recipes and following established procedures. If any of these are changed simply to satisfy individual customer requirements, then much of the experience acquired from earlier work ceases to be relevant and reliability becomes problematical. We therefore recommended that the prospective customer seeks contact with our development department as early as possible in the initial design phase of his system in order to avoid time consuming and possibly costly modifications later on.

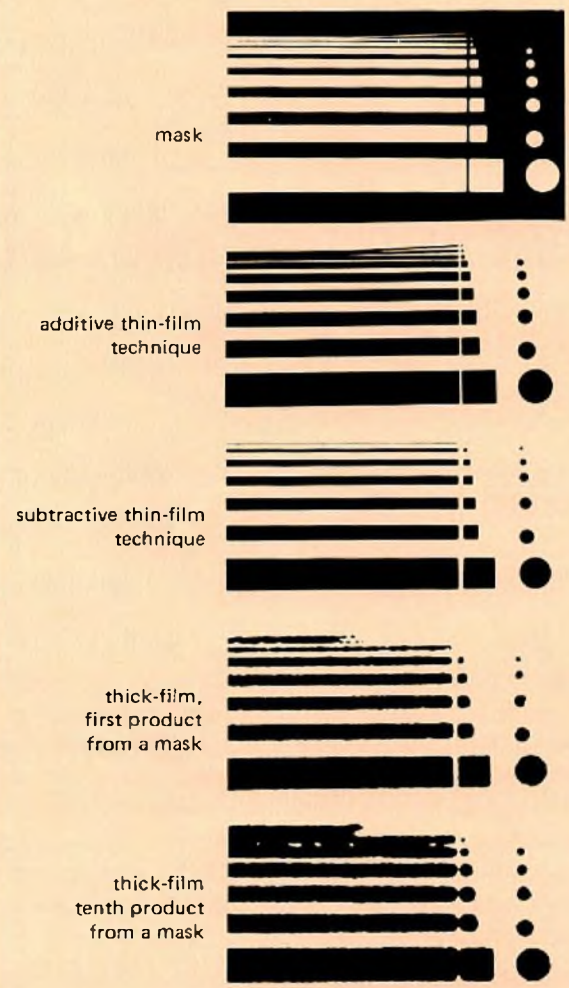

Fig.22 The test pattern can be used to check pattern accuracy. The patterns shown here were produced from the mask using various techniques

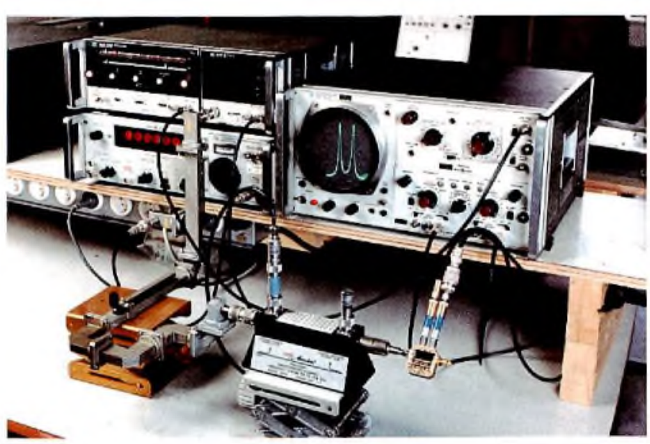

Fig.23 Part of the test pattern is used to provide an indication of Q-factor at radio frequencies. This can establish quickly whether or not a batch is acceptable

Fast and undistorted data transfer in noisy environments

J. EXALTO

The HEF4755 is a fast and versatile circuit for undistorted transmission or reception of data in electrically noisy environments. 8-bit parallel data can be serially transferred synchronously over three wires, or asynchronously over one wire. The maximum data transfer rate with a 10 V supply is 3.2 MBaud for synchronous operation and 125 kBaud for asynchronous operation. The circuit provides an optimum compromise between transmission accuracy and speed for applications such as:

- industrial measuring and remote control systems with long communication paths
- power stations and heavy current installations
- measuring and control systems in nuclear installations
- measuring and control systems in vehicles
- busses in buildings.

The main features of the HEF4755 are:

- start code generation and recognition
- data byte size protection
- redundancy check byte generation and checking
- bit check by bit level evaluation
- error type recognition
- 8-bit parallel I/O for data.

Maximum accuracy of the data and efficiency of transfer are ensured by features such as:

- short data blocks (7 bytes max.)
- data block synchronisation for reliable recognition of the beginning of a data block
- bit synchronisation for eliminating bit distortion due to data path delays
- data byte size protection to ensure that the information contains the correct number of data bytes
- cyclic redundancy check to protect the information content of a complete data block.

An important additional feature in the asynchronous mode, is a bit check which evaluates the logic level of every bit by sampling it 32 times in synchronism with the clock.

For maximum versatility, the circuit can be hard-wired to operate in one of 8 modes and can also be optimised for specific applications by pre-setting conditions such as:

- number of data bytes (fixed or variable)
- Hamming distance of the cyclic redundancy check (Hamming distance = 4 or 6)
- acceptable bit distortion.

Figure 1 is a block diagram of the HEF4755. The blocks marked A are only used in the asynchronous mode. The input/output names and functions are:

TT0, TT1	for setting acceptable bit distortion
RX	for setting the receiving mode
TX	for setting the transmitting mode
AS	for setting the asynchronous mode
START	input for start pulse or synchronisation signal
CLK	clock input
R	reset input

HD	for setting the redundancy check Hamming distance
ML0, ML1	for setting the number of data bytes
DIO0 to DIO7	parallel data I/O from/to data bus
TST	test input (must be at V_{SS} during normal operation)
MI	message input (serial data input from line)
MO	message output (serial data output to line)
MOS	synchronisation (data frame).

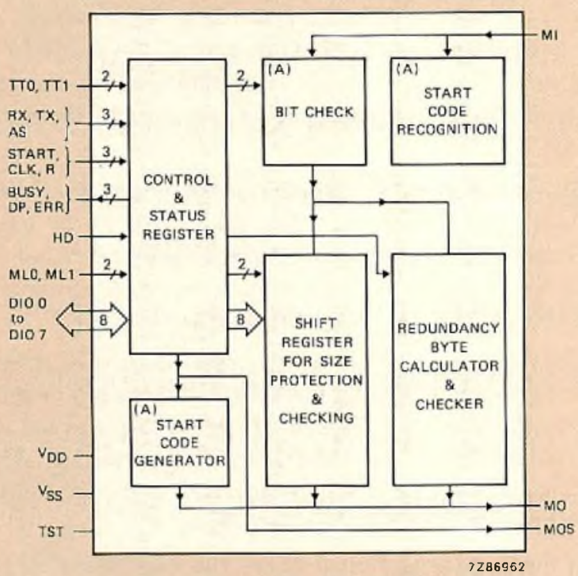

Fig.1 Block diagram of the HEF4755

MESSAGE STRUCTURE

Short data blocks containing up to 7 bytes are used to ensure efficiency and security of data transfer. A data block can contain more than one byte because some coherent statements must be transferred in one piece to ensure that one rejected byte does not corrupt the whole statement. A data block, together with its protection codes, is called a message and is constructed as shown in Fig.2.

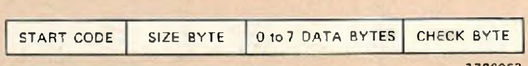

Fig.2 The message format

Start code

The start code is only used in the asynchronous mode to synchronise the data transfer. It is generated by the circuit and is easily distinguished from all other information because it contains an interval of 1.5 bit periods as shown in Fig.3

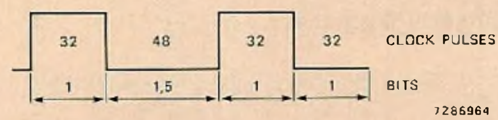

Fig.3 The start code

Size byte

The size byte is only used with variable data length. It indicates the number of data bytes on the data bus waiting for transfer and is constructed as follows:

$\overline{C}\ \overline{B}\ \overline{A}\ \overline{P}\ P\ A\ B\ C$

where: $n = C \cdot 2^2 + B \cdot 2^1 + A \cdot 2^0$

and $P = C \oplus B \oplus A$

The size byte also contains information n in a symmetrical code (Hamming Distance = 4).

Data bytes

The circuit can be hardwired to operate with a fixed number of data bytes (2,4 or 6), or with a variable number (0 to 7). There is no protocol for the data bytes, any code may be used but there must always be a whole number of 8-bit bytes. The maximum number of different messages that can be transferred is thus $2^7 \times 8 = 2^{56} \approx 10^{17}$.

Cyclic redundancy check (CRC) byte

The HEF4755 has an adjustable CRC byte generator/calculator. The CRC byte prevents the data byte from being incorrectly interpreted and is calculated in parallel with the data stream. It is transmitted as the last part of the message to the receiver where it is compared with the self-calculated check byte. In the event of discrepancy, the receiver reports a code error.

To accommodate various applications, the CRC byte calculator can be set for a minimum Hamming distance of 4 using 7 bits, or for a minimum Hamming distance of 6 using 15 bits. A Hamming distance of 4(6) means that if two data blocks differ by only one bit value, the corresponding CRC bytes differ by 4(6) bit values. Thus, all false bits up to 3(5) and any greater uneven number of false bits will be recognised. The error detection efficiency and the CRC generator polynomials are given in the Appendix.

OPERATING MODES

Synchronous modes of data transfer

In the synchronous mode (Fig.4), the transmitter and receiver share a single clock source. The bit rate is equal to the clock frequency, and the start of the message transfer is determined by an additional synchronisation signal MOS. Three wires are required for data + protection bytes, clock and message synchronisation signal. There are three modes of synchronous data transfer.

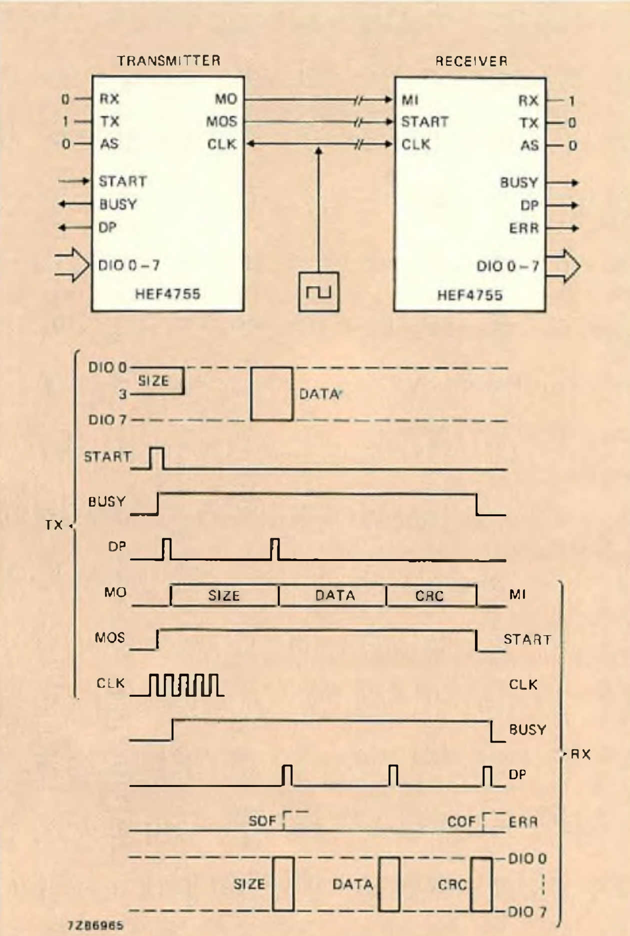

Fig.4 Pulse diagram for synchronous data transfer

Transmitting a Message with CRC byte. After accepting the START pulse, the transmitter generates the synchronisation signal at the MOS output and feeds it to the START input of the receiver. The number of data bytes in the message (2, 4, 6 or variable) is set at inputs ML0 and ML1. The size byte and the data bytes are fetched from the data bus using the DP signal. With a fixed number of data bytes, the size byte is omitted. The size and data bytes are parallel-to-serial converted and shifted out on output MO. Simultaneously, the CRC byte is calculated and shifted out after the data bytes.

Receiving a message with CRC byte. The synchronising signal at the START input reports the beginning of a message transfer. The data bytes are shifted in via MI, serial-to-parallel converted, and passed byte by byte to the data bus. The steady state of the bytes is reported by the leading edge of the data pulse (DP). The CRC byte is compared with the self-calculated check byte to detect transmission errors. If a size and/or code error is recognised, the error information is stored in the status register and output ERR is activated.

Receiving a message without CRC byte and transmitting the same message with CRC byte. A message without a CRC byte is received and handled like a message with a CRC byte except that, on being passed to the data bus, the message is simultaneously transmitted bit-serially via output MO. The calculated check byte is then added. If an error is found, the error information is stored in the status register. It is not possible to transmit a message without a CRC byte.

Figures 5 and 6 show the maximum delays in the synchronous mode. The following general remarks should be noted:

- The duration of the START pulse at the transmitter must always be shorter than the message to be transferred. A good procedure for achieving this is to use the BUSY signal to end the START pulse. A continuous START signal will cause malfunction.
- The recovery time between two messages must be at least one bit period. During this time, the message line must remain stable. A good way to achieve this is to use the trailing edge of the BUSY signal to generate a START signal. In practice, if data is delivered to the transmitter fast enough, START can be $\overline{\text{BUSY}}$.
- If the lines have different delays, the message line should have the longest delay. If it is not certain which line has the longest delay it is possible to phase shift the clock signal of the receiver by inverting it. This is only possible with point-to-point lines.

Asynchronous modes of data transfer

In the asynchronous mode, the transmitter and receiver have their own clock oscillators operating at 32 times the bit rate. Their frequencies must be matched to within ±2%. Since the data is synchronised by the start code, only one wire is needed. Figure 7 is the basic timing diagram for asynchronous data transfer.

Bit check. As shown in Fig.3, the start code includes an interval of 1.5 bit periods and is easy to recognise. After reception of the start code, blocks of 32 clock pulses are generated to perform the bit check which samples the amplitude of each bit 32 times. The bit check circuitry decides within the selected bit distortion tolerance (6/32, 8/32, 10/32 or 12/32) whether the bit is a certain 1, a

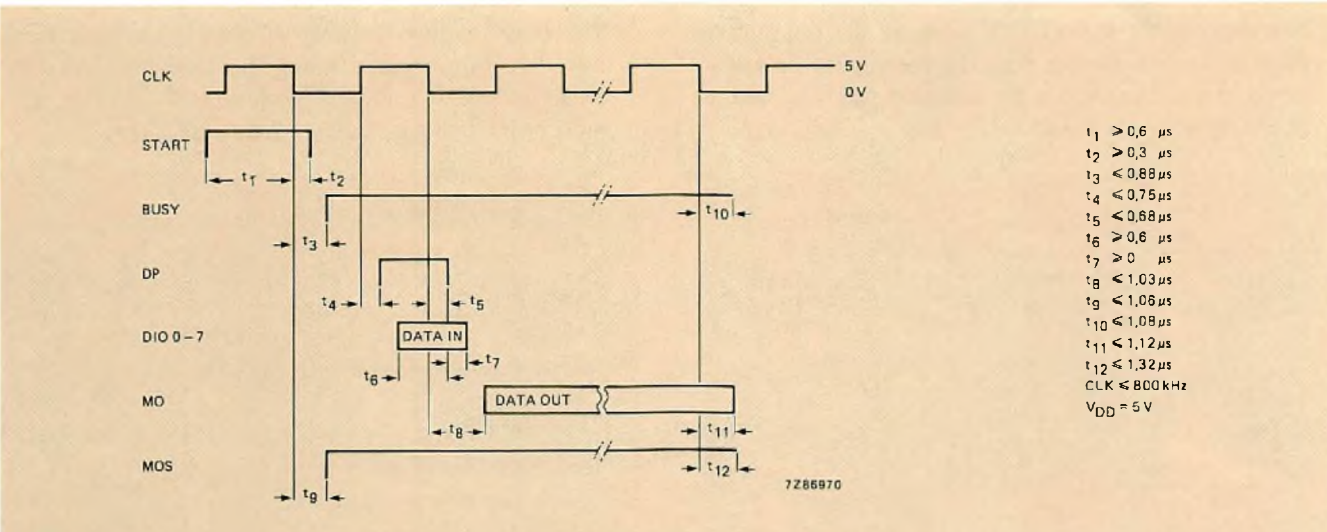

Fig.5 Maximum delay of output signals for synchronous data transmission

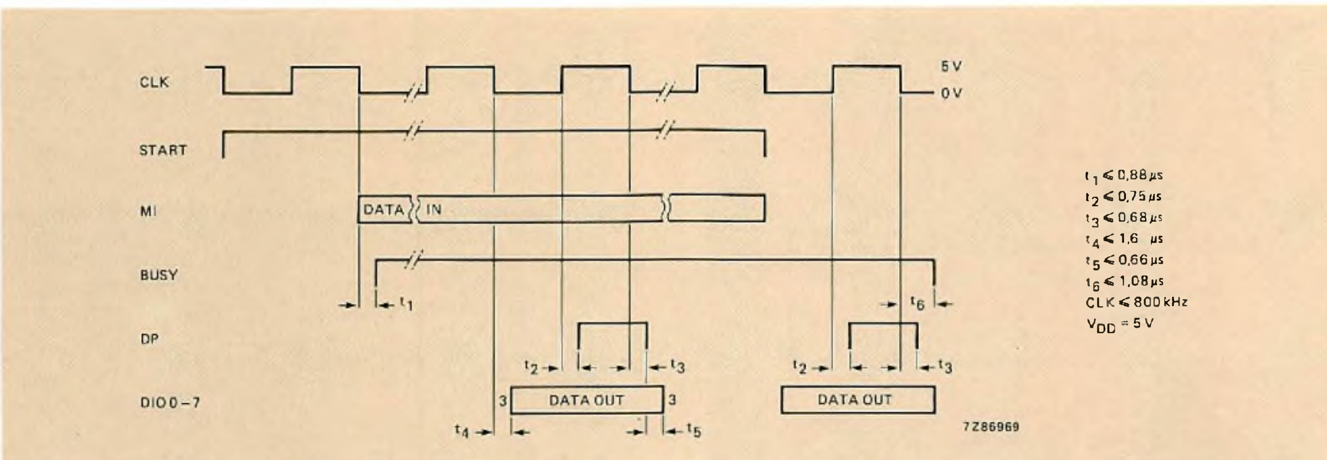

Fig.6 Maximum delay of output signals for synchronous data reception

certain 0 or if there is a bit error. After the first bit period in Fig.8, twenty-eight samples are a logical 1 (>1 V) and four samples are a logical 0 (<1 V). This means that, within all the possible distortion tolerances a certain logical 1 is evaluated. After the second bit period, eleven samples are a logical 1, and 22 samples are a logical 0. For distortion tolerances of 6/32, 8/32 or 10/32, the second bit will be reported as an error. For distortion tolerance 12/32, it will be evaluated as a logical 0.

Transmitting a message with CRC byte. After the START command, the transmitter generates the start code and shifts it out via output MO. The message is fetched from the data bus as previously described for the synchronous mode and shifted out bit-serially from MO together with the CRC byte.

Receiving a message with CRC byte. The start code informs the receiver that a message has started. The bit check circuitry evaluates the bits which are then serial-to-parallel converted as in the synchronous mode and passed on to the data bus.

Receiving a message with a CRC byte and transmitting the regenerated message. The message with check byte is received as previously described. In addition, the received message is regenerated and transmitted bit-serially via output MO.

Testing the data stream. The message is received as previously described but data is not passed to the data bus. Output MOS is set when a start code is recognised and must be reset by a RESET signal to initiate a new start code search. In this mode, the circuit not only checks for errors but also converts an asynchronous message into a synchronous message.

Figures 9 and 10 show the maximum delays in the asynchronous mode. The following general remarks should be noted:

– The duration of the START pulse at the transmitter must always be shorter than the message to be transferred. A good procedure for achieving this is to use the BUSY signal to end the START pulse.

– The recovery time between two messages must be at least two bit periods. During this time, the line must remain stable to prevent generation of an error. This must be ensured with external hardware/software.

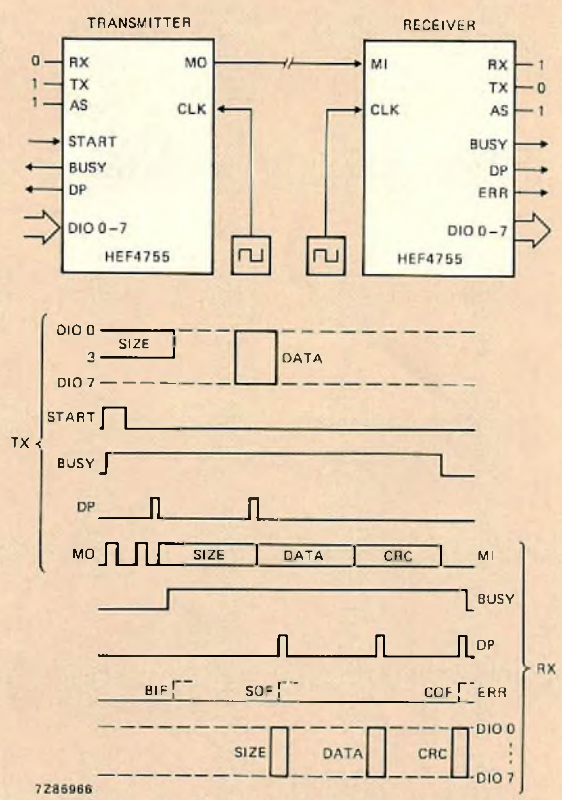

Fig.7 Pulse diagram for asynchronous data transfer. In this mode, the bit check causes a delay of one bit period

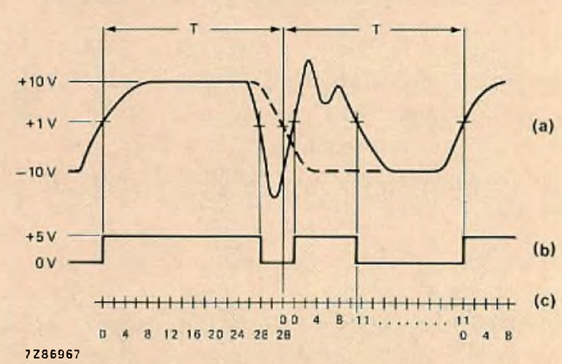

Fig.8 The bit check. (a) two distorted bits on the message line. The dashed line indicates undistorted bits; (b) the transmitted bits after passing through a line receiver; (c) number of samples evaluated as a logical 1

DATA FOR THE HEF4755

Ratings

See family specifications in the HE family data book.

Characteristics

As for the HE family except for:

V_{DD} = 4.5 V to 12.5 V

V_{OL} < 0.4 V at I_{OL} = 1.6 mA

V_{OH} > V_{DD} – 1 V at I_{OH} = –1 mA

Outputs can drive 1 TTL load = 4 LSTTL loads.

Package

28 pin DIL (Plastic or Cerdip)

Baud rates

The maximum allowable Baud rates and clock frequencies are:

V_{DD}	synchronous Baud rate (Mbits/s)	synchronous clock freq. (MHz)	asynchronous Baud rate (Mbit/s)	asynchronous clock freq. (MHz)
5 V	0.8	0.8	0.031	1
7 V	1.6	1.6	0.062	2
10 V	3.2	3.2	0.125	4

Pin names and functions

pin no. and name	*function*
1 TST	Test input. Must be connected to V_{SS} during normal operation
2, 3 ML0&1	Inputs for programming message length

ML 0	ML 1	Message length
1	0	2 data bytes
0	1	4 data bytes
1	1	6 data bytes
0	0	variable message length according to the size byte (max. 7 data bytes).

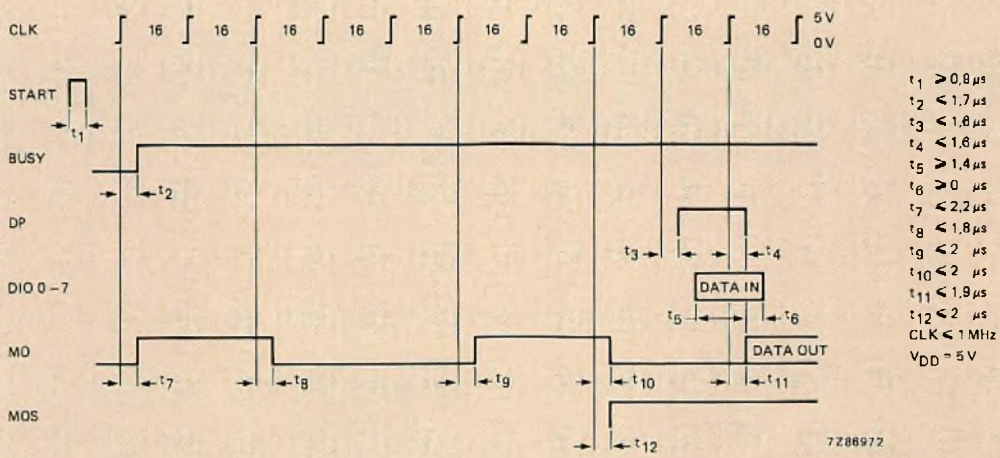

Fig.9 Maximum delay of output signals for asynchronous data transmission

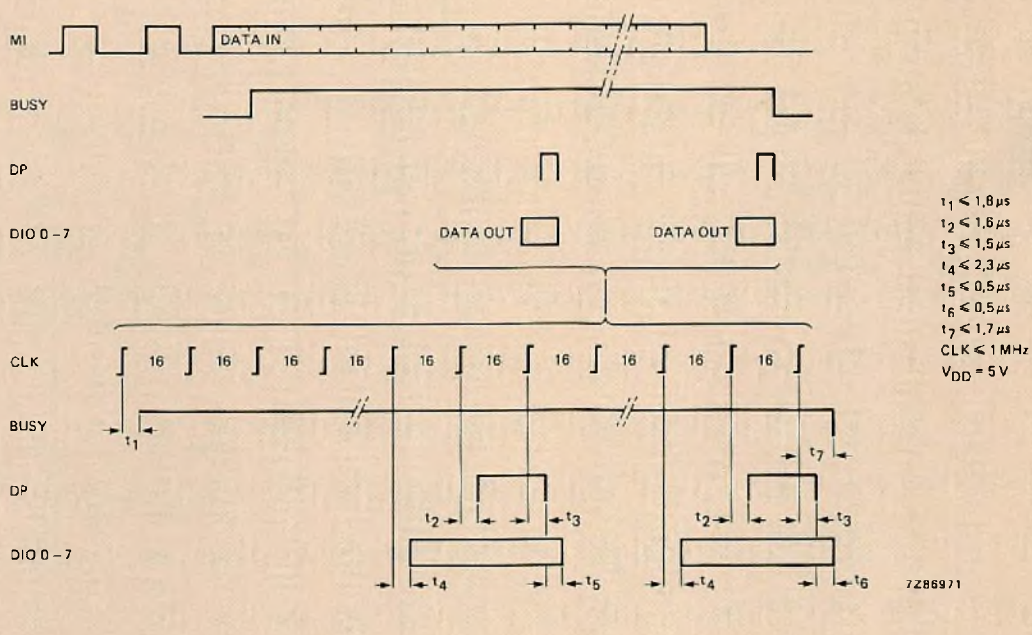

Fig.10 Maximum delay of output signals for asynchronous data reception

pin no. and name	*function*
4 to 11 D10 0 to D10 7	Data I/Os from or to the data bus The number of data bytes is coded in the size byte and transmitted as follows: D10 0 = C, D10 1 = B, D10 2 = A: $n = C \cdot 2^2 + B \cdot 2^1 + A \cdot 2^0$ D10 3 = P = C + B + A The remaining four data I/Os are not taken into consideration when reading the size byte.
12, 13, 15 RX, TX, AS	Inputs for programming the mode of operation

0 0 0 — The 8-bit status register can be read as follows for identifying the type of error and supervising the transfer via the data bus:

BIF bit errors	D10 0
SOF size errors	D10 1
COF code error	D10 2
BEF operating error*	D10 3
MES message input synchronisation	D10 4
FSO release size	D10 5
FDA release data	D10 6
FPZ release check byte	D10 7

Synchronous operation:

1 0 0
- Receiving a message with CRC byte
- Checking the message for transmission errors
- In event of errors, storing the error information in the status register
- Parallel data output to data bus

0 1 0
- Reading parallel data from the bus
- Calculating CRC byte
- Transmitting a message with CRC byte

1 1 0
- Receiving a message without CRC byte
- Parallel data output to data bus
- Calculating CRC byte
- Transmitting the same message with CRC byte

* BEF is set when the HEF4755 requires the data bus for I/O during reading of the status register.

Asynchronous operation

0 0 1
- Bit check
- No parallel data output on the data bus

1 0 1
- Receiving a message with CRC byte
- Bit check
- Checking message for transmission errors
- In the event of errors, storing the error information in the status register
- Parallel data output to data bus

0 1 1
- Reading parallel data from data bus
- Generating the start code
- Calculating the CRC byte
- Transmitting a message with CRC byte

1 1 1
- Receiving a message with CRC byte
- Bit check
- Checking the message for transmission errors
- In the event of errors, storing the error information in the status register
- Parallel data output to the data bus
- Transmitting a regenerated message with CRC byte

pin no. and name		*function*
16	R	A RESET pulse at this input erases all internal registers
17	START	For synchronous or asynchronous transmission, a START pulse is required at this input For reception in the synchronous mode, the synchronisation signal MOS is required at this input
18	BUSY	Output for the BUSY report during transmission or reception, BUSY = MES + FPZ, MES: message input synchronisation FPZ: release check byte
19	HD	Input for programming the CRC byte length (Hamming distance = $H_{d\,min}$) *HD* 0 = 7-bit CRC byte ($H_{d\,min} = 4$), 1 = 15-bit CRC byte ($H_{d\,min} = 6$)
20	MOS	Output for message synchronisation, serves for recognising start of message in synchronous mode

21	MO	Output for bit-serial message
22	MI	Input for bit-serial message
23	DP	Output for the takeover pulse for the data on the data bus
24	ERR	Output for the overall error report. ERR = 1 = BIF + SOF + COF + BEF BIF = bit error, SOF = size error, COF = code error, BEF = operating error.
25	CLK	Clock input. In the synchronous mode the clock frequency is equal to the Baud rate. In the asynchronous mode the clock frequency is equal to the 32-times the Baud rate
26, 27	TT 0	Inputs for programming the bit duration tolerance for the bit check

TT1	TT0	d(%)	t/T
1	1	37	12/32
1	0	31	10/32
0	1	25	8/32
0	0	19	6/32

14	V_{SS}	OV (corresponds with logic 0).
28	V_{DD}	Supply 4.5 V to 12.5 V (corresponds with logic 1).

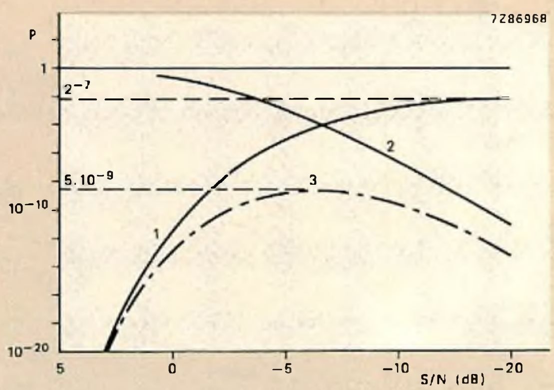

Fig.11 Probability of an unrecognised error as a function of signal-to-noise ratio
(Hamming distance = 4, bit check tolerance = 25%)

APPENDIX

Error detection efficiency

Figure 11 shows the error detecting efficiency of the bit check and the CRC byte as functions of the signal/noise ratio. Curve 1 shows the efficiency of the CRC byte and curve 2 shows the efficiency of bit check. Curve 3 shows the combined efficiency of both checks. The average time (t) between two successive undetected errors is an inverse function of probability P and is given by:

$$t = n/(V \times P)$$

where n is the number of bits in the message and V is the Baud rate.

Figure 11 shows that, with a Hamming distance of 4 for the redundancy check and a 25% tolerance for the bit check, the maximum value of P is 5×10^{-9}. For a Hamming distance of 6 for the redundancy check and a tolerance of 25% for the bit check, the maximum value of P is 6×10^{-13}.

For a Baud rate of 9.6 kbits/s and a Hamming distance of 4:

$$t = \frac{56 + 7}{9.6 \times 10^3 \times 5 \times 10^{-9}} = 15 \text{ days}$$

For the same Baud rate and a Hamming distance of 6:

$$t = \frac{56 + 15}{9.6 \times 10^3 \times 6 \times 10^{-13}} = 390 \text{ years.}$$

The error detection efficiency can be further improved at the cost of data transfer rate by sending a recived message back to the transmitter for checking.

CRC polynomials

The CRC generator polynomials by which the data bytes are divided are:

For a Hamming distance of 4

$$G(X) = X^7 + X^6 + X^2 + 1$$

For a Hamming distance of 6

$$G(X) = X^{15} + X^{14} + X^{10} + X^8 + X^7 + X^4 + X^3 + 1$$

ACKNOWLEDGEMENT

The author wishes to thank Dr. Ing. Gerd Uwe Paul for his cooperation during the preparation of this article.

High-resolution monitor tube

L. J. W. VERSNEL

Despite advances in rival technologies such as liquid crystals and LEDs, the cathode-ray tube remains the workhorse of high-density display. Early displays used tv picture tubes, and some of the less expensive ones still do. For the most part, however, these have now been superseded by tubes developed to meet the special requirements of dot-matrix display. Today's high-performance tubes are capable of resolving as many as a million pixels: sufficient for 8000 well-delineated 7 x 9 dot-matrix characters in 9 x 14 cells. For most alphanumeric displays this is quite adequate.

For some applications, though, it is still not good enough. To show a clear distinction between such symbols as 1, *l*, I and !, or to form characters with subscripts or superscripts, requires at least an 18 x 32 character cell. A full page display to high-resolution facsimile standards requires not one million but nearly four million pixels (1728 x 2288). Other applications calling for high resolution with no degradation of detail include photo-typesetting, complicated graphics, and medical-diagnostic imaging.

The M38-200, a new tube with an extremely small and uniform spot diameter, satisfies all these requirements. On its 200 mm x 270 mm useful screen area it is capable of displaying four million pixels.

TUBE CONSTRUCTION

In general features the M38-200 resembles any other monochrome monitor tube. The difference is in the electron optics.

To achieve a resolution of nearly four million pixels on a screen about the size of an A4 page, the beam spot must be as small as possible but also bright enough for easy viewing in normal surroundings. For practical reasons, the tube outline must be about the same as that of an ordinary monitor, and so must the drive currents and voltages. From the user's point of view the only significant difference, in fact, should be its resolution: the ability to delineate details too fine to be discerned by the unaided eye. It is this that determines the spot diameter requirement.

The most important parameters affecting the spot diameter d_s are the effective cathode temperature T, the diameter d_o of the cathode area from which the beam originates, the accelerating voltage V, the length L over which the beam converges linearly, and the diameter D from which it converges (Fig.1). In very simplified form,

$$d_s \sim \frac{d_o L}{D}\sqrt{\frac{T}{V}}$$

Here, T is fixed by the type of cathode used and 18 kV can be taken as a practical upper limit for V; that leaves only d_o, L and D with which to optimise d_s.

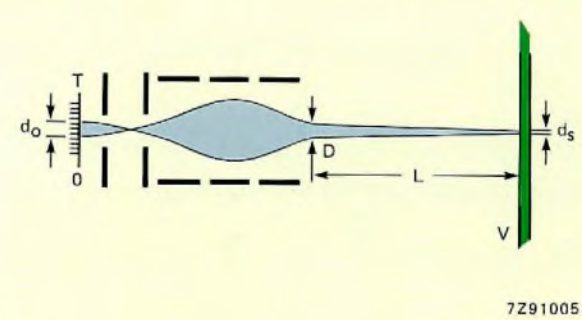

Fig.1 Parameters determining spot diameter d_s

The diameter d_0 of the cathode area from which the beam originates can be made small by using a first grid with a very small aperture. In the M38-200 the aperture diameter is 0.3 mm.

Although a small d_0 is good for obtaining a small spot diameter, it also increases the cathode loading. At the beam current needed for full brightness in the M38-200 the peak loading at the centre of the cathode active area approximates 2 A/cm^2. An impregnated cathode is therefore used to ensure long life.

A short convergence length L and a large convergence diameter D would also help to reduce spot size, but the benefit would be lost owing to deflection defocusing. The larger the deflection angle the worse the effect. In the M38-200 the deflection is limited to 70° to keep deflection defocusing to a minimum.

For focusing, an equidiameter, equipotential lens is used because of its inherently low spherical aberration. The larger the diameter of the focusing lens in relation to the beam, the less spherical aberration it introduces. Since the beam becomes quite wide inside the lens, 18 mm diameter electrodes are used. The gun assembly therefore consists of a miniature beam generating and modulating part followed by what seems a disproportionately large focusing part (Fig.2). Accurate alignment of such dissimilar parts calls for extreme care in fabrication and assembly.

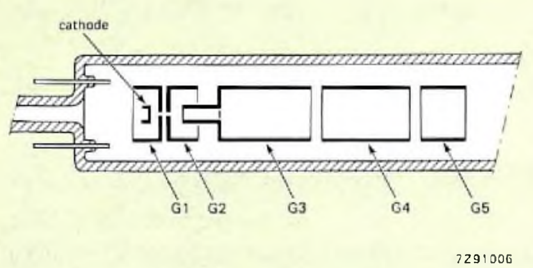

Fig.2 Electron gun of the MS38-200 high-resolution monitor tube

DEFLECTION UNIT

The AT1991 deflection unit designed for the M38-200 is a double-saddle type with sectionally-wound, 'flangeless' line and field deflection coils. Design criteria included optimum focus in the corners, minimum spherical aberration, and minimum astigmatism. The deflection fields are such that no raster correction is required: linear deflection currents give a straight-sided rectangular raster. There is some deflection defocusing, but it is easy to correct with dynamic focusing. The unit incorporates adjustable permanent magnets for centring the raster and eliminating residual astigmatism.

RESOLUTION AND SMALL-AREA CONTRAST

At screen centre the resolution of the M38-200/AT1991 combination is equivalent to more than 3000 tv lines. A more relevant criterion for evaluating the performance, however, is the CCITT Group III standard for digital facsimile. This specifies a line raster having 7.7 lines per millimetre, each consisting of 8 dots per millimetre.

The finest horizontal black line that can be traced on such a raster is obtained by omitting one raster trace; the finest vertical black line, by omitting one dot from each successive trace. To appreciate the facsimile performance of the M38-200 it is necessary to consider how such a line contrasts with its surroundings, and how the contrast depends on spot diameter and line spacing.

The luminance profile of the spot is approximately gaussian (Fig.3), as described by

$$\frac{B(y)}{B_{max}} \approx \exp\left(-\frac{y^2}{2\sigma^2}\right)$$

where

B_{max} is the peak luminance at the spot centre

$B(y)$ is the luminance at distance y from the spot centre

σ is the standard deviation of the gaussian luminance distribution.

All the customary definitions of line width and spot diameter are expressed in terms of σ:

d_{sr} = 2σ, shrinking-raster line width

d_{50} = 2.35σ, full width at half maximum

$d_{1/e}$ = 2.83σ, width or diameter at which $B(y) = (1/e)\,B_{max}$

d_{05} = 4.9σ, width or diameter ('thermal spot size') at which $B(y) = 0.05B_{max}$.

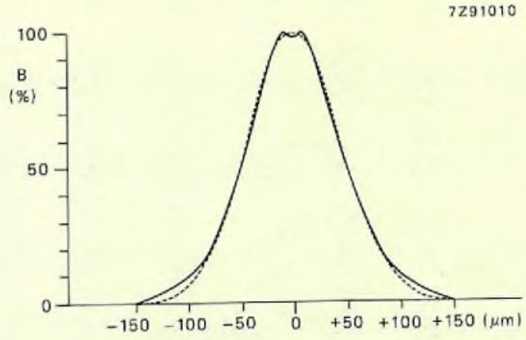

Fig.3 The spot luminance distribution of the MS38-200 (solid line) is approximately gaussian (dashed line)

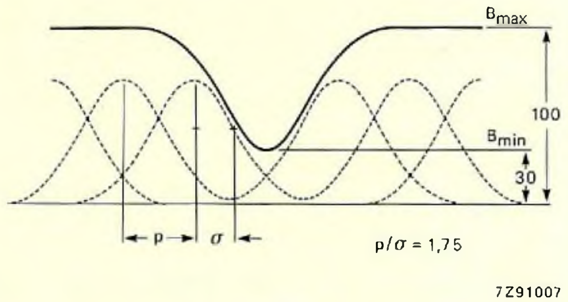

Fig.4 Luminance distribution to either side of a horizontal black line formed by omitting one raster trace

The worst-case spot diameter for the M38-200, d_{05} = 0.35 mm in the corners of the screen, gives $\sigma = 0.35/4.9 = 71.4\,\mu m$. Using this as a basis, the luminance gradient on either side of a single horizontal black line is easy to calculate and is plotted in Fig.4. A vertical black line can be treated similarly (Fig.5), provided spot velocity and video pulse duration are taken into account.

Commonly applied measures of the ease with which a dark feature can be distinguished from a light background are small-area contrast, C_s, and modulation depth, M; both are functions of the relative luminance of the feature and the background:

$$C_s = \frac{B_{max}}{B_{min}} \qquad M = \frac{B_{max} - B_{min}}{B_{max} + B_{min}} \times 100\%$$

From Fig.4 it can be seen that, for a black horizontal line traced on a CCITT facsimile raster, the M38-200 gives

$$C_s = \frac{1}{0.3} = 3.33 \qquad M = \frac{0.7}{1.3} \times 100\% = 53\%$$

and, from Fig.5, for the corresponding vertical line

$$C_s = \frac{1}{0.4} = 2.5 \qquad M = \frac{0.6}{1.4} \times 100\% = 43\%$$

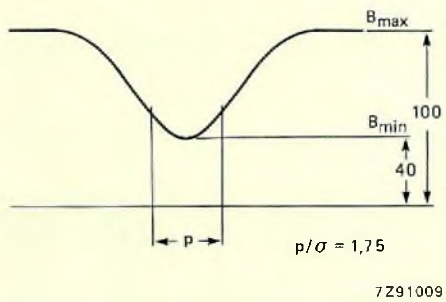

Fig.5 Luminance distribution to either side of a vertical black line formed by omitting one dot from each successive raster trace

Figure 6 shows how modulation depth varies as a function of the ratio of the line or dot pitch, p, to the standard deviation σ.

From the photos in Fig.7, which are tenfold enlargements of rasters with various line spacings but otherwise similar conditions, it is apparent that a modulation depth of 40% is sufficient for good delineation of detail.

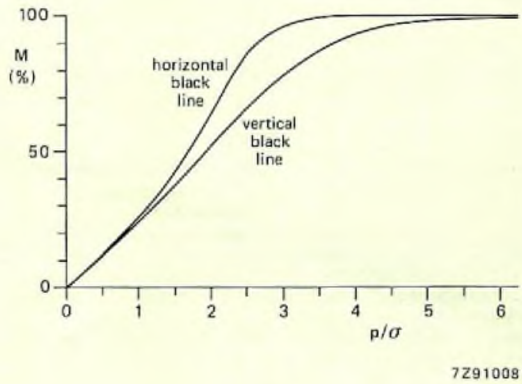

Fig.6 Modulation depth of a single horizontal line and a single vertical line as functions of p/σ, the ratio of raster pitch to the standard deviation of the dot luminance distribution

LUMINANCE AND BEAM CURRENT

The luminance of the screen, B_s, is a function of its area S, the beam current i, the accelerating voltage V, the phosphor efficiency η, and the glass transmission T_g:

$$B_s = \frac{\eta i V T_g}{\pi S}$$

For facsimile purposes the screen should be capable of duplicating the luminance of a white page in a well-lit room. The luminance of a page with a reflection coefficient of 0.9 illuminated by 500 lux is

$$B = \frac{500}{\pi} \times 0.9 = 143 \text{ cd/m}^2$$

The M38-200 has a useful screen area $S = 0.054\,m^2$ and a glass transmission $T_g = 0.5$; thus, with a phosphor efficiency of 45 lm/W and an accelerating voltage of 18 kV, the required beam current is

$$i = \frac{143 \times 0.054 \times \pi}{45 \times 18 \times 10^3 \times 0.5} = 59\,\mu A$$

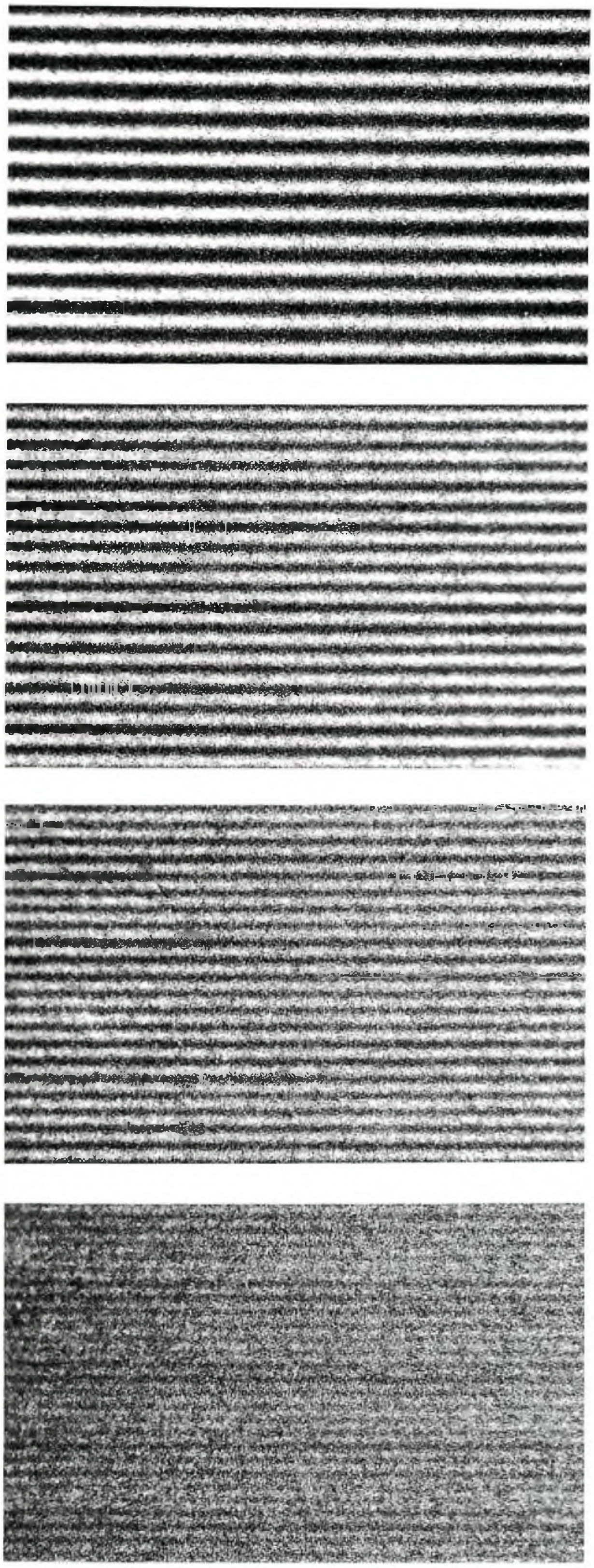

Fig.7 Tenfold enlargement of line rasters with (top to bottom) 92%, 62%, 40%, and 16% modulation depth

LARGE-AREA CONTRAST

Large-area contrast is the contrast between lit and unlit parts of the screen; e.g. between a 'black' square and the white background on which it is displayed. Owing to ambient illumination and the reflectivity of the phosphor, the luminance B_0 of the unlit parts of the screen is not zero:

$$B_0 = \frac{ET_g^2 r_p}{\pi}$$

Here, r_p, the reflection coefficient of the phosphor, is about 0.8, and E, the ambient illumination as it affects the screen, can be taken to be about 200 lux for a vertically disposed screen in a well-lit room. Thus,

$$B_0 = \frac{200 \times 0.5^2 \times 0.8}{\pi} \approx 13\ \mathrm{cd/m^2}$$

and the maximum large-area contrast is

$$C = \frac{B_S + B_0}{B_0} = 12$$

which corresponds to a modulation depth of about 85%.

Ambient light also degrades the small-area contrast; corrected for this, the effective small-area contrast becomes

$$C_s' = \frac{C}{1 + \frac{C-1}{C_s}}$$

For a horizontal black line ($C_s = 3.33$) this gives $C_s' = 2.79$; and, for a vertical line ($C_s = 2.5$), $C_s' = 2.22$.

M-38-200 monitor tube and AT1991 deflection unit

PHOSPHORS

The M38-200 is available with a choice of phosphors for specific applications. For document display, phosphor WA is best; its warm white facilitates eye adaptation when the screen and an actual page are viewed alternately. Medical workers may prefer standard white W (P4) because its bluish cast is easier to correlate with X-ray photos. If maximum brightness is a criterion and green is acceptable, GH is the phosphor of choice.

All these are short-persistence phosphors, necessitating a frame frequency of at least 75 Hz to avoid flicker. (Longer-persistence phosphors can be made available, but they tend to degrade image quality – 'picture smear' – and are not recommended.) In most applications a frame frequency of 75 Hz implies a line frequency of at least 100 kHz and a video bit rate of 200 Mbit/s. The demands on the electronics are therefore considerable, but well within present-day limits. Work stations of the Megadoc filing system (Ref.1, 2), for instance, use M38-200 tubes for document display, and Ref.3 gives details of an M38-200 display with 100 Hz frame frequency and 125 kHz line frequency that uses a number of standard tv parts, including a tv frame-frequency oscillator and a tv line-output transformer.

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Microcomputer peripheral IC tunes and controls a tv set

K. H. SEIDLER

The necessity for television set manufacturers to reduce costs, provide more features, simplify tuning and incorporate remote control has led to a need for all-electronic digital tuning and control circuits. Naturally enough, component manufacturers would prefer to meet the need with a dedicated integrated system which they can make in large quantities. This however is impractical because it would not allow the set manufacturers to satisfy the widely varying requirements of the tv market. The most suitable system is therefore one controlled by a standard microcomputer (e.g. one from the MAB8400 family), so that the variants can be accommodated by software. The only additional components than then need to be separately integrated are those required for interfacing and for performing functions that cannot be handled by the microcomputer because of speed, voltage or power consumption considerations. To minimise costs and maximise performance however, the partitioning of the remaining functions and their allocation to various integrated circuits peripheral to the microcomputer must be carefully considered.

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

The foregoing considerations have led to the design of the SAB3035 integrated Computer Interface for Tuning

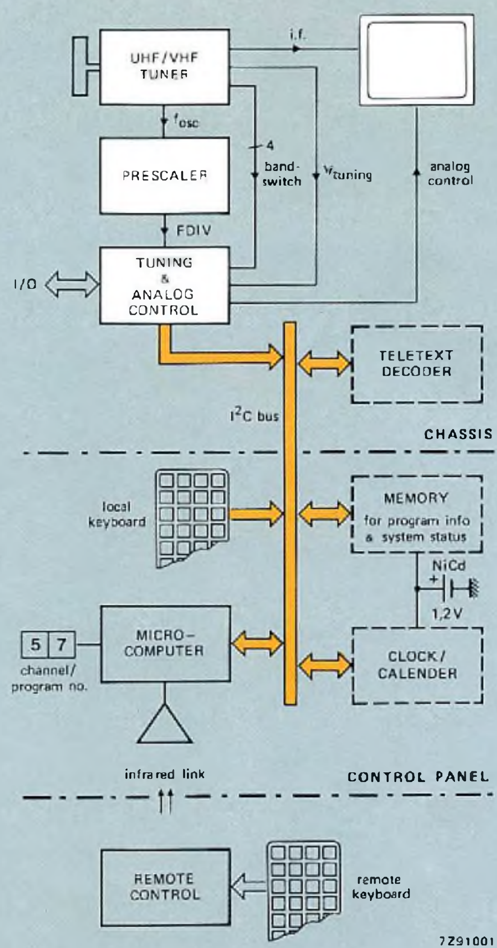

Fig.1 Basic tv control system

and Analogue Control (CITAC). The SAB3035 is an I²C bus-compatible microcomputer peripheral IC for digital frequency-locked loop (FLL) tuning and control of analogue functions associated with the tv picture and sound. As shown in block form in Fig.2, the IC incorporates a frequency synthesiser using the charge pump FLL principle and contains the following circuits:

- 15-bit frequency counter with a resolution of 50 kHz
- charge pump and 30 V tuning-voltage amplifier
- a.f.c. amplifier
- logic circuitry for programming the currents from the charge pump and a.f.c. amplifier
- four high-current band switches
- four general purpose I/O ports for additional control functions
- a one-pin crystal-controlled 4 MHz reference oscillator
- receiving/transmitting logic for the 2-wire I²C bus
- eight static DACs for control of analogue functions associated with the picture and sound.

FUNCTIONAL DESCRIPTION

I²C bus

The SAB3035 is microcomputer-controlled via an asynchronous, Inter-IC (I²C) bus. The bus is a two-wire, bidirectional serial interconnect which allows integrated circuits to communicate with each other and pass control and data from one IC to another. The communication commences after a start code incorporating an IC address and ceases on receipt of a stop code. Every byte of transmitted data must be acknowledged by the IC that receives it. Data to be read must be clocked out of the IC by the microcomputer. The address byte includes a control bit which defines the read/write mode.

Frequency synthesis tuning system

Figure 3 is the block diagram of the frequency synthesising system comprising a frequency-locked loop (FLL) and an external prescaler which divides the frequency of the voltage-controlled local-oscillator in the tv tuner by 64 or 256. The tuning section comprises a 15-bit programmable

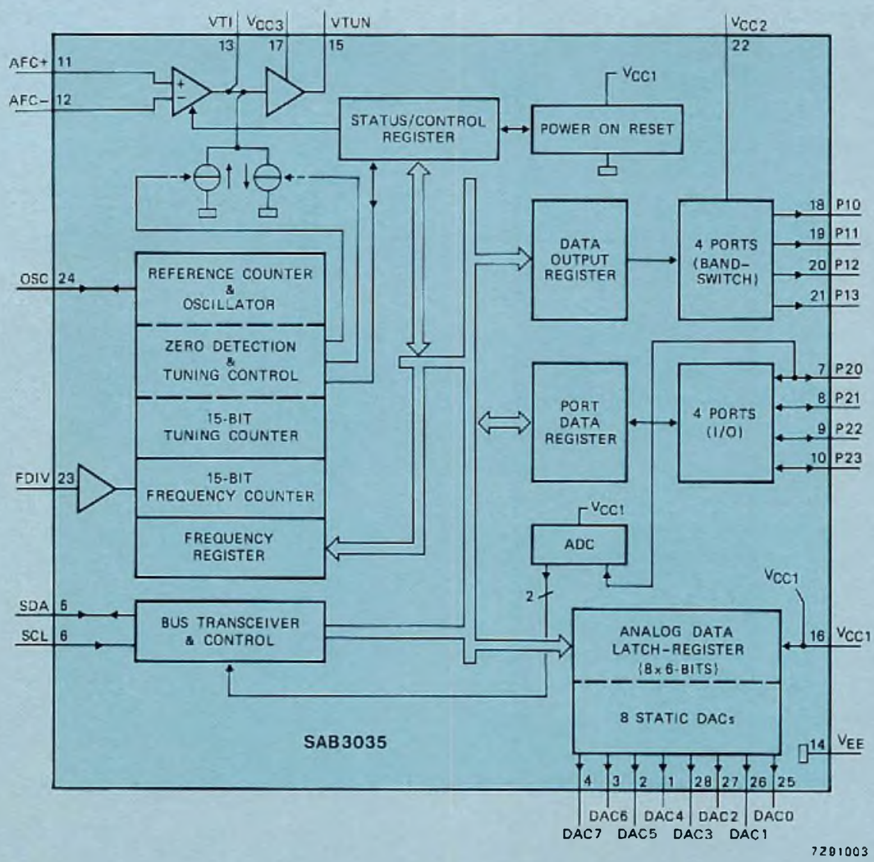

Fig.2 Block diagram of the SAB3035

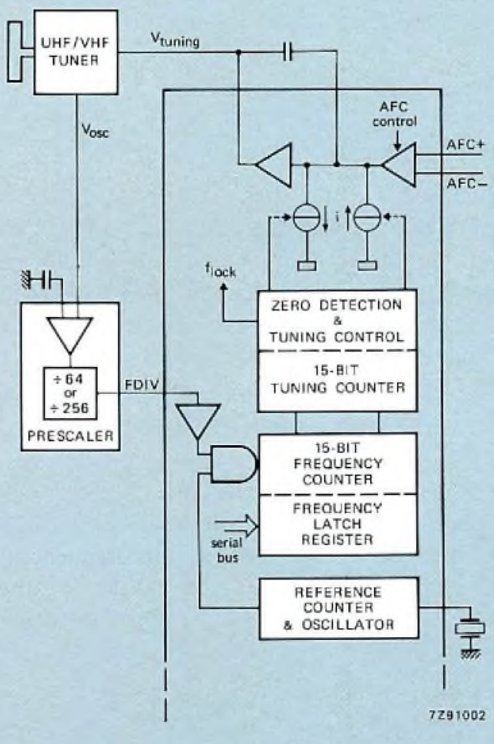

Fig.3 Block diagram of the SAB3035

frequency counter, a 15-bit tuning counter, tuning control and zero detection logic, a reference counter and a charge pump followed by a low-pass filter amplifier.

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

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

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

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

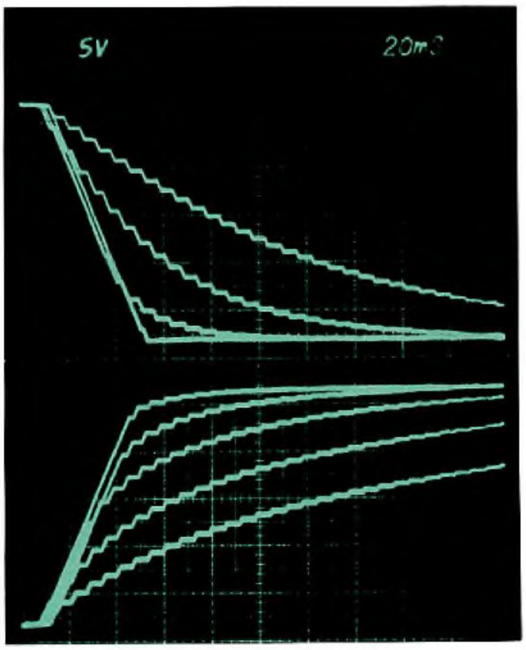

Fig.4 Using some of the selectable charge pump currents for making 50 kHz tuning steps in the u.h.f. band

decreasing frequency (top)
increasing frequency (bottom)

Bandswitching

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

I/O ports

There are four bidifectional ports P20 to P23 for additional control signals to or from the tv receiver. Typical examples of these additional controls are stereo/dual sound, search tuning and switching for external video sources. The output data for ports P20 to P23 is stored in the port data register. Input data must be present during the read cycle. Two of the inputs are edge-triggered. Each input signal transition is stored and can be read by the microcomputer via the serial data bus. The stored data is cleared after each read cycle.

Analogue controls

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

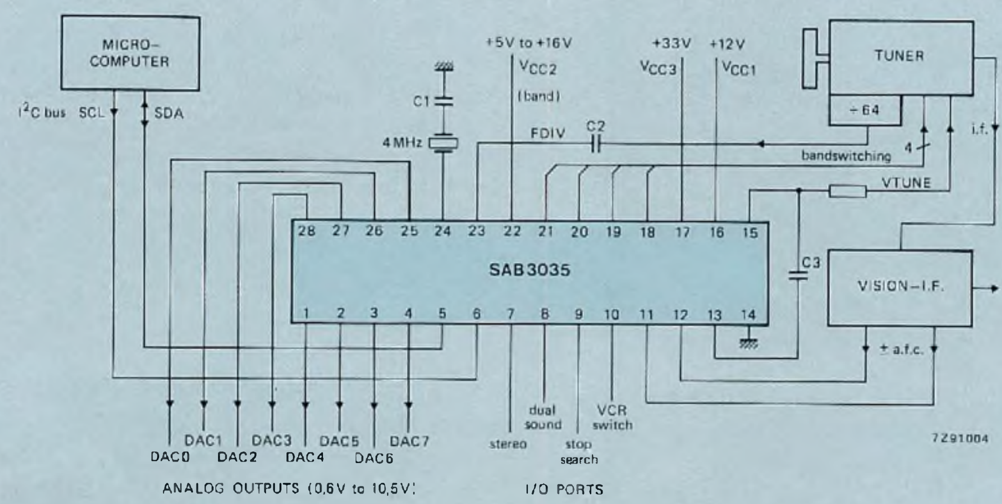

Fig.5 This typical example of the SAB3035 in a tv tuning and control system shows how the peripheral components have been reduced to three capacitors, a resistor and a 4 MHz crystal

ACKNOWLEDGEMENTS

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

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Next generation microcontroller

K. A. H. NOACH

The new 8X305 bipolar microcontroller from Signetics is a faster and more powerful version of the 8X300 which has established itself as the industry standard for 8-bit control in many application areas. And, designers who have already exploited the 8X300 will be particularly interested in the fact that the more powerful 8X305 is functionally interchangeable: both microcontrollers are in similar 50 pin packages.

The 8X305 has been designed together with a full supporting family of devices to make up a complete microsystem hardware package (Fig.1):

8X305	microcontroller
8X310	interrupt controller
8X320	bus interface array
8X330	floppy disk controller
8X350	256X8 RAM with address latches
8X360	memory address director
8X371	8-bit bidirectional port
8X372	8-bit bidirectional port (addressable)
8X374	I/O port with parity (addressable)
8X376	8-bit bidirectional port (addressable)
8X382	4-in/4-out I/O port addressable bus expander

Faster than the 8X300 although using the same low-power Schottky technology, the 8X305 will fetch, decode and execute a 16-bit instruction in a minimum of 200 ns. The power of the 8X305 is demonstrated by the fact that only a single instruction enables 8 bits of data to be read from the bus, latched, rotated, marked, combined with other data in an ALU operation, shifted, merged with original bus data and returned to the bus. The bit-oriented instruction set in the 8X305 has been expanded to give even greater flexibility and the number of general-purpose 8-bit working registers has been increased from 8 to 13 to further improve the overall throughput capability.

In conjunction with the new support chips, the 8X305 has extended the already impressive application base of the 8X300 to add new interrupt handling capabilities, larger working storage and parity support.

The 8X305, like the 8X300, is a microcontroller – a microprocessor whose architecture and instruction set are tailored to the requirements of the control system market. A microcontroller's task is to monitor a large number of status bit inputs and provide outputs of one or more bits directed to a given location; in contrast, a conventional microprocessor is essentially a 'number cruncher' providing 8-bit word outputs.

The 8X305 shares the many features which made the 8X300 so popular: bit-oriented instruction set, separate buses for instruction, instruction address and three state I/O, source/destination architecture, TTL input/output compatibility, on-chip oscillator and timing generation, and a single +5 V supply requirement.

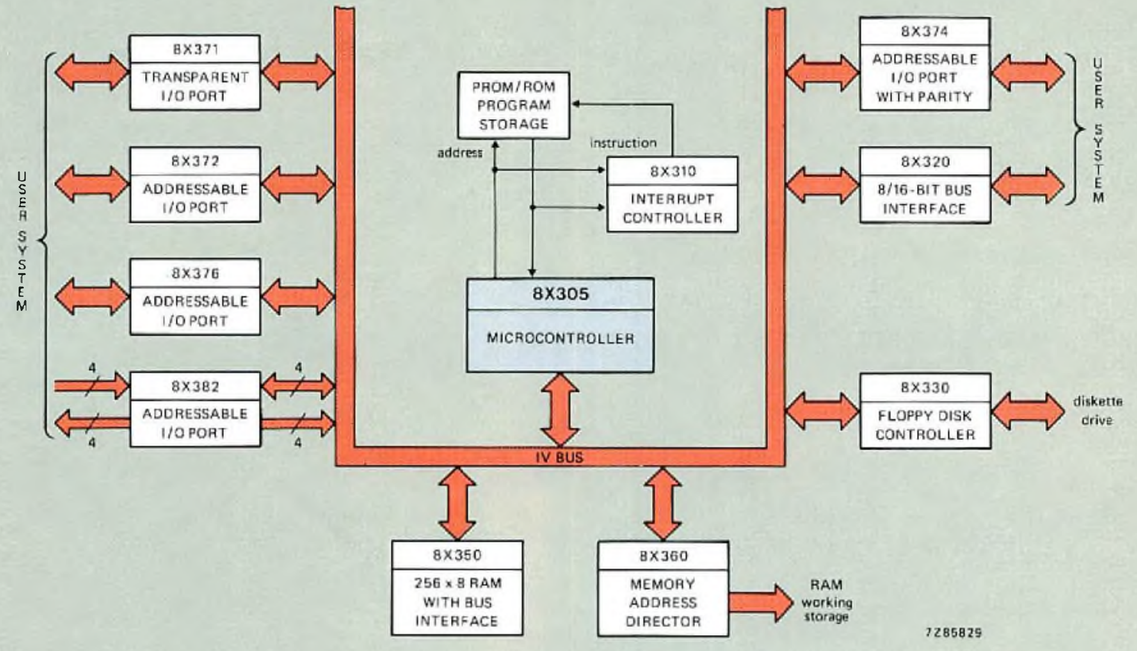


Fig.1 The 8X305 microcontroller and family

PERIPHERALS EXTEND CAPABILITY

Hardware design of an 8X305-based system is straightforward and is aided by the availability of tailor-made supporting devices. Here the design operations are only outlined, full details are given in the comprehensive data for each device in the family.

The first task is, of course, to select and interface with a suitable program storage device (ROM or PROM). Next, select the interfacing input/output devices (RAM, ports and other 8-bit addressable I/O devices). After this, decide on the type of system clock you will use (capacitor controlled, crystal controlled, or externally driven). Then, the only task that remains is the selection of a suitable off-chip series pass transistor for the voltage regulator circuit.

Because of its extremely high speed, the 8X305 is able to use software to perform many operations that would otherwise require additional hardware. For example, a complete diskette drive controller can be built using only ten ICs. The controller is programmable and capable of supporting multiple drive types and formats.

INTERRUPT CONTROL COPROCESSOR 8X310

The 8X310 coprocessor enables the 8X305 (and family) to be used efficiently in such interrupt-driven environments as real-time control systems. In addition, the address handling capabilities of the 8X310 can be used to support subroutine handling.

The device has three prioritised interrupt pins. On receipt of an interrupt signal, the address of the next sequential instruction to be executed (return address) is stored in a 4-level pushdown stack. At the same time, the 8X305 places a JUMP instruction on the microcontroller's instruction bus which forces transfer to an assigned memory location that corresponds to the interrupt pin that initiated the action. When the subroutine is completed, a coprocessor instruction forces a 'JUMP to the next address in the stack' instruction onto the microcontroller bus.

The 8X310 requires a single +5 V supply. The package is 40 pin DIL.

BUS INTERFACE REGISTER ARRAY 8X320

The 8X320 is a dual-port RAM designed to provide a convenient and economical interface between a host processor and a peripheral processor.

The host has 8-bit (byte) or 16-bit (word) user-selectable access to the primary port. The secondary port is bus compatible with the 8X300 and consists of eight input/output lines plus four bus control lines. The array also provides simple handshake control via two 8-bit flag registers, logic to facilitate DMA transfers, and write-protect for the primary port.

A single +5 V supply is required, and the package is 40-pin DIL.

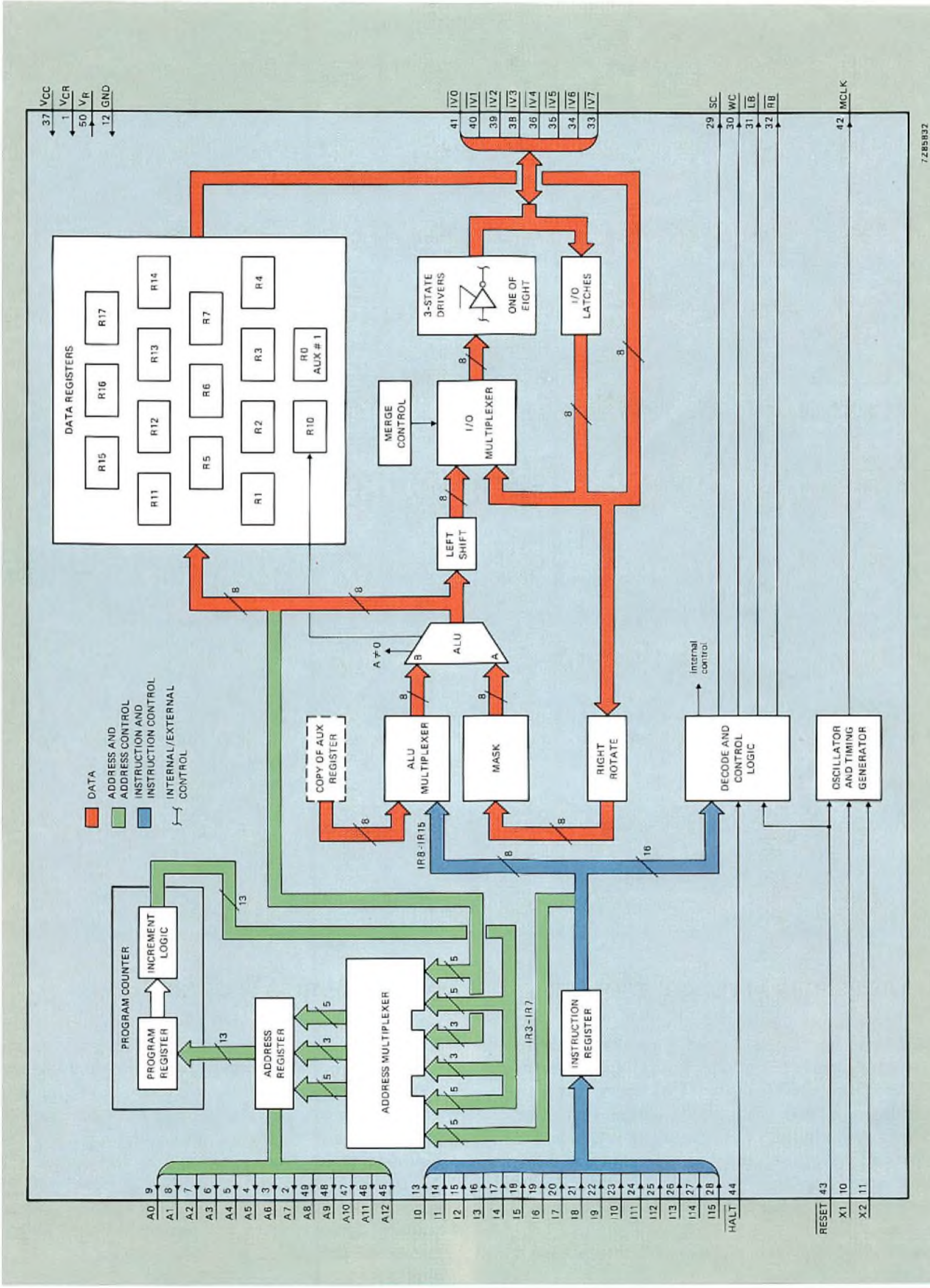


Fig.2 Architecture and pin designation of 8X305 microcontroller

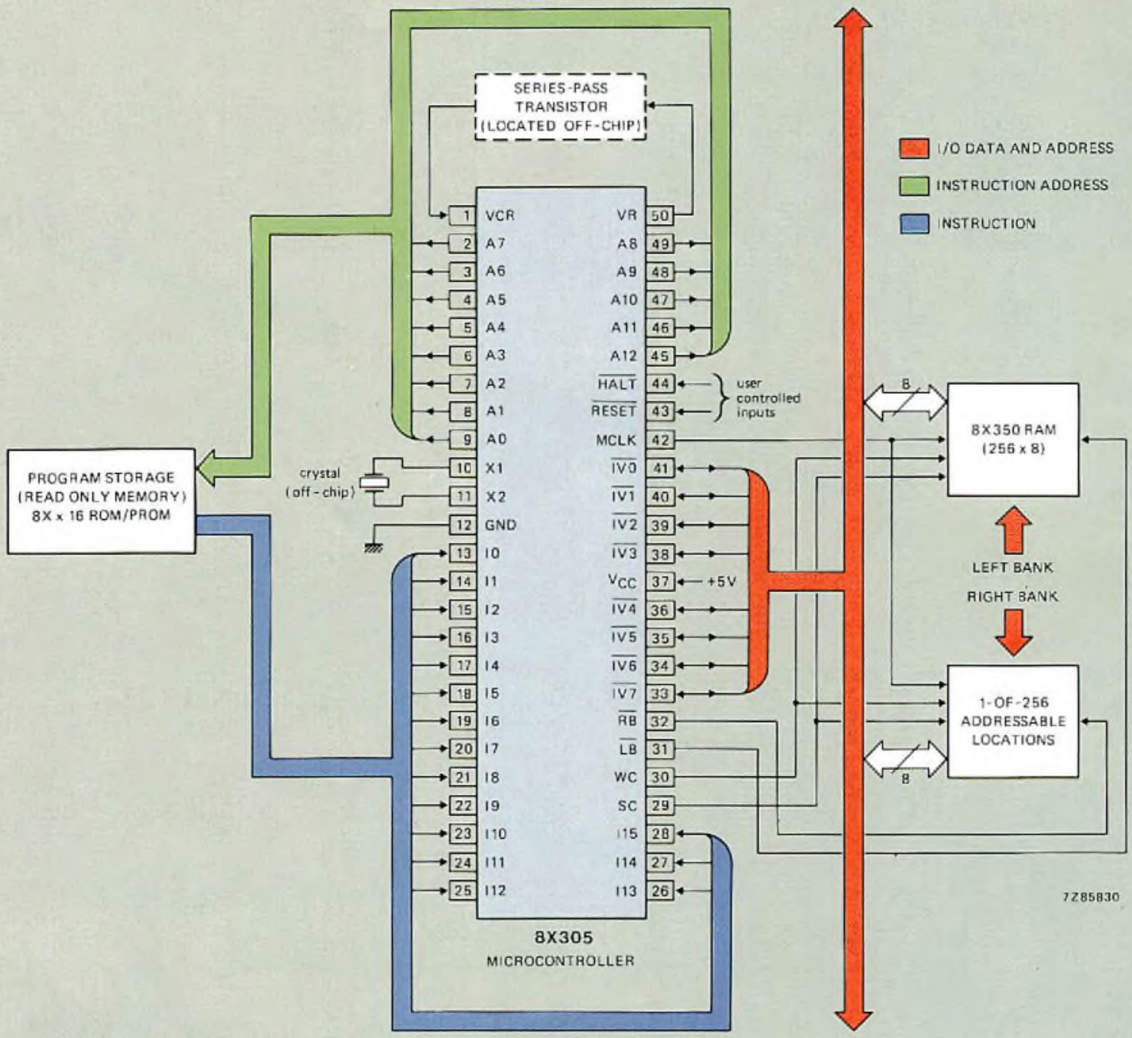


Fig.3 Typical 8X305 system hook-up

FLOPPY DISK FORMATTER CONTROLLER 8X330

The 8X330 contains all the processing circuits and the required control logic to encode/decode double-density (MFM/M^{2}FM) and single-density (FM) codes; the only items not included on the chip are the crystal, a capacitor, the series transistor for the on-chip voltage regulator and an active low-pass filter. Even the data separation and write-precompensation logic are on the chip; in addition, 16-bytes of scratch-pad RAM are provided for storage of control/status parameters.

A single +5 V supply is required and the package is 40 pin DIL.

MEMORY ADDRESS DIRECTOR 8X360

The 8X360 memory address director enables the microcontroller system to be operated at significantly higher data rates by handling address management tasks automatically and so freeing the processor and software for other tasks. It is used in applications demanding large working storage and high-speed data transfer.

All addresses contained in the 8X360 are 16 bits in length, thus permitting the attachment of up to 64K words of working storage. Address registers are automatically incremented or decremented at the user's option.

Supply is +5 V, package is 40-pin DIL.

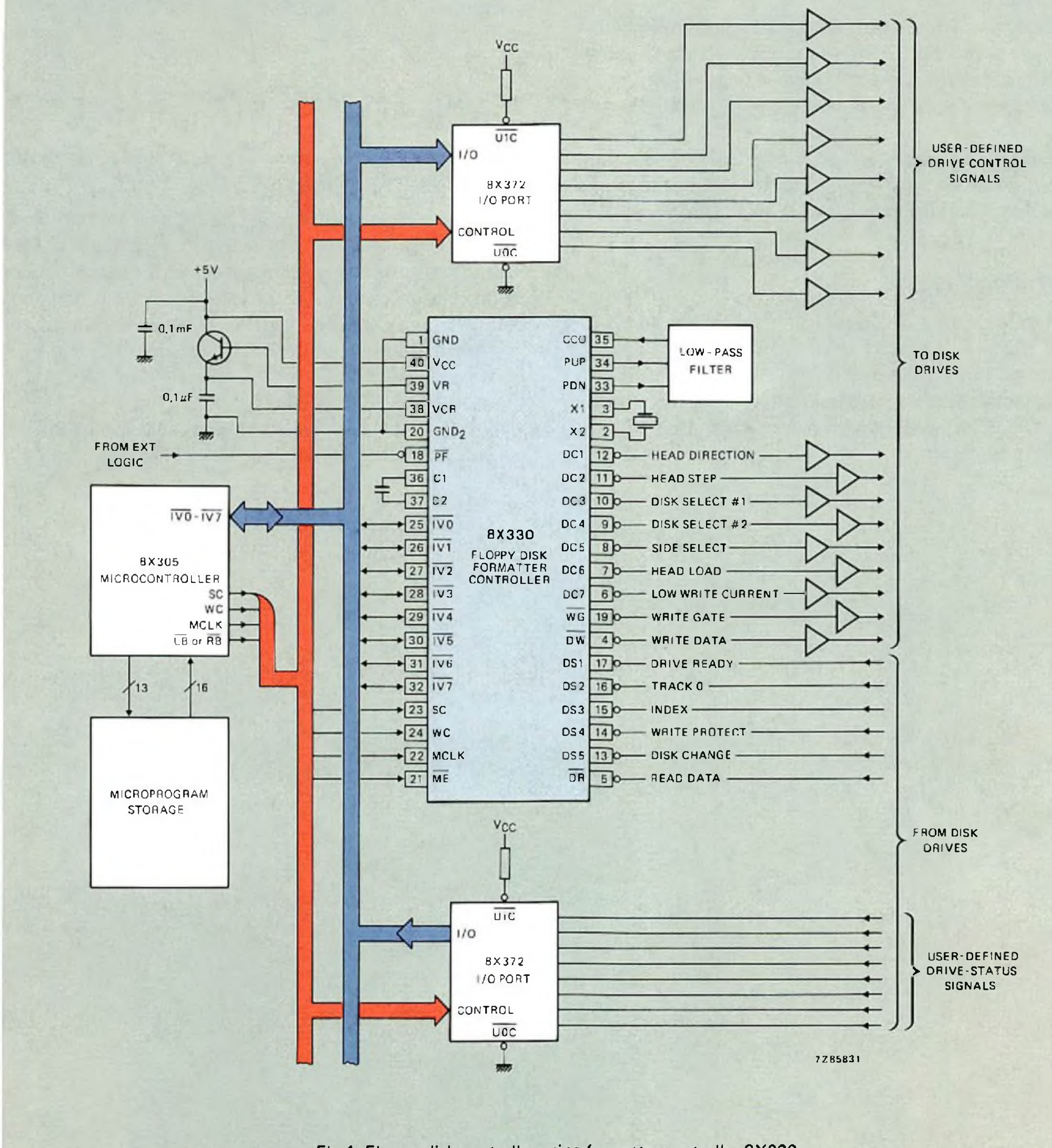


Fig.4 Floppy disk controller using formatter controller 8X330

I/O PORT 8X371

The 8X371 is an 8-bit bidirectional I/O port similar to the 8X372/8X376 but not programmable. In some applications it can be used to enhance the system performance by using a technique called extended microcode or fast IV select.

I/O PORTS 8X372 AND 8X376

The 8X372 and 8X376 are addressable bidirectional I/O ports for systems with TTL compatible buses. The 8X372 is synchronous, the 8X376 is asynchronous. Each consists of eight identical data latches. The latches are transparent and accessed through two bidirectional 8-bit buses. Both buses operate independently and the user bus has priority for data entry.

Both devices are available with pre-programmed addresses: either device can be field-programmed over the same address range (0_{10} through 255_{10}). A port is selected by putting its address on the IV bus; the port remains selected until a different port address is put on the bus. Data is accessible on the user bus at all times. The IV bus can be organised into two separate and independent banks of I/O devices by an Enable input. Packages are 0.4 in, 24-pin DIP.

I/O PORT 8X374

The 8X374 is an addressable bidirectional I/O port with parity which is identical to the 8X372 except for the addition of parity circuitry. This enables the input data to the microcontroller to be checked for parity and a parity signal to be included on the output data. Odd or even parity can be chosen.

I/O PORT 8X382

The 8X382 is a 4 input/4 output addressable I/O port for systems with TTL compatible buses. Data input is synchronous. The port consists of eight data latches accessed through either:

- an 8-bit bidirectional IV bus connected to the microcontroller
- a user data bus consisting of four dedicated inputs and four dedicated outputs.

This port is unique because it can be used for both dedicated input and output operations. This facility can be used to reduce the overall port package count; it can also be used to implement a handshake interface. Since both input and output bits reside in the same port, I/O operations can be performed without readdressing. Pre-programming, field programming and port selection operations are identical to those of the 8X372 and 8X376.

A single +5 V supply is required and the package is 0.4 in, 24-pin DIP.

ERRATUM

In the article in Vol. 4, No. 4 'Long-term reliability of linear bipolar ICs' we neglected to acknowledge the contribution of Mr. Schelski of Valvo, Germany to the exercise reported.

A complete fm radio on a chip

Until now, the almost total integration of an f.m. radio has been prevented by the need for LC tuned circuits in the r.f., i.f., local-oscillator and demodulator stages. An obvious way to eliminate the coils in the i.f. stages is to reduce the normally used intermediate frequency of 10.7 MHz to a frequency that can be tuned by active filters. The op-amps and resistors of the active filters can then be integrated. An i.f. of zero would seem to be ideal because it would also eliminate spurious signals such as repeat spots and image response, but the i.f. signal could not then be limited prior to demodulation, resulting in poor S/N ratio and no a.m. suppression. An i.f. of 70 kHz overcomes these problems and positions the image frequency about half way between the desired signal and the centre of the adjacent channel. In common with all conventional f.m. radios, there remains a need to suppress excessive noise when the radio is not tuned to a station or is tuned to a weak signal, spurious responses above and below the centre frequency of the desired station (side tuning), and harmonic distortion due to very inaccurate tuning.

We have now developed a revolutionary f.m. reception system which allows the almost total integration of a mono f.m. radio. It uses an active 70 kHz i.f. filter and a unique correlation muting circuit to suppress side responses caused by the flanks of the demodulator S-curve. With such a low i.f., distortion would occur with the usual ±75 kHz i.f. swing with maximum modulation. The maximum i.f. swing is therefore limited to ±15 kHz by negative feedback from the demodulator output to the local-oscillator in a frequency locked loop (FLL). This results in distortion of only 2% with $\Delta f_{mod} = \pm 75$ kHz. The combined action of the mute system and the FLL also suppresses image response.

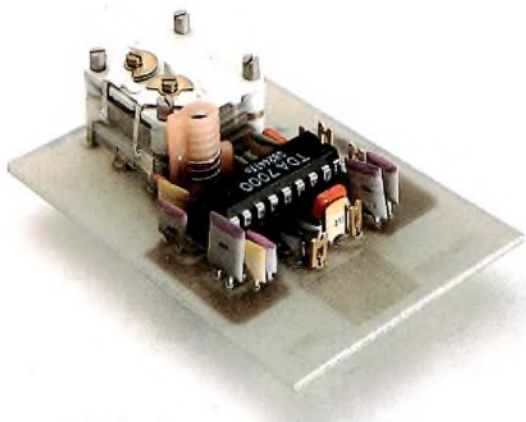
The new circuit is the TDA7000 which integrates a mono f.m. radio all the way from the aerial input to the audio output. External to the IC are only one tunable LC circuit for the local-oscillator, a few inexpensive ceramic plate capacitors and one resistor. The TDA7000 dramatically reduces assembly costs and totally eliminates the need for tuned circuit alignment during manufacture. The complete f.m. radio can be made small enough to fit inside a calculator, cigarette lighter, key-ring fob or even a slim watch. The TDA7000 can also be used as a receiver in equipment such as cordless telephones, CB radios, radio-controlled models, paging systems and the sound channel of a tv set.

Using the TDA7000 results in significant improvements for all classes of f.m. radio. For simpler portables, the small size, lack of coils, easy assembly and low power consumption are not the only attractive features. The unique correlation muting system and the FLL make it very easy to tune, even when using a tiny tuning knob. For higher-performance portables and clock radios, station presetting facilities using variable-capacitance diode tuning are often required. These are simple to provide with the TDA7000 because there are no variable tuned circuits in the r.f. signal path. Only the local-oscillator needs to be tuned, so tracking and distortion problems are eliminated.

The TDA7000 is available in either a 28-lead plastic DIL package (TDA7000), or in a 16-pin SO package (TDA7010T). Future developments will include reducing the present supply voltage (4.5 V typ.), and the introduction of f.m. stereo and a.m./f.m. versions.

BRIEF DATA

Typical supply voltage	V_P	4.5 V
Typical supply current	I_P	8 mA
R.F. input frequency range	f_{rf}	1.5 to 110 MHz
Sensitivity for −3 dB limiting e.m.f. with $Z_S = 75\ \Omega$, mute disabled	$V_{rf\text{-}3\,dB}$	1.5 μV
Maximum signal input for THD < 10%, $\Delta f = \pm$ 75 kHz e.m.f. with $Z_S = 75\ \Omega$	V_{rf}	200 mV
Audio output (r.m.s.) with $R_L = 22\ k\Omega$	V_o	90 mV



A laboratory model of the TDA7000 in a complete f.m. radio

CMOS gate arrays - the fast way to semi-custom logic

Fully custom ICs are compact, reliable and dissipate the minimum of power but take a long time to develop and are made in large quantities to obtain acceptable unit cost. Standard logic ICs cost less but occupy more sockets and board space, cost more to assemble and test, and the overall system dissipates more power and is less reliable. A semi-custom mask-programmable gate array bridges the gap because it is an off-the-shelf chip which can be programmed to perform exactly the logic functions you need. Design and development times are much shorter, and development costs much lower, than for fully custom logic. It also has the advantage of more design flexibility and higher gate utilisation than other types of programmable logic array.

Our PCF/PCC family of mask-programmable silicon-gate CMOS gate arrays comprises five types of pre-processed silicon chips consisting of a central matrix of 165, 224, 352, 558 or 864 cell units (two 2-unit NAND/NOR gate equivalents) and a periphery of 38, 26, 38, 66 or 84I/O devices with 40,28,40,68 or 86 bonding pads. They are backed by an extensive computer-aided design (CAD) package. Starting from a logic network description, which represents the overall logic circuit required by the customer, computer simulation program SIMON automatically calculates gate delays and accurately predicts the logical behaviour of the circuit. It also generates test patterns which, together with a pin assignment list, are used to generate an automatic test program. An automatic cell placement program PLACE, and an automatic interconnect program INGATE completely eliminate tedious hand layout and allow a system designer to also design his own IC. The output from the INGATE program directly generates the mask pattern tape for programming the array. Since the customer can participate at various stages of the design, he can retain control over the entire process of converting his logic diagram into a semi-custom IC.

To convert his logic design into a programmed PCF/PCC gate array, the logic circuit designer uses a comprehensive computer-based cell function library consisting of macro descriptions and pre-defined wiring and contact mask patterns which are ultimately used to connect the uncommited cells to perform the required logic functions. The cell library also contains a wide selection of I/O circuits and pull-up/down resistors which can be assigned to the bonding pads. Designing with a mask-programmable gate array is therefore just as easy as designing a printed-wiring board by using a standard set of interconnection patterns which are guaranteed to be correct for performing specific functions. Tedious and error-prone tasks like breadboarding, digitising the layout and defining interconnections by hand are completely eliminated.

Outstanding advantages of our PCF/PCC family of mask-programmable gate arrays are:

- Over 90% of available gates can be used in a functional circuit
- Designing is as easy as with standard SSI or MSI
- Design changes are easy to incorporate
- A wide selection of I/O types and a range of pull-up/down resistors allow interfacing with most other logic families
- The PCF/PCC family is backed by a comprehensive Computer–Aided Design (CAD) system. This dramatically reduces development time and allows the designer of the original logic circuit to retain control of his own system throughout the final integration phase instead of relying on the extertise of a MOS IC designer
- The network description is based on a computer-based library of fully defined cell functions. The input/output list, dynamic characteristics and topology list of the logic functions are part of the cell library macro descriptions. CAD program SIMON simulates the overall circuit and tests it to discover and correct any logical errors, race hazards, spiking conditions or timing uncertainties before interconnection and contact masks are made
- After functions have been assigned to the array cell units and I/Os, CAD routeing program INGATE calculates the correct interconnections
- INGATE uses the network description of SIMON which automatically specifies metallisation and contact mask patterns that are guaranteed to be correct for interconnecting bonding pads, I/Os, supply rails and cell transistors
- Fast turnaround for development samples
- The final semi-custom logic IC protects proprietary rights because it is very difficult to copy.

Research news

Fast-electron lithography - another step toward LSI

In the manufacture of integrated circuits using electron lithography, backscattering of electrons often causes problems in the drawing of fine, close details. At the Philips Research Laboratories, Redhill, England, this problem has been overcome by working with faster electrons.

The trend towards increasing miniaturisation of integrated circuits requires the development of increasingly advanced techniques for applying the required patterns to a silicon wafer. Details of about 2.5 μm can be made by currently used photolithographic methods; with very high-quality machines, such as the Philips wafer stepper, details as small as 1 μm can be obtained. If electron-beam lithography is used instead of photolithography, it is possible to obtain a line width of 0.1 μm. The electron beam writes the patterns directly on the sensitive layer on a silicon wafer, without using masks. However, the electrons focused onto the sensitive layer get scattered both in the layer and in the silicon substrate. As a result, there is also some exposure on either side of the written line, which causes blurring. When fine patterns are drawn the scattering causes deformation of details that are close together. This is known as the proximity effect.

At Philips Research Laboratories the proximity effect has been eliminated by using 50 keV fast-electron beams instead of the usual 20 keV beams. The electrons are then scattered deep in the substrate and hardly at all in the sensitive layer; the backscattered electrons are spread over such a large distance on either side of the written line that their contribution to the exposure level in the lacquer is negligible.

Figure 1 shows the difference between pattern features obtained in a thick lacquer film with 20 keV and with 50 keV lithography. Figure 2 shows a grid of 0.2 μm line windows at 0.4 μm pitch in a lacquer film.

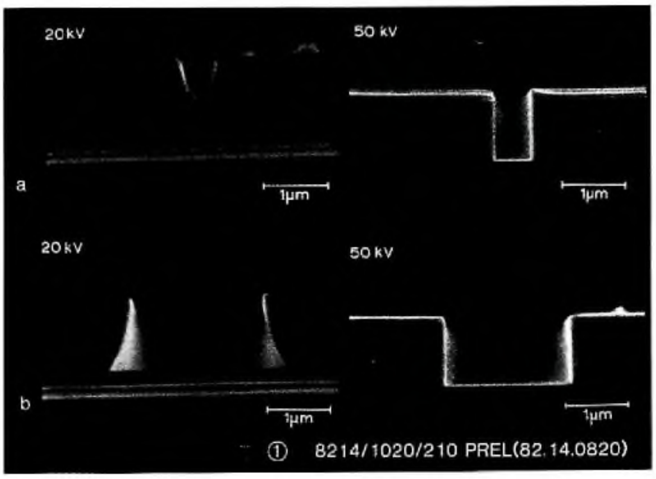


Fig.1

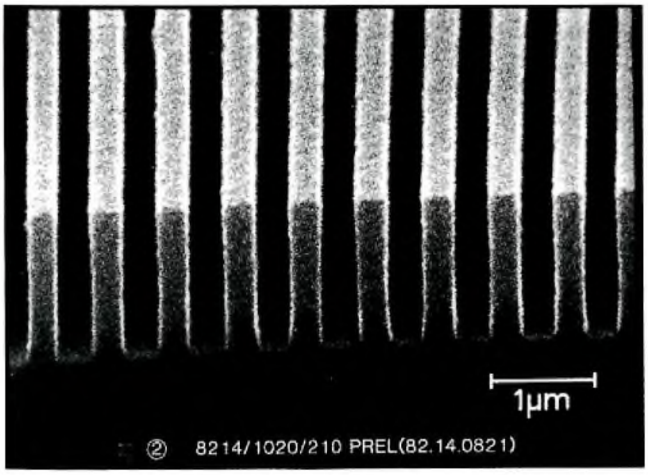


Fig.2

These notes report activities of Philips research laboratories and do not imply commercial availability of any product embodying the described results. For further information, written application should be made to the Publicity Department, Philips Research Laboratory, Eindhoven, The Netherlands

Abstracts

Bipolar IC merges telephone dialling and speech

This article describes a new bipolar IC which incorporates all the functions for an electronic telephone on a single bipolar chip. The functions performed are dialling tone generation, speech amplification, 4/2-wire conversion, dialling/speech switching and line adaptation. The dialling inputs can be directly driven by a simple rubber membrane keyboard or via a microcomputer if extended dialling features such as automatic redial or repetory dialling are required.

Circuits for modular design of consumer and industrial products – the CLIPS system

A newly introduced family of dedicated and general purpose ICs exploits the underlying similarities of systems that use 8-bit data transfer. Being designed to interface directly with the I^2C bus, these will enable system designers to go straight from block diagram to circuit design without concern for interfacing or compatibility problems. Identification of specific system functions with specific ICs will facilitate circuit board layout and trouble-shooting; it will also lead to the development of modular software that can be adapted easily to a variety of system requirements.

Microwave integrated circuits – design and realisation

Microwave integrated circuits (MICs) can provide considerable size reduction in microwave systems, and can, moreover, ease problems of interfacing between transmission-line systems and semiconductor devices. The smallness of MICs, however, brings problems of its own. In finding solutions for these, much expertise has been built up, both electrical and technological. Basically, reliability in MICs centres around using established recipes and following established procedures, so these must be maintained as far as possible whenever a new device is being contemplated or developed.

Fast and undistorted data transfer in noisy environments

A new circuit for secure and efficient serial transmission or reception of parallel data in electrically noisy environments is described. The data transfer can be either synchronous or asynchronous and the circuit includes a programmable bit level check, programmable cyclic redundancy check and byte size verification.

High-resolution monitor tube

A new 38 cm monochrome c.r.t. is able to resolve nearly four million pixels on a screen about the size of an A4 page. Operated as a facsimile terminal, it provides a modulation depth of better than 50% for a horizontal black line one pixel wide displayed against a white raster with a luminance of 143 nit. The large-area contrast at that luminance corresponds to a modulation depth of 85%.

Microcomputer peripheral IC tunes and controls a tv set

Recent technology advances and extensive experience of the digital control and tuning of tv sets have led us develop a new I^2C bus-compatible microcomputer peripheral IC. The circuit performs frequency synthesis together with on-chip band switching and tuning voltage generation. It also controls up to eight analog functions such as volume, tone, brightness, contrast, colour saturation etc. There are also I/O ports for additional controls such as stereo/dual sound, search tuning and switching of external video sources.

Next generation microcontroller

A new microcontroller using low-power Schottky technology will fetch, decode and execute a 16-bit instruction in 200 ns. Because of its high speed it is able to use software to perform many operations that would otherwise require additional hardware. The microcontroller is complemented by a family of supporting circuits which, like it, require only a single +5 V supply.

Ein neues bipolares Telefon-IC vereinigt Wähl- und Sprechkreis

Der Aufsatz beschreibt eine neue bipolare integrierte Schaltung, die alle Funktionen eines elektronischen Telefons zusammenfasst. Ein einziges Chip vereinigt die Komponenten für Wähltongenerator, Mikrofonverstärker, elektronische Weiche, Stummschaltung und Leitungsanpassung. Die Eingänge des Wähltongenerators können direkt von einer einfachen Gummimatten-Tastatur oder von einem Mikrokomputer angesteuert werden, so wie dies in Fernsprechapparaten mit automatischer Wählwiederholung oder Kurzwahl erforderlich ist.

Schaltungen für modularen Aufbau von Systemen der Unterhaltungs- und Industrieelektronik – das CLIPS System

Eine neu eingeführte Familie, bestehend aus speziellen sowie Mehrzweck – ICs, nutzt die grundlegenden Gemeinsamkeiten solcher Systeme aus, die mit serieller 8 bit-Datenübertragung arbeiten. Die bei jeder Schaltung vorhandene I^2C-Bus-Schnittstelle ermöglicht es dem Systementwickler, ohne irgendwelche Schnittstellen- oder Kompatibilitätsprobleme vom Blockschaltbild direkt zum Systementwurf überzugehen. Die Realisierung spezieller Systemfunktionen mit speziell hierfür entwickelten ICs erleichtert dabei sowohl den Entwurf der Schaltungsplatine als auch die Fehlersuche. Ausserdem erlaubt und fördert das Konzept die Entwicklung einer modular aufgebauten Software, die sich leicht an eine Vielzahl von System-Anforderungen anpassen lässt.

Integrierte Mikrowellenschaltungen – Entwurf and Realisierung

Integrierte Mikrowellenschaltungen (MICs) ermöglichen eine erhebliche Grössenreduzierung bei Mikrowellensystemen und verringern Probleme, die mit der Verknüpfung von Wellenleitersystemen und Halbleiterbauelementen zusammenhängen. Dagegen entstehen bei integrierten Mikrowellenschaltungen eigene Probleme, die von den geringen Abmessungen herrühren. Um Problemlösungen zu finden, ist auf diesem Gebiet (MICs) umfangreiche Erfahrung in elektrischer und technologischer Hinsicht erarbeitet worden. Grundsätzlich gilt: die Zuverlässigkeit von integrierten Mikrowellenschaltungen wird davon bestimmt in wieweit bewährte Rezepte und anschliessend bewährte Verfahren verwendet werden können. Diese sollen daher so weitgehend wie möglich für neue in der Diskussion oder Entwicklung befindliche Systeme beibehalten werden.

Schnelle und unverzerrte Datenübertragung in störbehafteter Umgebung

Es wird eine Schaltung für serielle Aussendung oder Empfang von parallelen Daten beschrieben, die in mit elektrischen Störungen beaufschlagter Umgebung sicher und mit hoher Leistungsfähigkeit arbeitet. Die Datenübertragung kann entweder synchron oder asynchron ablaufen. Die Schaltung verfügt über eine programmierbare Bitabtastung, programierbare CRC (Cyclic Redundancy Check) – Prüfung sowie programmierbare Wortlänge.

Hochauflösende Katodenstrahlröhre

Eine neue, einfarbige 38-cm-Katodenstrahlröhre bietet eine Auflösung von annähernd 4 Millionen Bildpunkten auf der Schirmoberfläche bei einer Grösse von etwa einer A4-Seite. Wird die Röhre als Faksimile-Terminal eingesetzt, ist die Modulationstiefe besser als 50% im Fall einer horizontalen schwarzen Linie (mit der Breite eines Bildpunktes) auf weissem Hintergrund-Raster mit einer Leuchtdichte von 143 cd/m². Der Grossflächenkontrast entspricht beim gleichen Leuchtdichtewert einer Modulationstiefe von 85%.

Mikrocomputer-Peripherie-IC zur Abstimmung und Steuerung von Fernsehempfängern

Neueste technologische Fortschritte und umfangreiche Erfahrungen auf dem Gebiet der digitalen Steuerung und Abstimmung von Fernsehempfängern haben uns dazu geführt, eine neue integierte Mikrocomputer–Peripherieschaltung für den Einsatz am I^2C-Bus zu entwickeln. Die Schaltung arbeitet nach dem Prinzip der Frequenzsynthese; Bereichsumschaltung und Abstimmspannungserzeugung befinden sich auf dem Chip. Ausserdem können bis zu acht Analogfunktionen gesteuert werden, z.B. Lautstärke, Klang, Helligkeit, Kontrast, Farbsättigung usw. Für zusätzliche Steuerfunktionen z.B. Stereo/Zweiton, Suchlauf und Schalten externer Videoquellen, sind weitere Ein-/Ausgänge vorhanden.

Ein Mikro-Controller der nächsten Generation

Der neue Mikro-Controller 8X305 in Low-Power Schottky-Technik arbeitet mit einer Zykluszeit von nur 200 ns für die Entgegennahme, Decodierung und Ausführung eines 16 bit-Befehls. Mit seiner hohen Arbeitsgeschwindigkeit kann er unter Programmsteuerung zahlreiche Operationen ausführen, die in herkömmlichen Konzepten zusätzliche Hardware erfordern. Der Mikro-Controller wird durch eine Familie peripherer Schaltungen komplettiert, die – ebenso wie dieser – nur eine Versorgungsspannung von 5 V benötigen.

Un nouveau circuit intégré bipolaire allie la numérotation et la parole en téléphonie

Cet article décrit un nouveau circuit intégré bipolaire incorporant toutes les fonctions propres à la téléphonie électronique, sur une simple puce bipolaire. Ces fonctions sont les suivantes: génération de la tonalité de numérotation, amplification de la parole, conversion 4/2 fils, commutation numérotation/parole et adaptation de ligne. Les entrées de numérotation peuvent être directement commandées par un simple clavier à membrane de caoutchouc ou par l'intermédiaire d'un microprocesseur, si l'utilisation requiert des exigences telles que la répétition d'indicatif automatique ou le rappel du numéro.

Circuits de modularité pour les produits grand public et industriels – le système CLIPS

Récemment présentée, une famille de circuits intégrés à usage général ou spécialisé met à profit les similitudes sous-jacentes des systèmes ayant recours au transfert des données à 8 bits. Conçus pour réaliser l'interface direct avec le bus I²C, ces circuits permettront aux concepteurs de systèmes de passer sans transition du diagramme schématique aux plans de réalisation du circuit, sans être préoccupés par des problèmes d'interface ou de compatibilité. L'identification des fonctions spécifiques du système, avec des circuits intégrés déterminés, va non seulement faciliter l'agencement des cartes de circuit et la détection des défectuosités, mais aussi mener au développement d'un logiciel modulaire aisément adaptable à un éventail d'impératifs inhérents au système.

Conception et réalisation de circuits intégrés pour micro-ondes

Les circuits intégrés pour micro-ondes peuvent contribuer à réduire considérablement les dimensions des systèmes à hyperfréquences et, par surcroît, faciliter la solution des problèmes d'interface entre des systèmes de ligne de transmission et des dispositifs à semi-conducteurs. Toutefois, les dimensions réduites de ces circuits créent des problèmes particuliers. Pour les résoudre, on a recours aux expériences acquises sur les plans électriques et technologiques. Essentiellement, la fiabilité de ces circuits intégrés repose sur des recettes et des procédures éprouvées qu'il s'agit donc de maintenir autant que possible, lorsqu'un nouveau dispositif est envisagé ou developpé.

Transfert rapide et sans distorsion des données en milieu bruyant électriquement

L'article décrit un nouveau circuit destiné à l'emission ou à la réception sérielle sûre et efficace de données parallèles, dans des conditions de bruit électrique considérable. Le transfert des données peut avoir lieu en mode aussi bien synchrone qu'asynchrone, tandis que le circuit englobe des contrôles programmables pour le niveau de bit et la redondance cyclique, ainsi que pour la vérification du format de multiplet.

Tube de monitorage à haut pouvoir séparateur

Ce nouveau tube cathodique monochromatique 38 cm permet de visualiser près de quatre millions d'éléments d'image sur un écran de format proche de celui d'un feuillet A4 courant. Fonctionnant à la manière d'un terminal de facsimilé, il offre une profondeur de modulation supérieure à 50% pour une ligne noire horizontale d'un élément d'image de large, s'inscrivant sur une trame blanche d'une luminance de 143 nit. Le contraste de zone large pour une telle luminance correspond à une profondeur de modulation de 85%.

Réglages et contrôles des téléviseurs par microprocesseur

Les récent progrès de la technologie et l'expérience considérable acquise en matière de commande numérique et d'accord des récepteurs de télévision ont abouti au développement d'un nouveau circuit intégré périphérique microprocesseur à compatibilité de bus I²C. Ce circuit réalise la synthèse de fréquence ainsi que la commutation de bande "sur puce" et la génération de la tension d'accord. Il assure en outre la régulation d'un maximum de huit fonctions analogiques telles que la puissance sonore, la tonalité, la luminosité, le contraste, la saturation des couleurs, etc. De plus, des portes entrée/sortie sont prévues pour des commandes supplémentaires, comme par exemple le son stéréo ou à deux voies, la recherche d'accord et la commutation de sources vidéo extérieures.

Un microcontrôleur de la génération suivante

Ce nouveau microcontrôleur recourant à la technologie Schottky basse tension, est capable de saisir, décoder et exécuter une instruction de 16 bits en 200 ns. Une telle rapidité permet l'utilisation d'un logiciel pour effectuer un grand nombre d'opérations qui, autrement, nécessiteraient un matériel supplémentaire. Le microcontrôleur est complété par une gamme de circuits auxiliaires qui, comme lui, se suffisent d'une simple alimentation de +5 V.

Un CI bipolar combina la conversación y el discado telefónicos

Este artículo describe un nuevo circuito integrado bipolar que incorpora todas las funciones necesarias para un teléfono electrónico. Las funciones que realiza son generación de tono para discado, amplificación de la voz conversión de 4/2 hilos, conmutación discado/ conversación y adaptación de línea. Las entradas de llamada pueden excitarse directamente mediante una sencilla membrana de goma o mediante un microordenador si se requieren más características de discado tales como discado automático o índice de llamadas.

Circuitos para diseño modular de productos industriales y de consumo – el sistema CLIPS

La familia de circuitos integrados específicos y de propósitos generales, recientemente introducida, hace uso de las similaridades comunes a los sistemas que utilizan transferencia de datos de 8 bits. Estando diseñados para acoplarse directamente con el I²C BUS, estos circuitos integrados capacitan al diseñador de sistemas, para ir directamente desde el diagrama de bloques al diseño de circuito, sin tener en cuenta los problemas de compatibilidad o de acoplamiento. La identificación de un sistema de funciones específico con circuitos integrados específicos facilitará la distribución del circuito impreso y la improbabilidad de fallo, y también conducirá al desarrollo de un software modular que puede ser fácilmente adaptado a una variedad de requerimientos del sistema.

Circuitos integrados para microondas – diseño y realización

Los circuitos integrados de microondas (MICs) pueden permitir una considerable reducción de espacio en sistemas de microondas y pueden, a su vez, facilitar problemas de acoplamiento entre sistemas de líneas de transmisión y dispositivos semiconductores. La pequeñez de los (MICs), sin embargo, trae problemas consigo. Para encontrar soluciones a estos problemas se ha empleado mucha pericia tanto en su parte eléctrica como en la tecnológica. Básicamente, acerca de la fiabilidad en centros de MICs, cesando fórmulas establecidas y siguiendo procedimientos establecidos, que deben ser mantenidos siempre que sea posible cualquiera que sea el nuevo dispositivo que se contemple o se desarrolle.

Transferencia de datos rápida y sin distorsión en ambientes ruidosos

Se describe un nuevo circuito para una transmisión serie o recepción de datos en paralelo eficiente y segura en ambientes eléctricamente ruidosos. La transferencia de datos puede ser síncrona o asíncrona y el circuito incluye verificación programable del nivel de bit, verificación programable de redundancia cíclica y verificación de la longitud del octeto.

Tubo monitor de alta resolución

Un nuevo TRC monocromo de 38 cm. capaz de definir cuatro millones de elementos de imagen en una pantalla del tamaño aproximado de una pagina A4. Cuando se opera como un terminal facsimil, suministra una profundidad de modulación mayor del 50% para una línea horizontal negra cuya anchura sea la de un elemento de imagen, visualizada sobre un fondo blanco de luminancia igual a 143 nits. El contraste de grandes áreas a esta luminancia corresponde a una profundidad de modulación del 85%.

Circuito integrado periférico de microordenador para sintonizar y controlar un aparato de TV

Los recientes avances tecnológicos y la amplia experiencia en la sintonía y control digital de los aparatos de TV nos han conducido a desarrollar un nuevo circuito integrado periférico de microordenador compatible con el bus I²C. Este circuito realiza la síntesis de frecuencia junto con la generación de la tensión de sintonía y la conmutación de bandas. También controla hasta ocho funciones analógicas tales como volumen, tono, brillo, contraste, saturación de color, etc. Además, tiene registros E/S para otros controles, como por ejemplo sonido estéreo/dual, sintonía automática y conmutación de las fuentes externas de video.

Nuevo microcontrolador

Un nuevo microcontrolador (8X305) diseñado con tecnología Schottky de baja potencia buscará, decodificará y ejecutará una instrucción de 16 bits en 200 ns. Debido a su alta velocidad, puede utilizar el programa para realizar operaciones que de otro modo requerirían circuitería adicional. El microcontrolador está complementado con una familia de circuitos de soporte que, incluido éste, requieren sólo una alimentación de +5 V.

Authors

Fred van Dongen, born at Schiedam, The Netherlands, in 1939, joined the magnetic components development group of Philips Electronic Components and Materials Division in 1961. In 1970 he moved to the professional subassemblies development department to work on both magnetic core and semiconductor memories. Since 1977 he has been a member of the central application laboratory, where he is now concerned with the design of ICs for electronic telephone equipment.

Tom Keve is the Strategic Product Marketing Manager for Philips CMOS ICs. He earned a first-class honours degree in physics at UMIST, Manchester, and a Ph.D. in crystallography at Imperial College, London. He spent two years at Bell Laboratories in New Jersey, followed by seven years at the Philips Research Laboratories in Redhill, England. Five years ago he moved to Eindhoven; since then he has marketed both Philips and Signetics ICs.

Jos Geboers was born in Valkenswaard, The Netherlands. After graduating in electronic engineering at Philips Technical School and Eindhoven Polytechnic, he joined the Central Application Laboratory of Philips Electronic Components and Materials Division where he worked on instrumentation and television; he now specialises in the development and application of telephony circuits.

Willem Goedbloed studied nuclear physics at the Free University of Amsterdam, receiving his PhD in 1970. He joined Philips the same year and was initially engaged, with S. D. Vriesendorp, in setting up the microwave-integrated circuits facility, Eindhoven. His main activities since then have been in development of microwave IC technology and of microwave solid-state devices and subassemblies, with particular interest in reliability and degradation phenomena. He is currently responsible for development and production of microwave solid-state subassemblies.

Jan Exalto, born at Eindhoven in 1942, graduated from Eindhoven Polytechnic in 1964. After military service he joined the Central Application Laboratory of Philips Electronic Components and Materials Division in 1966. Since then he has been principally engaged in digital electronics and is at present responsible for applications of standard and customised CMOS.

Kurt Noach was born in Amsterdam in 1939 and earned a degree in electrical engineering at Amsterdam Polytechnic. After four years' service in the Royal Dutch Air Force he joined Philips Electronic Components and Materials Division as a designer of manufacturing machinery. Six years later he moved to the commercial department where, since 1975, he has been concerned with European marketing of Signetics integrated circuits.

Henk J. M. Otten was born in Sint Oedenrode, The Netherlands, in 1946 and graduated in electronic engineering at the University of Technology, Eindhoven, in 1973. After joining the Central Application Laboratory of Philips Elcoma Division he became engaged in the development of integrated circuits for tv receivers and, later, of components for fibre-optic communication. At present he is concerned with the design and application of components for electronic telephone equipment.

L. J. W. Versnel, born at Hilversum, The Netherlands, in 1933, studied applied physics and mathematics at the University of Groningen, with special emphasis on electron optics. Upon graduation in 1967 he joined Philips' Electron Tube Division where he became engaged in the development of industrial and transmitting tubes and cathode-ray tubes. Transferred to Hyperelec at Brive, France, in 1973, he worked for some time on photomultiplier development but is now once again concentrating on cathode-ray tubes.

S. D. Vriesendorp was born in Haarlem, The Netherlands, in 1939, and graduated in electromagnetic theory from the University of Technology, Delft, in 1966. He joined Philips in 1968 and was initially involved in the setting up of the microwave-integrated-circuits facility, Eindhoven. From 1980 he worked on downconverter predevelopment for international satellite communications, and since 1982 he has been a member of the strategic marketing group for bipolar integrated circuits.

Karl-Heinz Seidler was born in 1937 at Hamburg. After graduating in engineering in 1962 he joined the application laboratory of Valvo where, as a member of the data processing group, he worked on logic modules and magnetic core memories. For the past several years he has been engaged in digital and remote-control system design, particularly for radio and television.

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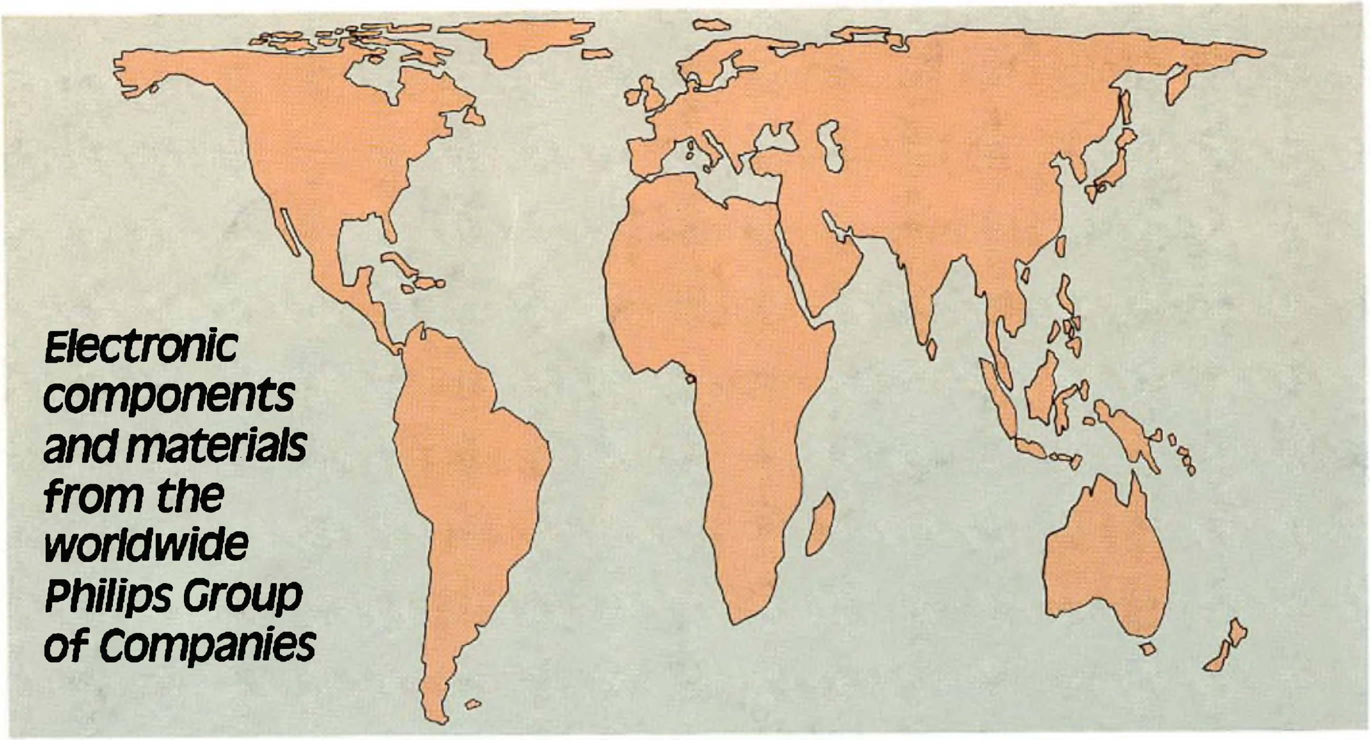

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