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The window on the cover is a window on tomorrow, a window on a new world of high-fidelity sound. The shimmering disc in the woman's hand contains an hour of digitally encoded music — music with a wider dynamic range, better stereo separation, less noise, and wider frequency range than has been possible until now. Repeated playing will not wear it out, dirt or fingerprints on its surface will not mar the purity of the sound. The first article in this issue opens a window on the technology behind tomorrow's digital high fidelity.

The general public had their first chance to see the Compact Disc digital audio system at the Festival du Son et de l'Image, Paris, in March. The system will be on the market later this year, backed by an extensive catalogue of record titles on Compact Disc. The principles and benefits of Compact Disc Digital Audio are described on this page. The article commencing opposite deals with a decoder for a Compact Disc player for which a set of dedicated LSI circuits has been developed to perform the complex signal processing and control tasks.

Compact Disc digital audio

Compact Disc digital audio represents the biggest step forward ever in sound reproduction.

The single-sided 12 cm dia. disc used in the system stores up to one hour of stereo sound. It has no grooves, the digitally coded recording lies beneath the disc surface, invulnerable to dirt and damage. The recording is read using a non-contacting laser pick-up which imposes no wear on the disc and suffers none itself. The pick-up is carried on a servo-controlled arm which tracks radially from the inside to the outside of the disc. The digital signals recovered from the disc are then used to accurately reconstruct the original sound recorded. Tracking, decoding and disc drive are all controlled by a central timing generator in the player.

The frequency response of the system is absolutely flat from 20 Hz to 20 kHz. Signal-to-noise ratio, channel separation and dynamic range are all more than 90 dB. Any wow or flutter generated by the mechanics of the system is removed by the decoder, which resynchronises the data to a crystal clock. Rumble is eliminated from the input data by the demodulator stage of the decoder. There is also no audible intermodulation distortion of the type that plagues analogue recordings. To fully appreciate these performance levels, compare the entries in the Table for Compact Disc and a 12-inch stereo long-playing record.

The Compact Disc player connects directly to existing hi-fi systems, just like another high-fidelity sound source. It is easy to operate and withstands vibrations and bumps, allowing its use in cars and other vehicles.

In short, Compact Disc digital audio offers unmistakably superior standards of sound reproduction.

System performance of Compact Disc and a 12-inch LP

	Compact Disc	12-inch LP
frequency range	20 Hz – 20 kHz	30 Hz – 20 kHz
dynamic range	>90 dB	<55 dB (1 kHz)
S/N ratio	>90 dB	≈60 dB
channel separation	>90 dB	25 – 35 dB
harmonic distortion	<0.01%	0.2%
wow and flutter	nil	0.03%
playing time	60 minutes	20 minutes/side

BENEFITS OF COMPACT DISC DIGITAL AUDIO

- The ultimate in sound reproduction
- Disc immune to dirt and scratches
- No wear on disc or pick-up
- Easy to use
- Compatible with existing hi-fi equipment
- Pocket-sized disc that can store 1 hour of stereo sound
- Insensitive to microphony, shock and vibration.

ICs for Compact Disc decoders

J. MATULL

One side of a 12 cm Compact Disc stores an hour of stereo music in the form of a helical track of tiny pits and flats representing more than five billion data bits. During playback, the pits scatter the light from a small semiconductor laser, the flats reflect it onto a photodetector. The data bits are recovered at a rate of 4.3 million bits/s and converted into a PCM signal from which the original audio waveform is accurately reconstructed without the degradation associated with analogue reproduction systems.

The change to digital signal processing leaves all analogue problems such as wow and flutter and crosstalk behind. Owners of a Compact Disc player will therefore be able to enjoy music of a quality unequalled by that available from disc or tape. Compact Disc players can be programmed to access individual pieces of music and reproduce them in any order desired by the listener. Some other possible features, not easily implemented in a conventional LP disc system, are the display of elapsed/remaining playing time, automatic switching of de-emphasis circuits and search and repeat functions.

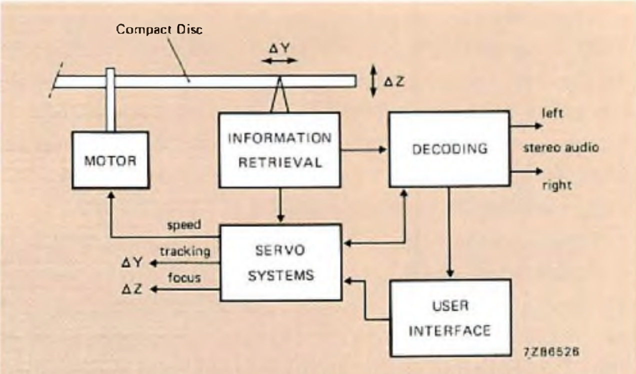
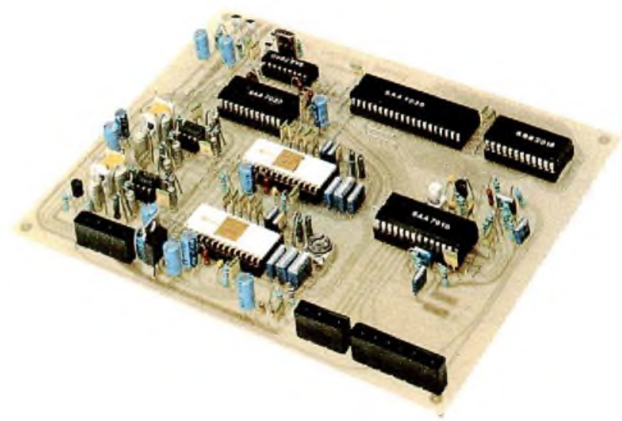


Fig.1 Functional schematic of a Compact Disc player.



Development decoder board which performs more than 4 million decoding, control and error-correction operations every second of playback of a Compact Disc. The four new NMOS LSI circuits at the heart of this mammoth processing task enable a complete decoder controlled by one master clock to be assembled on one small single-sided board ($\approx 220\ cm^2$)

Figure 1 shows the main components of a Compact Disc player. This article describes the decoding section for which four new NMOS LSI circuits have been developed:

- SAA7010, demodulator IC (28-pin DIL).
- SAA7020, error correction IC (40-pin DIL).
- SAA7000, interpolation and muting IC (18-pin DIL).
- SAA7030, digital oversampling filter IC (24-pin DIL) used with the TDA1540 DAC (28-pin DIL) in a unique 16-bit digital-to-analogue conversion system.

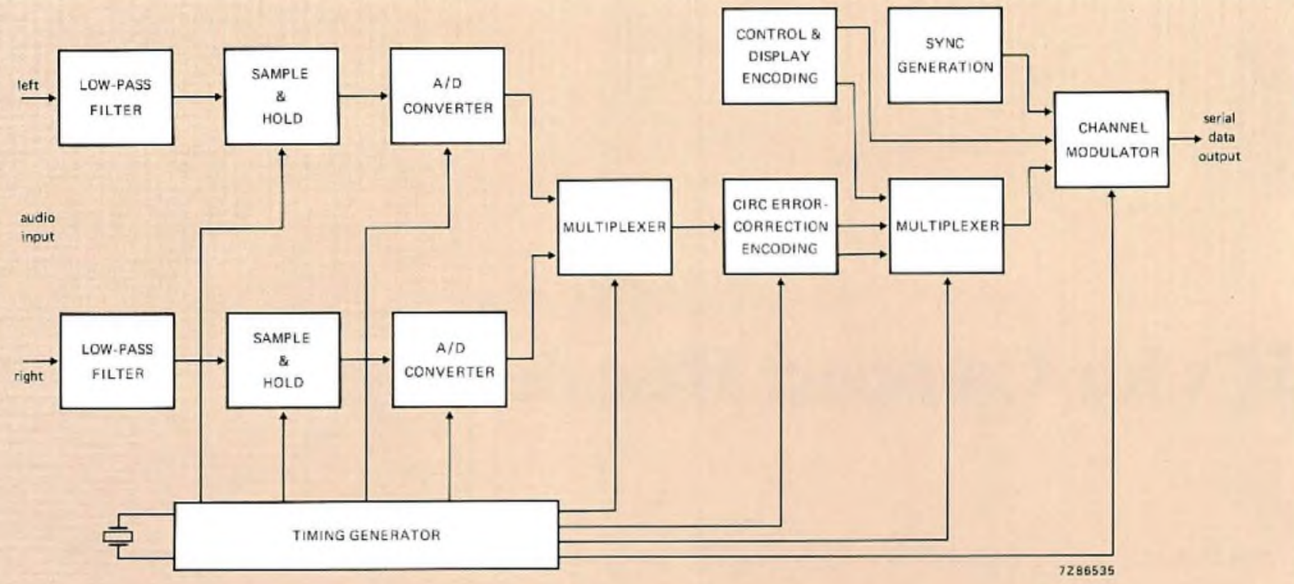


Fig.2 Compact Disc encoding system

These circuits are in standard DIL packages, require only a few peripheral components and are controlled by one master clock. Because of a high degree of integration, a complete decoder from h.f. input to audio output can be built on one small ($\approx 220\,cm^2$) single-sided printed circuit board.

The ICs are key components determining the performance of a Compact Disc player. Before describing them and the decoding process, it will be helpful to describe briefly the digitising and encoding of audio signals onto a Compact Disc. Figure 2 shows the encoding process.

ENCODING THE DIGITAL AUDIO

The first step is to sample the analogue audio and convert the samples to 16-bit PCM signals. Since the frequency spectrum of impulse-sampled audio signals is simply double sidebands of the audio signal spectrum repeated at multiples of the sampling frequency, the analogue input signals are first passed through a low-pass filter with sharp cut-off which limits their bandwidth to less than half the sampling frequency. This prevents intermodulation distortion due to aliasing. If required, the filtered analogue signals can be passed through a pre-emphasis network to improve signal-to-noise ratio (SNR).

The analogue signals are sampled at 44.1 kHz. The output of the sample and hold circuit is a flat-top PAM signal, whose pulse heights correspond to the audio signal amplitudes at the sampling instants.

Each sample is held long enough for the previous one to be converted to a 16-bit 2's complement number by an analogue-to-digital converter (ADC). Sixteen-bit quantisation gives extremely high resolution, the maximum SNR of a 16-bit PCM signal being ≈ 98 dB ($6n + 1.8$ dB, where n is the number of bits used to express each quantisation level).

Channel separation in Compact Disc digital audio is almost total, because the data of the left channel are completely independent of those of the right.

The 16-bit binary words from the ADC go to a multiplexer which sends a serial data stream of alternately 16 left channel bits and 16 right channel bits to an error-correction encoder.

Sixteen-bit quantisation solves one problem associated with any digital audio system – quantisation noise. Another – drop-out (the loss of single or several data bits in the recovered data) – is overcome in the Compact Disc system by the method used for encoding the digital data prior to recording the Disc. This enables any errors in the signal read from a Compact Disc, which could cause distortion, to be detected and corrected during playback. A powerful multiple-error-correcting code called the Cross Interleave Reed-Solomon Code (CIRC) is used. This code is based on parity bits and an interleaving (rearranging) of the digital audio samples in time.

This code can correct a drop-out in which up to about 3500 successive bits are lost (equivalent to a track length on the disc of about 2.4 mm). Beyond this, it enables the loss of up to about 12 000 bits, occupying a track length of 8.5 mm, to be compensated by interpolation. This, plus the track being beneath a protective layer, is why a Compact Disc is insensitive to scratches and dirt on its surface.

Figure 3 shows the principle of interleaving. In Fig.3(a), a sequence of signal processing events is shown without interleaving. An audio signal is sampled at instants 1, 2, 3 etc., digitised and the data representing the samples recorded onto disc. If there is a drop-out while reading the disc, some words will be missing in the received data. In Fig.3(a), three words are missing. If the length of the drop-out region is greater than the error correction ability of the decoder, the missing values cannot be reconstructed and the audio output has to be muted in order to avoid audible clicks in the reproduced sound.

In Fig.3(b), the same sequence of events is shown, this time with interleaving of data. The original audio signal is sampled, but the sequence of the samples is then rearranged prior to recording on disc. For the same drop-out as before, three words are again missing. After the original sequence of the data words is restored, the drop-out region is spread out in time, but now there are only single errors (dashed lines) which can be easily corrected.

The Compact Disc system enables many features, not possible with the microgroove record system, to be incorporated in the new players. For example, signals can be added to mark the pause between two successive pieces of music and used to implement search and repeat functions, or to indicate remaining/elapsed playing time, titles and composers. This control and display information (known as subcoding information) must be inaudible and so is encoded separately by the control and display encoder. The 8-bit output of this encoder means eight information channels can be recorded. Since the subcoding information is generated and recorded bit-serially, there must be provision for recognising the beginning of a block of subcoding data. There are two outputs from the encoder, one carrying the data, the other synchronising the data blocks.

The subcoding data and the 8-bit serial words from the error correction and encoding unit, which are either data or parity words, pass through a multiplexer to the channel modulator. The sync generator creates a unique pattern, not contained in the normal data, which identifies the beginning of each frame of data.

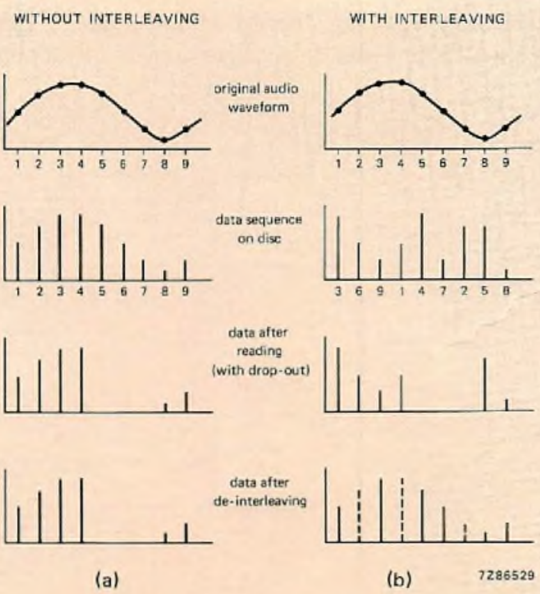


Fig.3 Principle of interleaving

The output data of both the error correction encoding unit and the control and display encoding unit are in non-return-to-zero (NRZ) format. Data in NRZ format are not suitable for recording onto disc, because the bit clock cannot be recovered from the data stream during playback. In addition, the data stream may contain low-frequency components which could interfere with the servo systems in the player, controlling the disc rotation and the tracking/focusing of the laser pick-up. The NRZ data is converted to another code in the channel modulator, whose output is then recorded on disc.

The requirements of a code for an optical audio disc system are:

- bit clock can be regenerated from the data
- low spectral power at low frequencies
- permits read-out at high information density
- small error propagation.

The encoding procedure used is called Eight-to-Fourteen Modulation (EFM). As the name implies, each group of 8 data bits (called a symbol) from the error correction encoder, or from the control and display encoding unit, is encoded into a group of 14 bits. To ensure that the bit clock can be regenerated from the data, and to permit the reading of high-density information, there are always at least two 0s between successive 1s and no more than ten consecutive 0s in a 14-bit EFM word. The transmitted information is contained in the transitions which are indicated by the 1s. There are 277 combinations of 14 bits which satisfy the constraint of at least two 0s. Deleting the 21 patterns with ten or more 0s in a row leaves 256 combinations, which gives a one-to-one correspondence between 8-bit NRZ data and 14-bit EFM data. Code conversion can therefore be done easily, via a T flip-flop, using a look-up table stored in a ROM.

The 14-bit blocks generated cannot, however, be concatenated without violating the constraint of the two to ten 0s at the block boundaries. Therefore, three merging bits are inserted between successive 14-bit blocks. These bits do not contain audio or control/display information and are skipped by the decoder. The d.c. content of the bit stream can be controlled by inserting a transition in the merging bits, if the constraints are not violated. The decision to insert is made using advance knowledge of symbols.

Because of the requirement for regenerating the bit clock, synchronisation is necessary. This is achieved by dividing the bit stream into frames and adding a unique pattern to each. Each frame contains:

- a synchronisation pattern
- twelve 16-bit data words representing 6 stereo samples
- four 16-bit error correction parity words
- one 8-bit control and display word.

merging bits →	3	3	12 x 3	4 x 3	12 x 3	4 x 3
	24 sync pattern bits	14 subcoding bits	12 x 14 audio data bits	4 x 14 error correction parity bits	12 x 14 audio data bits	4 x 14 error correction parity bits

← 1 frame (588 channel bits) →

7Z86530

Fig.4 Data stream after Eight-to-Fourteen Modulation (EFM)

The 16-bit data and parity words are split into 8-bit words before EFM encoding.

The total number of channel bits per frame after encoding is 588, comprising:

- 24 sync pattern bits
- 336 (12 x 2 x 14) data bits
- 112 (4 x 2 x 14) error correction bits
- 14 control and display bits
- 102 (34 x 3) merging and d.c. control bits.

Figure 4 shows the arrangement of the bit stream after EFM.

Since EFM is based upon a block structure of 8-bit input data, it is well suited for the CIRC error correction system, also based on blocks of eight consecutive data bits. Propagation of errors is limited to the eight bits in a symbol.

DECODER

The function of the decoding system of a Compact Disc player is to regenerate both analogue audio channels from the h.f. data stream retrieved from the disc by the laser pick-up. Figure 5 shows the block diagram of a decoder and the ICs used:

- SAA7010, for the signal processing tasks of demodulation. Converts the data stream from the laser pick-up back to its original format, regenerates the bit clock from the data stream, and separates the data representing the recorded sound from control and display data.
- SAA7020 and an SBB2016 RAM. De-interleave the demodulated data from the SAA7010, and detect and correct errors in the data stream. The SAA7020 generates an unreliable data signal if it is unable to correct an error. The SAA7020 also removes jitter in the data by resynchronising the data to a clock derived from a crystal oscillator.
- SAA7000, for interpolation and muting. Reconstructs audio data by interpolation should the SAA7020 be unable to correct any single samples in error; mutes consecutive erroneous samples.
- SAA7030 and two TDA1540s for digital-to-analogue conversion. SAA7030 increases the SNR of the 16-bit data stream containing the audio information by 13 dB so that 14-bit DACs can be used with no loss of quality in the reproduced audio.

Figure 6 shows the demodulation, error correction and muting section of the decoder in more detail. Figure 8 shows the 16-bit digital-to-analogue conversion section.

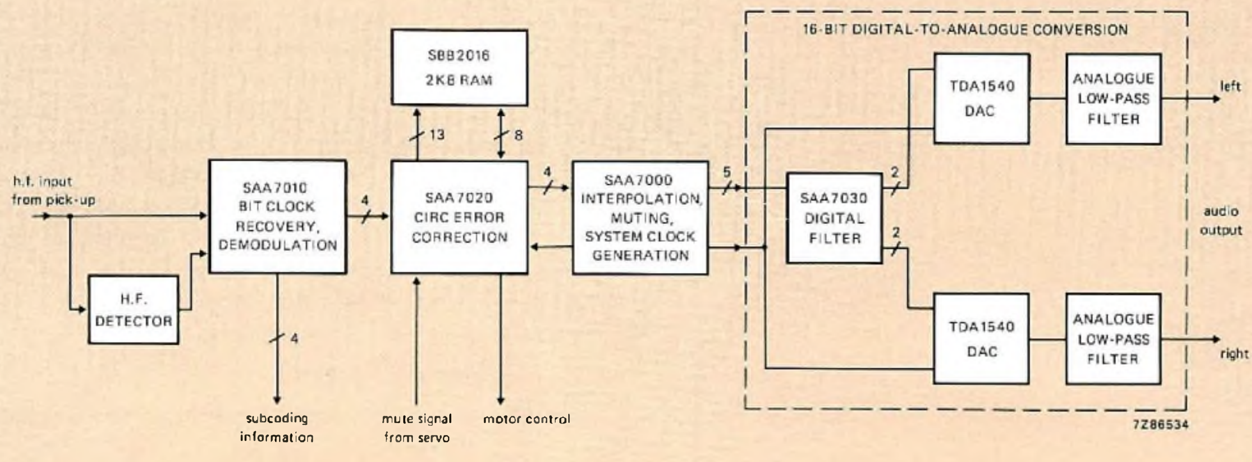


Fig.5 Block diagram of a Compact Disc decoder

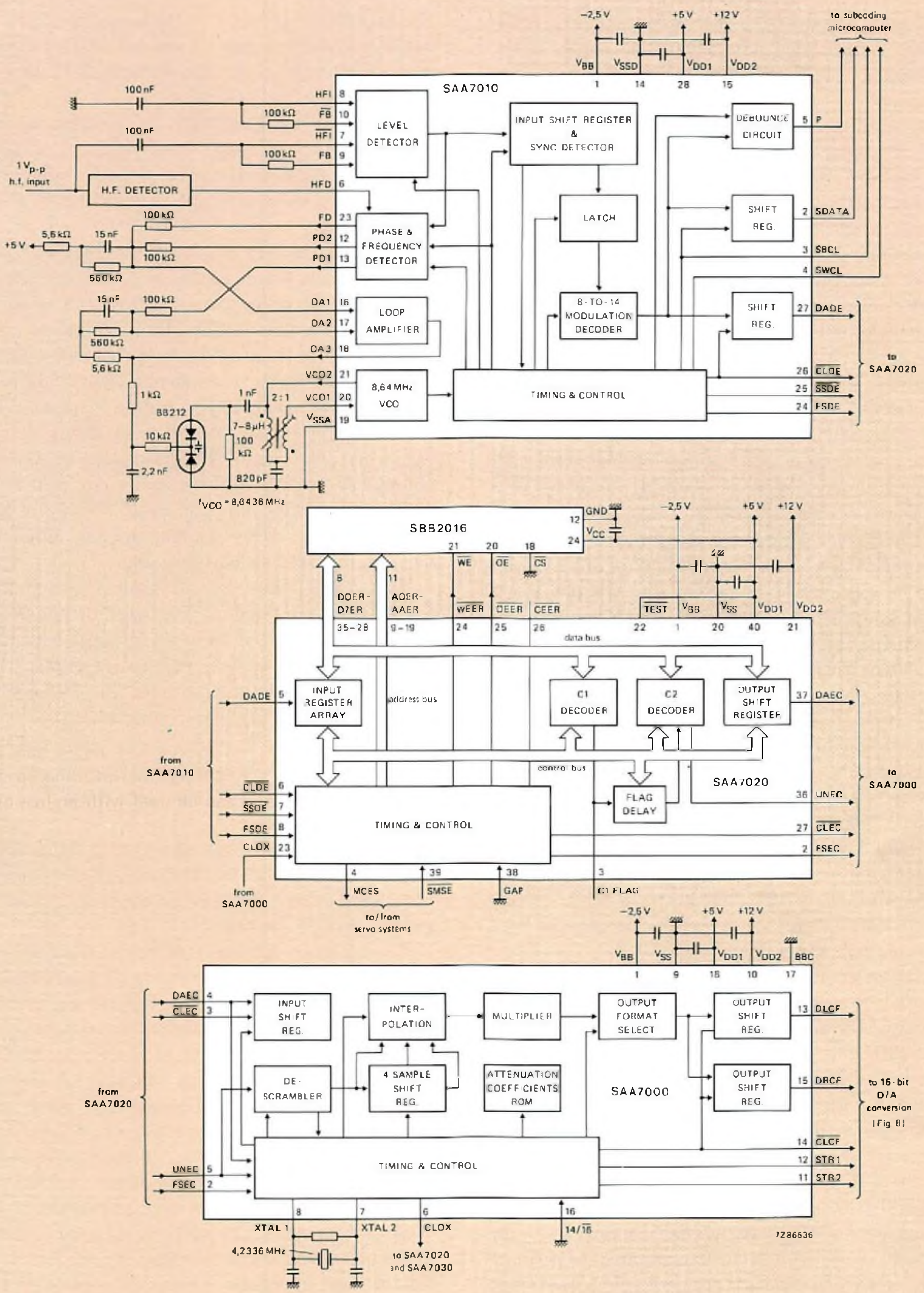


Fig.6 Demodulation, error correction and muting section of a Compact Disc decoder

SAA7010: demodulator IC

The SAA7010 forms the front-end of the decoding system and supplies demodulated data and timing signals to the error correction IC (SAA7020) and to the subcoding microcomputer.

The h.f. signal read from the Compact Disc by the laser pick-up is first amplified and filtered, then supplied to the input of the SAA7010 and to an optional h.f. threshold detector. The analogue h.f. signal is first digitised by a level detector and then passed to a phase-locked loop (PLL) which regenerates the bit clock from the incoming data.

There are six complete stereo audio samples in one frame (588 EFM channel bits) of the disc data. Therefore, one stereo sample comprises 98 channel bits. For a sampling frequency of 44.1 kHz, the data rate of the h.f. signal is 4.3218 Mbits/s. The voltage-controlled oscillator (VCO) of the PLL operates at twice the incoming data rate from which a 4.3218 MHz master clock signal used in all internal timing is derived. Frequency-sensitive and phase-sensitive detectors provide the coarse and fine tuning signals for the VCO. Both detectors are enabled/disabled using the HFD input.

The incoming data is clocked into a shift register and the sync pattern detected, which enables the start of each frame of data to be identified. Sync information is passed to the timing and control logic in order to synchronise the demodulation with the incoming data. The level detector has feedback which automatically adjusts to the optimum switching level. However, if the frequency of the incoming data falls below a set level, e.g. owing to a drop-out, the level detector switches to a nominal feedback level of half the supply voltage V_{DD1}. This prevents the switching level drifting away from its optimum.

Provided the timing generator is locked to the h.f. signal, each 14-bit word received is stored in a latch, then converted into an 8-bit word by the EFM decoder.

In the SAA7010, a logic array is used for code conversion, not a ROM look-up table as mentioned earlier. This saves chip area and reduces power dissipation.

Demodulated audio data are shifted out of the SAA7010 to the error correction IC SAA7020 on DADE, with the clock signal $\overline{\text{CLDE}}$ and the symbol and frame sync signals $\overline{\text{SSDE}}$ and FSDE respectively.

The seven bits Q to W of the subcoding data plus a sync bit are shifted out serially at the SDATA output with clock SBCL and sync signal SWCL, while channel P (pause bit) is output at P after passing through a debounce circuit.

Note, use of an h.f. threshold detector at the input HFD of the SAA7010 is optional. If one is not used, the HFD input is connected to V_{DD1} for normal operation of the demodulator. Using an h.f. detector can improve the performance of a Compact Disc player because it disables the phase and frequency detectors in the demodulator when the amplitude of the h.f. input is small. Thus, the PLL cannot lock onto noise in the absence of an h.f. signal and so clock jitter is prevented.

SAA7020: error correction IC

The SAA7020 detects, and where possible corrects, errors in the demodulated data stream and supplies these data together with a flag, indicating whether the data are correct or unreliable, to the SAA7000.

Data from the SAA7010 are arranged in frames of thirty-two 8-bit symbols. Twenty-four of these symbols contain 12 audio samples (i.e. 6 stereo samples), the remaining eight are parity symbols added for error detection and correction.

Data enter serially into a register array at DADE. This array comprises a shift register which accumulates symbols for parallel processing and a FIFO register which acts as a jitter reduction circuit. The FIFO register can compensate for deviation of up to ±2.25 frames from the nominal data rate. It is this register that eliminates wow and flutter in the Compact Disc system. The output data rate from the SAA7020 depends only on the clock signal CLOX derived from a crystal oscillator. Any discrepancy between the clock derived in the demodulator and that from the crystal oscillator generates an error signal MCES which controls the speed of the motor spinning the disc. MCES is a PWM signal with a range of 142 linear steps.

The CIRC makes use of interleaving and two Reed-Solomon codes C1 and C2. The data from the demodulator are de-interleaved by temporarily storing the thirty-two 8-bit symbols forming the input word of the C1 decoder and the output word of this decoder in a 2K8 RAM.

An 8-bit bidirectional bus is used for transferring data to and from the RAM, an 11-bit bus for addressing. Three bits control the RAM: write enable $\overline{\text{WEER}}$, output enable $\overline{\text{OEER}}$, and chip enable $\overline{\text{CEER}}$. The last is for operation with pseudo-static RAMs.

The C1 decoder of the SAA7020 is designed to correct one erroneous symbol in a 32-symbol frame. The C2 decoder is designed to correct one erroneous symbol, or two erasures in a group of 28 symbols.

The input word to the C1 decoder is checked for errors by multiplying this word with the C1 parity check matrix. This produces four syndromes. If there are no errors in the input word, all four syndromes are equal to zero and the 28 data symbols at the output of the C1 decoder (4 parity symbols being discarded) are written back into the RAM unchanged. In case of one erroneous symbol, this is corrected and the 28 corrected output symbols are written into the RAM. In case of two or more erroneous symbols, the 28 output symbols are written into the RAM unchanged and a flag is set which marks these 28 symbols as unreliable.

Since each symbol in a C1 output word is delayed by a different amount before reaching the C2 decoder, the C1 flag of each unreliable symbol is delayed by the same amount as the symbol to ensure that symbol and flag arrive at the input of the C2 decoder together.

The output symbols of the C1 decoder are de-interleaved further by means of the RAM. Then, the 28 symbols forming the input word of the C2 decoder are also checked for errors by examining the four syndromes resulting from multiplication of this word with the C2 decoder parity check matrix.

If there are no errors, all four syndromes are equal to zero, and the 24 data symbols at the output of the C2 decoder are written back into the RAM unchanged and the remaining four parity symbols discarded.

In case of one erroneous symbol, this symbol is corrected in the same way as for the C1 decoder. Then, the corrected 24 output symbols are written into the RAM. In case of two erroneous symbols, the sum of the error values is given by one of the syndromes. Information about the position of erasure symbols is given by the flags. With this additional information, both symbols can be corrected and the 24 corrected output symbols are written back into the RAM. In case of more than two erroneous symbols in a C2 word, all 24 symbols are rewritten into the RAM unchanged and a C2 flag is set which marks these 24 symbols as unreliable.

Data from the SAA7020 are output on DAEC in 16-bit bursts separated by 8-bit intervals. If the unreliable data signal UNEC is output simultaneously with a data symbol, it marks that symbol as unreliable. If UNEC is output during the interval between data bursts, it indicates that the sample arriving 30 sampling periods later is unreliable. This advance information of the reliability of symbols is used in the SAA7000 in a muting procedure.

The unreliable data signal is generated when:

- both C1 and C2 decoders detect an uncorrectable error pattern. In this case, UNEC marks one or more symbols of the C2 output data as unreliable.
- C1 fails to detect an error, but C2 detects it without being able to correct all erroneous symbols. In this case, the whole C2 output frame is marked unreliable.
- a mute signal $\overline{\text{SMSE}}$ is received from the servo system. UNEC is immediately activated and marks symbols as unreliable as long as $\overline{\text{SMSE}}$ is present.

The output data are shifted out serially with the data clock $\overline{\text{CLEC}}$ and the frame sync signal FSEC. The logic state of DAEC in the gaps between successive clock bursts, which is determined by the input GAP, is used to select 2's complement or offset binary format for the SAA7000.

The 8-bit input and output data format of the SAA7020 makes it also suitable for error correction in other applications.

CIRC decoding

Use of error correction is one of the fundamental reasons why digital audio is vastly superior to analogue audio. And from a practical point of view, the ability to correct errors means that the requirements for disc manufacture are not pushed to impossible limits.

A very powerful error-correcting code called the Cross Interleave Reed-Solomon Code is used in the Compact Disc system. This code, which is based on the use of parity bits and an interleaving of the digital audio samples, enables an error burst of up to about 3500 bits to be corrected. Beyond this, it enables a loss of up to about 12 000 bits to be compensated by interpolation. In systems not using interleaving, a drop-out means that several adjacent samples are lost, leaving no chance of reconstructing them.

Two Reed-Solomon Codes termed C1 and C2 are used for correcting erroneous audio samples. C1 is characterised by thirty-two 8-bit symbols, comprising 28 data symbols and 4 parity symbols. C2 is characterised by twenty-eight 8-bit symbols, comprising 24 data symbols and 4 parity symbols. The minimum distance of each code, i.e. the Hamming distance at symbol level, is five. Thus, in theory, 2 symbols in error, or four erasures (symbols whose position is known but whose value is not) can be corrected.

Figure 7 shows the principle of a CIRC decoder. Delays are used to rearrange (de-interleave) the digital samples back to their original sequence and are organised for efficient error correction.

Delays at the input of the C1 decoder are all of equal length and separate the even and odd numbered symbols. This enables the C1 decoder to correct small errors in adjoining symbols. Delays between the two decoders are of unequal length and longer than those at the input of the C1 decoder which enables the C2 decoder to correct burst errors.

The input to the C1 decoder is a frame of 32 symbols which is multiplied by the C1 parity check matrix to produce four syndromes. From the syndromes, errors can be detected and corrected. The C1 decoder of the SAA7020 is designed to fully correct one erroneous symbol out of a 32-symbol frame. If there is more than one error, each of the 28 output symbols (the four C1 parity symbols having been dropped) is marked with a flag, indicating that each may contain an error. If there is no flag associated with a symbol arriving at the C2 decoder, that symbol is correct.

The C2 decoder is designed to correct one erroneous symbol, or two erasures. When even the C2 decoder cannot correct all errors, it outputs the 24 data symbols (the four C2 parity symbols having been dropped) uncorrected, but marked with flags. Most of these output symbols are usually error-free and can be reconstructed by linear interpolation, because the combination of the C1 and C2 flags is used as an unreliable data signal for interpolation, or if necessary, muting of symbols in the SAA7000.

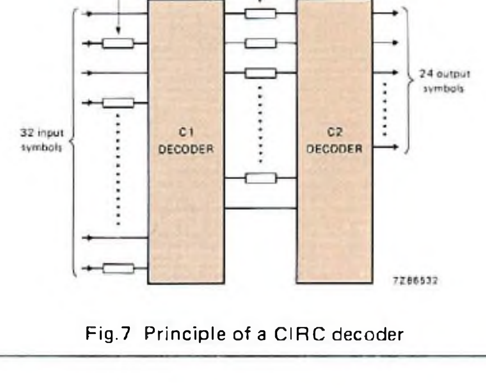


Fig.7 Principle of a CIRC decoder

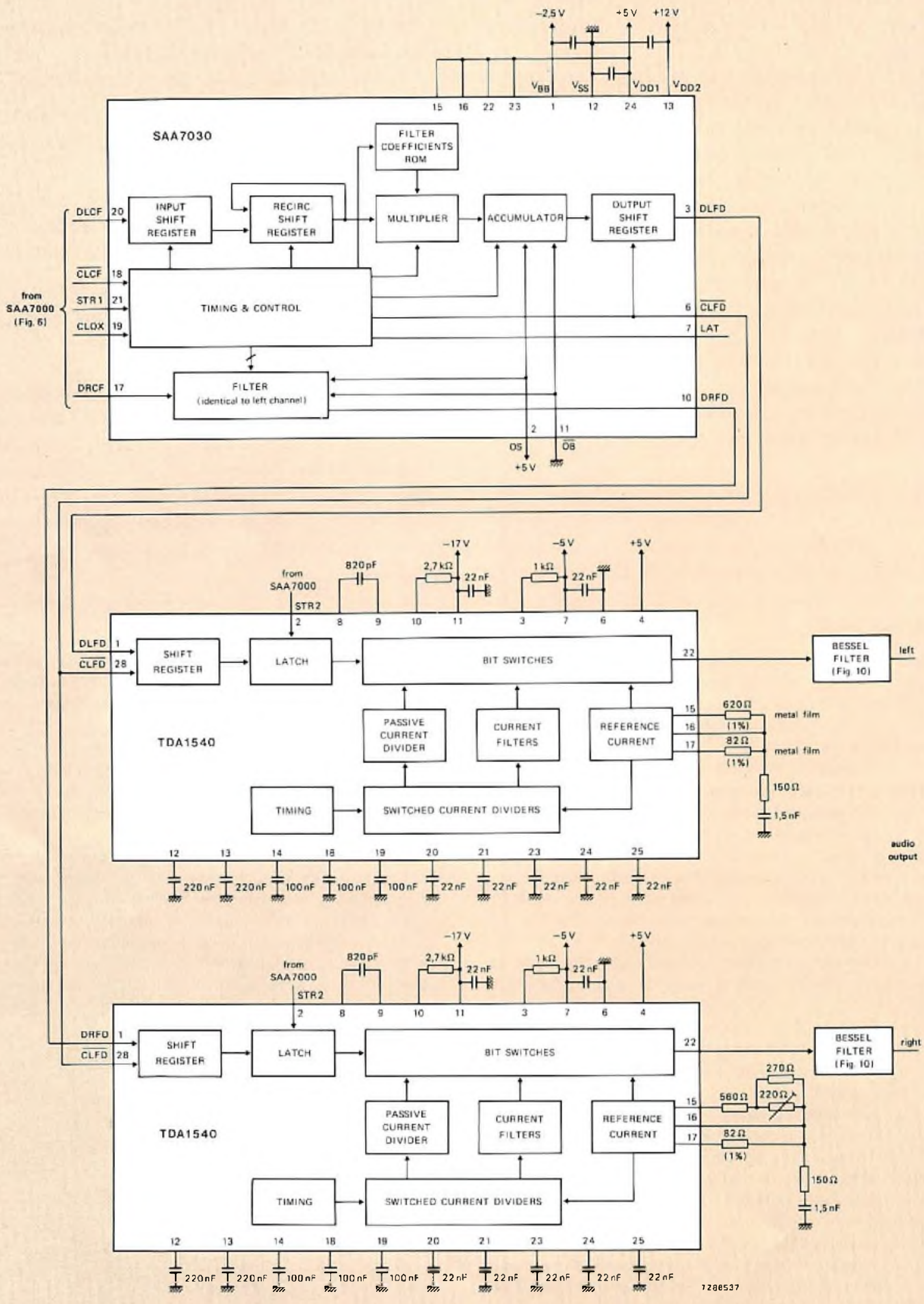


Fig.8 16-bit digital-to-analogue conversion section of a Compact Disc decoder

SAA7000: interpolation and muting IC

The SAA7000 eliminates the audible annoyance that could result if an erroneous symbol which managed to get through the error correction circuitry was processed further. The SAA7000 also generates the clock used for internal timing of the SAA7020 and SAA7030.

Serial data from the SAA7020 on DAEC are entered into a shift register using the clock $\overline{\text{CLEC}}$. Data are then descrambled and separated into left and right channel samples. A similar descrambling and separation is performed on the unreliable data flag UNEC. When there are no unreliable data flags, the data values of the audio samples are unaffected by the SAA7000.

When, for either left or right channel, a single unreliable sample is flagged between two correct samples, linear interpolation is used to replace the erroneous sample. If two or more adjacent samples are flagged as unreliable, they are muted. Starting thirty samples before muting (by making use of the advance information of UNEC), the correct samples are attenuated smoothly to zero according to a cosine curve. The levels of the first thirty samples following muting are then increased back to the normal level, again according to a cosine curve. The thirty-sample (5-frame) delay necessary in muting is obtained using the 2K8 RAM.

The data which are now either correct or processed are converted into 2's complement or offset binary depending on the digital-to-analogue converters used. Selection is made by the status of DAEC during the intervals between input samples which is controlled by the GAP input of the SAA7020. Left and right channel output data on DLCF and DRCF are clocked out by the shift clock $\overline{\text{CLCF}}$. Strobe signals STR1 and STR2 are generated for the digital-to-analogue conversion unit. A 14-bit or 16-bit output data format can be selected using the input $14/\overline{16}$. A full-performance 16-bit conversion system is described in the next section.

A crystal oscillator is used to generate internal timing signals and the clock signal CLOX for the SAA7020 and the SAA7030. By using the frame sync signal FSEC to reset the internal timing, the SAA7000 is automatically synchronised with the output of the SAA7020.

16-bit digital-to-analogue conversion

The digital-to-analogue conversion section (Fig.8) converts the digital output data from the interpolation and muting IC to analogue voltages and removes all unwanted frequency components above the audio band.

This conversion system is an outstanding feature of the Compact Disc decoder. It is based on a digital oversampling filter (SAA7030) followed by two (one for each channel) 14-bit DACs (TDA1540) and low-order analogue filters. This arrangement has the SNR of a system using 16-bit DACs and high-order filters, plus the following advantages over that approach:

- linear phase response in the audio band (0 – 20 kHz)
- reduced slew rate distortion
- requires only a 14-bit DAC
- less intermodulation distortion, because the oversampling frequency moves any intermodulation products well outside the audio band
- only a simple low-cost analogue filter is needed after the DAC to suppress any residual frequency components not removed by the digital filter.

SAA7030: digital oversampling filter IC

The SAA7030 consists of three main parts: oversampling section, transversal filter, and noise shaper. Circuitry in the SAA7030 is duplicated, one for each channel.

The two 16-bit data streams, at DLCF and DRCF are fed into shift registers which quadruple the sampling frequency from 44.1 kHz up to 176.4 kHz. Quadrupling the sampling frequency also quadruples the effective audio bandwidth, which is thus increased from 22 kHz to 88 kHz. The quantisation noise power, previously distributed uniformly across a 22 kHz bandwidth, is now distributed uniformly up to 88 kHz. Since 75% of the noise is now above the audio band, it can be suppressed by filtering.

The SAA7030 incorporates two identical filters, one for each stereo channel. Each filter is a finite impulse response transversal filter of length 24 with ninety-six 12-bit coefficients and using 16-bit input data words entering at 44.1 kHz. A recirculating shift register is used to store data required during the multiplication of input data with the filter coefficients which are stored in ROM. Each multiplication produces a 28-bit word which is stored in an accumulator, the 14 most significant bits of the words are then shifted out at DLFD for the left channel, DRFD for the right channel, with the clock $\overline{\text{CLFD}}$. Overflow protection is incorporated so that the output always limits cleanly in the unlikely event of accumulator overflow.

Oversampling and filtering add 6 dB to the SNR of the digital audio signal, bringing it to 90 dB for the 14-bit output words. A further improvement of 7 dB is obtained in the SAA7030 by means of a noise shaper. This circuit redistributes the quantisation noise, now uniformly spread across 0-88 kHz because of oversampling, reducing the noise in the audible region still further and increasing it above 22 kHz, see Fig.9.

A choice of offset binary or 2's complement output data can be made using $\overline{\text{OB}}$. The OS input enables a 3% d.c. offset to be added. This can be useful to reduce the effects of glitches at low output voltages with certain DACs. A 176.4 kHz strobe (LAT) is provided which can be used to latch data into the DAC in applications where a SAA7000 is not used.

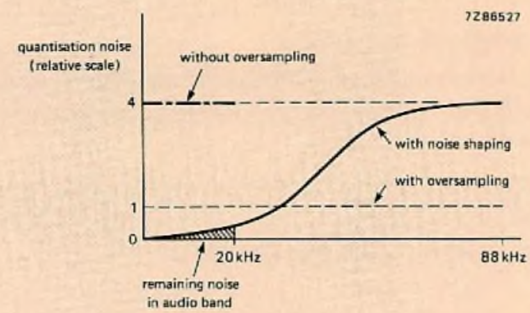


Fig.9 How 16-bit SNR is obtained from a system using 14-bit DACs. Oversampling reduces the in-band quantisation noise by spreading the total noise over a wide bandwidth. Only a quarter of the original quantisation noise remains in the audio band after oversampling. This increases the SNR of the words applied to the 14-bit DACs from 84 dB to 90 dB. Noise shaping adds another 7 dB, making the maximum SNR of the conversion system 97 dB, the same as that of a conventional 16-bit system

TDA1540: 14-bit DAC

Each TDA1540 converts 14-bit digital audio samples arriving at 176.4 kHz to an analogue output current. The output current is held between conversions by latching a flip-flop in the output circuitry of the DAC.

Data DLFD (left channel), or DRFD (right channel) are entered serially with the clock $\overline{\text{CLFD}}$. STR2 strobes the input data into a latch.

The TDA1540 uses a method of current division called dynamic element matching to achieve high-accuracy, binary-weighted currents with long-term stability. The input data are used to activate fourteen bit switches which determine the output current. Conversion to an output voltage is done in the analogue filters following the DAC.

The hold function of the DAC results in a (sin x)/x response with a first zero point at 176.4 kHz which suppresses the signal around that frequency.

If desired, one of the reference current sources in the TDA1540s can be adjusted, see Fig.8. This will compensate for slight differences in the DAC output currents for identical input data due to tolerances in the external components.

Analogue filters

The analogue filters suppress any remaining frequencies above the audio band. Since the first spectral lobe is at 176.4 kHz ± 20 kHz and has already been largely suppressed by the hold function of the DAC, third order low-pass filters provide sufficient attenuation. A Bessel filter with a cut-off frequency of 30 kHz is used because it has a linear phase characteristic. Figure 10 shows a suitable filter designed around the NE5532 dual opamp. The output of the filter is 1 V r.m.s. max. The filter components don't have to be adjusted and the de-emphasis circuit can be automatically controlled by information on the discs.

Both the Bessel filter response and the (sin x)/x response of the DAC have been taken into account when calculating and scaling the filter coefficients in the SAA7030 to give the digital-to-analogue conversion unit a flat frequency response in the audio band.

Note: the conversion system can be used at other sampling frequencies since the digital filter retains its cut-off frequency at 0.45 times the input sampling frequency. In addition, the (sin x)/x compensation is unchanged for different sampling frequencies, the first zero always occurring at the oversampling frequency. For a large frequency variation, only the analogue filter has to be changed to maintain an optimally flat amplitude characteristic.

Figure 11 shows how the original audio spectrum is recovered from the digital audio by oversampling and digital filtering followed by analogue filtering.

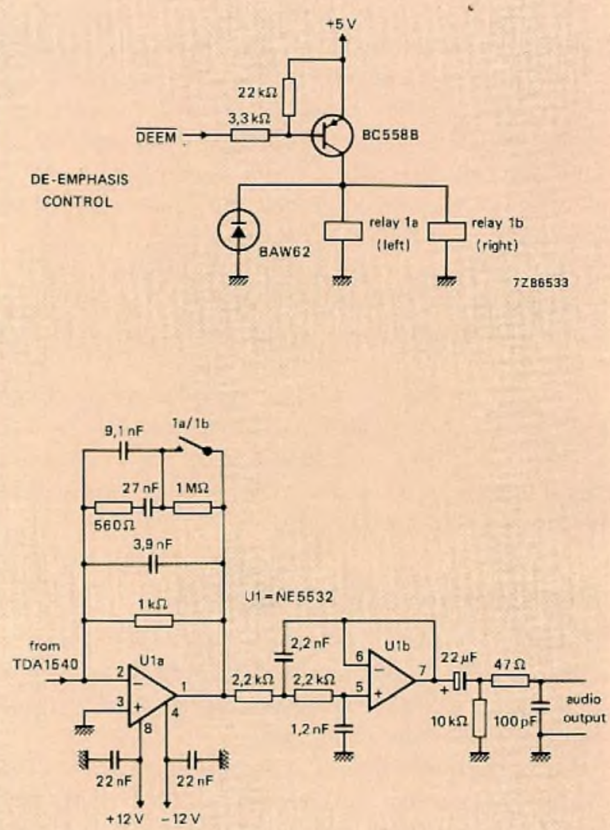


Fig.10 Third-order Bessel filter shown in Fig.8

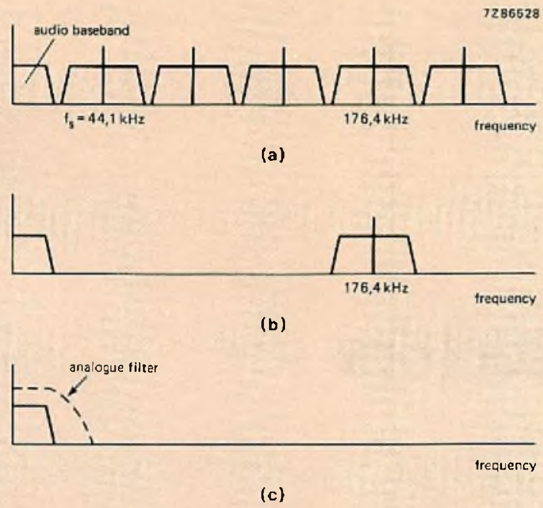


Fig.11 Oversampling and digital filtering remove the spectral lobes between the baseband and 176.4 ± 20 kHz, so that cheap low-order low-pass analogue filters can be used after the DACs instead of very-high-order ones necessary to suppress the 44.1 ± 20 kHz lobe in systems not using oversampling techniques. (a) Spectrum of an impulse-sampled audio signal sampled at 44.1 kHz. (b) Spectrum after oversampling at 176.4 kHz and digital filtering which suppress the lobes between the audio baseband and 176.4 ± 20 kHz. (c) Audio output spectrum. The lobe around 176.4 kHz is suppressed by the hold function of the DAC, which has a sin x/x characteristic with a first zero at 176.4 kHz, and by the analogue low-pass output filter with a cut-off frequency between 30 and 40 kHz

Power supply

Supplies for the decoder shown in Figs.6 and 8 are easily derived. Figure 12 shows an example.

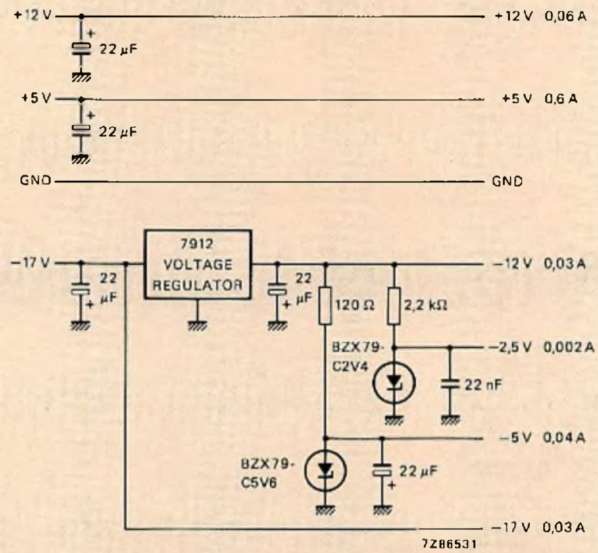


Fig.12 Power supply of decoder

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2. VAN DE PLASSCHE, R. J., 'Monolithic 14-bit DAC with 85 dB S/N ratio', Electronic Components and Applications, Vol. 2, No. 4, August 1980, pp. 234-241.

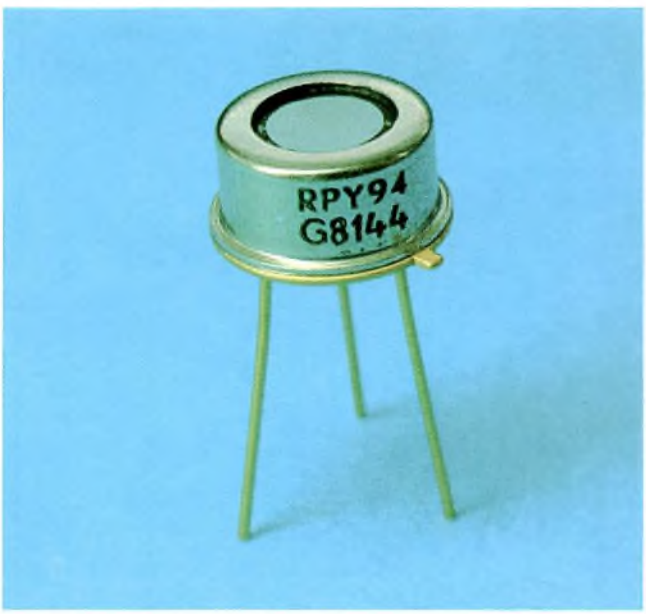
Pyroelectric ceramics that respond to changes in ambient infrared can be used in fire and intruder alarms, remote controls and level sensors, radiation pyrometers, and gas analysers. Suitable ceramic elements, together with impedance-matching networks, are now available in simple, windowed encapsulations.

Pyroelectric infrared detectors

M. A. ROSE

Early applications of infrared detection were principally of a scientific or military nature but, with the development of low-cost detectors, industrial and consumer applications are of increasing importance. Our capability as a manufacturer of infrared detectors dates from the 1940s, and we are now the largest European producer and offer the most comprehensive range of devices. This article concentrates on a portion of our range namely ceramic pyroelectric detectors. These devices are rugged low-cost components ideally suited for use in intruder detection systems, infrared radiometry, and similar applications.

Fig.1 TO-5 encapsulated ceramic pyroelectric detector



DEVICE DESCRIPTION

Each ceramic pyroelectric detector consists of an infrared sensitive element, a low-noise impedance-matching circuit, and an infrared window, all in a TO-5 encapsulation; see Fig.1. The sensitive element is an electrically-polarised ceramic slice with electrodes deposited on opposite faces. If the level of infrared radiation incident on the slice changes, then the temperature of the slice will change and, because of the pyroelectric nature of the ceramic, a voltage will be developed between the electrodes.

The detectors contain either a single or dual pyroelectric element, and each device can be represented electrically by one or two capacitors, an n-channel FET, and a non-linear network*, connected as shown in Fig.2. Single- and dual-element devices are shown in Fig.3.

The dual-element devices have differentially-connected sensitive areas with a single impedance-converting amplifier

*UK patent 1580403

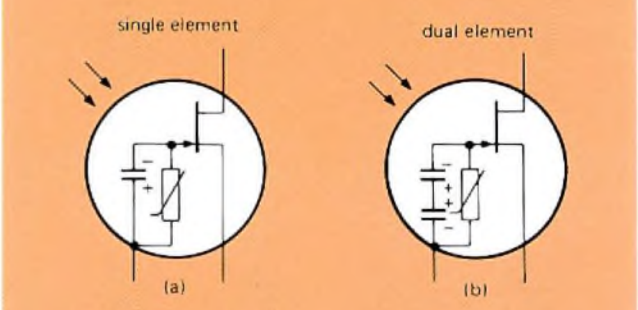


Fig.2 Symbols for single- and dual-element detectors

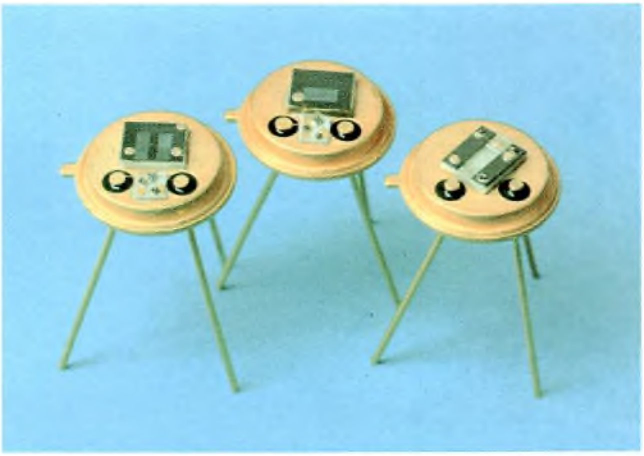


Fig.3 Detector interiors: top RPY86/87 single-element detector, bottom left RPY93 dual-element detector, bottom right RPY94/95 dual-element detector.

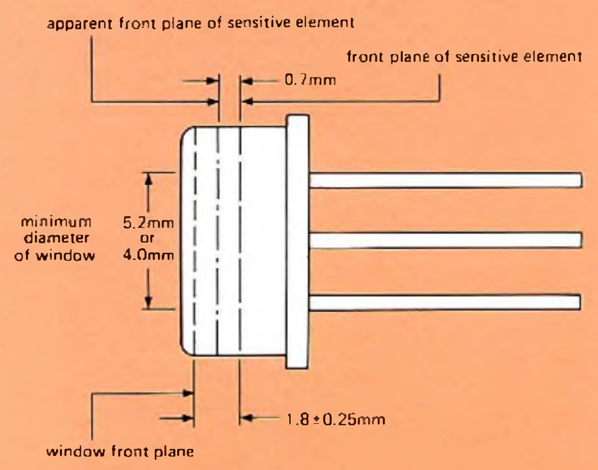


Fig.4 Section through a ceramic pyroelectric detector

to provide immunity from common-mode signals such as those generated by variations in ambient temperature, background radiation, and acoustic noise. An electrical signal is only produced when the two elements are receiving different levels of infrared radiation. They are thus invariably used with an infrared focusing system, the electrical signal being produced as the image moves across one or other of the sensitive areas.

The non-linear network is important: it protects the gate of the FET (which forms part of the impedance-matching circuit) from excessive negative voltages, and progressively limits the pyroelectric voltage resulting from large changes in the ambient temperature. Signals can therefore be obtained under conditions which would otherwise either overload the preamplifier or require it to have a very large dynamic range.

Two standard types of window are available: silicon, with a substantially flat transmission over the wavelength range from 1 μm to beyond 15 μm; or a 'daylight' filter, which transmits in the narrower range 6.5 μm to greater than 14 μm (thereby making the device insensitive to short-wavelength infrared, as emitted by the sun). Because the window material has a high refractive index, the apparent position of the front plane of the sensitive element is moved 0.7 mm towards the window front plane (as shown in Fig.4).

Details of our ceramic pyroelectric infrared detectors are given in Table 1.

PYROELECTRICITY

A pyroelectric ceramic is composed of a mass of minute crystallites, each of which behaves as a small electric dipole. In freshly manufactured material, the orientation of the dipoles is random, and the overall polarisation of the ceramic is zero (see Fig.5a). If the material is heated to just below a critical temperature, the Curie temperature, and an electric field is applied to it, the dipoles tend to line up with the applied field (see Fig.5b). After the material has cooled, and the applied field has been removed, the dipoles remain in the 'poled' position giving rise to a remanent polarisation. If the temperature of the ceramic is altered, for example by incident radiation, the polarisation will change. These changes in polarisation are reversible, provided the Curie temperature is not exceeded. Above the Curie temperature, however, the polarisation is irretrievably lost. (This is a somewhat simplified explanation that ignores the effect of domains in the crystallites.) To perform the poling process, electrodes are deposited on opposite faces of the material so that the poled ceramic forms a charged capacitor.

An increase in the incident radiation will raise the temperature of the ceramic, and will reduce the density of the captive charge within the ceramic material. This will leave an excess of induced charge on the electrodes, so that the potential difference across the ceramic will increase, giving a rising voltage signal at the gate of the impedance-matching FET. Similarly, a reduction in the level of the incident radiation will result in a corresponding fall in the FET gate voltage. Thus the detector will produce an output signal only when the level of the incident radiation is changing. To detect a stationary object with a constant infrared emission, it is necessary to chop the incident radiation.

ELEMENT MATERIAL

The pyroelectric ceramic material used in the detectors has been specially developed in our research laboratories. It is a lead zirconate titanate, doped to optimise the properties required for infrared detectors. It is insensitive

TABLE 1 Ceramic pyroelectric infrared detectors

Type	Number of elements	Element spacing (mm)	Element dimensions (mm)	Spectral response (μm)	Window diameter (mm)
RPY86	1	–	2 X 1	6.5 to > 14	5.2
RPY87	1	–	2 X 1	1.0 to 15	5.2
RPY88	1	–	2 X 2	6.5 to > 14	5.2
RPY89	1	–	2 X 2	1.0 to 15	5.2
RPY93	2	0.5	2 X 0.75	6.5 to > 14	5.2
RPY94	2	1.0	2 X 1	6.5 to > 14	5.2
RPY95	2	1.0	2 X 1	6.5 to > 14	4.0
RPY96	1	–	2 X 1	6.5 to > 14	4.0

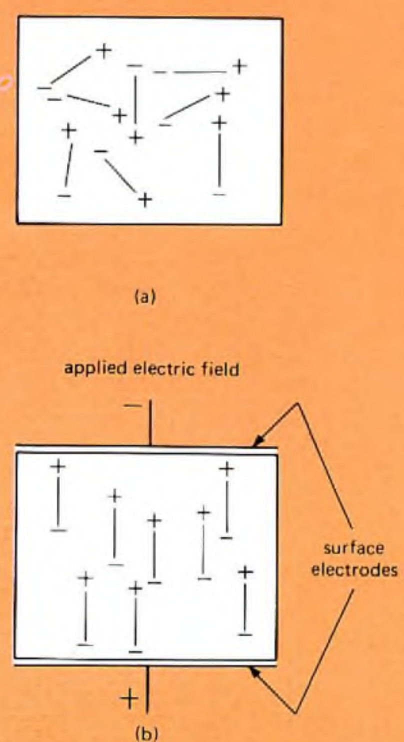


Fig.5 –Orientation of electric dipoles
a) unpoled b) poled

to water and is extremely rugged, and can be handled by mass-production techniques similar to those used in the manufacture of conventional semiconductor devices. The material has a high Curie temperature, and can be used up to 100°C. Furthermore, responsivity is only slightly temperature-dependent.

PERFORMANCE

The performance of a ceramic pyroelectric detector is defined principally by its responsivity and noise equivalent power (N.E.P.). Values of responsivity and N.E.P. for our range of devices are given in Table 2.

Responsivity

The responsivity is defined as the ratio of the r.m.s. signal voltage to the r.m.s. incident radiant power. The units of responsivity are volts per watt. In published data, values of responsivity are followed by figures in brackets. For example, in Table 2 the responsivity of the RPY86 is given as 600 VW^{-1} (10 μm, 10). The 10 μm denotes the wavelength of the infrared radiation generating the signal voltage, while the 10 indicates that the radiation is chopped at a frequency of 10 Hz.

Noise equivalent power (N.E.P.)

The N.E.P. is a measure of the minimum power that can be detected. It is the r.m.s. radiant power incident on the

TABLE 2 Performance data for ceramic pyroelectric infrared detectors

	Responsivity (typical values) VW^{-1}	N.E.P (typical values) $WHz^{-½}$
RPY86	600 (10 μm, 10)	0.9×10^{-9} (10 μm, 10,1)
RPY87	500 (6 μm, 10)	1.05×10^{-9} (6 μm, 10,1)
RPY88	300 (10 μm, 10)	1.65×10^{-9} (10 μm, 10,1)
RPY89	250 (6 μm, 10)	2.0×10^{-9} (6 μm, 10,1)
RPY93	800 (10 μm, 10)	1.4×10^{-9} (10 μm, 10,1)
RPY94	650 (10 μm, 10)	1.5×10^{-9} (10 μm, 10,1)
RPY95	450 (10 μm, 10)	2.1×10^{-9} (10 μm, 10,1)
RPY96*	130 (10 μm, 10)	3.5×10^{-9} (10 μm, 10,1)

*The values of typical responsivity and N.E.P. for the RPY96 are specified for a source-follower configuration (that is, unity gain). All other detectors are specified using the recommended preamplifier which has a gain of 4.8.

detector which produces an r.m.s. signal equal in magnitude to the r.m.s. detector noise. The units of N.E.P. are watts per root hertz, and as with responsivity, the relevant measuring conditions must be specified. Thus the N.E.P. of an RPY86 detector is given in Table 2 as 0.9×10^{-9} $WHz^{-½}$ (10 μm, 10, 1), where 10 μm is the wavelength of the incident radiation, 10 is the chopping frequency in Hz, and 1 is the bandwidth in Hz.

AMPLIFICATION

Since the sensitive element is effectively a capacitor, it has a very high impedance, and an impedance-matching circuit (the integral FET plus a preamplifier) is thus essential to match the sensitive element to subsequent amplifiers.

Preamplifier

The incorporation of a low-noise FET in the detector gives considerable freedom in the choice of the preamplifier, although in any design it is necessary to allow for the spread in the d.c. characteristics of the FET. The recommended preamplifier is shown in Fig.6 and is relatively independent of the FET characteristics. It has a gain of 4.8, set by the 1.8 kΩ and 470 Ω resistors connected to the source. The noise level at the output is typically 250 $nVHz^{-½}$ at 10 Hz and the output impedance is approximately 200 Ω. The d.c.

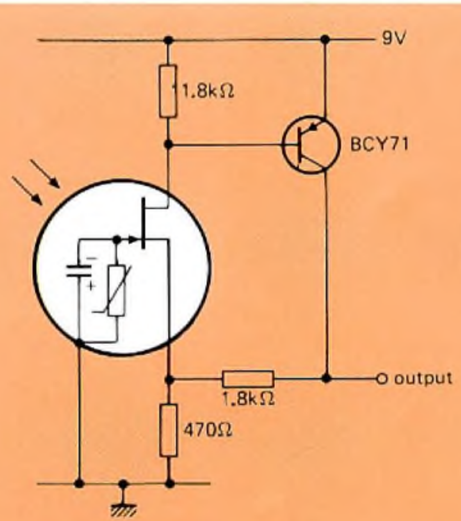


Fig.6 Recommended preamplifier

output, with zero input, will lie between 2 and 8 V. This spread is the main reason for limiting the gain, although higher gains can be achieved if the resistors are selected to match the FET. The gain provided by this preamplifier is important where low noise is necessary, since the noise introduced by subsequent standard operational amplifiers would otherwise prove obtrusive.

Amplifiers

If the amplifier stages subsequent to the preamplifier are to increase the noise level by less than 10%, their input noise voltage should be less than one-third of the preamplifier output noise; that is, their input noise should be less than 80 $nVHz^{-1/2}$ at 10 Hz. A suitable operational amplifier is the μA776 which has a typical input noise level of 50 $nVHz^{-1/2}$ at 10 Hz (for comparison, a standard μA741 has a noise level of 70 $nVHz^{-1/2}$ at 10 Hz).

If the system design can tolerate some degradation in the overall signal-to-noise ratio, a wider range of operational amplifiers can be considered. A particularly useful device is the LM324, a quad operational amplifier, which is ideally suited to single-supply operation. The typical noise voltage for the LM324 is 170 $nVHz^{-1/2}$ at 10 Hz. With four independent amplifiers in a single package, this device provides a compact and economical solution to a variety of application problems. A dual version, the LM358, is also available. A selection of amplifier circuits, using operational amplifiers, is given in Ref. 1.

APPLICATIONS

The detector responds to moving objects by detecting the flux change they produce in passing into and out of the field of view. For the detector to respond to stationary objects the flux must be chopped. The following application examples cover both cases.

Intruder alarm

In this application, flux variation is caused by an intruder moving across the detector's field of view. Recommended detectors for this application are types RPY93 to RPY95. These are dual-element devices incorporating 'daylight' filters. They are thus largely insensitive to environmental effects, such as variations in temperature and background radiation, so the risk of false alarms is reduced.

To ensure a well-focused image (essential for dual-element detectors) and an acceptable operating range, the infrared radiation must be focused by an optical system. This could be an inexpensive moulded plastic mirror, similar to the glass prototype shown in Fig.7. The prototype mirror is constructed from five segments, each being an off-axis section from a sphere of radius 80 mm. The segments are mounted on a circle of radius 40 mm so that their foci are coincident; see Fig.8. With this type of off-axis design, the detector can be mounted at the focus without obscuring the field of view. As an intruder moves across the field of view of the mirror, five separate images will be focused onto each element of the detector, ensuring multiple triggering of the alarm.

An alarm unit using the dual operational amplifier LM358 is shown in Fig.9. This is a low-cost design with a simple single-shot threshold detector as a trigger. Although the use of a dual-element detector provides a substantial immunity to false alarms, this can be further improved by a double-pulse triggering technique; an alarm circuit with double-pulse triggering is described in Ref.1.

Using the circuit of Fig.9 and a mirror of the type shown in Fig.7, it is possible to detect a man in shirt sleeves at a distance of 30 m.

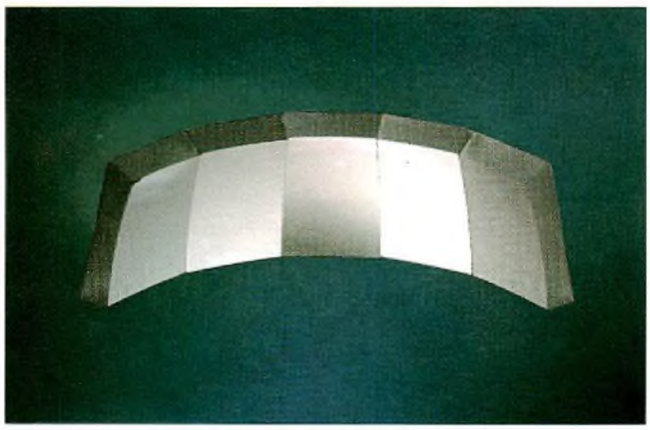


Fig.7 Multi-faceted mirror

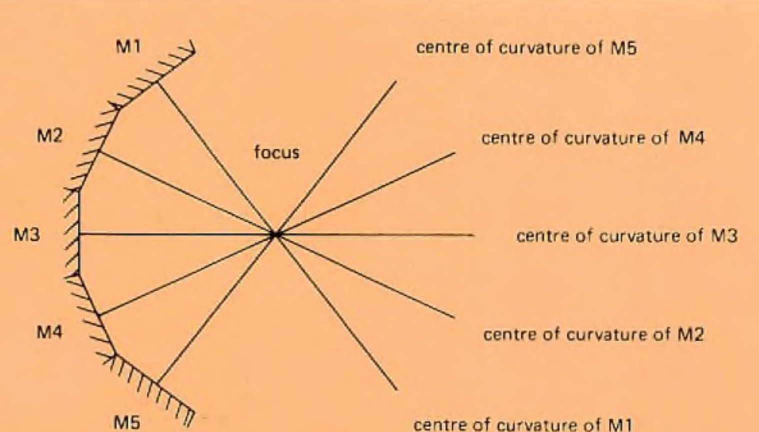


Fig.8 Section through mirror

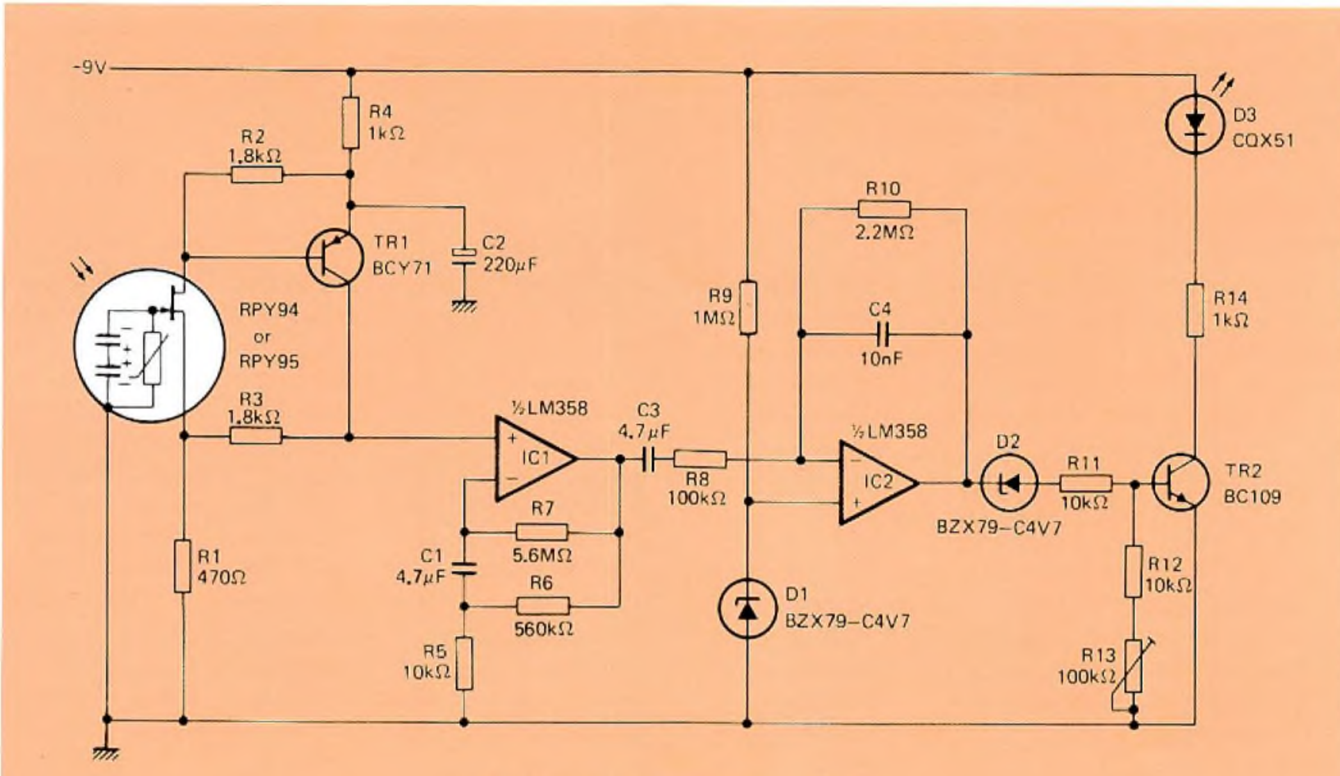


Fig.9 Circuit of low-cost intruder alarm

Low-cost remote switch

The intruder unit described above can be used as the basis for a low-cost remote switch. In its simplest form, this would use a single-element detector, a simplified circuit, and no collecting optics. While a switch of this type would have a restricted range, typically half a metre, and only a limited immunity to false triggering, it could form an attractive feature in many domestic appliances. The use of such a switch in a digital alarm clock is described in Ref.2. By simply waving a hand near the clock, the alarm can be muted and the display switched on.

By detecting the presence of someone in a dark corridor or staircase, the same system may also be used as a remote light switch. However, in comparison with the alarm clock example, such an application would generally require an improvement in sensitivity. This could be achieved via higher amplification or the use of collecting optics. The addition of a light-level sensor and a delay facility would provide a fully automatic and vandal-proof system.

Fire alarm

The detectors can also be used in fire alarm systems. The flickering of flames and hot gases provide the necessary modulation of the infrared radiation received by the detector. Since flames are copious sources of infrared, collecting optics are unnecessary. The false alarm immunity of such detector systems can be improved by using a window with a specific narrow transmission characteristic. Detectors with non-standard windows, having the appropriate transmission characteristics, can be supplied to customer requirements.

Sorter

The detectors can be used to sort objects with different temperatures, emissivities, thermal conductivities, or radiation transmission. For example, identical objects on a conveyor belt can be made to pass across the field of view of a detector. The detector output signal, which will be a measure of the temperature of each object with respect to the background, can control an accept/reject mechanism. A similar technique can be used to mark material in a continuous strip process.

Simple radiometer

A simple radiometer can be constructed by placing a mechanical chopper in front of the detector and defining the field of view with an aperture. The signal S from the detector is given by:

$$S = \beta \left(\epsilon_2 T_2^4 - \epsilon_1 T_1^4 \right)$$

where ϵ_1 is the emissivity of the chopper blade, T_1 is its temperature (in kelvin), ϵ_2 is the emissivity of the target,

T_2 is the temperature of the target, and β is a constant. Variations in T_1 can be compensated by using a linear temperature transducer. A calibrated gain control can be used to allow for variations in emissivity of the target.

Radiometer with compensation for ambient temperature

A low-cost remote-sensing radiometer which does not require chopper blade temperature compensation is described in Ref.3. It uses a null technique in which radiation from a defined area of the target is balanced against the radiation from an internal reference source. At balance:

$$\frac{\epsilon_t}{\epsilon_r} = \left(\frac{T_r}{T_t}\right)^4$$

where ϵ_t is the emissivity of the target, ϵ_r the emissivity of the internal reference source, T_r is the absolute temperature of the internal reference source, and T_t is the absolute temperature of the target. The value of T_r can be measured using a contact thermistor or thermocouple, so that provided ϵ_t and ϵ_r are known, T_t can be readily determined.

Radiometer with compensation for target emissivity

By splitting the target radiation into two spectral regions, it is possible to make a radiometer that compensates for variations in target emissivity. The detector samples each region alternately, and the ratio of the two output signals is proportional to the target temperature. The target radiation must be chopped prior to filtering so that the radiation emitted by the filter is eliminated as a source of error.

Radiometer with collecting optics

The use of collecting optics will increase the radiation energy incident on a detector, in the ratio of the area of the collecting aperture to the area of the sensitive element in the detector. As a result, the temperatures of areas of a remote radiant object can be measured provided its emissivity is known.

Level sensor

A dual-element detector can be used with a chopper and focusing optics to sense levels by detecting small temperature discontinuities. An application for such an instrument is in sensing solid or liquid levels in vertical storage tanks and silos. In containers of this type, there is usually a temperature difference between the full and empty parts of the tank. By employing a dual-element detector (elements in series opposition) and a chopper blade which masks and then exposes both elements simultaneously, a difference signal is produced. The instrument can thus be used to scan a field of view and produce a maximum difference signal at the level discontinuity.

Gas analysis

The presence of some gases can be detected by the amount of infrared radiation that they absorb at characteristic wavelengths. To detect a particular gas, infrared radiation at the characteristic absorption wavelength of the gas (the radiation is selected by narrowband filters) is passed through a reference chamber, which does not contain the gas, and a chamber which contains the sample. The relative transmission of infrared radiation through the two chambers is a measure of the amount of the sought-after gas. The absorption of the gas is likely to be small, so that a long radiation path-length is essential. This can be accommodated in a small space by the use of multiple reflection between mirrors. A system of this type could be used to detect alcohol in the breath ('breathalyser'), or to measure industrial pollution or car exhaust gases.

RELIABILITY

Ceramic pyroelectric detectors are manufactured and tested using standards and practices which correspond to those employed in the production of high-quality transistors. The incoming components are sampled for quality, and further sample checks are made after each stage in the production process. After final assembly, all detectors are tested for responsivity and N.E.P., and the dual-element detectors are tested for signal matching.

The measuring conditions used in the quality assurance programme for the detectors are based on those stipulated for the CECC 50 000 series of approved transistors. Details of the programme for the RPY93 are given in Table 3, and identical or very similar test conditions apply to the other types. The first four tests listed in Table 3 define the batch release procedure. Failure in any one of these tests will result in the rejection of a batch.

Samples of the devices have been subjected to continuous storage under a variety of conditions, including ambient, high temperature, and continuous electrical operation. Over one million device hours have been logged in these tests without a single failure.

HIGH-PERFORMANCE PYROELECTRIC DETECTORS

Ceramic pyroelectric detectors are not designed to meet the performance requirements of infrared spectroscopy or high-resolution radiometry. For such applications, we have developed a range of pyroelectric detectors, types

TABLE 3 Quality assurance specification for the RPY93

Test	Severity	Duration	Frequency	Corresponding CECC group
electrical parameter measurement	–	–	for each batch	A
moisture resistance, steady state – IEC 68-2-3 test Ca	+40 °C, 95% RH	168 h	for each batch	B
solderability – IEC 68-2-20 test T	+235 °C, 1.5 mm from header	5 s	for each batch	B
lead fatigue – IEC 68-2-21 test Ub	4 cycles	–	for each batch	B
high-temperature storage – IEC 68-2-2 test Ba	+70 °C	2000 h	quarterly	C
low-temperature storage – IEC 68-2-1 test A	–40 °C	2000 h	quarterly	C
change of temperature – IEC 68-2-14 test Nb	–40 °C to +70 °C	10 cycles	quarterly	C
vibration, swept frequency – IEC 68-2-6 test Fc(B4)	125 Hz to 2 kHz 196 ms^{-2}	2 h in each orientation	quarterly	C
acceleration, steady state – IEC 68-2-7 test Ga	196 000 ms^{-2}	60 s	quarterly	C
shock – IEC 68-2-27 test Ea	14 700 ms^{-2}	3 pulses 6 orientations	quarterly	C
continuous operation	–	–	quarterly	C
resistance to solder heat	+350 °C, 6 mm from header	3 s	annually	–
performance measurement over operational range (temperature and frequency)	–	–	annually	–

P1601, P1603, and P1604, which use deuterated L-alanine doped* triglycine sulphate (DLATGS) as the sensitive element. These devices retain the TO-5 encapsulation of the ceramic detectors, and also incorporate an impedance-matching amplifier. Three window materials are available: anti-reflection coated germanium (the P1601 detector, spectral response 1 to 15 μm), caesium iodide (the P1603 detector. spectral response 1 to 70 μm), and calcium fluoride (the P1604 detector, spectral response 1 to 9 μm). With a caesium iodide window, the responsivity is typically 5×10^3 VW^{-1} (500 K, 10) and the N.E.P. is 5×10^{-10} $WHz^{-1/2}$ (500 K, 10, 1).

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* UK patent 1297198

The SAA1057 is a new 18-pin LSI circuit which performs all the functions of a PLL controlled by a microcomputer in an electronically-tuned radio. Since the h.f. prescaler and tuning voltage amplifier are on-chip, this circuit and only a few peripheral components perform all the PLL functions that previously required the use of three integrated circuits and a host of peripheral components.

Single-chip synthesiser for radio tuning

J. MATULL and J. VAN STRAATEN

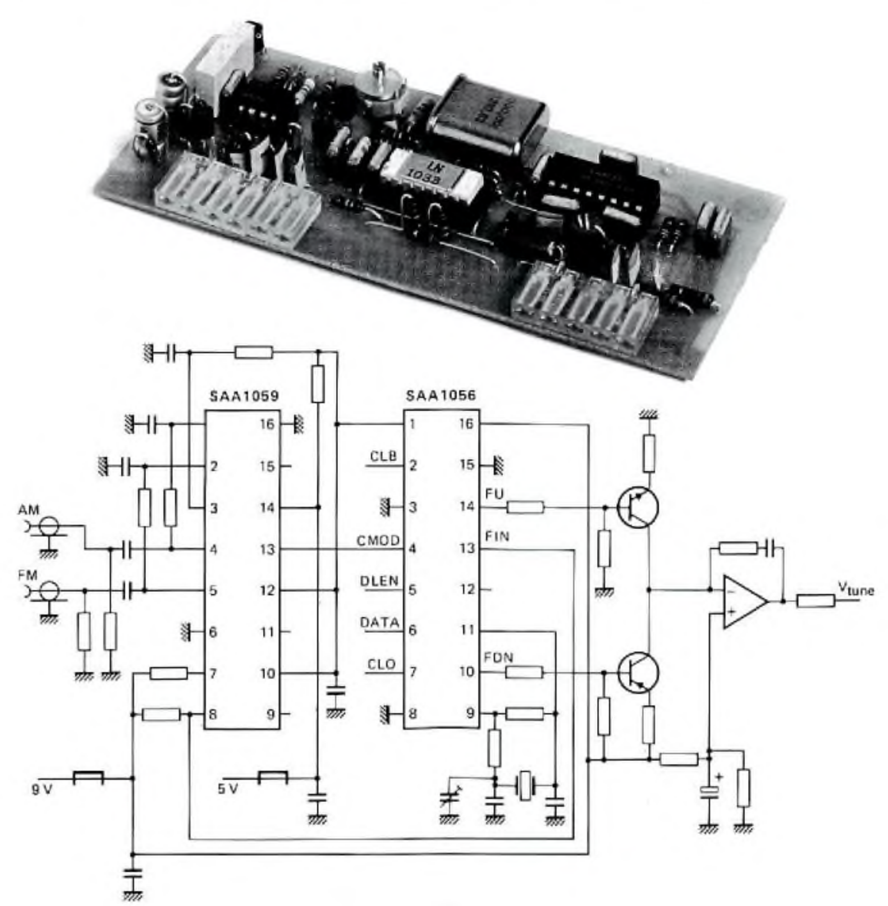
To remain competitive, manufacturers of domestic radios must not only produce a comprehensive range of reliable equipment with the required performance at the right price, but must also meet the needs of the market with regard to styling, ease of operation and available functions. Although the widespread use of integrated circuits has allowed vast improvements of performance and reliability and has increased the range of available facilities, the integrated circuits are not always optimally matched, resulting in partial redundancy and a large number of peripheral components. We foresaw this problem and were able to avoid it by using a total systems approach to manufacture our comprehensive range of ideally matched integrated circuits for signal processing and digital control of tuning, displays and analogue functions in all classes of radio. We can now therefore devote our design resources and considerable knowledge of integration technologies and techniques to reducing radio manufacturers' development and assembly costs by minimising the number of integrated circuits needed to implement the wide range of features and facilities required in today's radios.

If a radio must incorporate facilities such as search tuning and/or tuning by direct entry of frequency at a keyboard, variable-capacitance diode tuning must be used and a stable local-oscillator signal can be generated by indirect frequency synthesis with a phase-locked loop (PLL) controlled by a microcomputer. This system was fully described in Ref.1 which showed how integrated circuits SAA1059 (h.f. prescaler), SAA1056 (PLL frequency synthesiser) and an op-amp integrator (PLL loop filter and amplifier) were used in our radio tuning system (RTS). We have now used bipolar technology to combine analogue circuits with several types of logic (I^2L, ECL and miniwatt) so that all the functions previously performed by three integrated circuits can be performed by a single 18-pin LSI integrated circuit called synthesiser module SAA1057. The component economy afforded by the SAA1057 is amply illustrated by Fig.1 which shows that tuning synthesiser functions which previously required the use of three integrated circuits and a large number of peripheral components can now be performed by the SAA1057 and only 16 peripheral components.

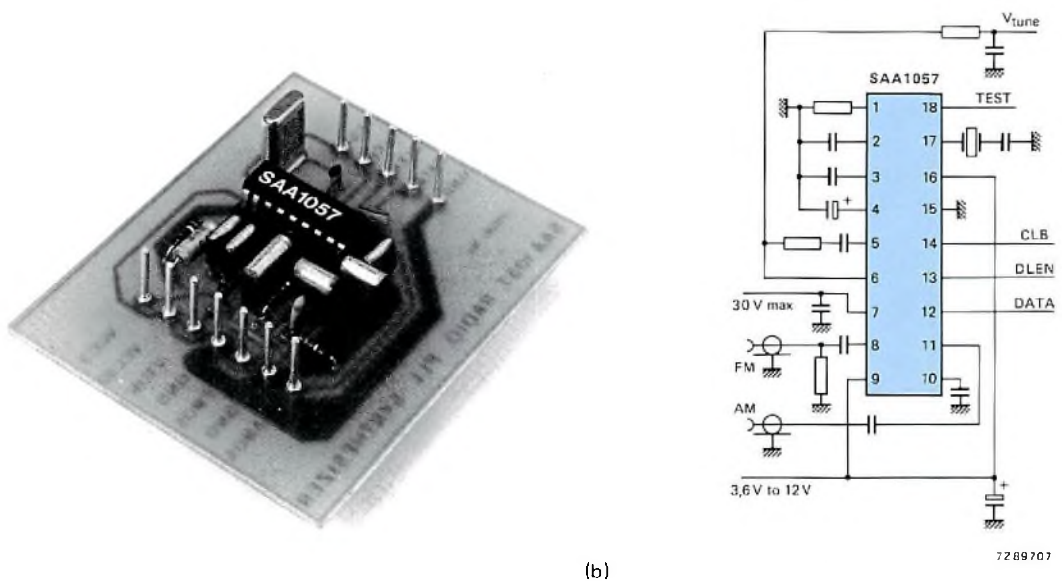
The SAA1057 is not only economical with regard to the required number of components. It also consumes very little current (<20 mA) and is able to meet the varied performance requirements of all classes of radio from battery-powerd portables to mains-powered hi-fi tuners. For example, a novel twin phase detector system in the PLL achieves the fast tuning often required for car radios and also ensures that, when the PLL is locked, the VCO signal has high spectral purity to ensure low distortion in hi-fi tuners. The wide frequency range (a.m. 512 kHz to 32 MHz, f.m. 70 MHz to 120 MHz) and high maximum tuning voltage (30 V) make the SAA1057 suitable for multi-waveband mains sets. The low current consumption combined with the wide supply voltage range (3.6 V to 12 V) due to internal stabilisation allow it to be used in battery-powered portables.

In addition to the basic function of tuning by direct entry of frequency, the SAA1057 can also provide the following software-controlled facilities:

- search tuning with muted interstation noise
- continuous up/down step tuning (manual tuning)
- accurate storage and automatic tuning to preset frequencies



(a)



(b)

Fig.1 Basic radio tuning synthesisers. Circuit (a) uses three integrated circuits and 36 peripheral components. Circuit (b) performs the same functions using synthesiser module SAA1057 and 16 peripheral components

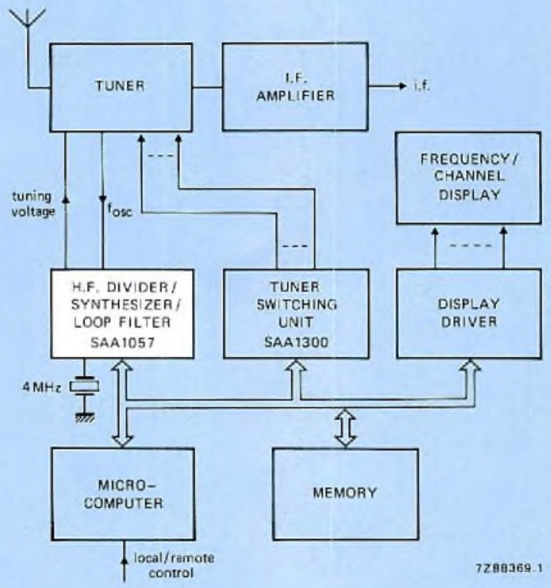


Fig. 2 Integrated circuits for tuning systems using SAA1057.

BIPOLAR CIRCUITS

Remote control

TDB2033	gain-controlled remote IR receiver amplifier

Frequency synthesiser

SAA1057	radio tuning PLL frequency synthesiser

Display drivers

SAA1060	32 segment LED
SAA1062/T	20 static outputs for LCD
SAA1063	32 segment FTD

Tuner switching

SAA1300	5-line switching circuit

MOS CIRCUITS

Display drivers

PCE2100	40 segment LCD	in duplex mode
PCE2110	60 segment LCD + 2 LEDs	in duplex mode
PCE2111	64 segment LCD	in duplex mode
PCE2112	32 segment LCD static	
SAA1061	16 static outputs for LED drive and switching functions	
SAB3044	2 digit LED	

Single-chip 8-bit microcomputers

MAB8021	with 1K byte ROM and 28-pin package
MAB8048	with 1K byte ROM and 40-pin package
MAB84XX	NMOS family with 1 to 4 Kbyte ROM and I^2C bus
MAB85XX	CMOS family with 0.5 to 4K byte ROM and I^2C bus

Memories

PCD8571	128 x 8-bit CMOS memory with serial I/O
PCB1400	100 x 16-bit EEPROM with serial I/O

Infrared remote-control receivers

SAB3023	receiver and analogue memory
SAB3033	receiver and analogue memory
SAB3042	receiver and decoder with C-bus
SAB3028	receiver and decoder with I^2C bus

Infrared remote-control transmitters

SAB3004	7 x 64 commands
SAB3021	2 x 64 commands
SAB3027	32 x 64 commands

- loading of frequency data in synchronism with the sampling frequency to prevent disturbance of the tuning lock
- feed out of a number of internal signals for alignment purposes
- adjustment of PLL current gain over 40 dB range (0.023 to 2.3) to eliminate switching of external loop filter components during waveband selection.

As the word 'module' in the name of the SAA1057 indicates, this new IC is part of a modular, data-bus compatible, digitally-controlled tuning system in accordance with the systems design philosophy followed for other circuits in our range of ICs for digital systems in radios. The modular approach minimises radiation and reduces wiring and screening costs because:

- all the sensitive signal processing circuits for the tuning systems are now in the SAA1057 which can be mounted in the ideal position close to the tuner
- internal h.f. dividers eliminate the need for an external prescaler
- two sensitive, internally switched VCO inputs to the SAA1057 allow direct connection of the f.m. and a.m. local-oscillator signals without additional impedance matching, amplification or switching
- the crystal-controlled reference oscillator for the PLL operates at the same frequency for the a.m. and f.m. waveband and causes little radiation because it generates a low-level sinewave
- the separate microcomputer and memory can be mounted close to the keyboard and their capacity can be tailored to meet the demands of specific radios
- the frequency display driver can be mounted close to its display.

As shown in Fig.2, the data-bus compatibility of tuning systems using the SAA1057 also allows the simple addition of circuits as required for waveband-switching and for driving LED, LCD or fluorescent displays of preset station number, waveband and channel number. Other facilities which can be simply and economically accommodated are analogue signal control, extra display functions, and remote control via an infrared data link.

OPERATING PRINCIPLES OF FREQUENCY SYNTHESIS

A basic digitally-controlled PLL for radio tuning is shown in Fig.3. The output from the voltage-controlled local-oscillator in the radio is converted into a pulse train, and frequency divided by a programmable divider, before being applied to one of the inputs of the phase detector. The output from the crystal-controlled reference oscillator is converted into a pulse train, and frequency divided by one of two ratios, before being applied to the other input of the phase detector. The phase detector output, which is proportional to the relative phase (and therefore the frequency) of the two input signals, is passed through the low-pass loop filter to remove the high-frequency components and fed back to the VCO as the tuning control voltage. The loop is locked, and the radio correctly tuned, when $f_{osc} = Nf_{ref}$ where N is the programmable division ratio determined by selecting the frequency of the required broadcast.

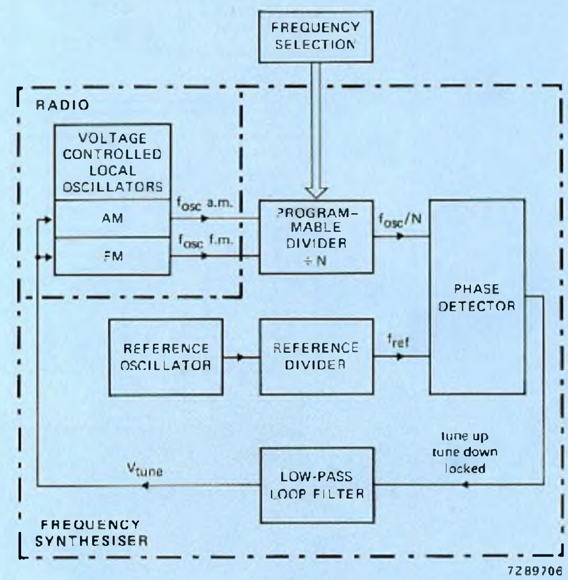


Fig.3 A basic digitally-controlled PLL for radio tuning

BRIEF DESCRIPTION OF THE FUNCTIONS OF THE SAA1057 (Fig.4)

Local-oscillator inputs

The local-oscillator signals from the radio are applied to inputs FFM for f.m. and FAM for a.m. Since these inputs have a sensitivity of 30 mV to 500 mV (a.m.) and 10 mV to 500 mV (f.m.), the local-oscillator signals can be directly applied without preamplification or buffering. A separate pin (DCA) allows the bias circuitry of the internal input amplifiers to be decoupled by an external capacitor. The input frequency range is 512 kHz to 32 MHz for a.m. and 70 MHz to 120 MHz for f.m., the f.m. signals being passed through an internal divide-by-ten h.f. prescaler which is switched off by software to minimise current consumption whilst tuning the a.m. band. Since the a.m. and f.m. local-oscillator signals are automatically selected by software, they need not be externally switched during waveband selection.

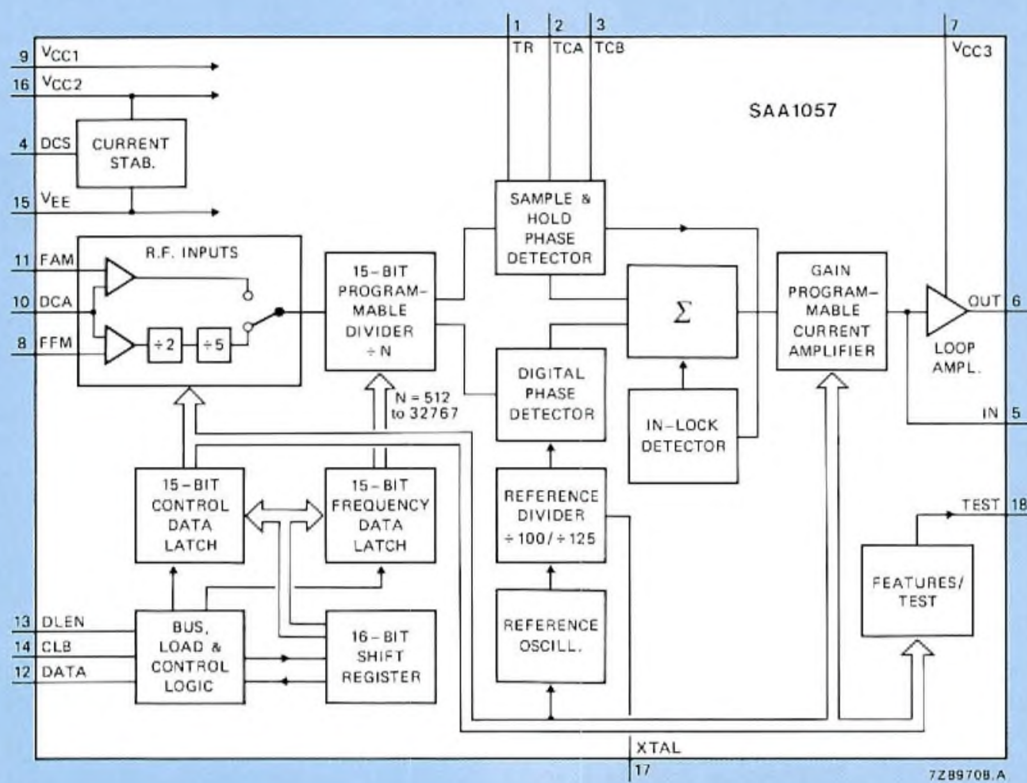


Fig.4 Block diagram of the SAA1057

Programmable divider

This 15-bit frequency divider, which is designed in a special manner to minimise current consumption, is programmed with a binary-coded divisor (N) to synthesise the required frequency for the voltage-controlled local-oscillator in the radio. The local-oscillator frequency (f_{OSC}) is usually the i.f. above the tuned frequency. The dividing number is $(32f_{OSC})/f_{ref}$ for a.m. and $(3.2f_{OSC})/f_{ref}$ for f.m., where f_{ref} is the output frequency from the reference frequency divider (40 kHz or 32 kHz). The minimum divisor is 512 and the maximum divisor is 32 767. The frequency-divided local-oscillator signal is applied as one of the inputs to a dual phase detector system.

Reference frequency oscillator

This stable, temperature-compensated oscillator is controlled by an inexpensive 4 MHz crystal (series resistance <150Ω) connected in series with a capacitor between pin 17 of the SAA1057 and the common return line. The reference frequency may alternatively be derived from a stable external source. In this case, a 4 MHz squarewave of 5 V peak to peak may be connected to pin 17 via a series-connected 10 nF capacitor and 22 kΩ resistors.

Reference frequency divider

This circuit divides the frequency of the signal from the reference oscillator by 125 or 100 to obtain a reference frequency of 32 kHz or 40 kHz for the dual phase detector system under the control of software. If the selected reference frequency is 32 kHz, the minimum tuning step is 1 kHz on a.m. and, due to the divide-by-ten h.f. divider, 10 kHz on f.m. If the selected reference frequency is 40 kHz, the minimum tuning steps for a.m. and f.m. are 1.25 kHz and 12.5 kHz respectively. If larger tuning steps are required, integer multiples of these tuning steps can be selected by software.

Phase detector system

To simplify the design of the PLL loop filter, the SAA1057 incorporates a novel dual phase detector system that uses the same reference frequency for a.m. and f.m. One of the phase detectors is a high speed digital memory (flip-flop) type, the other is a high gain analogue memory (sample and hold) type. The digital phase detector operates at the reference frequency, generates about 100 times as much tuning current as the analogue phase detector and provides high speed tuning over a wide frequency range. The analogue phase detector operates at 1/32 of the reference frequency, has no region of uncertainty in its transfer characteristic and provides increased spectral purity of the local-oscillator signal when the PLL is locked. The 'hold' voltage from the analogue phase detector is converted into a d.c. current and summed with the output pulses from the digital phase detector to provide a current proportional to tuning error. This current drives a gain-programmable amplifier to generate the tuning voltage output.

The analogue phase detector is always operating, but the digital phase detector can be switched on/off by setting/resetting the in-lock detector with features/test bits in the software (e.g. to minimise noise during step tuning). If the software does not include any features/test bits, the digital phase detector is automatically switched on if the tuning error exceeds the phase range of the analogue phase detector. This could occur, for example, as the result of executing a large frequency change. When the in-lock detector determines that the tuning error has been reduced to within the operating range of the analogue phase detector for three consecutive sampling periods, the digital phase detector is automatically switched off again.

Gain-programmable current amplifier

The sum of the output currents from the two phase detectors drives a gain-programmable bidirectional current source which replaces the normally-used resistor between the charge pump and loop amplifier of a PLL. This allows the loop gain of the PLL to be software programmed over a 40 dB range within the limits 0.023 to 2.3, thereby eliminating the need to switch loop filter components during waveband selection.

Loop amplifier

The loop amplifier is capable of providing a tuning voltage output of up to 30 V and only requires a series-connected RC network between its input and output to form an active low-pass loop filter. The supply voltage for the loop amplifier (V_{CC3}) need not be stabilised but it should be adequately filtered.

Reception of frequency and control data

Data for the SAA1057 consists of serially-transmitted 17-bit frequency setting and control words from a microcomputer. Both types of word incorporate a zero start bit which is tested to identify a correct transmission. Each word also contains a latch selection bit which is 0 for a frequency setting word and 1 for a control word. The incoming data is transmitted via an asynchronous data highway with separate data (DATA), clock (CLB) and enable (DLEN) lines. The logic levels on the lines are TTL compatible and are independent of supply voltage.

Sixteen bits of each incoming data word are loaded into a shift register. The bus, load and control logic then checks that the transmission is valid by checking that the first bit is zero and that the word length is correct during the HIGH period of the DLEN line. If valid, the data word is then transferred to the appropriate latch by the next pulse on the clock line.

A frequency-setting word includes fifteen bits which define the required frequency expressed as a 15-bit binary-coded divisor (512 to 32 767) for the programmable divider.

A control word includes fifteen bits for the following purposes:

- one bit (FM) to control the switch to select the required input from the a.m. or f.m. local-oscillator. If the a.m. input is selected, the divide-by-ten prescaler is switched off to conserve power
- one bit (REFH) to program the divisor for the reference frequency divider
- four bits (CP0 to CP3) to set the gain of the gain-programmable current amplifier
- one bit (SB2) to determine whether the remaining eight features/test bits should be used or not
- one feature bit (SLA) which determines whether frequency setting data is loaded into the programmable divider immediately after reception (asynchronous loading) or synchronised with the sampling frequency (synchronous loading). Synchronous loading is for minimising noise during manual tuning without muting
- two features bits (PDM0 and PDM1) which set the operating mode of the digital phase detector as previously described
- one feature bit (BRM) which sets the bus receiver into an automatic mode so that it is switched off to conserve power after a data transmission
- four test bits (T0 to T3) which can route the reference signal, the output from the programmable divider or the output level from the in-lock detector to the TEST pin for alignment purposes.

TECHNIQUES USED TO OBTAIN THE HIGH PERFORMANCE OF THE SAA1057

Many new circuit techniques have been used in the SAA1057 to achieve the high performance, application flexibility and low power consumption. A description of the techniques listed here is beyond the scope of this article but further information can be found in the references:

- travelling-wave dividers in the divide-by-ten prescaler ensure low current consumption and high sensitivity for the r.f. inputs

- a tail-end divider is used to increase the speed of the digital phase detector
- a rate-select technique in the programmable divider minimises phase jump in the digital phase detector
- current consumption is minimised by using stacked logic for the three different types of digital circuits (I^2L, ECL and miniwatt). In this way, many of the logic circuits act as current sources for other logic circuits
- use of a bandgap current reference ensures that the current consumption remains constant over a wide range of supply voltage and operating temperature
- the op-amps at the r.f. inputs have an input bias current of less than 10 nA and also have a very high slew rate
- the tuning voltage is derived from a 30 V op-amp with a low bias current and a high slew rate.

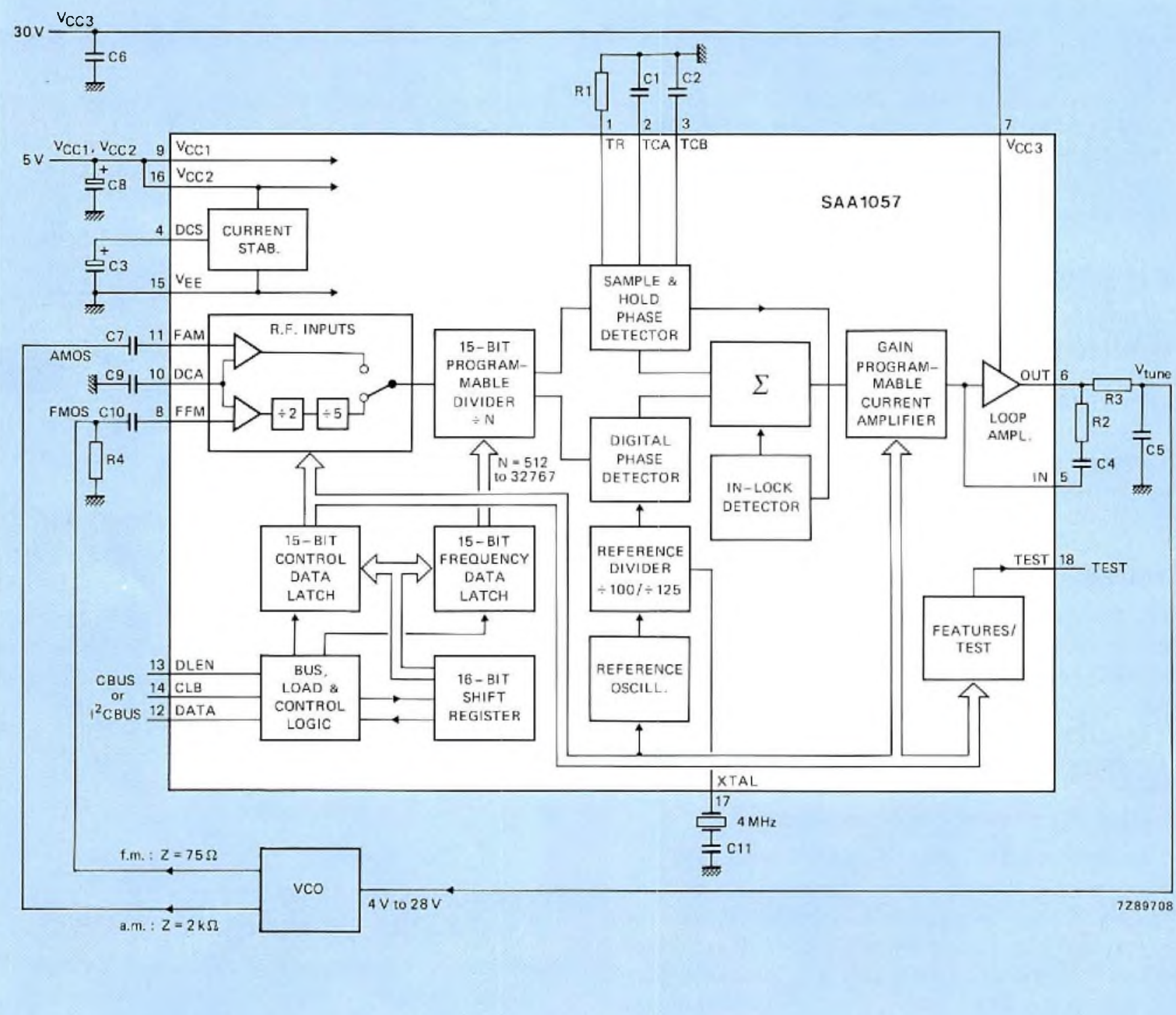


Fig.5 An a.m./f.m. frequency synthesiser using the SAA1057

BASIC APPLICATION OF THE SAA1057

Figure 5 is the circuit diagram of a complete frequency synthesiser using the SAA1057. The functions and values for each component in the diagram are as follows:

ref.	function	value
R_1	defines the current in the analogue phase detector	180 Ω
R_2	loop filter resistor (value depends on VCO)	18 kΩ
R_3	low-pass filter resistor (value depends on VCO)	100 Ω min. 10 kΩ typ.
R_4	matching resistor for 75 Ω f.m. input	180 Ω
C_1	sample capacitor (low leakage type)	2.2 nF typ.
C_2	hold capacitor (low leakage type)	10 nF typ.
C_3	decoupling of internal reference voltage	47 μF
C_4	loop filter capacitor (value depends on VCO)	330 nF typ.
C_5	low-pass filter capacitor, normally located in the tuner (value depends on loop frequency)	100 nF typ.
C_6	power supply filtering	100 nF
C_7	d.c. blocking	1 nF
C_8	power supply filtering	100 μF
C_9	decoupling of r.f. input stages	10 nF
C_{10}	d.c. blocking	11 nF
C11	series capacitor for crystal (value depends on crystal)	33 pF

PERFORMANCE OF THE CIRCUIT FOR F.M.

tuning range	87.5 (88) to 108 MHz
tuning steps	10 kHz or 12.5 kHz
intermediate frequency	10.7 MHz (variable in steps of 10 kHz or 12.5 kHz)
tuning voltage of the VCO	4 to 28 V
VCO gain	0.3 to 3 MHz/V
ref. frequency	32 kHz
prog. divider ratios	9820 (9870) to 11870
time to tune across band	<400 ms
gain of current amplifier	0.3
loop filter time constant	1 ms
r.m.s. ripple on tuning voltage	
noise (20 Hz to 20 kHz)	5 μV
1 kHz	<1 μV (0.3 μV)

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Comparison of the three available types of small and miniature electrolytic capacitor clearly highlights the superiority of solid aluminium types for many professional and industrial applications which were hitherto considered to be the exclusive domain of wet electrolytics or tantalum.

Small electrolytic capacitors compared

E. H. L. J. DEKKER and P. S. FRIEDRICH

Three basic types of small and miniature electrolytic capacitor are available:

- wet aluminium electrolytic (aluminium anode and a liquid electrolyte)
- solid tantalum electrolytic (tantalum anode and a solid electrolyte)
- solid aluminium electrolytic (aluminium anode and a solid electrolyte)

The two types of solid electrolytic are each made in two versions. A compact resin-dipped version with radial leads (such as our 122-series of solid aluminium 'pearls'), and a version in a tubular metal can with axial leads (such as our 123-series of solid aluminium electrolytics). Figure 1 shows a selection of comparable capacitors from the three basic types.

Price is only one factor determining which type of electrolytic is most suitable for a specific purpose. Consideration must also be given to reliability, the permissible range of operating conditions, size (especially important in high density circuits), and any need for current limiting resistors like those which increase the cost of using tantalums for many applications. Making the correct choice therefore requires a comprehensive knowledge of the merits and limitations of all three types. Furthermore, if the correct choice was made some time ago, it may not still be valid if one of the manufacturing technologies has undergone significant improvement. For example, use of a recently developed deeply-etched foil allows our 123-series of axial-lead solid aluminium (SAL) electrolytics to achieve the high CV density which makes them less expensive replacements for tantalums in a wide variety of professional and industrial equipment.

Since the characteristics of SAL electrolytics are less well-known than those of wet electrolytics and tantalums, this article compares the three types of small electrolytic and thereby demonstrates that there are indisputable reasons why SAL electrolytics are ideal for many uses hitherto considered to be the exclusive domain of wet aluminium electrolytics or tantalums.

COMPARISON OF SMALL ELECTROLYTIC CAPACITORS

A survey of the avaiable rated voltage (U_R) and nominal capacitance ranges for the three types of small and miniature electrolytic is given in Fig.2. Their characteristics are compared in the Table (page 160) and in the remaining illustrations.

Measuring conditions for the comparisons

The level of applied voltage can have a marked influence on the results of measurements of the electrical characteristics of electrolytics, especially the leakage current. For example, reducing the test voltage for 40 V solid aluminium electrolytics to 35 V halves the measured leakage current. To obtain comparable results for all three types of electrolytic, a uniform test voltage (for example 35 V) should be used. However, for some capacitors, the maximum specified surge voltage would then be exceeded. The comparison curves in Fig.3, 4 and 5 have therefore been measured with applied voltages equal to the rated voltages. The comparisons are thus restricted to capacitors with identical or very similar rated voltages.

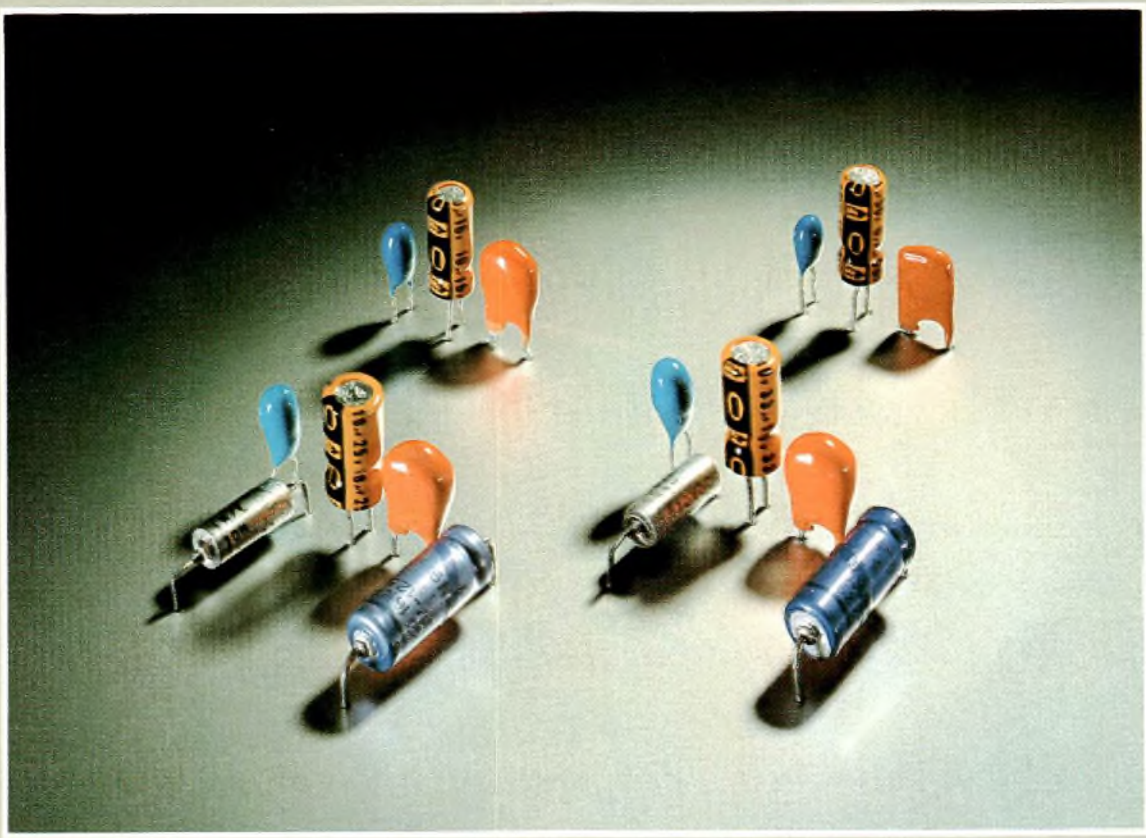


Fig.1 Comparison of small and miniature electrolytic capacitors. The two groups at the rear are, from left to right, solid tantalum, wet aluminium, solid aluminium with radial leads. The two groups in the foreground are, from left to right, solid tantalum with radial leads, solid tantalum with axial leads, wet aluminium, solid aluminium with radial leads, solid aluminium with axial leads. Each group shows capacitors of similar CV product.

Wet aluminium electrolytics

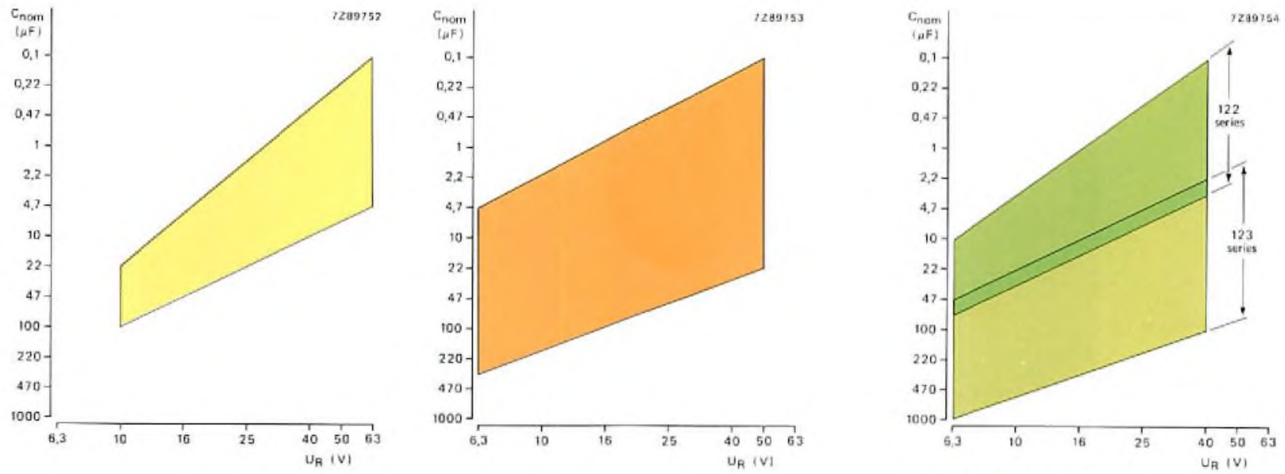
These are inexpensive but generally larger than their 'solid' counterparts. The 'tantalum replacement' types have rated voltages up to 63 V but can only withstand short-term reverse voltage of up to 1 V. If they are used at the higher end of their limited operating temperature range, their life expectancy, which is normally only one-tenth of that of solid electrolytics, is even more severely reduced due to drying-out of the electrolyte. After long storage, their leakage current increases and can only be restored to its original value by post-forming. They are used mainly in equipment for the consumer market.

Solid tantalum electrolytics

These are small, but the high and fluctuating price of tantalum makes them expensive, especially the higher capacitance types. They have a very long life expectancy but their reliability can only equal that of solid aluminium electrolytics if they are specially selected and series resistors are used to limit their charge/discharge rate and thereby avoid the possibility of internal short-circuits due to field crystallisation. Their a.c. performance is poor at high frequencies and is even further degraded by the current limiting resistors which increase their impedance. If they are used at high temperatures, their reliability is clearly

Comparison of electrolytic capacitor characteristics

	wet aluminium (Ta replacement types)	solid tantalum axial leads	solid tantalum radial leads	solid aluminium axial leads	solid aluminium radial leads
• temperature range (°C)	−40 to +85	−55 to +125	−55 to +85	−80 to +175	−80 to +175
• voltage derating above 85 °C	not usable	2/3 of U_R	not usable	not required	40 V types only
• leakage current max specified	0.002 CV	0.02 CV	0.05 CV	0.1 CV	0.05 CV
measured after	1 min	5 min	1 min	1 min	15 s
typical (% of max)	10 to 30	1 to 10	1 to 10	5 to 10	1 to 10
• inrush current	large	small	small	small	small
• allowed ripple current (mA) (33 μF, 10 V) 100 Hz/85 °C	90	60	60 (estimated)	170	170
100 Hz/125 °C	not usable	30	not usable	85	85
• allowed reverse d.c. level	1 V	0.5 V	0.5 V	30% of U_R	30% of U_R
duration	brief	brief	brief	continuous	continuous
• 50/100 Hz voltage allowed (without applied d.c.)	≤1 V	≤0.5 V	≤0.5 V	0.8 U_R	0.8 U_R
• charge/discharge limiting	not required	3 Ω/V	3 Ω/V	not required	not required
• effect of transients (peaks)	none	risk of s/c	risk of s/c	none	none
• allowed surge voltage (forming V/rated V)	1.15 to 1.3 U_R	1.15 to 1.3 U_R	1.15 to 1.3 U_R	1.15 U_R	1.15 U_R
• effect of increasing V on C	C rises ≤3%	C drops ≤7%	C drops ≤7%	C rises ≤1.5%	C rises ≤1.5%
• stability of parameters with U_R applied at 85 °C	good upto 2000 hr	good	good	good	good
at 125 °C	not usable	risk of s/c	not usable	good ≤175 °C	good ≤175 °C
• guaranteed life (h) at 85 °C	2000	2000	1000	5000	5000
at 125 °C	not usable	2000	not usable	5000	2000
• life expectancy (kh) at 85 °C	5 to 10	>50	>25	>100	>100
at 125 °C	not usable	>25	not usable	>50	>50
• failure mechanism	drying-out of electrolyte	field crystallisation	field crystallisation	none known	none known
• type of failure	degradation of parameters, e.g. C, Z, tan δ	short-circuit	short-circuit	degradation of I_{leak}	degradation I_{leak}
• effect of temp. derating on life and failure rate	significant	significant	significant	negligible	negligible
• effect of voltage derating on life and failure rate	negligible	significant	significant	slight	slight
• failure rate (rated conditions, 60% confidence)	10^{-6}/h ≤2 kh	10^{-5}/h ≤2 kh	10^{-5}/h ≤2 kh	10^{-8}/h ≤20 kh	2 × 10^{-7}/h ≤10 kh
• field failure rate	10^{-7}/h	10^{-8}/h	10^{-7}/h	10^{-9}/h	10^{-8}/h
• mechanical shock	≤40 g	≤30 g (MIL: 100 g)	≤30 g	≤40 g (MIL: 10 000 g)	≤30 g
• vibration	10 g/500 Hz	15 g/2 kHz (MIL: 20 g/2 kHz)	15 g/2 kHz	50 g/2 kHz	15 g/2 kHz
• IEC stability category during humidity tests (40 °C, 90-95% RH, no load)	56 days	56 days	21 days	56 days	56 days
• cleaning solvent proof	no	yes	some	yes	yes



(a) wet aluminium electrolytics (tantalum replacement types)

(b) solid tantalum electrolytics (radial and axial leads)

(c) solid aluminium electrolytics (radial and axial leads)

Fig.2 Electrolytic capacitor ranges

inferior to that of solid aluminium electrolytics unless the applied voltage is derated. They can only withstand short-term reverse voltage up to 0.5 V and are mainly used in equipment for the industrial and professional market.

Solid aluminium electrolytics

These are very slightly larger than comparable tantalums but their lead pitch allows them to fit into the same print-board holes. Due to the low price and plentiful supply of aluminium, their price is lower and more stable than that of equivalent tantalums, but slightly higher than that of comparable wet aluminium electrolytics. Their rated voltage remains valid throughout their very wide operating temperature range and they can withstand continuous reverse voltage up to one-third of their rated voltage, even at temperatures as high as 125 °C. Unlike the other two types, they have no inherent wear-out or catastrophic failure mechanism so their characteristics remain constant throughout a very long life which is not shortened if they are operated at the maximum permitted temperature. Unlike tantalums, their charge/discharge rate is unrestricted, so additional current-limiting resistors are not required. Their impedance falls more steeply with increasing frequency (up to their resonant frequency) than that of the other types and their resonant frequency is higher. Their minimum impedance is thus lower, as clearly shown by Fig.3. This, combined with their ability to withstand 50/100 Hz sinusoidal voltage with a level of up to 80% of their rated d.c. voltage, makes them suitable for many uses including a.c. coupling, filtering, smoothing and energy storage. Present applications include telecommunications, space projects and nuclear power systems. Their small size and low cost, combined with their electrical and mechanical robustness also makes them ideal for the automotive industry. Since they are subject to continuing improvement, it is anticipated that their future applications will be even more extensive and diverse.

DELIVERED QUALITY OF OUR SAL ELECTROLYTICS

Quality control is not limited to a rigorous final inspection but is a complete quality assurance programme integrated into the production process. Inspections are made at all stages of manufacture. The final inspection is a 100% check of the main electrical parameters followed by statistical tests to determine the level of quality immediately prior to delivery. Since defective batches are rejected, the quality level of the components delivered to our customers is very high. Information from customers is fed back to the factory and, if necessary, used as a basis for modification of the manufacturing process. These changes, combined with the basic quality assurance programme, ensure that the very high quality level is maintained. This is confirmed by a report from a large manufacturer of telecommunications equipment which states that the failure rate for 122-series solid aliminium 'pearls' received from our factory is less than 60 ppm (mechanical and electrical failures combined).

THE CASE FOR SOLID ALUMINIUM ELECTROLYTICS

The foregoing comparison shows that the performance of SAL electrolytics far exceeds that of wet aluminium electrolytics and is comparable, and in some respects superior, to that of tantalums. In view of this, combined with the high price of tantalum and the plentiful supply of aluminium, there is a clear case for a more widespread use of solid aluminium electrolytics.

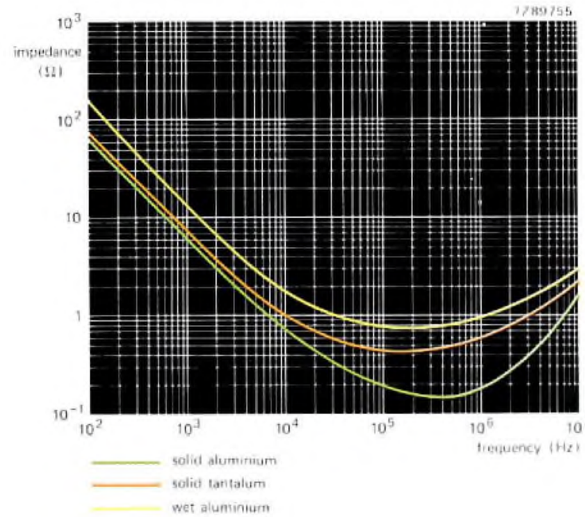


Fig.3 Impedance as a function of frequency with capacitor type as a parameter. All capacitors are 33 μF, 10 V

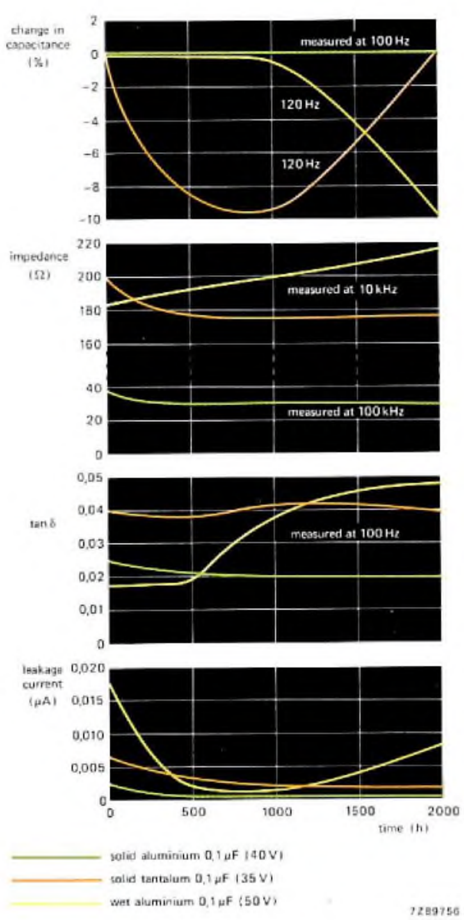


Fig.4 Stability of the main electrical parameters of the three types of electrolytic as a function of time. The curves are for a 0.1 μF capacitor (rated voltage = 35 to 50 V) and are measured at a temperature of 85 °C with the rated voltage applied

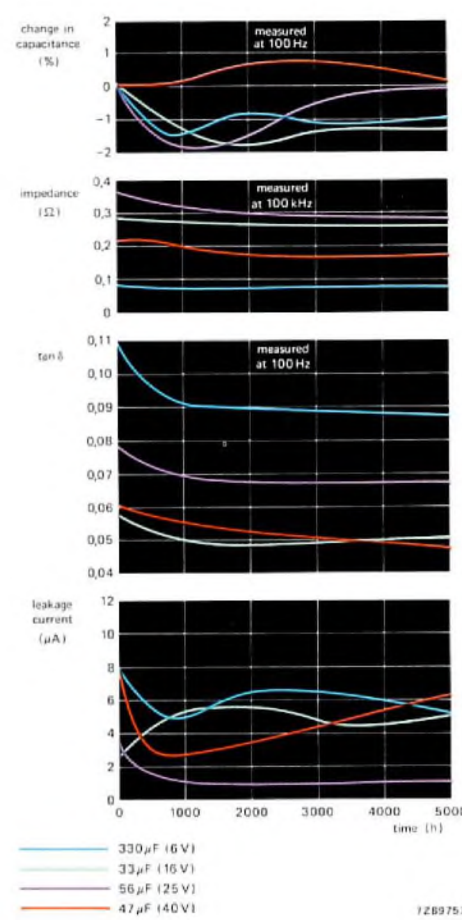


Fig.5 Stability of the main electrical parameters of axial-lead solid aluminium electrolytics as a function of time with capacitor value as a parameter. The curves are measured at a temperature of 175 °C with the rated voltage applied

Designers of video display units can profit from hardware and circuit technology developed for tv receivers. The full-page display described here uses tv transformers and ICs; the deflection yoke is wound by the pin-indexed method developed for colour tv, and the c.r.t. is a refinement of a balck-and-white picture tube.

Full-page display using television components

A full-page data-graphic display, with upwards of 5000 characters on an upright screen, puts severe demands on deflection, resolution, and raster geometry(Ref.). To meet those demands it has been usual to specify expensive monitor tubes and deflection components designed expressly for the purpose. Now, however, that is no longer necessary. The laboratory prototype shown in the photo more than satisfies the usual specifications for full-page display, and it does so with parts mainly derived from normal television practice.

The cathode-ray tube is a refinement of a black-and-white television tube, adapted to the requirements of data-graphic display. Its 38 cm diagonal screen is capable of displaying 10^6 pixels. The deflection assembly, though designed for the adapted tube, is constructed according to principles first developed for colour television. In addition, the line-driver, line-output, and picture-shift transformers are standard television parts, unaltered, as are the two integrated circuits used in the line and field timebase circuitry.

The use of television-derived technology benefits not only the cost and reliability of the display but also its size. Thanks to the 110° deflection angle, the 38 cm tube occupies no more depth than a 31 cm tube with 90° deflection. This and its television-style push-through mounting afford the designer more than ordinary latitude in the styling of an attractive, slim-line enclosure for a production version of the display.

Laboratory prototype full-page display. 110° deflection and push-through tube mounting afford generous styling latitude for a production version

PERFORMANCE SPECIFICATIONS

Table 1 lists the main specifications. The line and field deflection frequencies given there are for a non-interlaced raster of 1066 lines, just over 1000 of which are displayed. Non-interlacing precludes jitter, and at the 60 Hz field deflection frequency flicker is not a problem, even at the high brightness used with inverse video. Higher field frequencies may be used, as may an interlaced raster, but for thermal and other reasons the line deflection frequency should not exceed 64 kHz.

The prototype shown in the photo incorporates the elements to the right of the dashed line in the block diagram (Fig.1). The amplifier to the left of the line is required for video inputs at lower than TTL level. The power supply used with the prototype was a mains-isolated switched-mode supply operating at 30 kHz and stabilised against mains variations between 185 V and 225 V r.m.s.

The excellent raster geometry of the display is due largely to the AT1039/00 deflection unit. Both the line and field deflection coils are saddle-type and are wound by the pin-indexed method in which each coil is divided into groups of turns that can be positioned to close tolerances. Since its introduction more than ten years ago to solve some of the electron-optical problems peculiar to 110° colour television, pin-indexed winding has gained wide acceptance and has opened the way to a number of improvements in deflection coil design. The AT1039/00 embodies one of the latest of these: 'flangeless' winding, in which the narrow end of each coil is not bent up into the usual flange but lies flat along the neck of the tube and is enclosed by the ferrite ring. The shape of the coils permits even closer control of their turns distribution, and their total enclosure by the ferrite ring prevents radiation of line and field frequencies. In a datagraphic display it also prevents ringing effects and the vertical bars that are often seen when a white page is displayed.

The electrical characteristics of the AT1039/00 coils are:
line deflection coil (parallel connected)

$L = 0.23$ mH $\qquad R = 0.48\ \Omega$

$I = 6.18$ A for edge-to-edge scan at 17 kV e.h.t.

field deflection coil (series connected)

$L = 9.1$ mH $\qquad R = 10.1\ \Omega$

$I = 1.3$ A for edge-to-edge scan at 17 kV e.h.t.

Built-in adjustable permanent magnets correct residual raster errors. No raster correction circuitry is required.

The AT1039/00 is for displays with upright page presentation. A unit with equivalent characteristics for horizontal page presentation is the AT1039/01; owing to the longer line scan, the line frequency with this unit is limited to about 32 kHz.

TABLE 1
Brief specification of the display

cathode-ray tube	M38-328
deflection unit	AT1039/00
line deflection frequency	64 kHz
line flyback time	3 μs
field deflection frequency	60 Hz
field flyback time	0.6 ms
e.h.t. voltage	17 kV
line linearity error	⩽ 3%
field linearity error	⩽ 3%
raster size variation ('breathing') for 0 to 100 μA beam current variation	1%
video bandwidth at 30 V input	60 MHz
sync inputs	positive-going composite sync or separate line and field sync at TTL level
supply	330 mA at 105 V 90 mA at 12 V

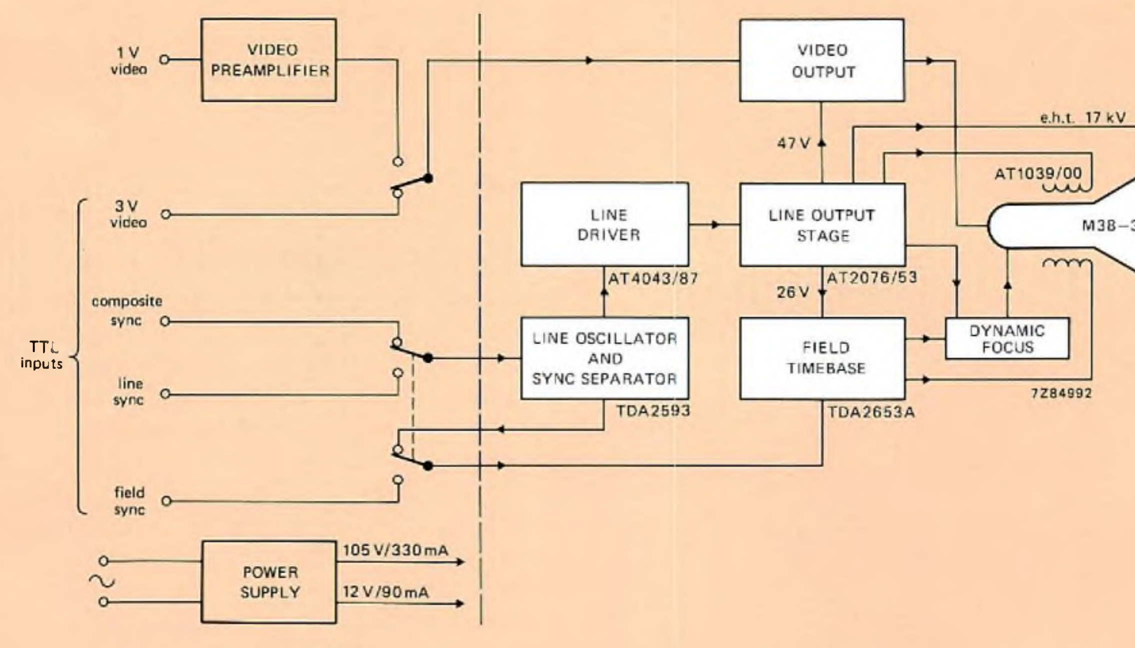


Fig.1 The photographed prototype includes the elements to the right of the dashed line

LINE OSCILLATOR AND LINE DRIVER

The TDA2593 in Fig.2 is a television IC that combines the functions of coincidence detector, sync separator, and line oscillator. Some functions provided for television are not used: for instance, a phase comparison loop for synchronising the oscillator voltage with the flyback pulse; the storage time of the line output transistor is sufficiently stable that there is no significant raster drift with respect to the video information during warm-up.

The AT4043/87 transformer that couples the line driver transistor to the line output stage is also a standard television part, chosen for the tight coupling between its windings. To minimise the effects of outside interference, particularly picture-tube flashover, the line driver is placed as close as possible to the line output transistor on the printed wiring board.

TABLE 2
Line driver stage operating conditions

supply voltage	105 V
supply current	40 mA
transformer voltage	63 V
transformer voltage ripple	3 V
peak collector voltage	150 V
peak collector current	120 mA

LINE OUTPUT STAGE

The line output stage (Fig.3) uses a parallel efficiency diode, as in television practice, with a 2.2 nF capacitor in parallel with the diode to give 3 μs flyback time. The small leakage inductance of the line driver transformer, aided by the speed-up capacitor and resistor in the base circuit, ensures suitable storage and zener times for correct turn-off of the transistor; the base-emitter resistor damps ringing of the base drive waveform.

The 220 nF capacitor in series with the AT1039/00 line deflection coil provides S correction. The AT4043/29 shift transformer and the potentiometer associated with it enable the raster to be centred laterally; the direction of shift can be reversed by reversing the connection of the flying lead shown dotted in the drawing. Since at the 64 kHz line fre-

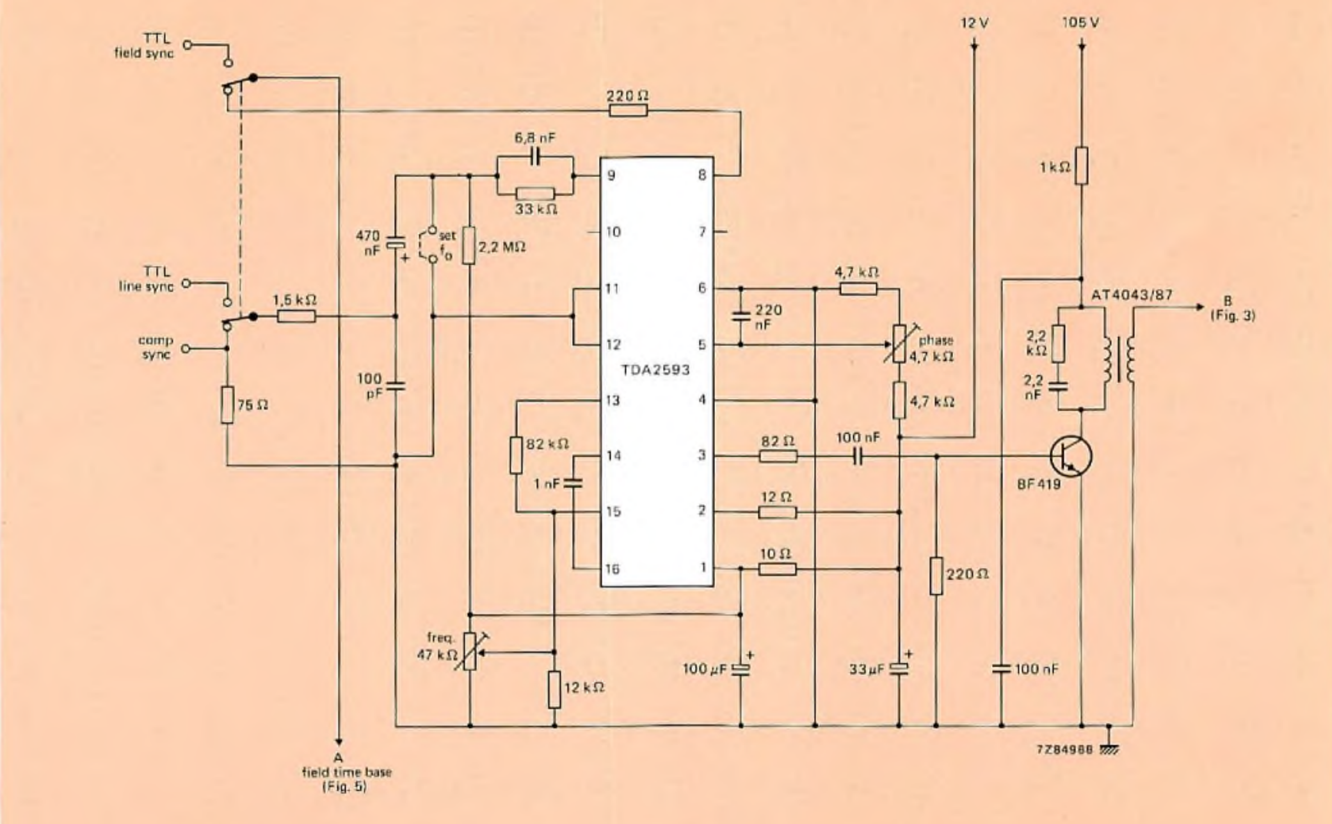


Fig.2. Line oscillator and line driver

quency the resistance of the deflection coil is negligible compared with its impedance, no asymmetric linearity correction is required. Figure 4 shows the line-scan linearity.

Like a television receiver, the display obtains its e.h.t. and housekeeping voltages from a transformer driven by the line output stage. It is a type AT2076/53 'diode split' transformer intended for large-screen colour sets with 25 kV e.h.t. and 15.625 kHz line frequency. Although working here at slightly more than four times that frequency, it is quite capable of supplying 17 kV at the lower current required by the smaller, single-gun tube. Because of the very tight coupling between its primary and e.h.t. windings, the AT2076/53 can operate over a wide frequency range without distorting the line-scan waveforms. A harmonically tuned line output transformer, as used in some television receivers, would not be suitable.

The e.h.t. winding is in three layers connected by built-in diodes. The cable connecting it to the picture-tube anode incorporates a bleeder resistor to prevent raster width variation ('breathing') at low beam current.

Other windings supply 26 V for the field timebase, 47 V for the video amplifier, and −140 V for brightness (grid 1) and focus (grid 3) control potentiometers. Three turns of insulated wire on the core limb opposite the primary winding supply the heater.

The grid 2 potential is obtained by peak rectification of the line output transistor collector voltage. To ensure the best possible resolution, the maximum allowable grid 2 to grid 1 voltage is used: 700 V. The upper end of the voltage divider containing the focus control potentiometer is also connected to this voltage to provide for positive as well as negative control voltages.

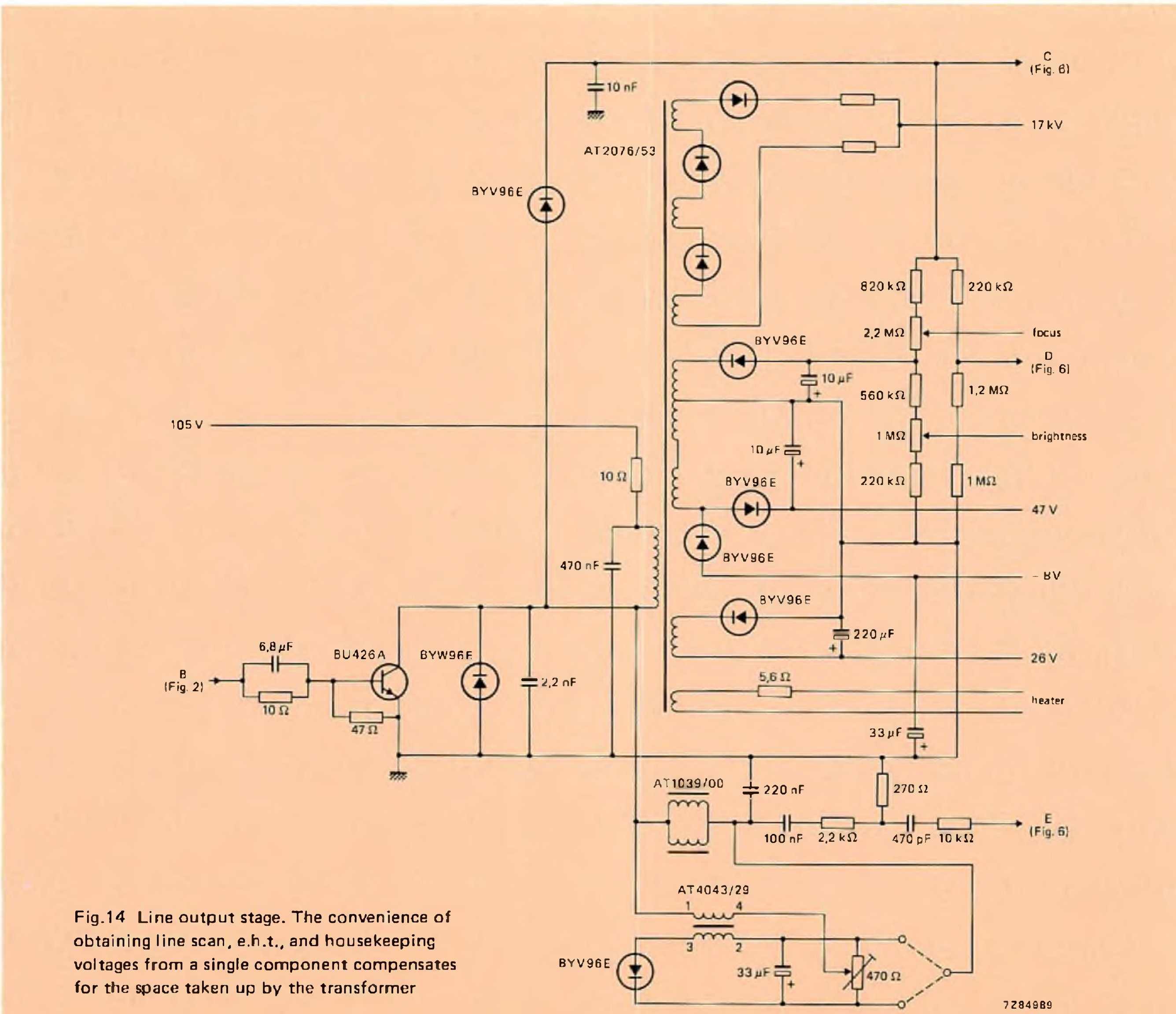


Fig.14 Line output stage. The convenience of obtaining line scan, e.h.t., and housekeeping voltages from a single component compensates for the space taken up by the transformer

TABLE 3
Line output stage operating conditions

supply voltage	105 V
supply current	290 mA
supply voltage ripple, 64 kHz	1 V
100 Hz	0.1 V
peak collector current	3.2 A
peak collector voltage	720 V
base current at end of scan	0.6 A
reverse base-emitter voltage	3 V
deflection coil current, peak-to-peak	6 A
scan-width/screen-width ratio	87 %

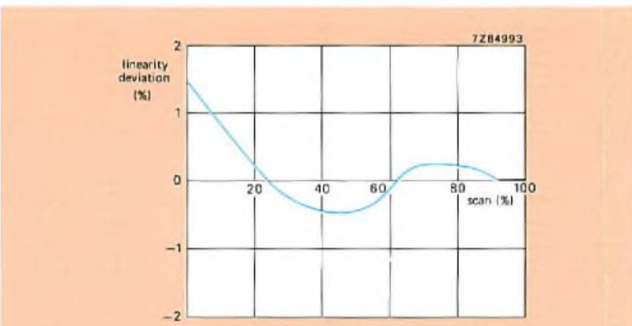


Fig.4 Line-scan linearity

FIELD TIMEBASE AND WAVE-SHAPING CIRCUIT

The TDA2653A field timebase circuit (Fig.5) is another television IC. It can accept separate sync pulses at TTL level or pulses obtained from a composite sync input by the sync separator of the TDA2593 in the line oscillator circuit. For interlaced operation, separate line and field sync pulses are recommended.

The TDA2653A has a class B output stage and thermal and short-circuit protection.

The wave-shaping circuit integrates the sawtooth voltage across the feedback resistor R to produce a positive-going field-frequency parabola for dynamic focus.

TABLE 4
Field timebase operating conditions

supply voltage	26 V
supply current	232 mA
deflection coil current, p-p	1.25 A
flyback time	0.6 ms
pin 5 voltage	25 V
pin 5 voltage ripple	2 V

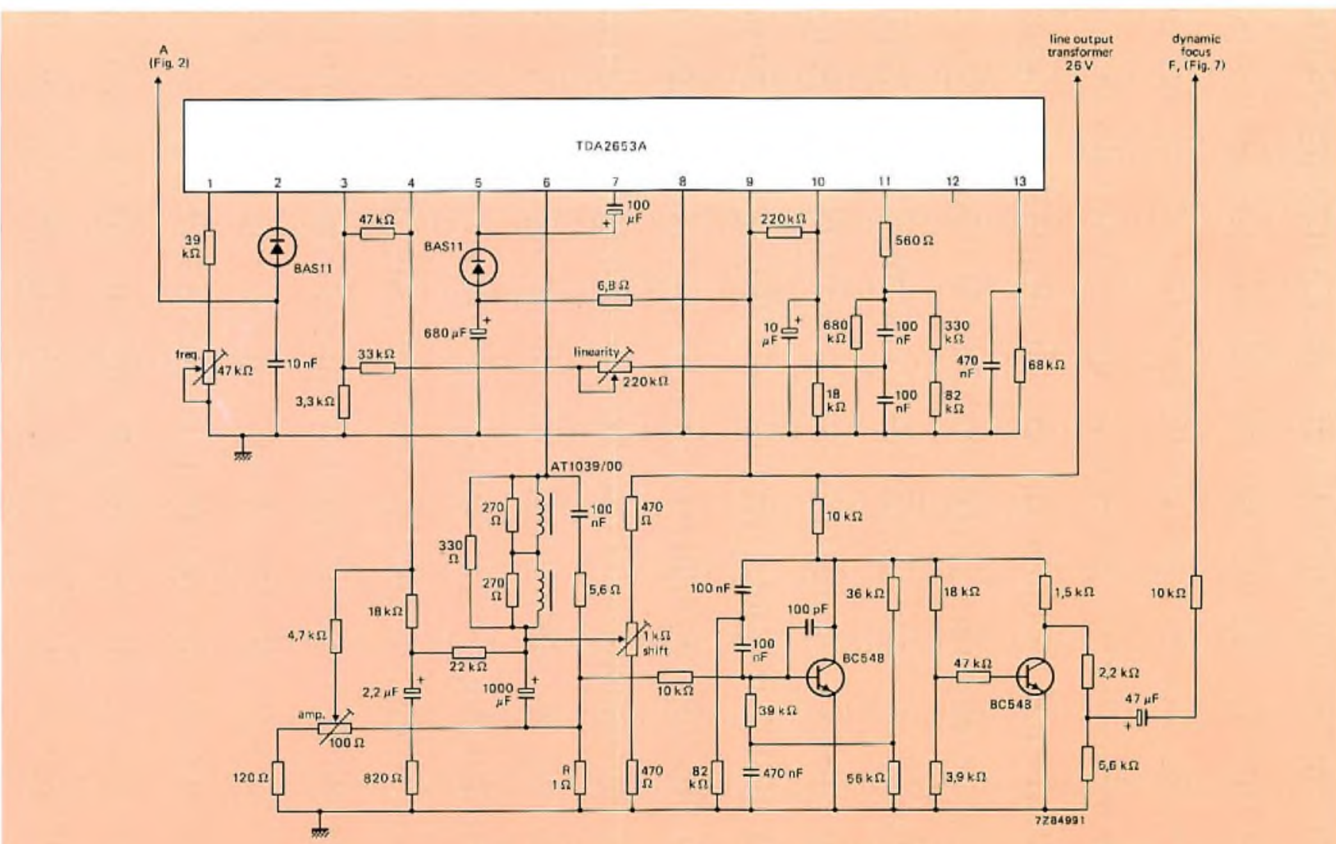


Fig.5 Field timebase and wave-shaping circuit

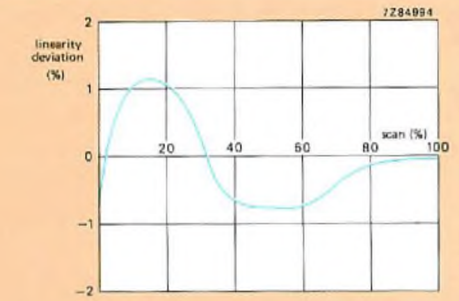


Fig.6 Field-scan linearity

VIDEO AMPLIFIER AND DYNAMIC FOCUS

The video amplifier (Fig.6) requires an input of about 4 V for a maximum output of 30 V. The small load resistance and peaking coil in the output transistor collector circuit, together with some compensation in the emitter circuit, ensure a bandwidth of 60 MHz. No provision is made for blanking; if it is required, it can be introduced in the logic circuits, as part of the video information.

Although the design of the deflection coils minimises deflection defocusing, there is some residual error. To correct it, line and field-frequency parabolas of about 250 V each are superimposed on the grid 4 focus voltage. The dynamic focus circuit uses two transistors in series to accommodate the 500 V excursion that occurs when the field-frequency parabola is maximum. Because the tube has a unipotential electron gun, the focus characteristic is fairly flat and it is unnecessary to provide for separate adjustment of the dynamic focus.

TABLE 5
Dynamic focus operating conditions

supply voltage	730 V
supply current	3.2 mA
input voltage, line frequency, p-p	5 V
input voltage, field frequency, p-p	5 V
output voltage, line frequency, p-p	250 V
output voltage, field frequency, p-p	250 V

CONSTRUCTION

Complete details of construction, setting up, and adjustment, together with a parts list and oscillograms of waveforms at all significant points, are available as a separate publication: Technical Publication M81-0084 from Mullard Ltd., Mitcham, England; or Technical Publication 026 from N.V. Philips', Electronic Components and Materials Division, Eindhoven, The Netherlands.

REFERENCE

'Cathode-ray tube requirements for data-graphic displays' Electronic Components and Applications, Vol. 4 No. 1, Nov. 1981.

ACKNOWLEDGEMENT

This article is based on the work of D. J. Gent of Mullard Application Laboratory, Mitcham, as more fully reported in the publication cited under 'Construction'.

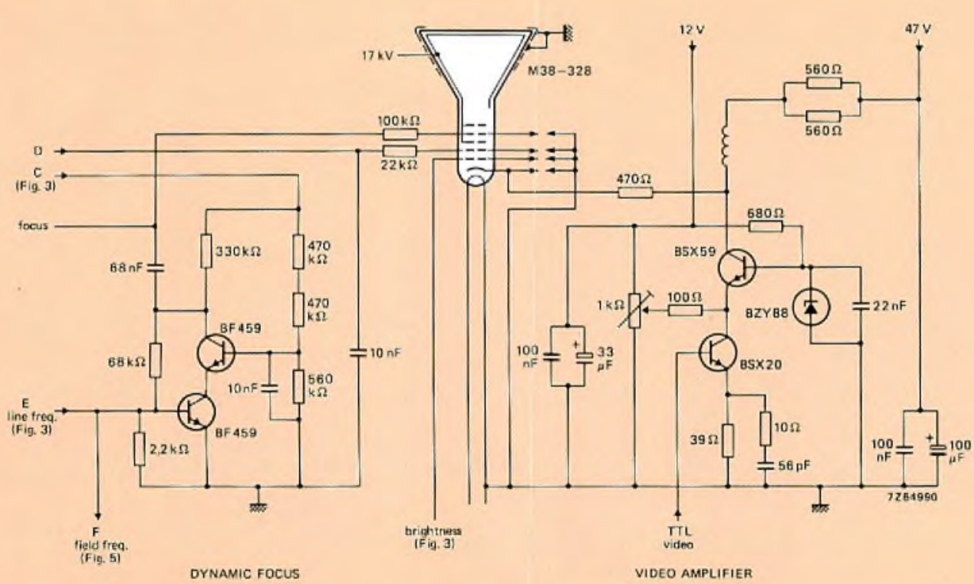


Fig.7 Video amplifier and dynamic focus

The magneto-optical properties of bismuth iron garnet make it possible to construct a fast, high-resolution light-switching array for use in optical line printers.

Light-switching array for high-resolution pattern generation

B. HILL and K. P. SCHMIDT

LiSA 512 is an electronically controlled light-switching array developed for high quality pattern generation in optical (electro-photographic) line printers. It possesses all the advantages inherent to optical printers – high spatial resolution, very high speed, noiseless operation – plus several important features that give it an edge over other optical printing systems.

Take laser printers for example. These are currently the most widely used optical printers, yet by their very nature, laser printers suffer from some rather severe limitations. For instance, solid-state lasers emit radiation predominantly in the near-infrared region, yet the photoconductive drum on which the latent image is formed in an optical printer, responds most readily to blue/green light. This is not a problem for gas lasers, but the added complexity of these devices means that they are rather costly. Moreover, laser printers need a mechanical deflection system, adding greatly to complexity and costs.

LED-based systems also emit radiation predominantly in the near-infrared/red region, but at least they don't need a mechanical deflection system. And they have the advantage of being IC compatible. They do, however, have several disadvantages of their own. For instance, a typical system may have 2000 or more closely spaced elements that have to be switched at high speed. So the system may have to dissipate a substantial amount of heat, say 20 W or more, which for such small systems makes cooling difficult. In addition, the current state of LED technology limits the degree of integration to about 100 elements/chip, so LED-based systems can be costly.

LiSA on the other hand, has none of these disadvantages. It has no moving parts, it operates from any standard light source, and it has no cooling problems. Moreover, with LiSA, the degree of integration is already five times that of LED technology, and future developments will probably increase that margin.

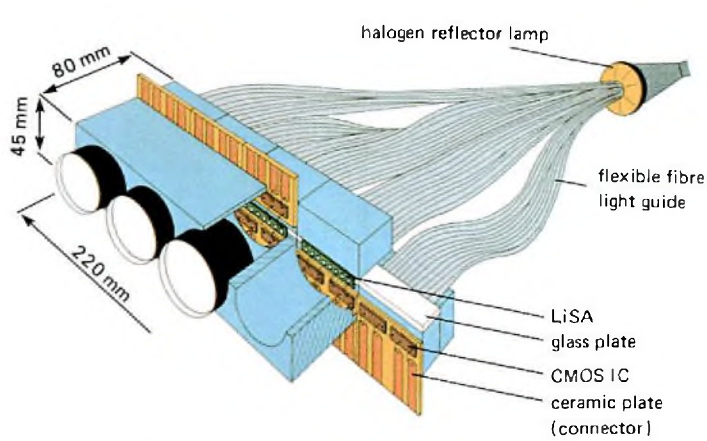


Fig.1 LiSA (Light Switching Array) for high resolution pattern generators has no moving parts and can operate from a simple light source

OPERATING PRINCIPLES

The LiSA chip consists of a magneto-optic bismuth iron garnet film, grown epitaxially on a gadolinium gallium garnet substrate and etched into separate cells (Fig.2). This film magnetises spontaneously along an axis normal to the plane of the substrate. Moreover, it exhibits the Faraday effect, so that plane polarised light passing through the cells is rotated, the sense of rotation depending on the direction of magnetisation. In combination with an optical polariser and analyser therefore (Fig.3), the cells effectively function as light valves controlled by the direction of magnetisation.

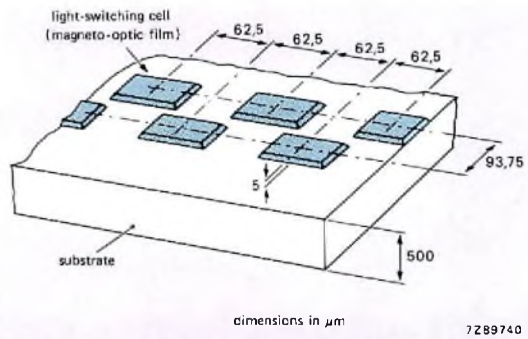


Fig.2 The LiSA chip consists of a bismuth iron garnet film, grown epitaxially on a substituted gadolinium gallium garnet substrate and etched into separate light switching cells. Each cell is 65 X 65 μm^2, smaller than the smallest possible magnetic domain (about 100 to 200 μm diameter). An opaque metallic film covers the space between the cells

Cell switching

LiSA uses a combination of heat and magnetic field pulses to switch the cells. To understand how this works consider the magnetic behaviour of bismuth iron garnet in more detail.

As we said, the film magnetises spontaneously in a direction normal to the plane of the substrate. In an extended film, the magnetic domains will align themselves either parallel or anti-parallel to the normal, and, depending on the prevailing magnetic field, their alignments are free to change at random. So the film will have a random arrangement of parallel and anti-parallel domains.

But if the film is etched into cells that are smaller than the smallest possible domain, each cell will contain only one domain and its magnetisation will be uniform. Not only that, but up to about 70 °C, the cells will resist changes in their magnetisation and will remain stable against even high magnetic fields.

We therefore have a ready made switching mechanism. By applying a heat pulse to the cells (or a portion of the cells), and following this with a magnetic field pulse, the cells can be switched from parallel to anti-parallel (and vice versa).

Figure 4 shows the basic setup for switching the cells. A thin film resistive layer deposited on top of each cell supplies the heat pulse, and a separate winding around all the cells supplies the magnetic field pulse.

Figure 5 gives the pulse sequence. The heat pulse heats a corner of the cell above 100 °C. This creates a region of magnetic instability that propagates throughout the cell, allowing the magnetic field pulse to switch the cell into the new direction. Within about 10 μs from the end of the heat pulse, the cell cools and stabilises in the new magnetic direction.

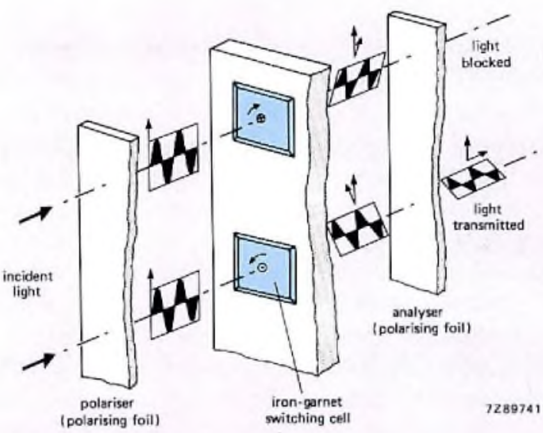


Fig.3 In combination with an optical polariser and analyser, the iron garnet cells act as light valves controlled by the direction of magnetisation

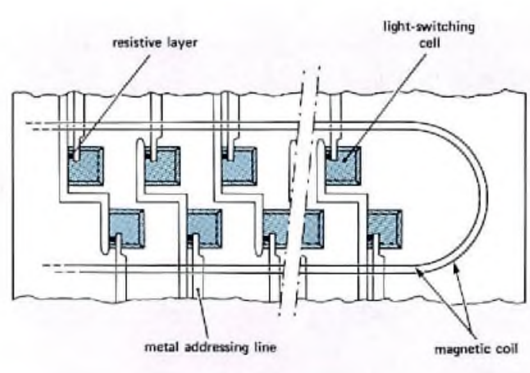


Fig.4 Basic setup for switching the cells. A thin film resistive layer on top of each cell supplies a heat pulse, heating the cell corner to about 100 °C. This sets up a region of magnetic instability, and the cell's magnetisation can then be reversed by an external magnetic field

Transmission characteristics

The transmission coefficient of a LiSA unit depends on the angle through which the plane polarised light is rotated. For a wavelength of 550 nm this angle is about ±20°, and it decreases towards the red. Apart from the magneto-optic effect, bismuth iron garnet also exhibits optical absorption

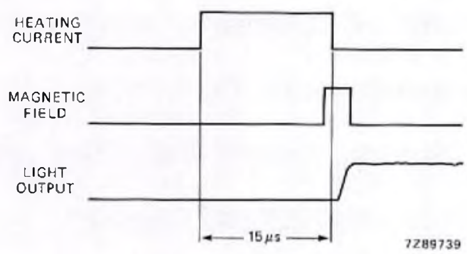


Fig.5 Pulse sequence for cell switching

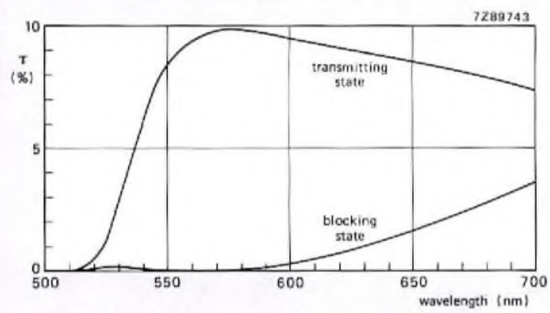


Fig.6 Transmission coefficient (τ) of a LiSA unit as a function of wavelength (with the polariser and analyser inclined at 15° to each other)

that decreases towards the red. Figure 6 shows how the overall transmission coefficient varies with wavelength for both transmitting and blocking states.

It should be noted that the transmission coefficient varies with wavelength for both states of magnetisation, and that total blocking is only possible for monochromatic light of a specific wavelength.

LiSA 512 UNIT

Figure 7 shows the LiSA 512 unit. It contains a LiSA chip of 512 switching cells arranged in a meander array. The chip is cemented on a glass support between two additional metal supports that also function as heatsinks. A sandwich of glass plates perpendicular to the chip forms the input window, and between this and the chip are located the polariser and magnetic coil. The units normally come without an analyser (which must be cemented to the glass support) but if the application is specified, a suitable analyser can be fitted before delivery.

The meander array of cells in the LiSA chip improves character definition and eases the logic functions. The chip is effectively divided into upper and lower banks of cells, only one bank being operative at any given time. When LiSA is installed in an optical printer, therefore, the bank switching rate must be matched to the speed of the photo-conductive drum to form interlacing linear patterns on the drum. These patterns overlap slightly to minimise discontinuities in the printed line.

Besides the optical hardware, the LiSA 512 unit comes with four 64-bit MOS shift register ICs, two on each side of the chip. The outputs of these shift registers each connect with two switching cells, one in the upper bank and one in the lower. And since only one bank is selected at any time, the total of 256 bits available in the shift registers is just sufficient to operate the 512 elements of the LiSA chip.

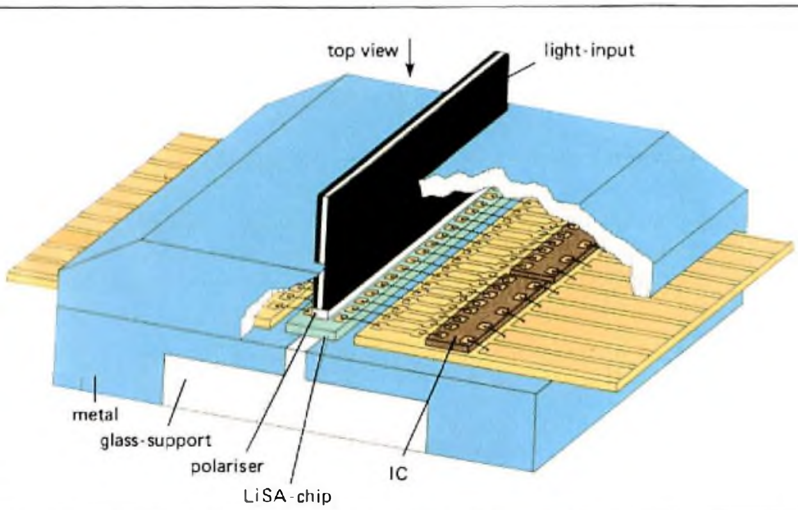


Fig.7 Cutaway drawing of a LiSA 512 unit. This contains 512 switching cells arranged in a meander array, and four 64-bit CMOS shift registers. Light enters via the input window at the top of the unit, and exits via the glass support at the bottom. The analyser (not shown) is cemented to the glass support

The table below gives the specifications of LiSA 512, and Fig.8 gives an example of a pattern generated by a single LiSA 512 unit. The unit can accept input data at the rate of one Mbit/s, which gives a maximum pattern generating speed of about 2000 patterns/s. This is comparable with the speed of the latest computer line printers, but the quality is much higher.

In a practical printing system, several LiSA 512 units would be installed side-by-side, the exact number depending of course on the line width required. The units would need a fibre-optically coupled light source, and a lens system (as shown in Fig.1) to focus the pattern onto the photo-conductive drum.

In the LiSA system all switching operations are controlled by hybrid ICs external to the LiSA unit, so cooling is never a problem.

Specifications of LiSA 512

number of light valves	512
optical aperture of a light valve	65 X 65 μm^2
pitch of light valves	62.5 μm
"optical" length	32 mm
total length	42 mm
optical switching time	<2 μs
thermoelectronic switching delay	18 μs
cycle time (typical)	500 μs
electronic input data rate (typical)	1 Mbit/s
drive voltage	12 V/30 V
power requirements, at 2000 patterns per second	ca. 5 W

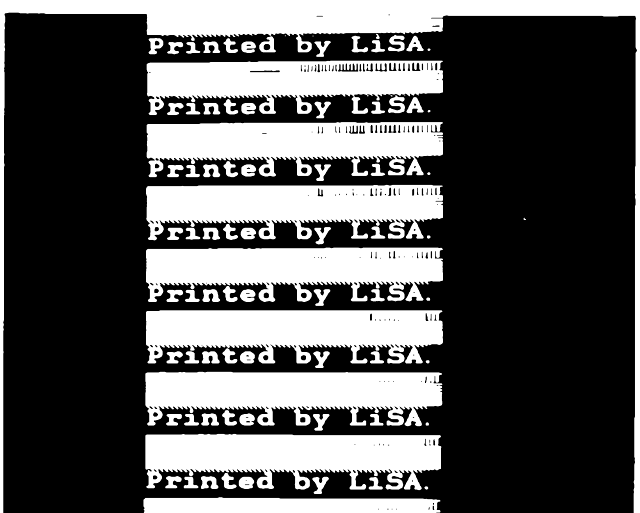


Fig.8 Pattern generated by a single (experimental) LiSA unit generating 12 dots/mm

EXTERNAL CONTROL AND INFORMATION INPUT

Figure 9 shows the control inputs to the LiSA unit, and Fig.10 shows how the external control functions are divided. The external control comprises three circuits: the control pulse generator, the heat switch and the magnetic coil switch. The first two can be assembled on one board, so only two control boards are needed. These supply all voltages and signals needed to drive LiSA, and no additional connections need be made to the LiSA unit itself.

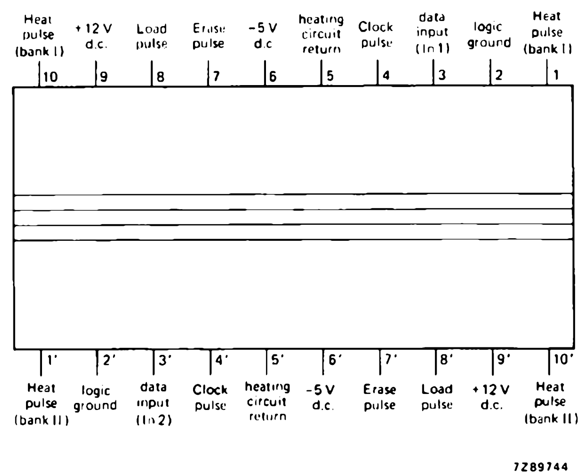


Fig.9 Control inputs of the LiSA 512 unit (view from the top)

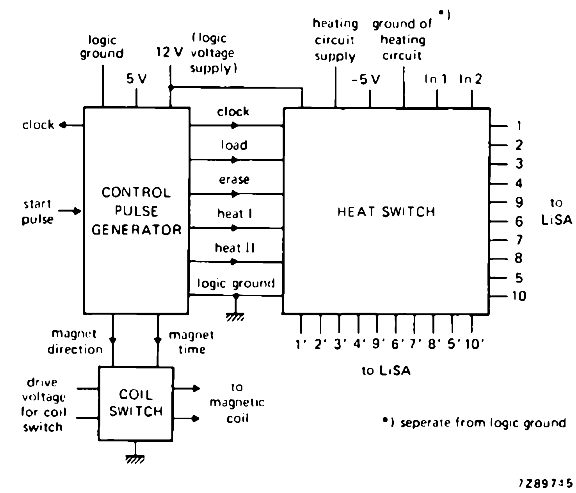


Fig.10 Division of external control functions

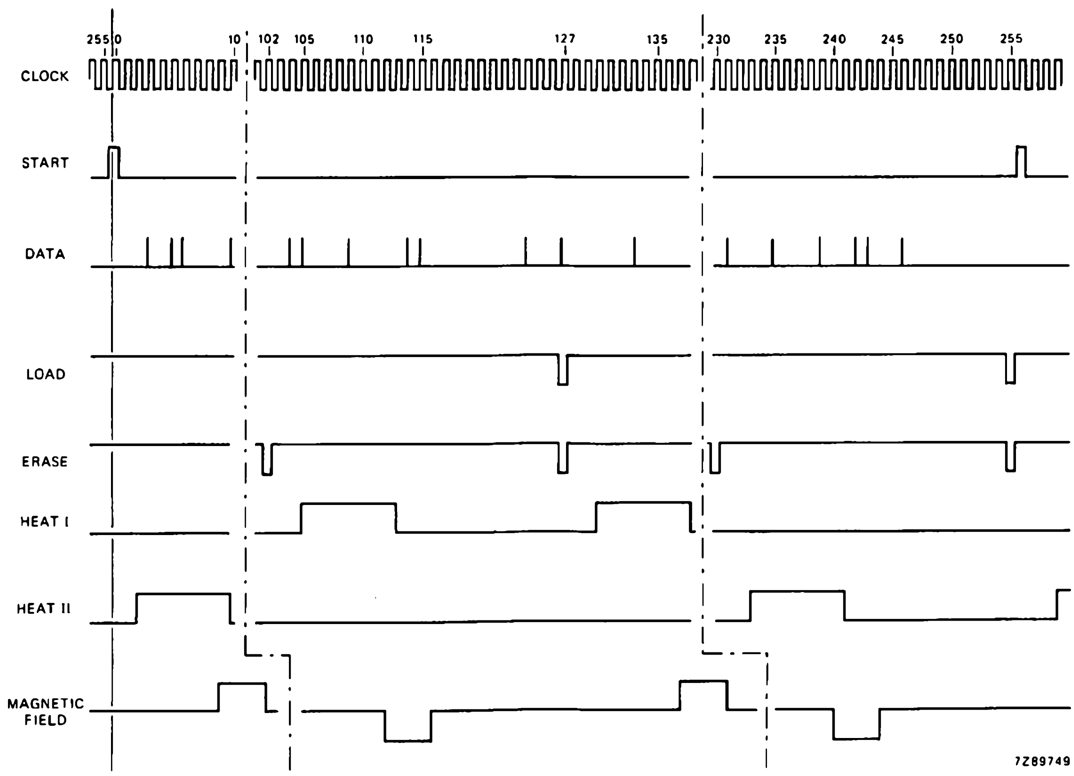


Fig.11 Pulse sequence for the LiSA 512 unit. Data input frequency 500 kbits/s

Control-pulse sequence

To understand how the system works, consider the pulse diagram shown in Fig.11, intended for a data frequency of 500 kbits/s. This diagram is best considered with reference to Fig.12, which shows the switching cell layout and signal input in more detail.

The Start pulse must be generated externally, e.g. by the external shift register that supplies the digital information to the LiSA unit. This pulse resets the internal clock in the control pulse generator and starts it counting. The Start pulse also activates the bank selector within the control pulse generator, which then selects the bank of cells (upper or lower) to be switched.

A reset cycle then starts. This sets all the switching cells into the blocking state, in preparation for receiving fresh information. First an Erase pulse is applied to the shift registers, making all output stages conductive. This is followed by a Heat pulse, along say line I (Fig.12), which heats all the cells in the upper bank. Finally, the Magnetic Field pulse, which starts just before the end of the Heat pulse, switches all the cells in the upper bank into the blocking state.

During the reset cycle, the first 256 bits (of the line to be printed) are read into the shift registers. By the 127th count this process is complete and the upper bank of cells can be loaded with the line pattern. To do this, a Load pulse is applied to the shift registers at the same time as the Erase pulse, making some of the output stages – those connected to cells that must be switched to the transmitting mode – conductive. The Heat pulse is then again applied to line I, followed by the Magnetic Field pulse (in the opposite direction to the earlier pulse), switching the required cells in the upper bank into the transmitting state.

From the 230th count on, this whole process is repeated for the lower bank of cells.

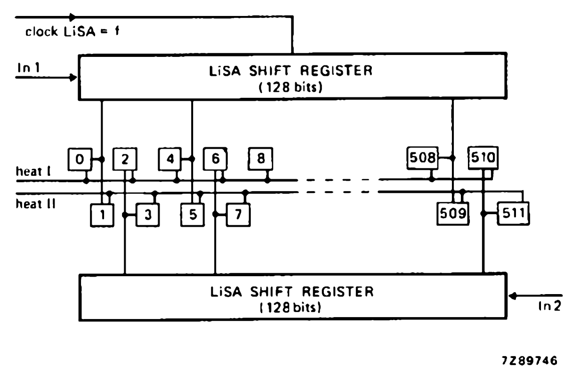


Fig.12 LiSA 512 switching cell layout and signal input

Data encoding

Figure 13 shows how the serially arranged data supplied by the external shift register should be encoded.

The demultiplexer separates the data into that required for the upper and lower cell banks. Following this, the data must undergo second transformations T_1, T_2 tailored to the internal logic of the four 64-bit shift registers. Figure 14 shows how to carry out these transformations, and Fig.15 suggests a way this can be done in practice.

A present aim of LiSA development is to simplify the internal logic of the shift registers so that these second transformations will become unnecessary.

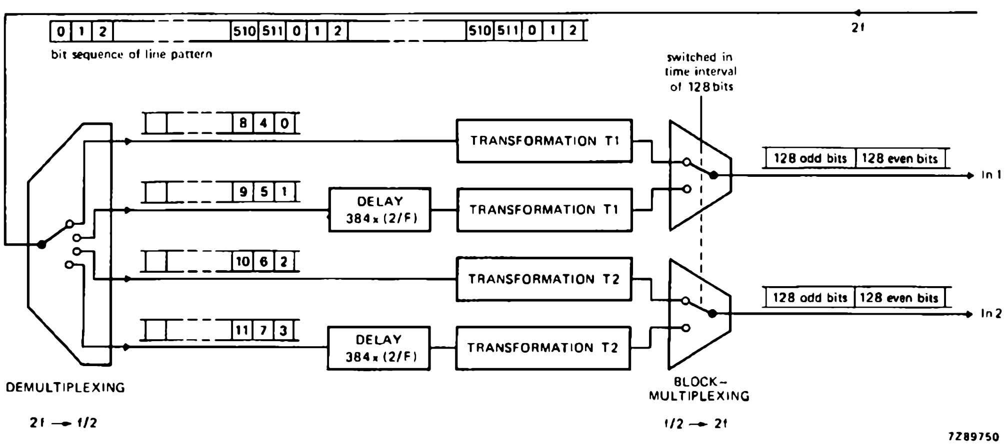


Fig.13 Encoding the serially arranged data for supply to the LiSA 512 unit

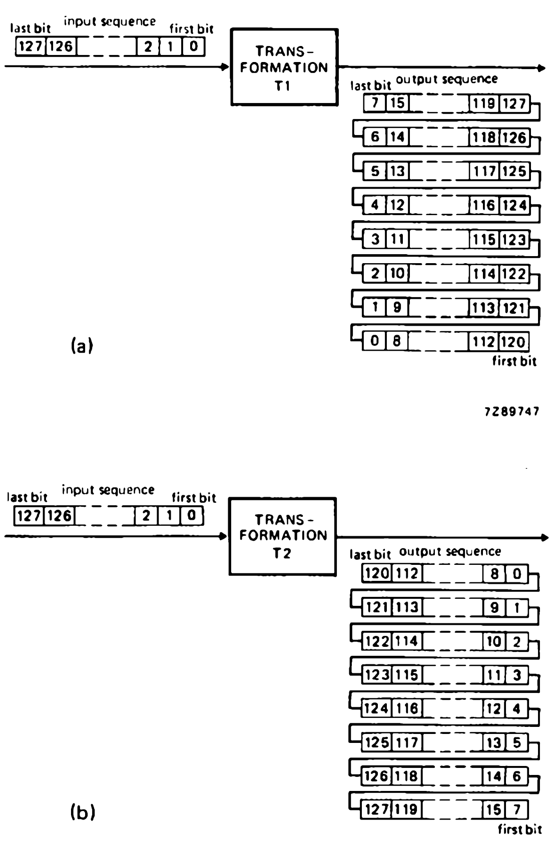


Fig.14 (a) Transformation T_1 and (b) transformation T_2 necessary for adapting the input data to the internal logic of the LiSA shift registers

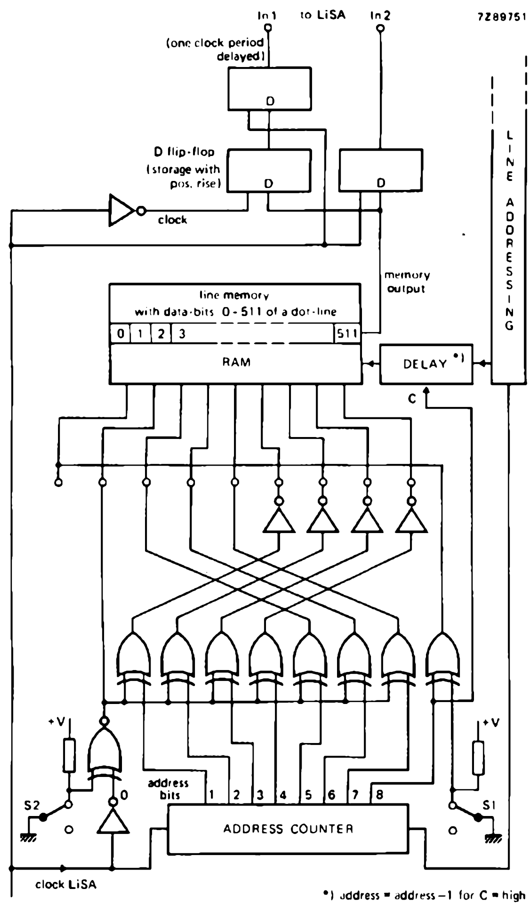


Fig.15 Practical data encoding circuit. For direct or reversed imaging of the LiSA dot pattern, S1 exchanges right and left shift registers, and S2 exchanges upper cell bank with lower

It has long been thought that silicon diodes are unsuitable for ratio detectors for f.m. signals because of their high threshold voltage. It has now been shown that this is not the case: noise can be utilised to pre-bias a silicon diode so that its performance in a ratio detector is just as good as that of a germanium diode.

Silicon ousts germanium for ratio detector diodes

A. GARSKAMP

Ideally, the detector in an f.m. radio must convert the frequency deviations of the modulated i.f. carrier into amplitude variations identical to those that originally caused the frequency modulation at the transmitter. The conversion should be efficient and linear. In addition, although the i.f. amplifier can limit the amplitude of higher level signals, the f.m. detector should be insensitive to any spurious amplitude modulation caused by noise on weak signals. This latter point is of particular importance in car radios which must often receive signals with low or varying signal strength.

Four types of f.m. detector are in common use. They are phase-locked loop (PLL), pulse counter, quadrature and ratio detector. The first two are not suitable for car radios because of their inadequate weak signal performance. In general, the quadrature detector, when driven by a limiting i.f. amplifier, is suitable but it cannot suppress noise caused by spurious amplitude modulation which is particularly offensive when the radio is receiving weak signals below the limiting level. This shortcoming of the quadrature detector can be mitigated by using it in conjunction with dynamic limiting and muting circuits.

The ratio detector achieves highly efficient demodulation of f.m. and has low efficiency with a.m. (good a.m. suppression, even of signals below the i.f. amplifier limiting level) and simple overall circuitry compared with a quadrature detector with ratio detector performance. Furthermore, the ratio detector uses only passive components and two diodes so that the only possible noise sources are the diodes and resistors. It is commonly believed however, that to enable a ratio detector to demodulate weak signals, it must use diodes with a low threshold voltage. The assumption is therefore that silicon diodes cannot be used because their high threshold voltage (>450 mV) must be offset by applying a d.c. bias which entails additional component costs and makes it difficult to derive a.f.c. from the detector because of the d.c. component in the audio output signal. The only apparent alternative is to use germanium diodes which have a lower threshold voltage (about 200 mV). Since germanium is rare and expensive, and the spread of characteristics of germanium diodes requires the use of a potentiometer to correct imbalance in the ratio detector, cost-conscious car radio manufacturers are seeking alternatives to the ratio detector for their products.

This article presents the results of investigations that prove these assumptions to be incorrect. Our silicon diodes BA281 not only replace germanium diodes in ratio detectors without any reduction of weak signal performance, they also eliminate the need for the balancing potentiometer.

WHAT CAUSED THE INCORRECT ASSUMPTION

Figure 1 is the circuit of a ratio detector driven by an integrated f.m. i.f. system TEA5560. Figure 2 shows the signal-to-noise ratios of this circuit as measured between the i.f. input and output. One pair of curves is for the circuit when germanium diodes are used in the ratio detector. The other pair of curves is for the circuit when silicon diodes are used in the ratio detector.

In Fig.2, the effect of the threshold voltage of the silicon diodes, which is higher than that of the germanium diodes, is clearly visible. With germanium diodes, good f.m. detection is achieved, even when the input voltage is reduced to

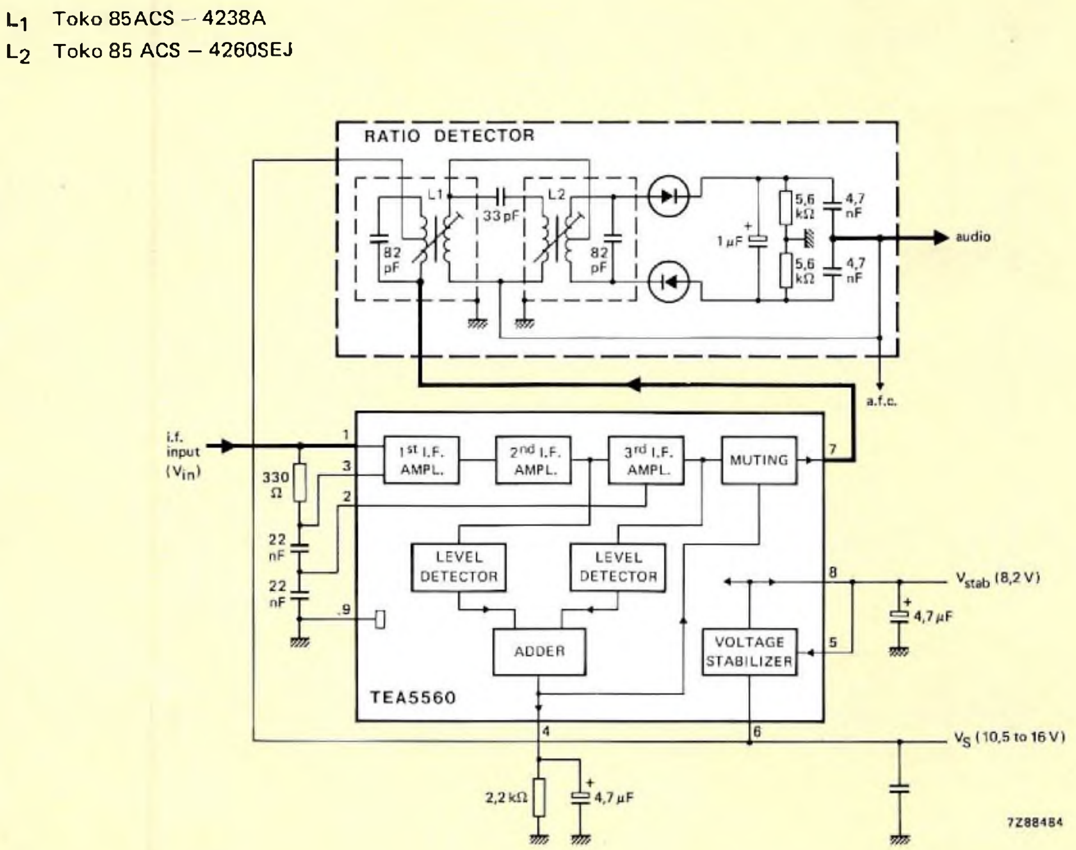


Fig.1 Ratio detector with integrated f.m. i.f. system TEA5560

1 μV. With silicon diodes, the noise increases with inputs below 100 μV, and both the signal and the noise levels rapidly fall with input levels of less than 15 μV as the diodes cease to conduct. Detection of weak signals is therefore apparently impossible with a ratio detector using silicon diodes.

This conclusion, although perfectly valid for the circuit of Fig.1 driven by a signal generator (virtually noiseless), does not apply to a practical radio. The i.f. circuit then receives its signal via a front-end and an i.f. preamplifier both of which contribute some noise. The remainder of this article shows that, if the overall gain of the circuit is correctly chosen, the noise content of the input signal to the ratio detector is sufficient to provide a d.c. bias for the diodes without degrading the signal-to-noise ratio at the output. There is then no discernible difference between the performance of ratio detectors using germanium or silicon diodes.

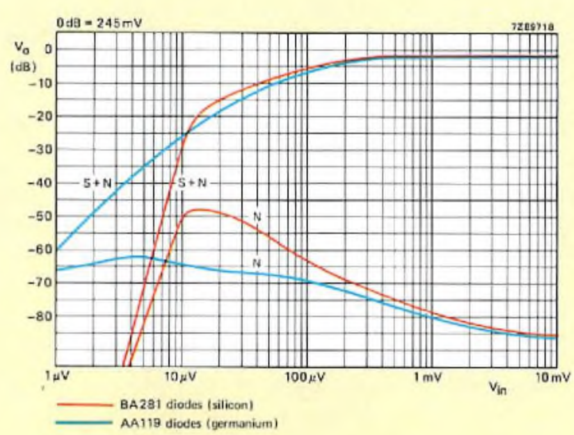


Fig.2 Signal-to-noise curves for the circuit of Fig.1 using germanium or silicon diodes in the ratio detector

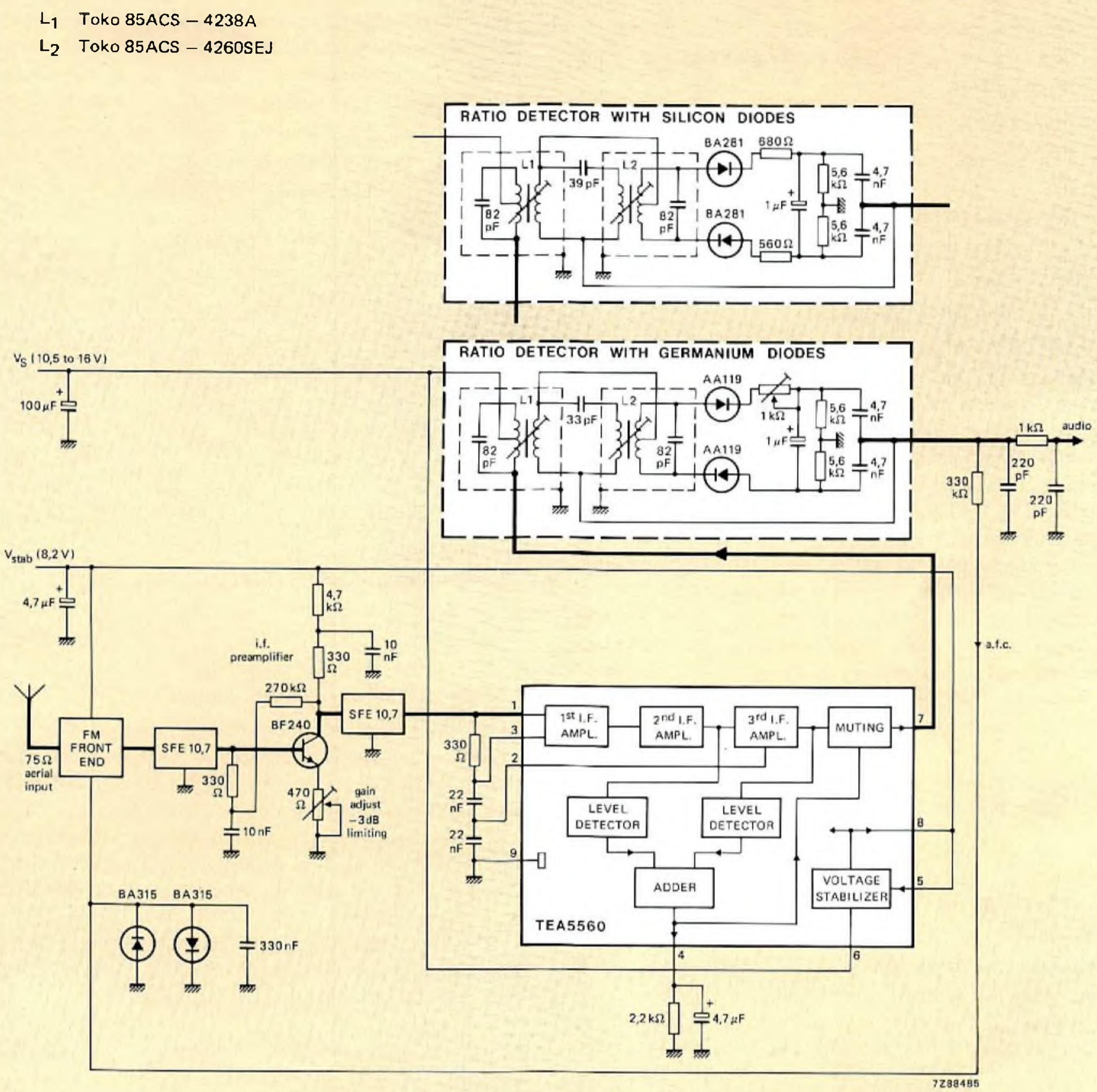


Fig.3 Small signal section of an f.m. radio using integrated f.m. i.f. system TEA5560 and a ratio detector with germanium or silicon diodes

HOW MUCH NOISE IS NECESSARY

The characteristics of the circuit in Fig.3 when using silicon diodes and with the i.f. preamplifier gain control set for maximum gain are:

noise figure of front-end	5 dB
aerial input for S/N = 0 dB	0.66 μV r.m.s. (1.7 μV p-p)
−3 dB bandwidth of i.f. amplifier	150 kHz
i.f. input to TEA5560 for output 3 dB before limiting	150 μV
aerial input for 3 dB before limiting (gain set to max.)	10 μV

Without an r.f. input signal to the circuit, 20 mV d.c. is developed across the 1 μF electrolytic capacitor in the ratio detector. It can therefore be concluded that the ratio detector diodes are conducting due to a rectified noise-induced bias of more than the threshold voltage of the diodes (about 450 mV). When the potentiometer in the emitter circuit of the i.f. preamplifier transistor is adjusted (gain reduced) until the voltage across the 1 μF electrolytic capacitor in the ratio detector is just reduced to zero, the sensitivity at the aerial input for 3 dB before limiting becomes 30 μV. Under these conditions, the diodes are receiving a noise-induced bias exactly equal to their threshold voltage (about 450 mV) and the overall noise gain between the aerial input and the ratio detector output is 450 mV/1.7 μV = 108 dB. From the foregoing, the requirements for obtaining sufficient noise to pre-bias silicon diodes in a ratio detector for an f.m. radio are:

	requirement	*typ. for modern car radio*
noise figure of f.m. front end	⩾5 dB	6 dB
gain between aerial and ratio detector output	⩾108 dB	115 dB
3 dB bandwidth for i.f. channel	⩾150 kHz	>170 kHz

In practice, these requirements can be met by using several different combinations of input noise, i.f. bandwidth and amplification.

PERFORMANCE OF RATIO DETECTORS

To minimise distortion in a ratio detector, the relationship between the coupling coefficient (kQ) of the ratio detector coils and the maximum a.m. suppression must be optimised to obtain a detection curve which is as linear as possible.

Circuit with germanium diodes

In the ratio detector circuit using germanium diodes in Fig.3 the coupling coefficient of the coils is set by the 33 pF capacitor and spreads of the diode characteristics (V_F at low I_F, C_d and R_d as functions of I_F, and the capacitance when reverse biased) are compensated by the potentiometer. The potentiometer is adjusted for maximum a.m. suppression with the aerial input at the −3 dB limiting level.

The following performance figures and the curves in Fig.4 were measured with f_0 = 93 MHz, Δf = ±22.5 kHz, f_m = 1 kHz, dummy aerial as shown in Fig.4.

aerial input (V_{in}) for 3 dB before limiting	15 μV
output for V_{in} = 1 mV	180 mV
noise at output with V_{in} = 0 V (0 dB = 180 mV)	−33 dB
THD at V_{in} = 1 mV	
Δf = ±22.5 kHz	0.24%
Δf = ±75 kHz	1.2%
(S + N)/N	
V_{in} = 4 μV	30 dB
V_{in} = 15 μV	44 dB
a.m. suppression, m = 0.3 (0 dB = output for V_{in} = 1 mV)	
V_{in} = 4 μV	−32 dB
V_{in} = 15 μV	−40 dB

Circuit with silicon diodes

In the ratio detector circuit using silicon diodes in Fig.3, the coupling coefficient is set by the 39 pF capacitor. Since the spread of the characteristics of BA281 silicon diodes is very small, the potentiometer used in the germanium diode circuit can be replaced by two fixed resistors. The values of the resistors are unequal because, in this circuit, the ratio detector coils and their capacitances to the common rail are not symmetrical.

The following performance figures and the curves in Fig.5 were measured with f_0 = 93 MHz, Δf = ±22.5 kHz, f_m = 1 kHz, dummy aerial as shown in Fig.5.

aerial input (V_{in}) for 3 dB before limiting	15 μV
output for V_{in} = 1 mV	204 mV
noise at output with V_{in} = 0 V (0 dB = 204 mV)	−37 dB
THD at V_{in} = 1 mV	
Δf = ±22.5 kHz	0.22%
Δf = ±75 kHz	1.3%
(S + N)/N	
V_{in} = 4 μV	32 dB
V_{in} = 15 μV	44 dB
a.m. suppression, m = 0.3 (0 dB = output for V_{in} = 1 mV)	
V_{in} = 4 μV	−36 dB
V_{in} = 15 μV	−43 dB

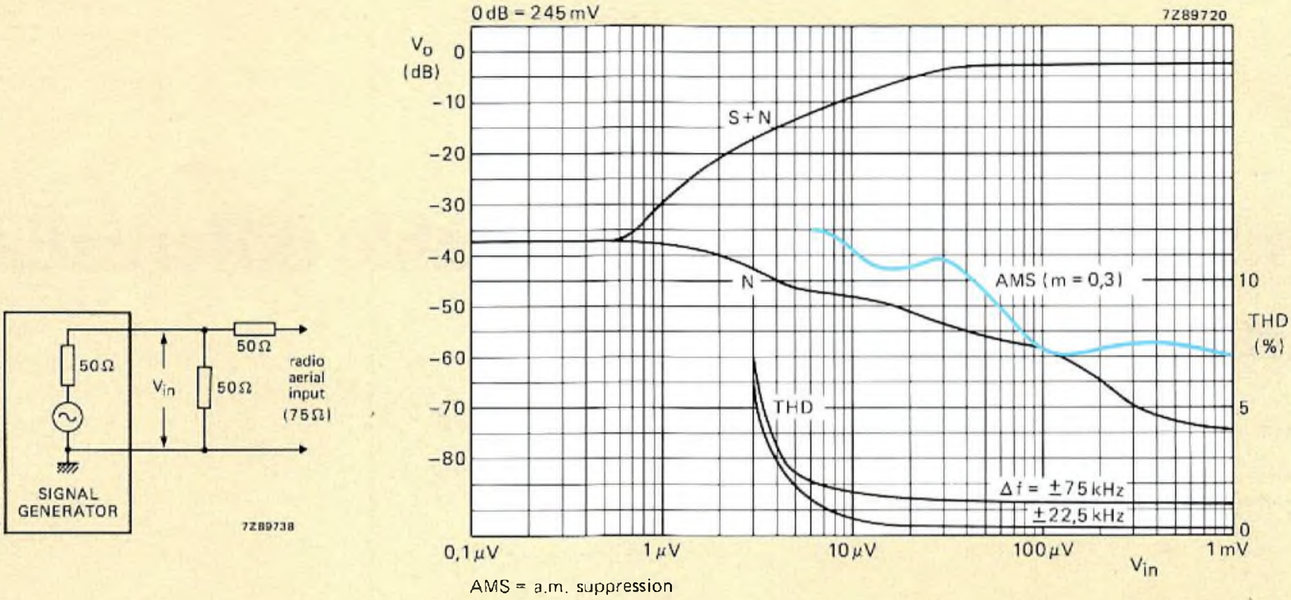


Fig.4 Performance of the circuit in Fig.3 using germanium diodes in the ratio detector

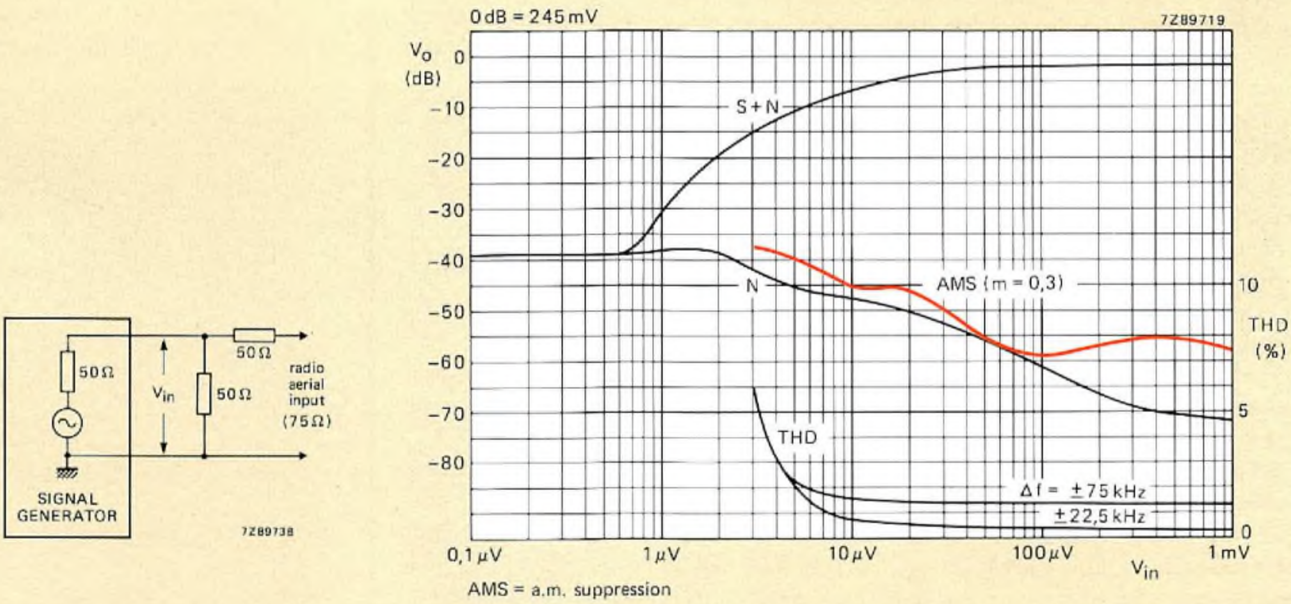


Fig.5 Performance of the circuit in Fig.3 using silicon diodes in the ratio detector

Linear variable differential transformers are simple, reliable displacement transducers that produce an a.c. output whose amplitude is proportional to the amount of movement and whose phase indicates the direction. Driving such a transformer and translating its output into a usable d.c. signal can now be done by a single IC.

Interface IC for linear variable differential transformers

L. HADLEY and E. HERCEG

LVDT DISPLACEMENT TRANSDUCERS

A linear variable differential transformer (LVDT) is an electromechanical transducer that measures very small linear movements in a structure or a mechanical device. Mechanical motion is translated into an electrical signal that contains position information. These accurate transducers help with the measurement of position, pressure, weight, and even acceleration. Extreme ruggedness makes the LVDT ideal for machine-tool applications, and an almost total lack of friction gives it unlimited mechanical life. High resolution and excellent repeatability round out its measuring repertoire.

An LVDT is essentially a differential transformer whose primary coil and two secondary coils are symmetrically spaced on a cylindrical form, and whose iron core is moveable. A cross-section of an LVDT and a plot of its operating characteristics are shown in Fig.1.

When the primary coil at the centre of the LVDT is energised by an a.c. source, voltages are induced in the two secondary coils. These coils are connected in series in such a way that the induced voltages are 180° out of phase. Therefore, the net output of the transducer is the difference between these voltages, which is zero when the core is at the centre, or null, position.

When the core departs from the null position, the induced voltage in the coil that it moves toward increases, while the induced voltage in the opposite coil decreases. This produces a differential voltage output that varies linearly with changes in core position. The phase of this differential output voltage changes abruptly by 180° as the core moves from one side of the null to the other. Linearity is well within ±0.25% and can extend in some cases down to ±0.05% of full scale.

Because the core is separated from the cylindrical body of the transformer, and because primary and secondary windings are totally separated from each other, the LVDT can provide certain advantages over other types of measurement transducers: first, there is no physical contact between the movable core and the coil structure, so an LVDT is inherently frictionless. This means long mechanical life, and good utility in certain tests such as dynamic deflection or vibration tests on delicate materials. Besides being extremely rugged, LVDTs can be sealed for immersion in fluids or for operation at extreme temperatures.

The high sensitivity of the LVDT makes it possible to measure small travels with almost infinite resolution and with

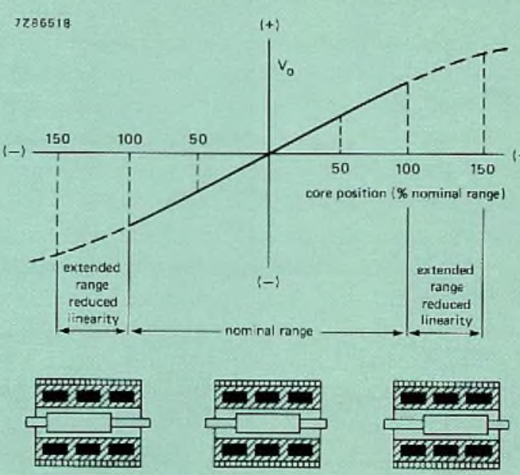


Fig.1 Linear variable differential transformer: output characteristic and schematic cross-section

excellent null repeatability. Because the LVDT is a transformer there is complete isolation between input excitation voltage and output voltages from the secondaries. So, while the LVDT is an asset to high-performance measurement and control loops, it does impose certain drive constraints.

For one thing, the interface circuitry must provide a highly stable excitation voltage for the primary. Although the excitation frequency may be anywhere from 60 Hz to 1 MHz, performance is usually best in the range 2 to 10 kHz. LVDT transfer characteristics are affected by both excitation frequency and temperature. Frequency affects the primary reactance; temperature change results in a positive-coefficient change in the winding resistance of the coils. (Figure 2 shows output characteristics of a typical LVDT for various loads and excitation frequencies.)

In addition to a.c. excitation of the primary, some form of synchronous demodulation of the output is needed to extract the directional information. While a full-wave rectifier (with filtering) will provide voltage-amplitude information, a synchronous demodulator is required to segregate useful phase information from the signal.

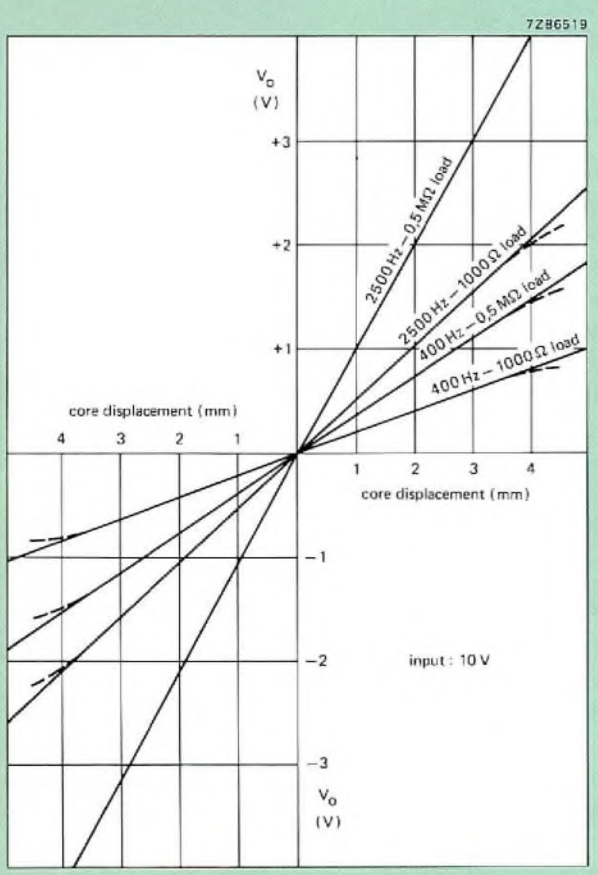


Fig.2 Effect of frequency and load on LVDT output characteristic

However, this signal requires additional filtering to eliminate the second harmonic of the carrier and other higher-order components. A low-pass active filter will do the trick. The d.c. voltage that remains will be a true representation of the exact position of the transformer core with respect to the null or zero position. The higher the voltage, the further the core from its null position; a change in polarity indicates movement to the other side of the null position.

NE5520 INTERFACE CIRCUIT

Signetics' NE5520 is a single-chip signal-conditioning circuit that simplifies the job of interfacing LVDTs with industrial measurement and control circuitry. It can work from single supply voltages in the range of 5 to 25 V or dual supply voltages from ±5 to ±12 V, and deliver a linear d.c. output up to two-thirds of the supply voltage. Operating current is typically of the order of 10 mA with a 10 V single supply. The output indicates the position of the LVDT's movable core over a range as large as 25 cm or as small as ±0.125 mm with nonlinearities less than 0.2% of full scale. The combination LVDT and chip can resolve differences in linear position as small as 0.5 μm.

The NE5520 includes a low-distortion sine wave oscillator with programmable frequency, a synchronous demodulator, accommodation for phase-shift correction, and an auxiliary amplifier that provides a buffered voltage output (Fig.3). Because of its low power consumption and small size, it is easy to mount next to the LVDT or on a data-acquisition system's control board. Measurement resolution is limited only by the sensitivity of the transducer and the gain of the on-chip demodulator. The chip's auxiliary gain stage adjusts the overall system sensitivity. A low-pass filter smooths ripple from the demodulator output. (Note that the design examples to follow assume a simple low-pass filter that allows gain and cut-off to be adjusted independently.)

Supply

Use of a single supply with the transducer chip produces a d.c. common-mode voltage at the output, equal to half the reference voltage on pin 12. This voltage may equal but not exceed the supply voltage on pin 14 (Fig.3). Dual supplies, however, require somewhat different biasing. One way is to eliminate the loop through resistor R, from pin 3 to the V_{ref} input, pin 12, and simply tie pin 12 directly to the V_{CC} positive input at pin 14. Pin 7, here, is no longer grounded, but is connected to the negative supply line. Pin 8 should be grounded instead, to provide the most easily manageable offset-nulling system.

For single-supply systems, the offset of the NE5520 system can be trimmed out with an adjustment potentiometer located between the two R-value resistors, with its

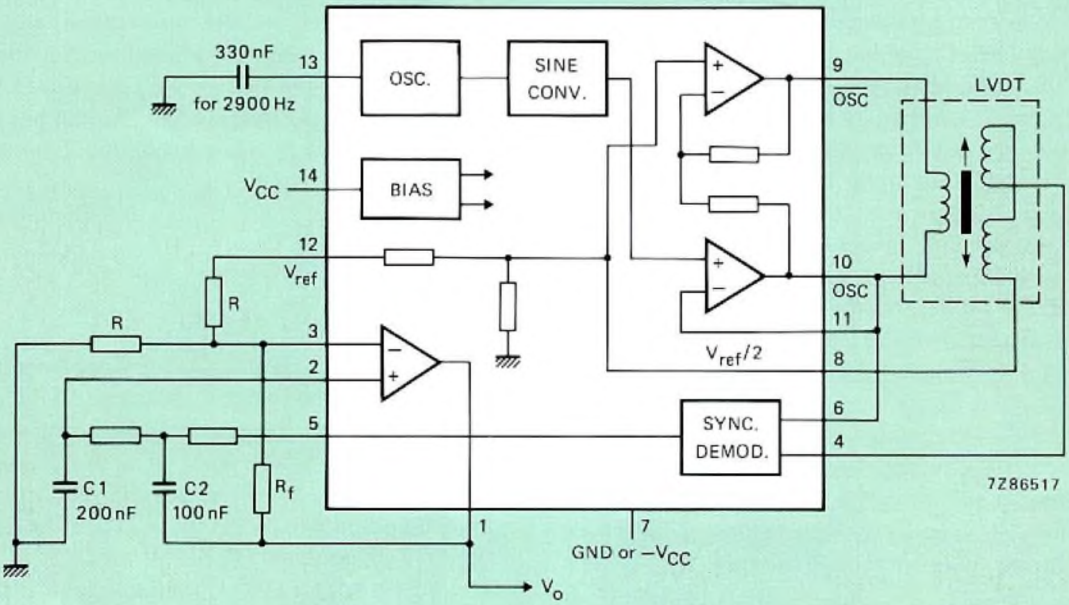


Fig.3 NE5520 interface circuit for LVDTs

central wiper connected to terminal 3. If each R-value resistor is 10 kΩ, the adjustment pot or trimmer can be 1 kΩ. The value of the feedback resistor (R_f) controls the overall gain of the chip's auxiliary op-amp.

For dual-supply systems, the best approach is the use of a dual-tracking voltage regulator as a supply source.

Oscillator

The oscillator, which is adjustable from 1 to 20 kHz by means of external capacitors, consists of a triangle-wave generator, a current source/sink circuit that switches when an off-chip capacitor is charged to ¼ and ¾, respectively, of the reference voltage. The reference voltage can be anything from 5 to 25 V, and is supplied to the V_{ref} input terminal. The total voltage swing of the triangle-wave oscillator is $V_{ref}/2$. The triangle wave is fed to a sine wave converter the output of which is buffered by two op-amps (pins 9 and 10), which are capable of delivering a sinewave with less than 5% harmonic content into a 1 kΩ load.

Demodulator

The second major functional part of the NE5520 is the synchronous demodulator, which performs full-wave rectification in phase synchronism with the oscillator output. To extract true position and direction information, the phase relationship of the LVDT secondaries relative to the primary voltage must be found. This is done by switching a unity-gain op-amp from non-inverting to inverting in sync with the primary a.c. voltage. The output signal is a pulsing d.c. containing second harmonics of the carrier frequency. Its polarity depends on which side of the null point the core is on.

The residual harmonics are removed by filtering. The demodulator output appears at pin 5; additional filtering can be done with the aid of the on-chip auxiliary amplifier, whose inputs are at pins 2 and 3, and whose output is at pin 1.

Biasing

The overall internal structure of the NE5520 must be biased to ensure its dynamic operating range. The bias voltage is normally $V_{ref}/2$, which appears at pin 8. When operating from single or dual supplies, V_{ref} (pin 12) is normally tied to the V_{cc} line (pin 14) in ratiometric mode. With single-supply operation, this means that $V_{ref}/2$ will be equal to half the supply voltage. For dual-supply operation, $V_{ref}/2$ is ground-referenced. In all cases, $V_{ref}/2$ must be less than or equal to V_{cc}, but not greater.

The NE5520 has an extremely linear transfer function with good repeatability. The full-scale output depends on the supply voltage, but the linearity is constant.

APPLICATIONS

Indication and control

In a simple measuring circuit using a single 10 V supply voltage, pin 1 output of the NE5520 floats above ground at about half the supply voltage. A d.c. microammeter, connected between pin 1 and a simple resistive divider connected between V_{CC} and ground, provides a ratiometric output as long as the supply voltage remains absolutely constant.

A more accurate indicator scheme uses a four-digit panel meter. Where the LVDT system displays information continuously – for example, in weighing or strain monitoring – a simple analog meter might not provide the necessary resolution. A three or four-digit panel meter can be constructed using a CMOS integrating analog-to-digital converter and a standard multidigit decoding panel meter.

In an industrial data acquisition system the NE5520 can feed a microprocessor instead of a display (Fig.4). Parallel binary information from the a-d converter can go directly to an 8-bit processor, can be stored for future use, or can be evaluated in real time for specific control situations. Real-time closed-loop control also requires a d-a converter to feed analog information back from the microprocessor to the actual control loop. Although the processor will require specialised software to provide optimum control of the loop and to maximise damping, it will provide real-time decision-making capability and, at the same time, handle a number of multiplexed LVDT inputs.

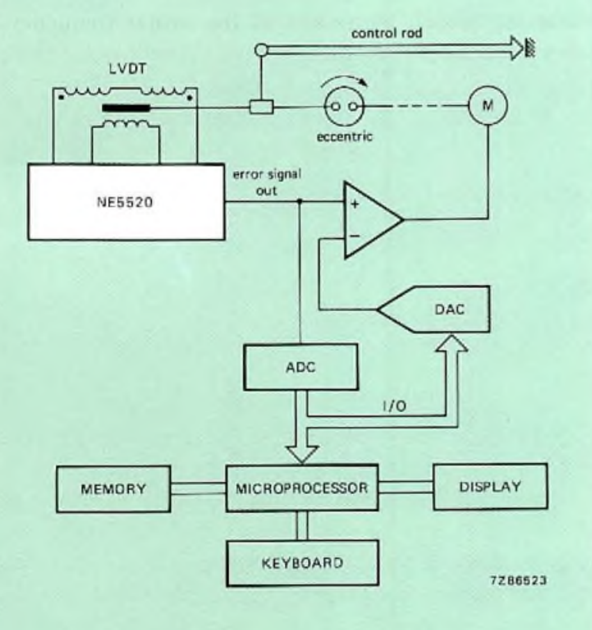


Fig.4 Microprocessor controlled servo with LVDT error sensor

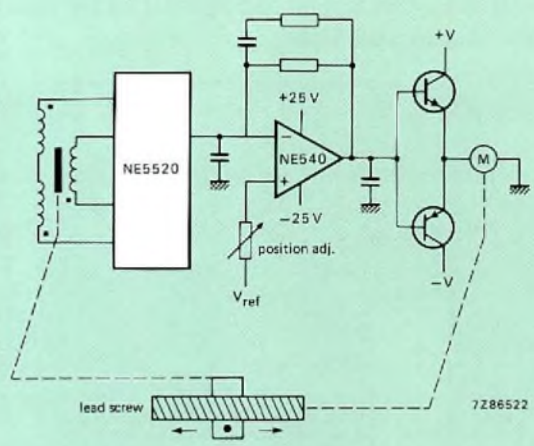


Fig.5 Manually controlled servo with LVDT error sensor

In other kinds of servo control systems – pressure rollers, hydraulic drivers, and other motor-driven systems – an LVDT signal-conditioning loop can provide real-time feedback and control without a microprocessor. A system like the one in Fig.5 can control linear position down to several millimeters.

Current loops

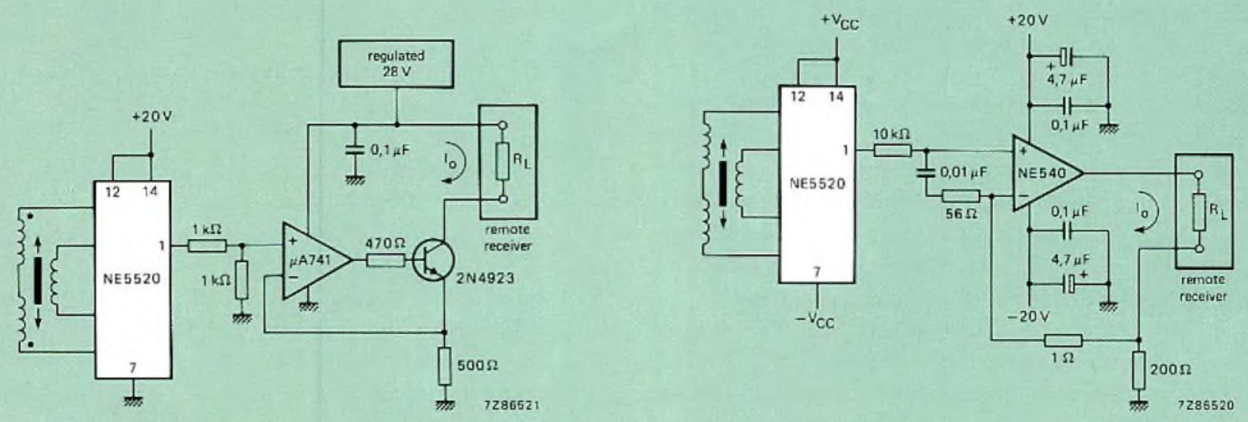
For industrial control in an electrically noisy environment, a 4-20 mA d.c. current loop is a very popular method for transmitting an analog signal from a remote transducer to a data conversion and intelligence center (Fig.6). Mounting the NE5520 in the same housing as the LVDT (with power supply separate) allows the transducer-signal and a conditioning output to be run over long lines.

In Fig.6(a), an auxiliary power supply near the transducer facilitates the construction of a unipolar current loop using an op-amp and a darlington-transistor pair or any power device with a current gain of about 500 and a current output around 150 mA. (Although Fig.6(a) shows 28 V, any supply voltage from 40 to 50 V can be used to power the system.)

Figure 6(b) shows a simplified bipolar current loop using an op-amp and a local power supply. The op-amp swings over ±10 V and, through the load resistor, produces bipolar loop currents.

Systems with a large number of LVDTs may require special provisions to avoid cross-coupling effects. When oscillation frequencies from separate signal conditioners are fed through long cable lines along with demodulation signals, beat frequencies may appear if the signal frequencies differ by only a few cycles. One way to overcome this difficulty is to drive several NE5520 and LVDT pairs from the same synchronised input.

A master-slave arrangement makes the NE5520 the master oscillator and all additional units slaves. For synchronous operation, the timing capacitor for the master oscillator should feed all other NE5520 oscillator inputs. Output signal lines, however, are available from each chip.



(a) Unipolar current loop for remote LVDT signal processing or read-out

(b) Bipolar current loop for remote LVDT

Fig.6

ACKNOWLEDGEMENT

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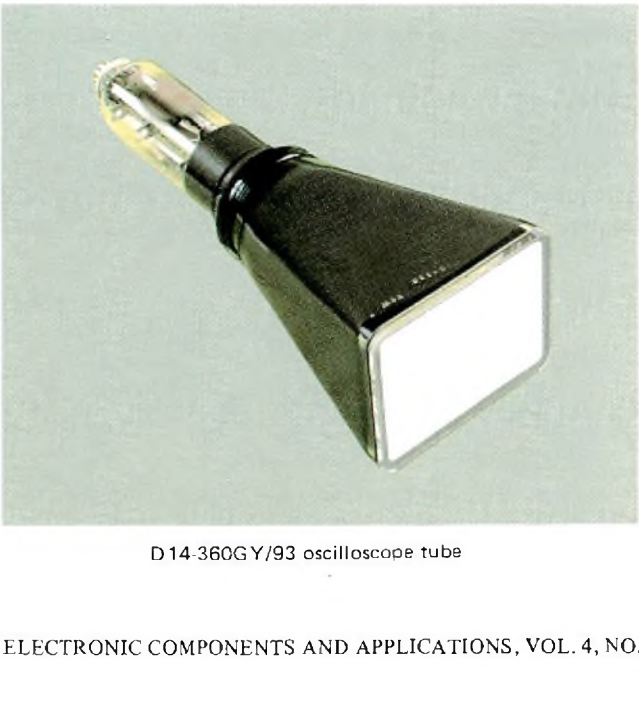
The oscilloscope tube, parent of the tv picture tube, has learnt a useful dodge from its successful offspring: how to use internal permanent magnets to manipulate the electron optics and thereby simplify the circuitry. Scope manufacturers with an interest in cost/performance will also welcome other novel aspects of the tube described here – namely, a newly designed envelope that is easier to mount and shield, and a brighter, easier-to-light screen graticule.

Oscilloscope tube with magnetic correction and scan magnification

K. ZEPPENFELD

The D14-360GY/93 is a 14 cm mono-accelerator tube for frequencies up to about 25 MHz that enables oscilloscope manufacturers to offer improved performance and, at the same time, to reduce manufacturing costs without compromising reliability or features. Greater and better-calibrated deflection sensitivity, more accurate beam centring, stigmatic focusing, better deflection orthogonality, and a brighter and more easily illuminated graticule are some of the advantages it has over other 14 cm tubes. In addition, its shape simplifies construction of the mu-metal shield, and reference points built into the faceplate automatically align the graticule with the mounting panel.

The novel features of the D14-360GY/93 are also applicable to tubes with post-deflection acceleration mesh, and the development of a family of such tubes for frequencies up to more than 100 MHz is now nearing completion.



D14-360GY/93 oscilloscope tube

ELECTRON GUN

An innovation in the D14-360GY/93 is the use of permanently magnetised wire rings in the electron gun for certain functions which in other tubes are performed electrostatically, if at all.

Magnetised rings in the electron gun were pioneered in colour television tubes, where they superseded adjustable external multipole magnets for static convergence correction (Ref.1). The rings are magnetised from outside the tube during a late stage of manufacture, and experience over more than three years has shown that the magnetisation withstands all conditions of operation and storage.

A colour television tube uses only one ring for convergence correction. The D14-360GY/93 uses two: one for scan magnification, beam centring and orthogonality correction, and the other principally for beam shaping. Mu-metal shielding outside the tube does not affect the internal fields developed by the rings.

Scan magnification

As shown in Fig.1, a permanent-magnet ring can be magnetised with a combination of dipoles and quadrupoles to give a variety of effects and corrections. In the D14-360GY/93 a ring on grid 5 (the interplate shield electrode between the horizontal and vertical deflection plates) produces a quadrupole field capable of giving a scan magnification factor of 1.25.

This magnification potential is not fully utilised for sensitivity increase. Part is held in reserve for calibration improvement, thereby reducing the tolerances that have to be reckoned with in designing the vertical deflection amplifier. The vertical deflection coefficient of the D14-360GY/93 is 11.5 ± 0.5 V/cm, an improvement of 15% over the 13.5 V/cm of the D14-250 series of tubes it replaces. However, the deflection coefficient of the D14-250 tubes can be as high as 15 V/cm, whereas that of the D14-360GY/93 never exceeds 12 V/cm. Thus, besides giving a 15% sensitivity increase, the benefit of the magnification is reflected in a 25% decrease of the maximum output voltage swing required of the deflection amplifier.

The ring on grid 5 gives two additional benefits. First, it centres the undeflected spot within ±2 mm (instead of ±7 mm as in previous tubes), raising to 30% the total reduction of the required voltage swing. The scope designer can use this margin either to increase bandwidth or, by raising the gun voltage, to increase brightness. And second, it makes it possible, during manufacture of the tube, to adjust the orthogonality of the horizontal and vertical deflection to within less than 30′, eliminating the need for costly and complex external adjustments.

The beam centring is accomplished by dipole fields, and the orthogonality correction by a weak quadrupole field, superimposed on the scan magnification field.

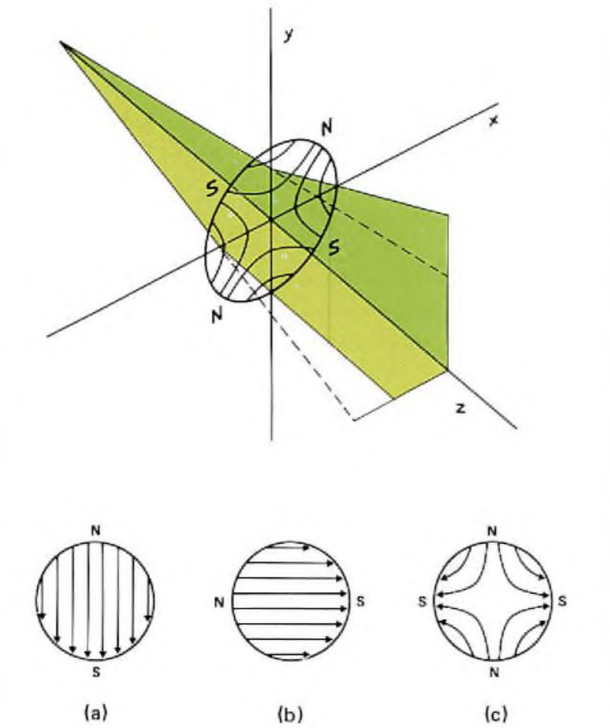


Fig.1 Different field configurations can be superimposed in a single magnetic ring. A quadrupole field in the plane of grid 5 magnifies the vertical scan as shown in the perspective drawing. Dipole fields as in (a) and (b) are used for beam centring. A quadrupole field as shown at (c) rotates the vertical scan. Quadrupole fields in the plane of grid 4 correct astigmatism

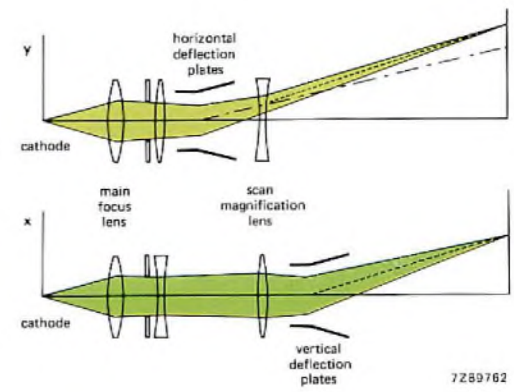


Fig.2 Vertical scan magnification and overall focusing with two weak quadrupoles and a common main lens. Top, vertical deflection; bottom, horizontal deflection

Stigmatic focusing

An additional magnet ring on grid 4, near the main focus lens, produces a second quadrupole field, somewhat weaker and of opposite sign, which focuses the beam to a circular spot on the screen. As this field is adjusted for a mean vertical deflection potential equal to anode potential, separate astigmatism adjustment is no longer required.

As shown in Fig.2, the crossed quadrupoles produce a beam in the drift space whose cross-section is non-circular, with an aspect ratio of about 1.5. The cross-sectional area of this beam is somewhat larger than would be the case without scan magnification, which means that the sensitivity increase is not gained at the cost of writing speed or brightness. What is sacrificed is the safety margin in horizontal deflection plate spacing; however, with the supplementary corrections that can be introduced via the magnetisation of the rings, that margin is no longer needed.

In theory, even a weak electron-optical system without intermediate crossover requires at least three quadrupole fields – rather than two plus a round lens – for scan magnification (Ref.2). Indeed, tubes such as the D13-500 and L14-140 (Ref.3), which use three electrostatic quadrupoles to obtain overall focusing with the same lateral crossover magnification in the x and y directions, still require double focus control. In the D14-360GY/93,

A magnetic ring of the kind used in the D14-360GY/93, and grids 4 and 5 with rings mounted. The elongated aperture in grid 5 is to accommodate beam displacement due to deflection however, equal magnification is not important, for cross-over size scarcely affects spot size. A more relevant consideration is that once the relative strengths of the two quadrupoles have been adjusted to give the optimum circular spot, it should be possible to reproduce that spot at any beam current with the usual single focus control voltage on grid 3 only. Because of the elliptical cross-section of the beam in the drift space, this means that to properly counteract space-charge repulsion the focusing action of the grid 3 voltage has to be different in the horizontal and vertical directions. This difference can be observed by playing with the focus voltage: for both under and over-focusing the spot expands mainly in the vertical direction, but when it is in focus it becomes circular and of the same size as it would be without scan magnification.

ENVELOPE

In a typical oscilloscope, the tube alone may account for as much as 10% to 20% of the total cost; moreover, the provisions for mounting, adjustment, shielding, and graticule illumination can easily double that amount. The D14-360GY/93 has been designed with an eye to minimising these secondary costs.

Bulb

The cone has a V-shaped taper throughout (see photo) and is cut and ground to close tolerances at both ends before being frit sealed to the faceplate and the neck. Both seals are made in a single heat treatment and at a low enough temperature not to affect the dimensional tolerances of the glass.

In previous tubes the end of the neck was blown to a larger diameter and fused to a shorter cone. The neck enlargement was for a post-deflection acceleration mesh and, in the interest of standardisation, was also used in tubes that had no mesh. The resulting shape made construction of a close-fitting mu-metal shield difficult and expensive.

Although the D14-360GY/93 is not fitted with a post-deflection acceleration mesh, the same bulb will be used with later tubes that are; the electron-optical problems of adapting such a mesh to a conical instead of a cylindrical surround have been solved. In fact, the moulded cone gives better definition of the mesh-lens region. It also has better x-ray absorption.

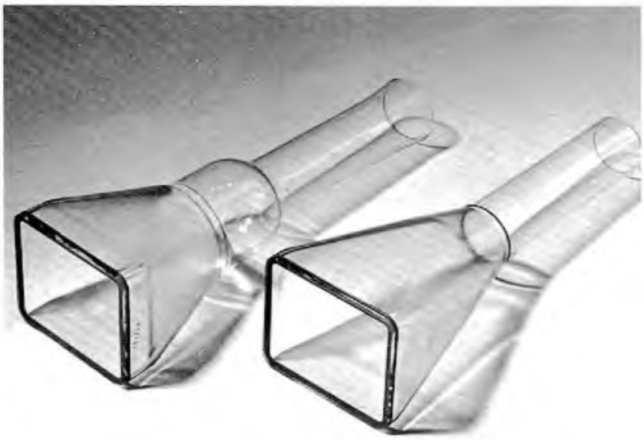
As the new envelope together with the trace rotation coil fits into a straight tangential line, it can be shielded by a plain rolled sheet of mu-metal.

Faceplate

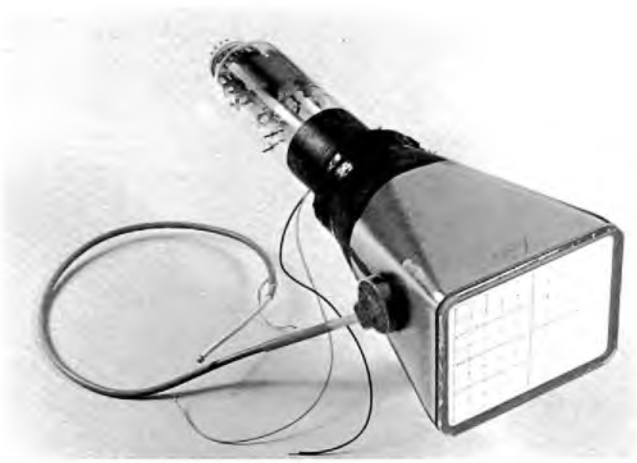
Many oscilloscope tubes have a moulded faceplate with a turned up rim fused to the cone, like black-and-white television tubes. This construction does not result in a screen-graticule combination that is optically as good as possible or as easy to illuminate.

The faceplate of the D14-360GY/93 is optical-quality glass of uniform thickness and very high transmission. Its edges are flat and the optical isolation afforded by the frit seal facilitates side illumination. Also, because of the flatness of the inner surface, it is possible to screen-print the graticule and bake it out before attaching the faceplate to the cone. Screen-printing gives better optical contact than powder-depositing from a photosensitive suspension as was formerly done; the light output efficiency is doubled. Because less light input is required, dirt and scratches incurred on the outside of the faceplate during use are much less noticeable.

Three reference points on adjoining edges of the faceplate make it possible to mount the tube accurately to the front panel without further adjustment.



Envelope of the D14-360GY/93 (right) alongside that of the D-14-250 it replaces



A tube with post-deflection acceleration using the same envelope and magnetic corrections as the D14-360GY/93

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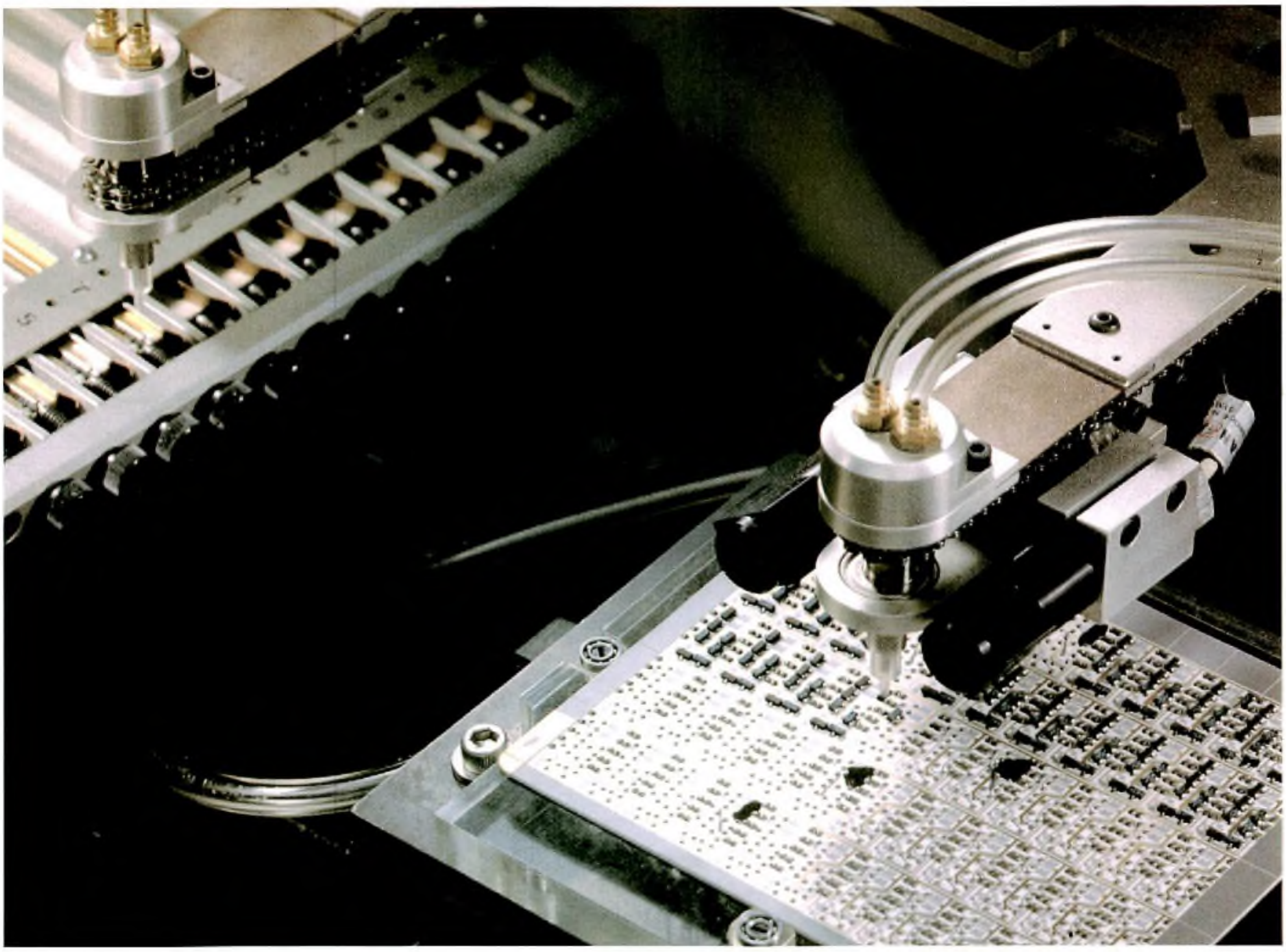
With their small size, high reliability, and flexibility of configuration, hybrid circuits bridge the gap between discretes and monolithics. When a conventional printed wiring board would be too bulky or unreliable, and the planned production run is too short to justify a custom IC, a hybrid is very often the answer.

A high degree of automation in today's manufacture of hybrids keeps the cost low and the quality high. Here, a microcomputer-controlled component placer capable of handling up to 3000 components per hour is placing components on a plate of 55 hybrid-circuit substrates. The placer has two heads working in alternation – one picks while the other places. Feed trays on either side of the plate can supply up to 24 different types of components. Patches of black lacquer mark defective substrates; optical sensors on the head seen above the plate ensure that these are passed over.



Abstracts

ICs for Compact Disc decoders

The Compact Disc Digital Audio system reproduces audio signals of the highest quality. Noise is reduced to a negligible level by using 16-bit digital audio samples, and encoding the data on Compact Discs in such a way that errors due to data drop-outs can be detected and corrected. In view of the extensive signal processing in error correction, a set of LSI circuits has been developed for cost-effective manufacture of decoders for the Compact Disc system. This article describes such a decoder with a special D/A conversion system, having 16-bit performance without using a 16-bit D/A converter.

Pyroelectric infrared detectors

Pyroelectric infrared detectors respond to changes in the infrared radiation in the scene to which they are exposed. They are therefore well suited to use in intruder and flame detectors and some passive control applications. With the addition of an optical chopper they can also be used in industrial radiometers; another potential application is gas analysis and identification of pollutants.

Single-chip synthesiser for radio tuning

If a radio must incorporate facilities such as search tuning and/or tuning by direct entry of required frequency at a keyboard, a stable local-oscillator signal can be generated by indirect frequency synthesis with a PLL controlled by a microcomputer. This is the function of the new 18-pin LSI circuit SAA1057 described in this article. The SAA1057 and 16 peripheral components perform the tuning functions which previously required the use of three integrated circuits and 36 peripheral components. A description and the performance of the basic circuit are given.

Small electrolytic capacitors compared

Recent advances in manufacturing techniques for small solid aluminium electrolytic capacitors have allowed a considerable increase in their volumetric efficiency. Although this improvement has led to their increased use in professional and industrial applications, the characteristics of this type of capacitor are generally less well known than those of wet electrolytics and tantalums. This article presents a comparison, thereby highlighting the clear advantages of solid aluminium electrolytics.

Full-page display using television components

The described display accepts video input at TTL level and presents an upright page of more than 5000 characters on a flicker-free, noninterlaced, thousand-line raster. The 38 cm c.r.t. is an adaptation of a television picture tube, and the deflection yoke uses television-derived pin-indexed winding to ensure a distortion-free image. Other circuit components, including the line and field oscillator ICs, are unmodified television parts.

Light-switching array for high resolution pattern generation

LiSA 512 is an electronically controlled light-switching array for high-quality pattern generation in electro-photographic printers. The LiSA 512 chip consists of a meander array of 512 light switching cells controlled by an external magnetic field. All switching and control functions are performed by ICs external to the chip. The system can accept input data at the rate of 1 Mbits/s, corresponding to a maximum pattern generating speed of 2000 pattern/s.

Interface IC for linear variable differential transformers

The NE5520 integrated circuit incorporates a variable-frequency oscillator for energising a linear variable differential transformer, and a synchronous demodulator for obtaining an output whose polarity and amplitude indicate the direction and amount of displacement of the transformer's movable core.

Silicon ousts germanium for ratio detector diodes

There is a common belief that germanium diodes must be used in a ratio detector because of their low threshold voltage. This article shows that this belief is unfounded and has only come about because investigators have hitherto only considered the i.f. amplifier and detector when assessing the relative merits of germanium and silicon diodes. When the total performance of a radio is measured between aerial and audio output, it has been revealed that a ratio detector operates just as well with silicon diodes as it does with germanium diodes.

Oscilloscope tube with magnetic correction and scan magnification

Permanent-magnet rings incorporated in the electron gun of a new 25 MHz mono-accelerator c.r.t. increase the vertical deflection sensitivity, ensure the roundness of the beam spot, and improve its centring. The newly designed cone of the tube envelope, which greatly simplifies the fitting of magnetic shielding, can also be used for higher-frequency tubes with domed-mesh post-deflection acceleration.

Integrierte Schaltungen für Compact-Disc-Decoder

Mit dem digitalen Compact-Disc-System (Digital-Schallplatte) lassen sich Tonfrequenzsignale mit höchster Qualität wiedergeben. Bei diesem System kann das Rauschen auf einem vernachlässigbar kleinen Pegel gehalten werden, wenn man das NF-Signal als 16-bit-Digitalsignal auf der Platte speichert und die Decodierung der Compact-Disc-Daten so durchführt, dass Fehler, die durch Ausfall von übertragenen Daten entstehen, erkannt und korrigiert werden. Zur kostengünstigen Herstellung von Compact-Disc-Decodern wurde eine Reihe von LSI-Schaltungen entwickelt. Dieser Artikel beschreibt einen derartigen Decoder mit einem speziellen Digital/Analog-Wandlersystem, das eine Auflösung von 16 bit besitzt, ohne einen 16-bit- Digital/Analog-Wandler zu verwenden.

Pyroelektrische Infrarot-Detektoren

Pyroelektrische Infrarot-Detektoren reagieren auf Änderungen der Infrarot-Strahlung der Szene, denen sie ausgesetzt sind. Daher sind sie gut geeignet zum Entdecken von Eindringlingen, als Flammendetektoren und in einigen passiven Steuerungsanwendungen. Mit einer zugeschalteten optischen Choppervorrichtung können diese Bauelemente auch in industriellen Strahlungsmessgeräten eingesetzt werden. Andere mögliche Anwendungen betreffen die Gasanalyse und die Erkennung von Schadstoffen.

Einchip-Synthesizer für Abstimmsysteme von Rundfunkempfängern

Wenn ein Rundfunkempfänger mit dem Bedienkomfort einer Suchlaufabstimmung und/oder einer direkten Frequenzwahl über Tasten ausgestattet werden soll, lässt sich das dazu erforderliche stabile Empfängeroszillatorsignal durch eine Frequenzsynthese mit einer mikrocomputergesteuerten PLL-Schaltung erzeugen. Das ist die Funktion der neuen, in einem DIL18- Gehäuse untergebrachten LSI-Schaltung SAA1057, mit der sich der vorliegende Artikel befasst. Die Synthesizerschaltung SAA1057 führt die Abstimmfunktion mit 16 peripheren Bauelementen aus, wozu früher drei integrierte Schaltungen und 36 externe Bauelemente erforderlich waren. Es wird die Grundschaltung beschrieben und deren Eigenschaften angegeben.

Vergleich von kleinen Elektrolytkondensatoren mit nassem und mit trockenem Elektrolyten

Die Anwendung neuer Techniken bei der Herstellung von kleinen Aluminiumkondensatoren mit festem Elektrolyten erbrachte eine beträchtliche Verminderung der räumlichen Abmessungen. Obgleich diese Verbesserung zu einem steigenden Einsatz bei professionellen und industriellen Anwendungen geführt hat, sind die characteristischen Eigenschaften dieser Kondensatoren allgemein weniger gut bekannt als die der Kondensatoren mit nassem Elektrolyten und derjenigen auf Tantalbasis. In diesem Artikel wird ein Vergleich durchgeführt, wobei die klaren Vorteile von Aluminiumkondensatoren mit festem Elektrolyten herausgestellt werden.

Ganzseiten-Display mit FS-Bauteilen

Das beschriebene Display kann mit Video-Signalen bei TTL-Niveau betrieben werden und stellt eine Seite im Hochformat mit mehr als 5000 Zeichen in einem flimmerfreien 1000-Zeilenraster ohne Zeilensprung dar. Die 38 cm-Katodenstrahlröhre stellt eine Spezialausführung einer FS-Bildröhre dar; das Ablenksystem ist in der von den Fernsehverfahren abgeleiteten Strangwickeltechnik hergestellt, um ein verzerrungsfreies Bild sicherzustellen. Die übrigen Bauteile sind nichtmodifizierte FS-Teile.

Eine Lichtschaltzeile zur Erzeugung optischer Punktemuster mit hoher Auflösung

LISA 512 ist eine elektrisch gesteuerte Lichtschaltzeile zur Erzeugung optischer Punktemuster mit hoher Auflösung für elektrofotografische Drucker. Der Chip dieses Bauelements enthält in versetzter Anordnung 512 Lichtschaltzellen, die durch ein äusseres Magnetfeld beinflusst werden. Alle Schalt- und Steuerfunktionen benötigen zu ihrer Ausführung externe integrierte Schaltungen. Das System kann Eingangsdaten mit einer Rate von 1Mbit/s verarbeiten. Dies entspricht einer maximalen Erzeugungsrate von 2000 Punktmustern pro Sekunde.

Silizium verdrängt Germanium als Halbleitermaterial in Ratiodetektor-Dioden

Es herrscht weitgehend die Meinung vor, dass in Ratiodetektoren Germaniumdioden wegen ihrer niedrigen Schleusenspannung verwendet werden müssen. Dieser Artikel zeigt, dass diese Meinung unbegründet ist und nur dadurch aufkommen konnte, weil bisher in die Untersuchungen zur Bewertung der relativen Vorzüge von Germanium- und Siliziumdioden immer nur der ZF-Verstärker und ZF-Demodulator mit einbezogen worden ist. Wenn dagegen die zwischen Antenne und NF-Ausgang gemessenen Gesamteigenschaften eines Rundfunkempfängers den Betrachtungen zugrunde gelegt werden, dann zeigt sich, dass ein Radiodetektor ebensogut mit Siliziumdioden arbeitet wie mit Germaniumdioden.

Interface-IC für linear variable Differential-Transformatoren

Die integrierte Schaltung NE5520 enthält einen Oszilator mit variabler Frequenz zum Betrieb eines linear variablen Differential-Transformators sowie einen Synchrondemodulator, bei dem über die Polarität und Amplitude des Ausgangssignals die Richtung und der Betrag der Verschiebung des beweglichen Transformatorkerns angezeigt werden.

Katodenstrahlröhre mit magnetischer Korrektur und Abtastvergrösserung

Durch Permanentmagnetringe, die in das Elektronenstrahlsystem einer 25-MHz-Katodenstrahlröhre eingebaut sind, kann die Ablenkempfindlichkeit in vertikaler Richtung erhöht werden, ferner wird die Rundheit und Zentrierung des Leuchtflecks gesteigert. Das neue Röhren-Konussystem, mit dem die magnetische Abschirmung verbessert wird, kann auch bei Röhren höherer Betriebsfrequenz mit gewölbten Nachbeschleunigungs-Netzelektroden verwendet werden.

Circuits intégrés pour décodeurs de "Compact Disc"

Le système digital "Compact Disc" reproduit des signaux audio de très haute qualité. Le bruit est réduit à un niveau négligeable grâce à un échantillonnage du signal audio à 16 bits et par un codage des données sur le "Compact Disc" permettant la détection et la correction à la reproduction des erreurs dues à des pertes de données. Etant donné l'importance du traitement des signaux nécessité par la correction des erreurs, un jeu de circuits LSI a été mis au point en vue de la fabrication avec un bon rapport coût-efficacité de décodeurs pour le système "Compact Disc". Cet article décrit un tel décodeur équipé d'un système de conversion digital/analogiques spécial à précision de 16 bits sans utiliser de convertisseur digital/analogique de 16 bits.

Des détecteurs pyroélectriques d'infrarouges

Les détecteurs pyroélectriques d'infrarouges réagissent aux variations du rayonnement infrarouge auquelles ils sont exposés. Ils conviennent donc bien à l'utilisation dans les détecteurs d'intrusion et d'incendie et dans certains dispositifs de contrôle. Complétés par un chopper optique, ils sont également utilisables dans des radiomètres industriels; une autre application possible est l'analyse de gaz et l'identification de polluants.

Synthétiseur à "puce" unique pour le réglage d'accord radiophonique

Lorsqu'un récepteur radiophonique permet la recherche automatique des émetteurs ou l'accord après affichage par clavier de la fréquence requise, il est possible d'engendrer un signal d'oscillateur local stable par synthèse de fréquence indirecte à l'aide d'un PPL commandé par un micro-ordinateur. C'est la fonction du nouveau circuit LSI à 18 broches SAA1057 décrit dans le présent article. Le SAA1057 et 16 composants périphériques assurent les fonctions de réglage d'accord, qui nécessitaient auparavant trois circuits intégrés et 36 composants périphériques. Une description du circuit de base est donnée et ses performances sont indiquées.

Comparaison de petits condensateurs electrolytiques

Les progrès récents des techniques de fabrication des petits condensateurs aluminium à électrolyte solide ont permis d'augmenter considérablement leur capacité par unité de volume. Bien que cette amélioration ait conduit à l'intensification de leur emploi dans des applications professionnelles et industrielles, les caractéristiques de ce type de condensateur sont généralement moins bien connues que celles des condensateurs tantale et aluminium à électrolyte liquide. Cet article présente une comparaison qui est nettement en faveur des condensateurs aluminium à électrolyte solide.

Affichage de pages complètes à l'aide de composants de téléviseurs

L'affichage décrit reçoit un signal d'entrée vidéo au niveau TTL et présente une page verticale de plus de 5000 caractères sur une trame de mille lignes non-entrelacée, sans papillotement. Le tube cathodique de 38 cm est un tube de téléviseur adapté et la bobine de déviation est équipée d'un bobinage "pin-index" assurant une image sans distorsion. D'autres composants des circuits notamment les circuits intégrés de l'oscillateur de ligne et de trame, sont des composants standard de téléviseurs.

Nouvelle mosaïque photo-électrique pour la génération de configurations à haute résolution

La "puce" LiSA 512 est une mosaïque de cellules photocommutatrices à commande électronique pour la génération de configurations de haute qualité dans les imprimantes électrophotographiques. Elle se compose d'une mosaïque en zig-zag de 512 cellules photocommutatrices commandées par un champ magnétique extérieur. Toutes les fonctions de commutation et de commande sont assurées par des circuits intégrés extérieurs à la puce. Le dispositif peut recevoir des données d'entrée à la cadence de 1 Mbits/s, correspondant à une cadence maximale de génération de configurations de 2000 configurations/s.

Le silicium supérieur au germanium pour les diodes de détecteurs de rapport

On croît communément qu'il faut utiliser des diodes au germanium dans un détecteur de rapport, à cause de leur faible tension de seuil. Le présente article montre que cette croyance est sans fondement et ne s'est instaurée que parce que les chercheurs n'ont jusqu'ici considéré que l'amplificateur et le détecteur de fréquences intermédiaires pour établir les mérites comparés des diodes au germanium et au silicium. L'étude des performances globales d'un récepteur de radio entre l'antenne et la sortie audio a montré qu'un détecteur de rapport fontionne tout aussi bien avec des diodes au silicium qu'avec des diodes au germanium.

Circuit intégré d'interface pour transformateurs différentiels linéaires variables

Le circuit intégré NE5520 comprend un oscillateur à fréquence ajustable pour l'excitation d'un transformateur différentiel linéaire variable, et un démodulateur synchrone permettant d'obtenir un signal de sortie dont la polarité et l'amplitude indiquent la direction et la distance de déplacement du noyau mobile du transformateur.

Tube cathodique à correction magnétique et amplification du balayage

Les aimants permanents annulaires intégrés au canon à électrons d'un nouveau tube cathodique mono-accélérateur à 25 MHz augmentent la sensibilité de la déviation verticale, assurent la rotondité du spot et ameliorent son centrage. Le cône de l'enveloppe du tube de conception nouvelle simplifie grandement l'ajustement du blindage magnétique et peut également servir dans des tubes à fréquence élevée à post-accélération par grille bombée.

Circuitos integrados para decodificadores de diso compacto

El sistema digital de audio de disco compacto reproduce señales de audio de la más alta calidad. El ruido se reduce a un nivel despreciable utilizándo muestras de audio, digitales de 16 bits, y codificando los datos en discos compactos de forma que los errores debidos a la separación de los datos pueden ser detectados y corregidos. En vista de que se necesita un extenso procesado de la señal para la corrección de errores, se ha desarrollado una familia de circuitos LSI para poder fabricar a un precio efectivo decodificadores para el sistema de disco compacto. Este artículo describe un decodificador de este tipo con un sistema especial de conversión D/A, que trabaja con 16 bits sin utilizar un convertidor D/A de 16 bits.

Detectores de infrarrojos piroeléctricos

Los detectores de infrarrojos piroeléctricos responden a variaciones de la radiación de infrarrojos en la zona a la que están expuestos. Por lo tanto son muy adecuados para ser utilizados en detectores de incendios y de intrusos y en algunas aplicaciones pasivas de control. Añadiéndoles un troceador óptico también pueden ser utilizados en radiómetros industriales. Otra aplicación potencial es en el análisis de gas e identificación de polución.

Circuito integrado sintetizador para sintonía de radio

Si un receptor de radio tiene que llevar incorporada la posibilidad de búsqueda automática de sintonía y/o sintonizar entrando directamente en un teclado la frecuencia requerida, se puede generar una señal estable del oscilador local mediante síntesis indirecta de frecuencia con un PLL controlado por microordenador. Esta es la función del nuevo circuito LSI de 18 terminales SAA1057 que se describe en este artículo. El circuito SAA1057 y 16 componentes periféricos realizan las funciones de sintonía, que antes requerían el uso de tres circuitos integrados y 36 componentes periféricos. Se da una descripción y las características de funcionamiento del circuito básico.

Comparacion de pequeños condensadores electrolíticos

Los recientes progresos en las técnicas de fabricación de condensadores electrolíticos de aluminio sólido han logrado un considerable aumento de su eficiencia volumétrica. Aunque esta mejora ha logrado que se utilicen cada vez más en aplicaciones profesionales e industriales, las características de este tipo de condensador son generalmente menos conocidas que las de los electrolíticos húmedos y los de tántalo. Este artículo presenta una comparación, resaltando de este modo las claras ventajas de los condensadores electrolíticos de aluminio sólido.

Visualizador de página completa utilizando componentes de television

El visualizador descrito acepta una entrada de video a nivel TTL y presenta una página escrita con más de 5.000 conectores sobre una trama de mil líneas, no entrelazada y sin parpadeo. El TRC de 38 cm. es una adaptación de un tubo de imagen de televisión, y el yugo de desviación utiliza una técnica precisa de colocación de espiras del bobinado para asegurar una imagen sin distorsión. No se modifican los demás componentes de televisión, incluyendo los circuitos integrados oscilador de campo y de línea.

Sistema de conmutación de luz para generación de imágenes de alta resolucion

El circuito integrado LISA 512 es un sistema de conmutación de luz para la generación de imágenes de alta calidad en impresoras electrofotográficas. El circuito LISA 512 consta de un sistema de 512 células conmutadoras de luz controlada por un campo mágnetico extremo. Tanto la conmutación como las funciones de control son realizadas por circuitos integrados externos al chip. El sistema puede aceptar datos de entrada a una velocidad de 1 Mbits/s, que corresponde a una máxima velocidad de generación de 2000 imágenes/s.

Silicio en lugar de germanio para diodos detectores de relación

Es una creencia común que los diodos de germanio deben utilizarse en un detector de relación debido a su baja tensión umbral. Este artículo muestra que esta creencia es infundada y que ha surgido debido a que hasta ahora los investigadores sólo consideraban el amplificador de F.I. y el detector al declarar los méritos relativos de los diodos de germanio y silicio. Pero al medir el comportamiento total de una radio entre la antena y la salida de audio, se ha visto que un detector de relación funciona tan bien con diodos de silicio como con diodos de germanio.

Circuito integrado de acoplamiento para transformadores lineales diferenciales variables

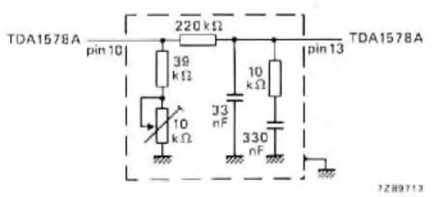
El circuito integrado NE5520 incorpora un oscilador de frecuencia variable para activar un transformador lineal diferencial variable, y un demodulador síncrono para obtener una salida cuya polaridad y amplitud indican la dirección el desplazamiento del núcleo móvil del transformador.

Tubo de rayos catodicos con cerreccion magnetica y amplificacion de exploracion

Los anillos de imán permanente incorporados en el cañón electrónico de un nuevo TRC mono-acelerador de 25 MHz aumentan la sensibilidad de desviación vertical, aseguran la redondez del punto luminoso del haz, y mejoran su centrado. El cono de la carcasa del tubo ha sido diseñado de nuevo, de modo que se simplifica en gran manera el posicionado del apantallamiento magnético.

CORRECTION

In the article 'Integrated circuits for hi-fi radio' in Vol. 4 No. 1 some of the component values in the economy PLL filter in Fig.9(d) on page 21 were incorrect. This drawing shows the correct values.



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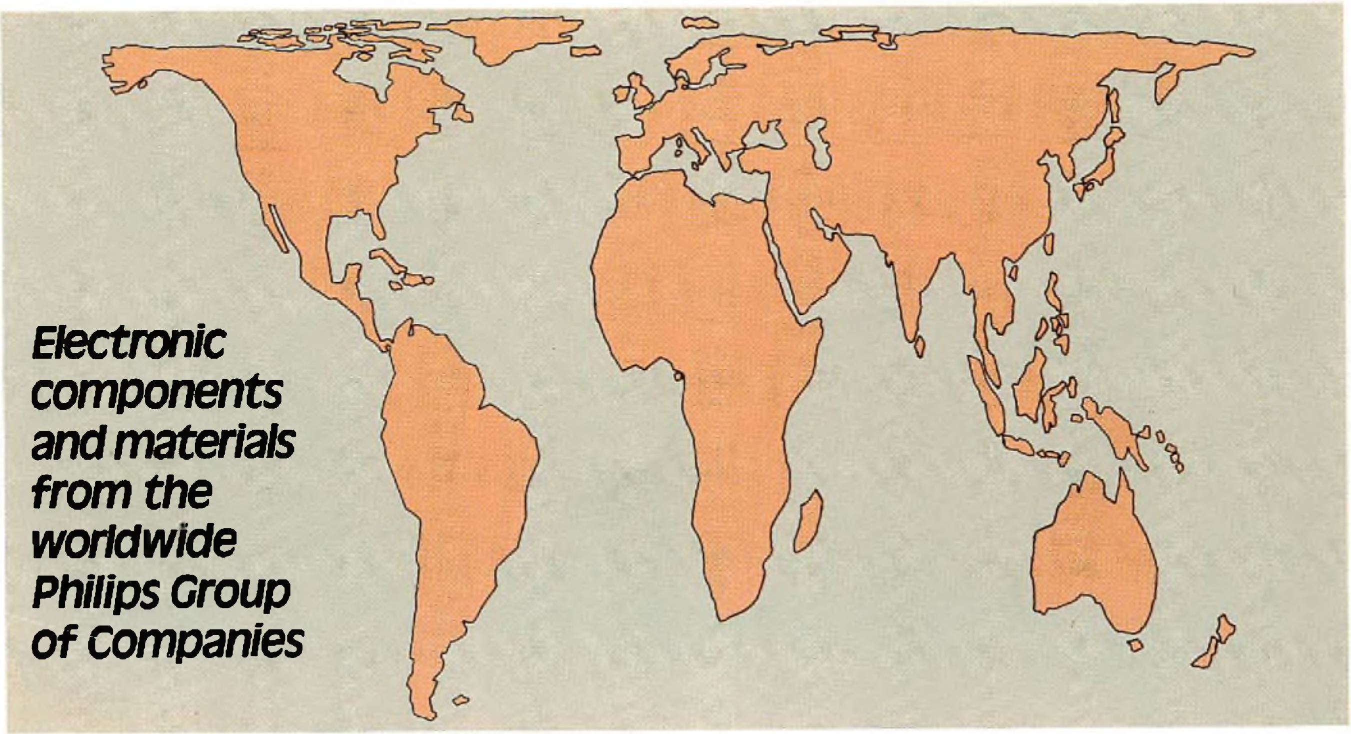

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