Electronic components & applications

Vol. 3, No
August 198

ISSN 0141–6219

Electronic components & applications

Volume 3, No. 4 August 1981

Contents

In the past half-century, and especially since the advent of the transistor, high-frequency radio communication has become an indispensable feature of industrial society. Without it, emergency vehicles would not reach emergencies in time, harbour and airport traffic would come to a standstill or worse, even the mundane taxi might not get you to the airport on time – no calamity, perhaps, since the plane would probably not leave on time either. Fortunately for all these essential services, the technology of reliable, high-frequency power transistors continues to make steady progress. Illustrative of recent advances, an article in this issue describes a transistor-equipped linear amplifier capable of delivering 400 W peak envelope power in the 1.6 to 30 MHz communications band.

Until now stereo broadcasting has been confined to frequency-modulated v.h.f. Amplitude-modulated medium wave is more accessible to car radios, however, and a standard soon to be adopted in the U.S.A. will provide for medium-wave a.m. stereo broadcasting. This article tells how a car radio described in EC&A Vol. 3 No. 2 can be adapted to decode stereo transmissions broadcast according to the expected standard.

A.M. stereo - a new dimension for car radios

W. JANSEN and W. KANOW

In many regions of the U.S.A., f.m. stereo radio broadcasts attract a wider audience than competing a.m. mono broadcasts. This is despite the poor penetration into fringe areas and multipath/fading problems which are particularly annoying to the many people who listen to the f.m. broadcasts on their car radios. It is not therefore surprising that car manufacturers, radio stations and car radio manufacturers all have an active interest in an early introduction of a.m. stereo broadcasts.

During the last four years, the Federal Communication Commission (F.C.C.) in the U.S.A. have been evaluating a variety of a.m. stereo transmission and reception systems proposed by Belar, Harris, Kahn/ Hazeltine, Magnavox* and Motorola. The results of this evaluation are expected to be announced during the next few months, and it is widely anticipated that the Magnavox system will be selected as the national standard for a.m. stereo radio.

An a.m. stereo radio should be compatible with present a.m. mono broadcast standards, and should achieve good stereo channel separation. It should also have an indicator lamp that rapidly and automatically identifies a.m. stereo broadcasts, and an a.g.c. characteristic which imparts the same tuning 'feel' as that experienced with a.m. mono radios. Furthermore, the components for the a.m. stereo circuits must be available as soon as the F.C.C. decision is announced, and the new circuitry must be inexpensive and easily adapted to existing radio designs.

* A North-American Philips company.

This article describes how one of the a.m./f.m. car radio concepts described in the February 1981 issue of EC&A † can be simply and economically adapted to receive a.m. stereo transmissions. The performance of the adapted radio is optimised for reception of a.m. stereo broadcasts according to the transmission standard proposed by Magnavox. A future article will describe how the system can be applied to hi-fi radios.

A.M. STEREO SYSTEMS

Transmission standards

All the proposed a.m. stereo systems use amplitude modulation for transmitting the mono L + R audio information so that the stereo transmissions can still be received, in mono, by conventional a.m. radios. The stereo L – R audio information and stereo identification signal (pilot tone) are additionally phase or frequency modulated onto the carrier. In future, the pilot tone may also be pulse-coded with slow-speed digital data regarding station identification, temperature and time. In an a.m. stereo radio, the amplitude-modulated L + R information is recovered in the normal way and the L – R signal and pilot tone are separately demodulated. The resulting stereo channel sum and difference audio signals are then dematrixed to produce the stereo audio outputs $0.5[(L+R)+(L-R)] = L$, and $0.5[(L+R)-(L-R)] = R$.

† Available as a separate publication.

The five proposed systems listed in Table 1 differ mainly in the way that the L – R information is modulated onto the carrier, and the frequency and modulation method used for the pilot tone. In the Magnavox system, the carrier is linear phase-modulated with the L – R signal (maximum modulation index = 1 radian) and with a sub-audible 5 Hz pilot tone which causes a carrier-wave frequency deviation of 20 Hz (constant modulation index $\Delta f_c/f_m = 4$ rad).

TABLE 1
Proposed a.m. stereo systems

system	L + R	L – R	pilot tone
Belar A.M./F.M.	a.m.	f.m. with 100 µs pre-emphasis	10 Hz
Harris CPM	a.m.	compatible phase multiplex	80 Hz
Kahn/Hazeltine ISB	a.m.	independent sideband	15 Hz
Magnavox A.M./P.M.	a.m.	linear p.m.	5 Hz
Motorola C-QUAM	a.m.	compatible quadrature a.m.	25 Hz

Comparison of f.m. and linear p.m.

Figure 1 compares the characteristics of f.m. with those of linear p.m. for two levels of modulation amplitude. Frequency modulation causes a carrier frequency deviation Δf_c which is directly proportional to the modulation amplitude and independent of its frequency. Linear phase modulation causes a carrier phase shift $\Delta\varphi$ which is also directly proportional to the modulation amplitude and independent of its frequency. However, since the carrier phase shift due to p.m. is $\Delta\varphi = \Delta f_c/f_m$, and it is independent of modulation frequency f_m, it must also cause a carrier frequency deviation Δf_c which is directly proportional to the carrier phase shift, and therefore also directly proportional to the modulation amplitude. It therefore follows that, for modulation of constant frequency, f.m. and linear p.m. are indistinguishable because they both cause a carrier frequency deviation which is directly proportional to the modulation amplitude. In both cases, the frequency deviation will be detected by an f.m. demodulator. If a band of modulation frequencies is considered however, the carrier frequency deviation caused by f.m. remains constant at all modulation frequencies, whereas that caused by linear p.m. increases with increasing modulation frequency. The treble frequencies of the f.m. demodulated linear p.m. signal will therefore be overemphasised. To achieve a flat audio frequency response, the signal resulting from demodulating linear p.m. with an f.m. detector must therefore be passed through a circuit with gain that decreases at 6 dB/octave with increasing audio frequency. An integrator is such a circuit.

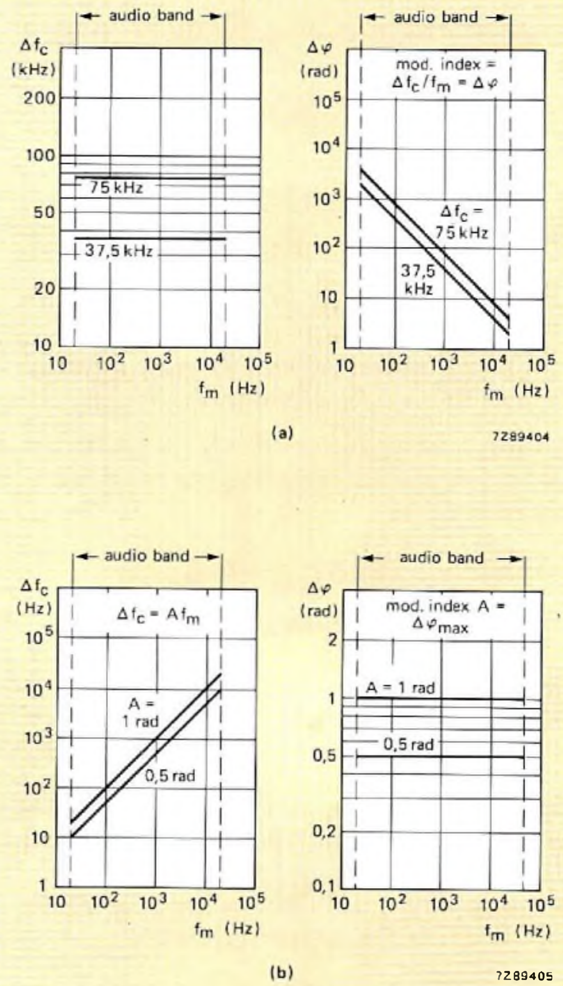


Fig.1 Comparison of f.m. and p.m. characteristics

(a) Frequency modulation characteristics for two amplitudes of modulation. The frequency deviation is directly proportional to the modulation amplitude and is independent of modulation frequency. The frequency modulation index is directly proportional to the modulation amplitude and inversely proportional to the modulating frequency.

(b) Linear phase modulation characteristics for two amplitudes of modulation. The carrier phase deviation ($\Delta\varphi_{max}$ = phase modulation index A) is directly proportional to the modulation amplitude and is independent of modulation frequency. The frequency deviation is directly proportional to both the frequency and amplitude of the modulation.

Detecting p.m. with an f.m. demodulator

If the phase angle of a transmitted carrier-wave varies in direct proportion to the amplitude of the a.f. modulation, the carrier is linearly phase-modulated according to the general expression:

$$e_c = \cos\,[\omega_c t + \Delta\varphi(t)]$$

in which $\omega_c t$ is the phase angle of the carrier and $\Delta\varphi$ is the phase angle deviation due to the modulation. For constant amplitude sinusoidal modulation, this can be expressed as:

$$e_c = \cos\,(\omega_c t + A \cos \omega_m t) \qquad (1)$$

where

$\omega_m = 2\pi f_m$, in which f_m is the modulation frequency
$\omega_c = 2\pi f_c$, in which f_c is the carrier-wave frequency
A = phase-modulation index
= peak instantaneous phase shift caused by the peaks of the constant amplitude sinusoidal modulation.

The associated carrier frequency deviation can be derived by simply differentiating the terms within the brackets of Eq.(1).

$$\omega_c(t) = d/dt\,(\omega_c t + A \cos \omega_m t)$$
$$= \omega_c - A\omega_m \sin \omega_m t \qquad (2)$$

From which the carrier frequency deviation that will be detected when the constant amplitude linear p.m. is f.m. demodulated is $\Delta f_c = Af_m$. Conversely of course, by integrating Eq.(2) for f.m., Eq.(1) for p.m. can be obtained. This confirms that the integral of the signal level obtained by f.m. demodulating linear p.m. is proportional to the carrier phase shift caused by the linear p.m. Linear p.m. can therefore be accurately demodulated by an f.m. detector followed by an integrator.

The Magnavox a.m. stereo transmission

As previously explained, the Magnavox proposal is for an a.m. stereo transmission in which the carrier-wave is amplitude-modulated by the L + R signal, and linear phase-modulated by the L – R signal and 5 Hz pilot tone. The equation for the transmitted signal is:

$$e_c = [1 + m\{L(t) + R(t)\}] \times$$
$$\times \cos\,[\omega_c t + A\{L(t) - R(t)\} + B \cos \omega_p t]$$

where

m = amplitude-modulation index for L + R, m_{max} = 100%
A = maximum phase-modulation index for L – R, A_{max} = 1 rad
B = phase-modulation index for constant frequency pilot tone
= $\Delta f_c/f_p = 20/5 = 4$ rad
$\omega_p = 2\pi f_p$, in which f_p is the pilot tone frequency (5 Hz)
$\omega_c = 2\pi f_c$, in which f_c is the carrier-wave frequency.

ADAPTING A CONVENTIONAL A.M./F.M. RADIO TO RECEIVE A.M. STEREO

An a.m./f.m. car radio with quadrature detector for f.m. was described in EC&A Vol.3, No.2, February 1981, pp 75-80. The remainder of this article shows how the following features of the integrated circuits used in that radio (TDA1072, TDA1575 and TDA1578) allow it to be very easily adapted for the reception of a.m. stereo transmissions according to the standards proposed by Magnavox.

A.M. channel TDA1072

- 90 dB a.g.c. control range holds the level of the a.m. i.f. signal almost constant over a broad range of r.f. input levels
- low audio output impedance (3.5 kΩ) allows simple connection of audio output to stereo matrix
- extremely low noise local-oscillator which is essential for eliminating spurious p.m. during a.m. stereo reception. The oscillator output level is controlled at a typical level of 140 mV up to a frequency of 50 MHz
- small spread of sensitivity.

F.M. channel TDA1576

- 4-stage limiter-amplifier provides 50 dB suppression of a.m. over most of the input signal range.
- signal-to-noise ratio 85 dB (Δf_c = 75 kHz). This low-noise performance is essential for reception of the p.m. stereo difference signal which causes only small deviations of the carrier frequency
- wide-range, logarithmic, field-strength-dependent control voltage with low spread can be used for automatic mono/stereo switching

- symmetrical f.m. demodulator allows simple addition of a.m. i.f. filter and only requires an integrator connected to its output to convert it into a p.m. demodulator for the a.m. stereo difference signal. The demodulator has two independent outputs, one of which can be used for the demodulated f.m. MPX signal and the other for the demodulated p.m. stereo difference signal. The additional integrator will not then influence the f.m. audio output.

PLL stereo decoder TDA1578

- the VCO of the PLL has a high output impedance, so it can be easily switched from 76 kHz (4 x f.m. pilot frequency) to 20 Hz (4 x a.m. stereo pilot frequency) by using a low-tolerance foil capacitor to lengthen the time-constant.
- pilot level detector with separate output allows recovery of possible low-speed digital data (time, temperature, station identification) pulse-coded onto the a.m. stereo pilot tone. The pilot level detector is separated from the output matrix by a muting circuit which is switched on during a.m. reception so that the pilot tone cannot reach the audio output
- two internal output op-amps with input and output pins can be used for simple active de-matrixing of the a.m. stereo signals by the addition of a few passive components
- the same stereo audio outputs and pilot lamp drive can be used for a.m. stereo and f.m. stereo so that switching of outputs during band selection is not necessary.

Required modifications

Figure 2 is a block diagram of the modified car radio. Only one quadruple op-amp, two FETs and a few passive components have been added to the original circuit to perform the following main functions:

- an op-amp increases the sensitivity of the a.g.c. applied to the TDA1072 to ensure good channel separation at all aerial input levels
- the ceramic i.f. filter originally used with the TDA1072 is replaced by a 3-stage LC filter to achieve a broad, flat response in the i.f. passband, good phase linearity and high stopband attenuation
- an op-amp integrator is connected to one of the outputs of the f.m. demodulator in the TDA1576, and a 455 kHz i.f. filter is connected in series with the 10.7 MHz f.m. i.f. filter. The TDA1576 is therefore a p.m. demodulator for the 455 kHz i.f. signal from the TDA1072 and an f.m. demodulator for the 10.7 MHz i.f. signal from the f.m. front-end

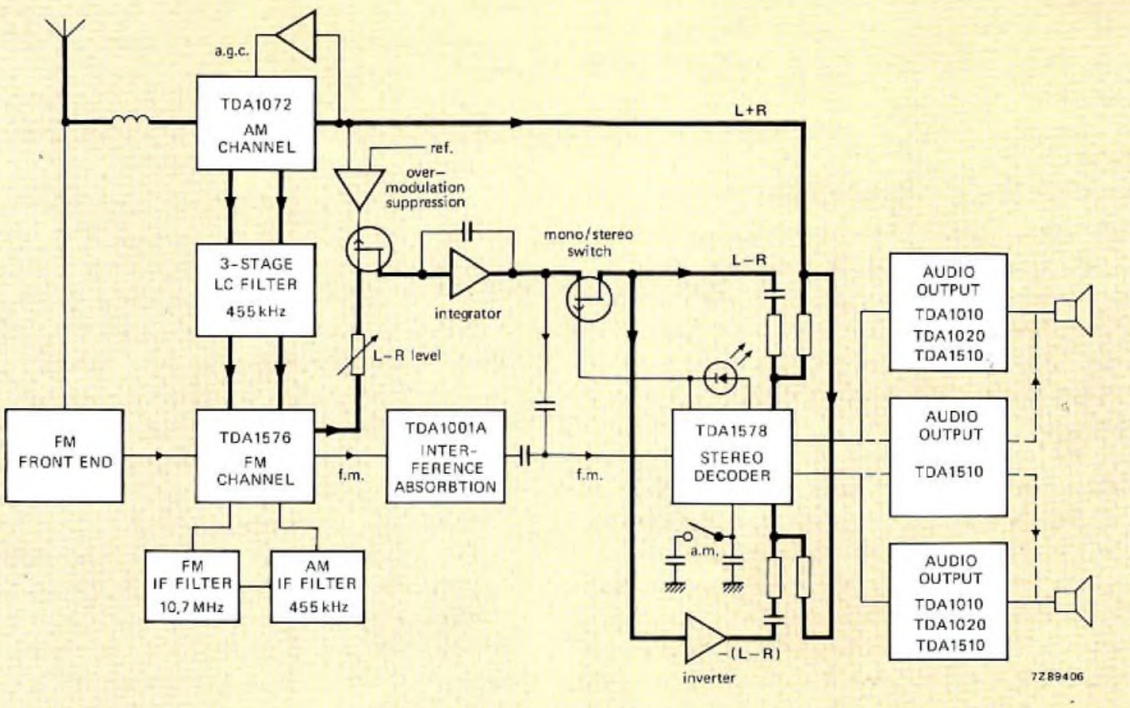


Fig.2 Block diagram of a car radio for f.m. stereo and a.m. stereo

- an op-amp controlled FET switch is inserted in the a.m. stereo difference output line from the TDA1576. The switch is controlled by the a.m. stereo sum signal from the TDA1072 and is adjusted to switch off if the depth of amplitude modulation exceeds 95% (negative)
- a FET-switch in the a.m. stereo difference output line from the op-amp integrator is controlled by the stereo lamp drive signal to effect automatic mono/stereo switching. Since the pilot level detector in the TDA1578 is controlled by a field-strength dependent signal from the TDA1576, this also influences the automatic mono/stereo switching. The reference level for the field-strength dependent signal is adjusted so that automatic mono to stereo switching for a.m. occurs when the signal-to-noise ratio is 30 dB
- an op-amp inverts the a.m. stereo difference signal so that it can be used for dematrixing the audio information for the right-hand channel
- a capacitor and a switch decrease the frequency of the VCO for the PLL pilot tone detector in the TDA1578 from 76 kHz to 20 Hz. A second capacitor and switch lengthen the time-constant of the pilot level detector
- the output amplifiers of the TDA1578 are used in an active matrix circuit which recovers the left and right-hand audio signals from the a.m. stereo sum and difference signals
- level adjustment of the demodulated stereo difference signal from the TDA1576 allows matching of the stereo sum and difference signal levels to minimise crosstalk

R.F. section and L + R demodulator

In Fig.3, the ceramic i.f. filter in the original peripheral circuit of a.m. channel TDA1072 has been replaced with a triple-tuned LC filter. For optimum stereo channel separation, this filter has flat response in the 10 kHz passband, good phase linearity and high stopband attenuation. A secondary winding on the final stage of the filter couples the 455 kHz i.f. signal to the input of the f.m. i.f. and demodulator circuit TDA1576 which demodulates the a.m. stereo difference signal. Since the TDA1072 has a wide a.g.c. control range, the high a.g.c. sensitivity required for a.m. stereo has been achieved by inserting an op-amp between the audio output at pin 6 and the a.g.c. input at pin 7. The amplified a.g.c. ensures that the relative amplitudes of the a.m. stereo sum and difference signals remain almost constant over a wide range of aerial input levels. This ensures accurate dematrixing of the a.m. stereo audio signals and consequent good channel separation. The time-constant of the low-pass audio filter at pin 6 (18 kΩ, 1.2 nF) has been carefully selected to compensate the a.m. stereo sum signal for the delay of the a.m. stereo difference signal caused by its additional processing in the p.m. channel. The delay compensation ensures that the channel separation remains independent of modulation frequency throughout the audio frequency range. To prevent spurious phase modulation of the i.f. signal, the local oscillator components between pins 11 and 12 of the TDA1072 must be isolated from possible mechanical vibration.

L – R demodulator

As previously explained, it is only necessary to pass the output from an f.m. demodulator through an integrator to convert it into a p.m. demodulator. This has been done in Fig.4 by passing the audio output from pin 9 of f.m. i.f. amplifier and quadrature f.m. detector TDA1576 through op-amp integrator OP1. During a.m. stereo reception, the composite 455 kHz i.f. stereo signal is derived from the secondary winding on the final stage of the a.m. i.f. filter connected to the TDA1072 (Fig.3) and fed to the input of the limiting amplifier at pin 15 of the TDA1576. The limiting amplifier suppresses the amplitude-modulated stereo sum signal, thereby ensuring that the TDA1576 only processes the p.m. stereo difference signal. The quadrature demodulator in the TDA1576 is tuned to the intermediate frequencies of both the f.m. and a.m. channel by the addition of a 455 kHz tuned-circuit in series with the existing 10.7 MHz tunedcircuit. Switching of the tuned circuits is not necessary because the power supply to the f.m. front-end is switched off during a.m. reception so that f.m. i.f. signals cannot reach the TDA1576.

Automatic mono/stereo switching and noise muting of the f.m. audio signals are functions of the field-strength dependent voltage level from pin 13 of the TDA1576. This level is referred to the voltage at pin 14. A switch is incorporated so that, for f.m., the level at pin 14 is adjusted so that the noise muting is -3 dB for an aerial input of 15 μV (signal-to-noise ratio = 50 dB). For a.m. stereo, the level at pin 14 is adjusted so that the mono to stereo switching, controlled by FET switch TR_1 (Fig.5), occurs when the aerial input level reaches 100 μV (signal-to-noise ratio = 30 dB).

To minimise crosstalk between the a.m. stereo audio channels, the level of the a.m. stereo difference signal is matched with that of the a.m. stereo sum signal by a variable resistor connected to the audio output line at pin 9 of the TDA1576. A d.c. bias network connected to the f.m. audio input to pin 1 of interference absorption circuit TDA1001A/V inhibits the input to the f.m. audio channel at this point during a.m. stereo reception.

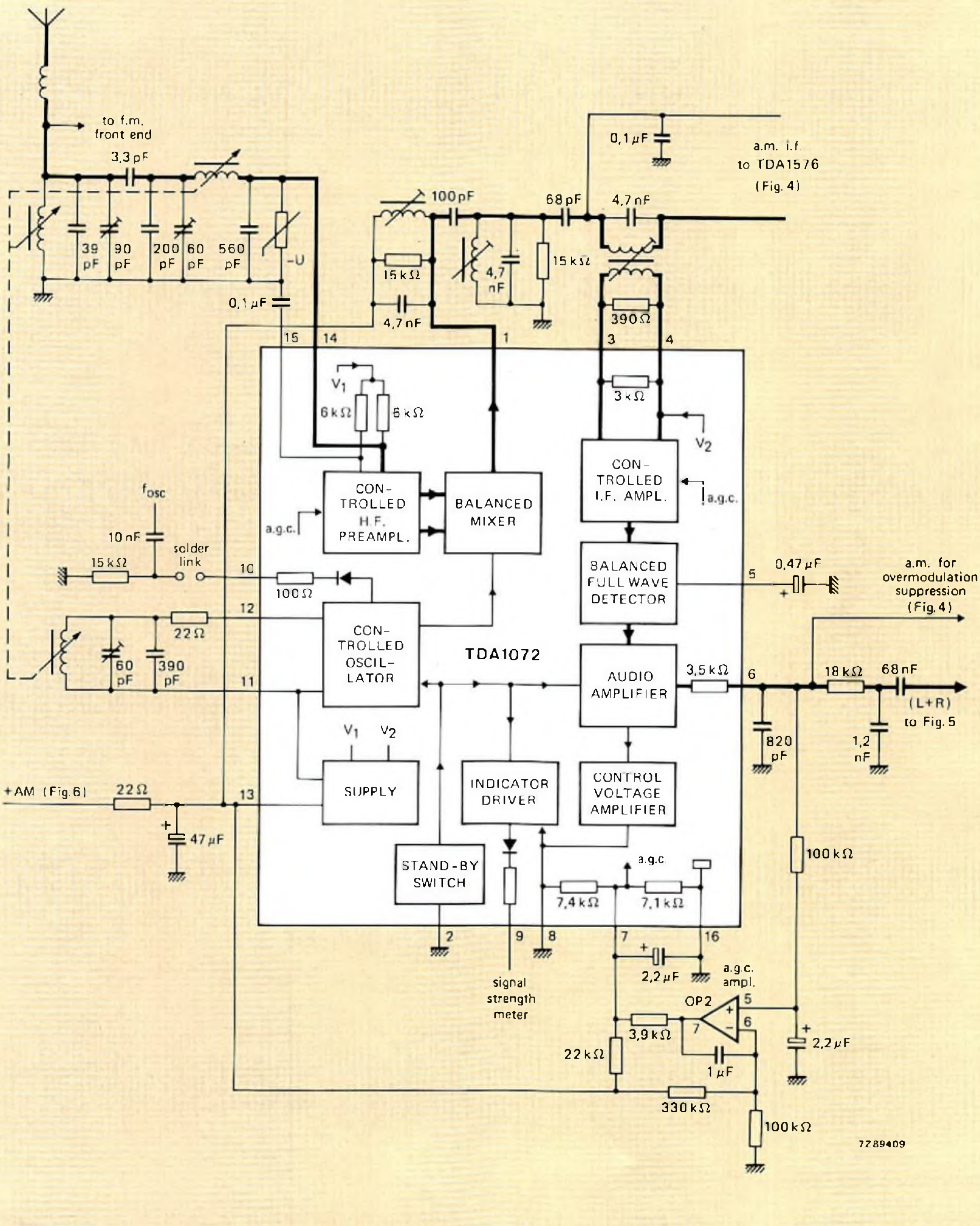


Fig.3 R.F. section and L + R demodulator

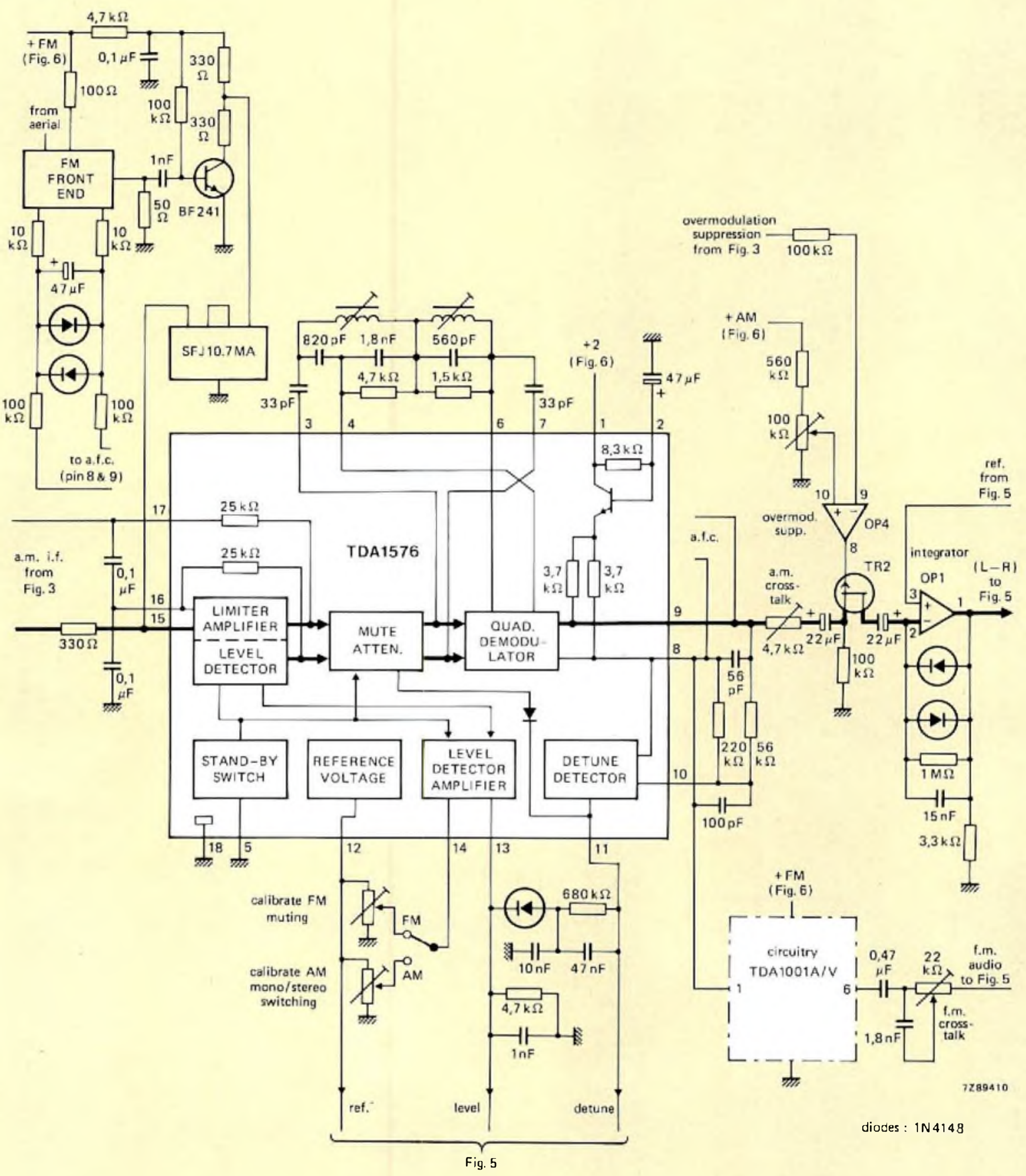


Fig.4 F.M. demodulator with output integrator for linear p.m. demodulation

Suppression of noise on the L – R audio signal

Selective fading, rapid tuning or overmodulation of the a.m. signal can cause considerable noise on the a.m. stereo difference signal at pin 9 of the TDA1576. It is essential to suppress this noise. As shown in Fig.4, a FET switch (TR_2) is connected between the stereo difference audio output at pin 9 of the TDA1576 and the input to integrator OP1. TR_2 is driven by comparator OP4 which causes the switch to open if the peak negative amplitude of the stereo sum audio signal at pin 6 of the TDA1072 (Fig.3) exceeds a predetermined level set by the 100 kΩ potentiometer connected to the non-inverting input of OP4. The potentiometer is adjusted so that TR_2 ceases to conduct when the stereo sum signal amplitude is sufficient to modulate the carrier to a depth of 95% (negative). TR_2 therefore interrupts the stereo difference signal for the duration of the noise and, assuming the interruption is short, the output from integrator OP1 remains constant during this period.

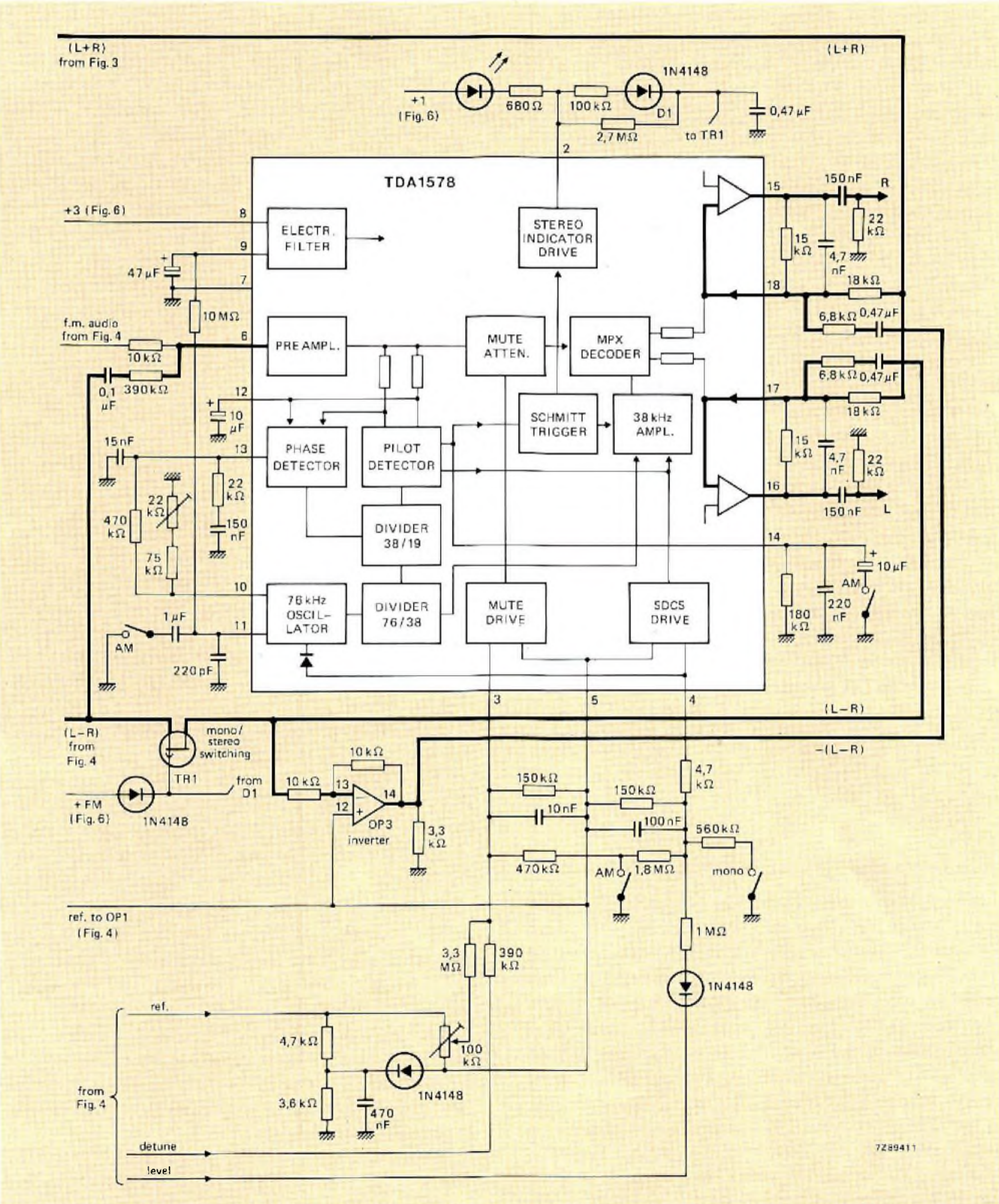


Fig.5 F.M. stereo decoder with automatic mono/stereo switching and dematrixing circuits for a.m. stereo

Pilot tone detection

For identification of a.m. stereo transmissions and for automatic mono/stereo switching, the transmitted a.m. stereo signal proposed by Magnavox incorporates 5 Hz phase-modulation with a constant modulation index of 4 radians. This is equivalent to an f.m. pilot tone which causes a carrier-wave frequency deviation of 20 Hz. Figure 5 shows how the peripheral circuitry of the stereo decoder TDA1578 has been modified so that the PLL pilot detector can detect the conventional 19 kHz f.m. pilot tone or the 5 Hz p.m. pilot tone. The pilot tone signal is connected to pin 6 of the TDA1578. For a.m. stereo, the time constant of the pilot level detector is increased by the capacitor and switch at pin 14. The frequency of the VCO is decreased from 4 x 19 kHz = 76 kHz to 4 x 5 Hz = 20 Hz by the addition of the switch and capacitor at pin 11. Due to the high impedance of the internal VCO, it is possible to use a low-tolerance foil capacitor instead of an electrolytic at pin 11 so that trimming of the VCO frequency for a.m. reception is not necessary.

During a.m. reception, the mute attenuator in the TDA1578 is activated by the switched resistor at pin 3 to prevent the a.m. stereo pilot tone reaching the output and causing interference (2 x 5 Hz). The a.m. stereo pilot tone will, however, operate the internal schmitt trigger to drive the stereo lamp connected to pin 2. The same lamp can therefore be used to indicate the reception of a.m. or f.m. stereo transmissions.

If, in future, the a.m. stereo pilot tone is pulse-coded with slow-speed digital data, this can be detected at the pilot level detector output at pin 14 of the TDA1578.

Automatic mono/stereo switching

Figure 5 shows that, during f.m. reception, automatic mono/stereo switching is controlled by the signal-dependent channel separation (SDCS) block in the TDA1578 which indirectly controls the MPX decoder. Since the MPX decoder in the TDA1578 is not used for a.m. stereo decoding, an external automatic mono/stereo switch must be provided. This is the function of FET switch TR_1 connected in the a.m. stereo difference input line to op-amp inverter OP3. During f.m. reception, TR_1 is switched off by the FM power supply. During a.m. stereo reception, the pilot detector in the TDA1578 switches TR_1 on via the stereo lamp drive from pin 2. The stereo difference signal is then fed directly to the stereo combining network for the left-hand channel at pin 17 of the TDA1578 and, via inverter OP3, to the stereo combining network for the right-hand channel at pin 18 of the TDA1578. Capacitive coupling of the signals to pins 17 and 18 suppresses the 5 Hz pilot tone. Since the stereo sum signal is also present at pins 17 and 18 of the TDA1578, the radio operates in the stereo mode. When TR_1 is switched off (pilot tone not detected), only the stereo sum signal is applied to the stereo combining networks and the radio operates in the mono mode. Since the pilot detector in the TDA1578 is also controlled by the field-strength dependent voltage from the TDA1576, this mono/stereo switching is also a function of the level of the received signal. The gate drive to TR_1 is applied via a double RC and diode network which causes stereo to mono switching (positive-going drive) to be less delayed than mono to stereo switching (negative-going drive). This minimises noise caused by frequent mono to stereo switching due to field-strength fluctuations.

Extracting the left and right channel audio signals

To extract the left-hand and right-hand audio signals, the stereo sum and difference signals must be applied to two active combining networks. Figure 5 shows how these combining networks are easily formed by using the two output op-amps of the TDA1578 together with a few passive components.

The required transfer function for dematrixing the left-hand audio channel signal is $0.5[(L+R)+(L-R)] = L$. To achieve this, the L + R signal from pin 6 of the TDA1072, and the L – R signal from the drain of TR_1, are summed vectorially at the amplifier input at pin 17 of the TDA1578. The gain of the amplifier is adjusted by external resistors to derive the left-hand channel signal at pin 16.

The required transfer function for dematrixing the right-hand audio channel signal is $0.5[(L+R)-(L-R)] = R$. To achieve this, the L – R signal from the drain of TR_1 is inverted by op-amp OP3 to produce a $-(L-R)$ signal. This signal and the L + R signal from pin 6 of the TDA1072 are then summed vectorially at the amplifier input at pin 18 of the TDA1578. The gain of the amplifier is adjusted by external resistors to derive the right-hand channel signal at pin 15.

Since the power supply to a.m. channel TDA1072 is switched off when the f.m. band is selected, the a.m. signals are inhibited and there is no need to switch the output combining networks during a.m./f.m. band selection.

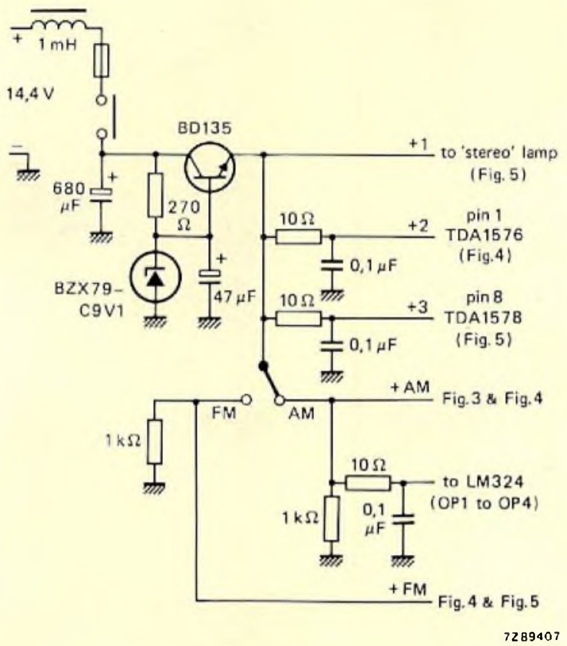


Fig.6 Power supply for the f.m. stereo/a.m. stereo radio

PERFORMANCE OF THE A.M. CHANNEL

General

supply voltage range	10.5 to 16 V
operating ambient temperature range	−30 to +80 °C
frequency range	510 to 1620 kHz
intermediate frequency	455 kHz

A.M. characteristics

V_{supply} = 14.4 V, T_{amb} = 25 °C, f_0 = 1 MHz, m = 0.3, f_{mod} = 400 Hz unless otherwise specified. Dummy aerial as shown in Fig.7.

aerial input voltage (mono mode)	
for (S + N)/N = 6 dB	6 µV
for (S + N)/N = 20 dB	27 µV
for (S + N)/N = 26 dB	64 µV
signal-to-noise ratio (V_{in} = 1 mV)	
mono mode	47 dB
stereo mode	42 dB
total harmonic distortion (m = 0.8)	
over most of the a.g.c. range	0.5%
r.f. signal handling capability (m = 0.8)	
with t.h.d. = 10%	2.1 V
stereo channel separation	
300 Hz to 2 kHz	30 dB

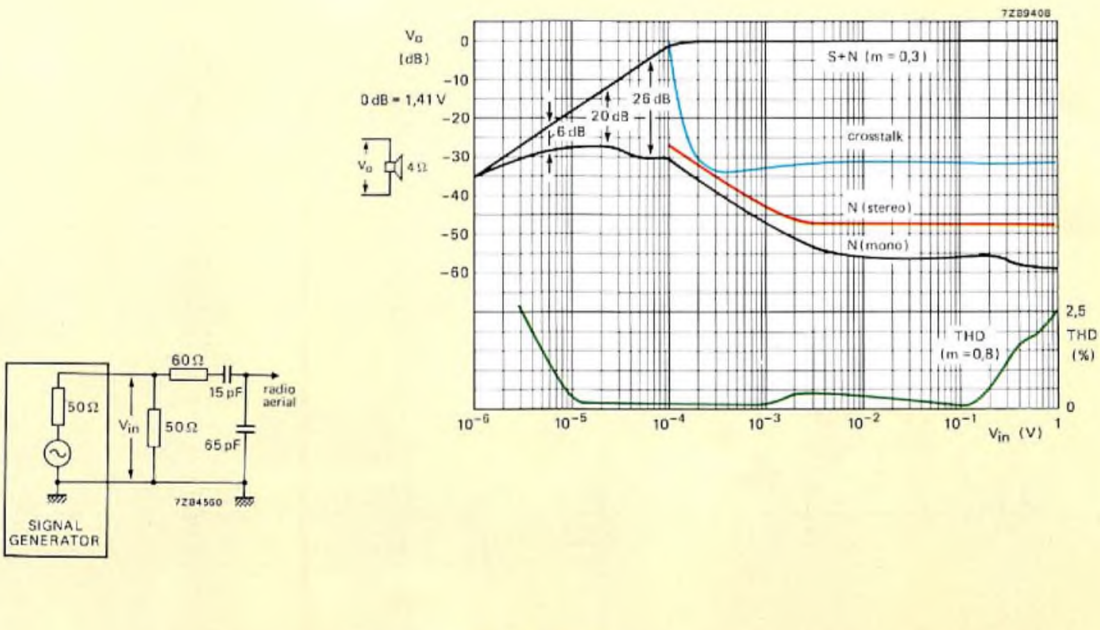
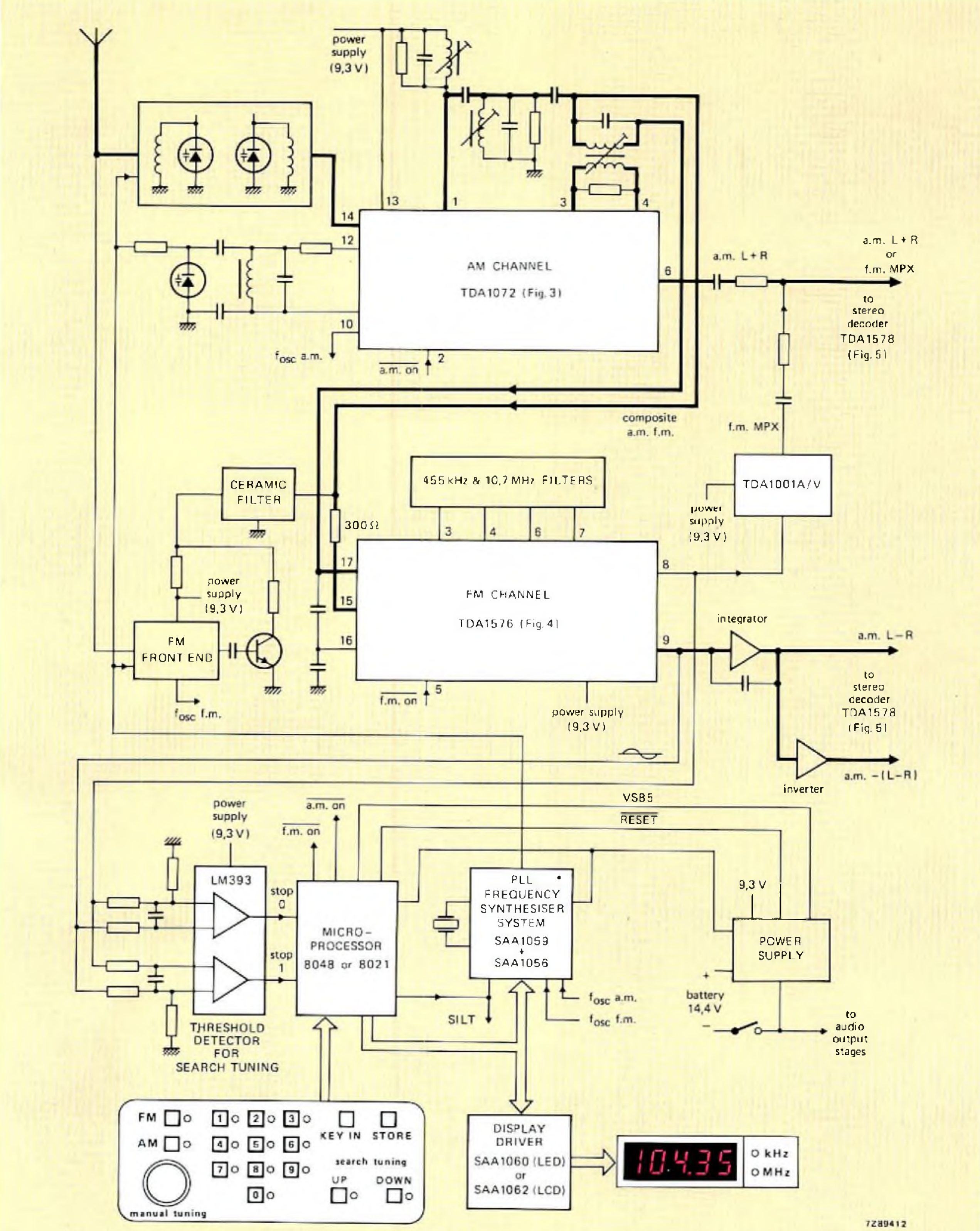


Fig.7 A.M. channel performance



*SAA1059 + SAA1056 will be replaced by SAA1057 (single-chip PLL frequency synthesiser)

Fig.8 Microprocessor-controlled car radio for f.m. stereo and a.m. stereo with simplified signal path

MICROPROCESSOR-CONTROLLED CAR RADIO FOR A.M. STERO AND F.M. STERO

Figure 8 shows how the previously described car radio can be microprocessor controlled. The tuning system uses a phase-locked loop frequency synthesiser to maintain stable, accurate tuning of the a.m. or f.m. channel. The local-oscillator signals from the a.m. channel TDA1072 and the f.m. front-end are passed to a frequency divider with a programmable division ratio in the PLL synthesiser. The output from the divider is compared with a crystal-controlled reference frequency. The result of the comparison, which represents the tuning error, is amplified and filtered before being used to modify the tuning voltage applied to the variable-capacitance diodes in the appropriate local-oscillator tuned circuit. The listener can change the tuned frequency by keying-in the required broadcast frequency, or by manual or search tuning. These commands cause the microprocessor to calculate the appropriate division ratio for the programmable divider after adding or subtracting the i.f. of the a.m. or f.m. channel. The stop pulses which are required when a station is detected during search tuning are generated in a very simple manner for both a.m. and f.m. by passing the S-curve at the output of the TDA1576 through the LM393 threshold detector.

FURTHER INFORMATION

For further information regarding the integrated circuits discussed in this article, refer to our data handbook 'Bipolar ICs for radio and audio equipment'. Loose data sheets are available for the more recently released integrated circuits.

ACKNOWLEDGEMENT

The authors wish to express their thanks for the assistance given by their many colleagues during this project. Particular acknowledgement is due to Bahne Peter Bahnsen of the Valvo application laboratory, Dr. Wolfgang E. Nolde of the development group for bipolar ICs for radio and industrial applications, and Dietmar Beer of Signetics.

For large capacitance in a small volume, electrolytics are the only choice. Amongst the solid electrolytic capacitors preferred for industrial applications, tantalum has held the lead for compactness. Thanks to recent advances in etching technology, aluminium, which has practical and cost advantages over tantalum, is now in a position to challenge that lead.

Advances in solid-aluminium electrolytic capacitor technology

E. H. L. J. DEKKER and H. SCHMICKL

The technology of 'solid' aluminium electrolytic capacitors is backed by several decades of manufacturing experience. Because of their well-earned reputation for reliability, long life, and resistance to temperature extremes, they have captured a secure position in professional and industrial electronics. Our 121-series axial-lead solid aluminium capacitors comply with CECC detail specification 30 302-001 and have gained the approval of British Telecom, FOA/FTL Sweden, and the Ariane space project. In comparison with more expensive tantalum capacitors their only shortcoming is that they are bulkier.

New developments in aluminium etching have now eliminated that shortcoming. Our 123-series solid aluminium electrolytics are roughly half the size of the 121-series; comparable, that is, to axial-lead tantalum capacitors of the same CV product. Moreover, the foil-winding technique used in the 123-series holds promise of extending solid aluminium capacitors into the still higher CV range that until now has been the exclusive domain of wet electrolytics.

CONSTRUCTION

In a solid aluminium capacitor a capacitive element consists of five layers (Fig.1). Layer 1, the anode, is high-purity aluminium foil about 100 μm thick. Layer 2, on which the capacitance and other electrical properties critically depend, is the deeply etched and oxidised surface of layer 1; the aluminium oxide that lines the etched pores is the dielectric. Layer 3 is a layer of manganese dioxide (MnO_2), embedded in a glass-fibre web, that penetrates and interconnects the pores. Layer 4 is similar to layer 2 but less deeply etched and oxidised. Layer 5, the cathode, is the aluminium foil on which layer 4 is etched.

In the equivalent circuit shown in the figure,

R_1 = resistance of the anode foil	*layer 1*
R_2 = resistance of the anode oxide coating C_A = capacitance of the etched anode surface	*layer 2*
R_3 = resistance of the MnO_2 matrix	*layer 3*
C_C = capacitance of the etched cathode surface R_4 = resistance of the cathode oxide coating	*layer 4*
R_5 = resistance of the cathode foil	*layer 5*

The equivalent circuit applies not only to a single cross-sectional element but also to the capacitor as a whole.

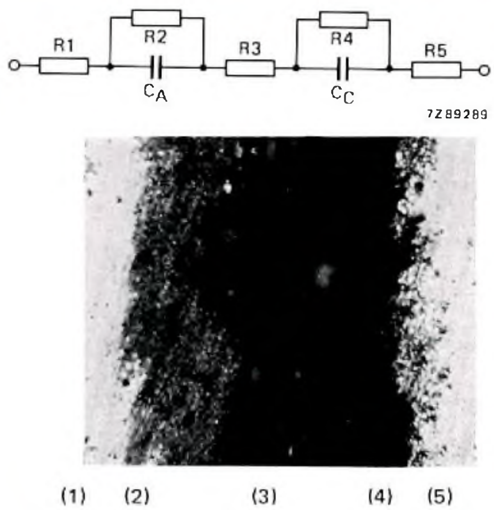


Fig.1 Photomicrograph (about 100x) of a single cross-sectional element of a solid aluminium capacitor, and its equivalent circuit: (1) anode foil; (2) oxide layer; (3) manganese dioxide; (4) oxide layer; (5) cathode foil

The construction of the 123-series capacitors (Fig.2) is the same as for the 121-series. The anode and cathode foils, together with the glass-fibre web, are wound on a solid aluminium pin to which the anode foil is welded; the tinned copper anode lead is welded to the end of the pin. Connection between the cathode foil and the case is made by an aluminium strip, welded at both ends; the cathode lead is welded to the outside of the case. Experience has shown that welding gives better reliability and long-term stability than any other means of connection.

To withstand the high temperatures that occur during manufacture (several hundred degrees Celsius), the disc insulator that seals the anode end of the case is ceramic; the same consideration governed the choice of glass-fibre for the internal spacer web.

Table 1 lists the capacitance, rated voltages, and corresponding case sizes of the 123-series. Note that at a rated voltage of 6.3 V a capacitor of 1000 μF can now be had in a size 6 case, as compared with only 330 μF in the 121-series or in axial-lead tantalum. Table 2, together with Fig.3, gives the outline and detail dimensions of the cases.

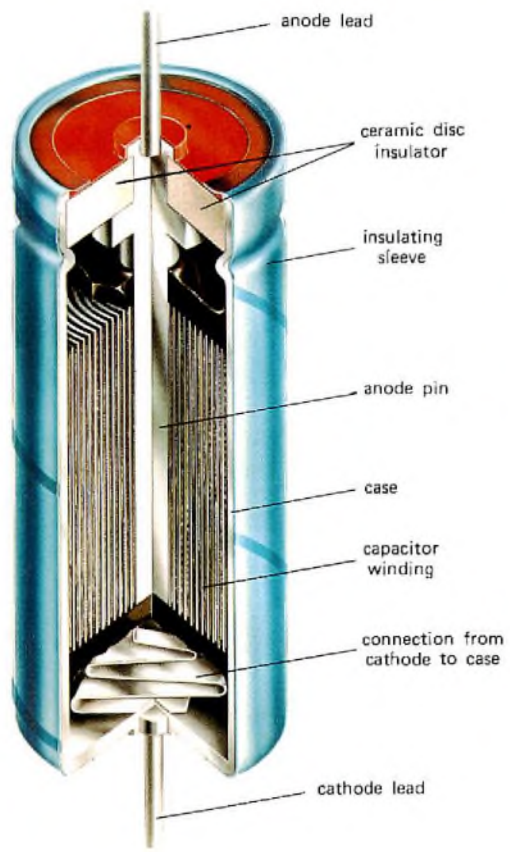


Fig.2 Cut-away view of a 123-series capacitor

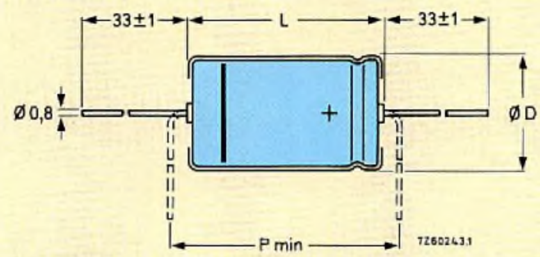


Fig.3 Case outline, 123-series capacitors; see also Table 2

TABLE 1
Capacitances, rated voltages and case sizes of 123-series capacitors

C_{nom} (μF)	U_R (V) 6.3	10	16	25	40
2.2					1
3.3					1
4.7					1
6.8					1
10			1	1	2
15			1		2
22			1	2	3
33		1	2	3	4
47	1	2	3	3	4
68		2	3	4	5
100	2	3	4	5	6
150	3	4	5	6	
220		4	6		
330	4	5			
470	5	6			
680	6				
1000	6				

TABLE 2
Case dimensions (see Fig.3)

size	D_{nom} (mm)	L_{nom} (mm)	P_{min} (mm)
1	6.5	17	20
2	6.5	22	25
3	8	22	25
4	10	22	25
5	10	31	35
6	12.5	31	35

Fig.4 Electron photomicrograph (about 60 000x) showing oxide layer in one of the anode pores

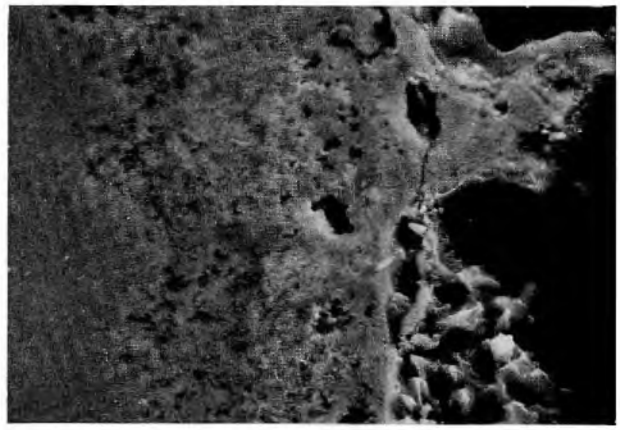


Fig.5 Electron photomicrograph (about 1600x) of a section of the anode foil of a fully formed capacitor

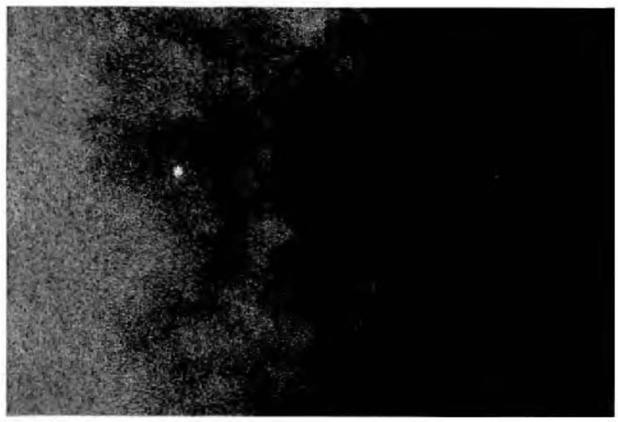


Fig.6 Elemental distribution map showing the aluminium distribution in Fig.5

Fig.7 Elemental distribution map showing the manganese distribution in Fig.5

MANUFACTURE

The etched surfaces of the foils are oxidised by anodising, the thickness of the oxide layers being controlled by controlling the forming voltage. Formation of the oxide layer necessarily shrinks the pores (Fig.4), an effect that has to be allowed for during initial etching.

The manganese dioxide layer is formed by pyrolysis of manganese nitrate. The first phase of pyrolysis releases water of crystallisation:

$$Mn(NO_3)_2 \cdot 4H_2O \rightarrow Mn(NO_3)_2 + 4H_2O\uparrow$$

The second phase decomposes the nitrate and oxidises the bivalent manganese, liberating nitrogen dioxide:

$$Mn(NO_3)_2 \rightarrow MnO_2 + 2NO_2\uparrow$$

After pyrolysis the oxide layer on the anode is re-formed by a current that flows preferentially through low-resistance spots to restore the layer there to the required thickness.

The 'dry' capacitance at this stage is appreciably smaller than the 'wet' capacitance after initial forming. This is because the manganese dioxide occupies less volume than the manganese nitrate from which it is derived, and therefore makes less contact with the pores (Fig.5, 6, 7). The lessened contact also increases the equivalent series resistance (and hence impedance) and the loss angle. Pyrolysis is therefore repeated several times until the manganese dioxide is fully taken up in the pores. This requires close control of such process parameters as heating and cooling rates and air circulation.

DATA

Except for size, the data of the 123-series are similar to those of the 121-series. In terms of the rated voltage U_R,

- permissible peak voltage is 1.15 U_R
- permissible reverse voltage is 0.3 U_R
- permissible a.c. voltage (50 to 100 Hz, sinewave) is 0.8 U_R.

At frequencies above 100 Hz the permissible a.c. voltage depends on the current. As impedance decreases, the heat developed in the capacitor sets a limit to the permissible a.c. load; despite the smaller size, this limit is the same as for the 121-series. Measurements to support even more generous a.c. limits are still in progress. The loss angle at 100 Hz is the same as for the 121-series; it depends mainly on the thickness of the anode oxide layer, and hence the rated voltage. At r.f. the impedance is determined mainly by the manganese dioxide in the pores; the deeper the etching the higher the impedance. Leakage current is the same as for the 121-series; it depends on the CV product, and hence on the quality of the oxide coating.

Solid aluminium electrolytics are noted for long life; the 123-series is no exception. In contrast to tantalum capacitors which are susceptible to field crystallisation, there is no known intrinsic breakdown mechanism. Electrical data remain unchanged throughout life. Products that are good to begin with remain good. Catastrophic failures – either short or open circuit – do not occur. The following high-temperature life expectancies based on many years' experience with the 121-series can be taken as indicative for the 123-series:

$\geqslant$20 000 h at 125 °C

$\geqslant$5 000 h at 150 °C

$\geqslant$2 000 h at 175 °C

The capacitors can be safely used at 125 °C without voltage derating. The permissible operating temperature range from –80 °C to 175 °C exceeds all known specifications for capacitors of comparable CV product.

APPLICATIONS AND FUTURE DEVELOPMENTS

123-series capacitors are particularly suited to applications calling for long life and high reliability under arduous conditions: telephony, aeronautics and astronautics, and the motor car industry. As coupling, decoupling, filtering and smoothing capacitors they can replace the 121-series with considerable savings of space, and tantalum capacitors with considerable savings of cost. Their reverse voltage and a.c. voltage characteristics are advantageous in electronic measuring equipment. Unlike tantalum capacitors, they do not require a 3Ω/V current limiting resistor.

The technology used in the 123-series has considerable further potential; extension of the capacitance range can be expected. Although solid aluminium capacitors cannot yet match the CV/volume ratio of wet electrolytics, their high reliability, long life, and resistance to temperature extremes will continue to open new fields of application to them. And the fact that the materials of their manufacture are comparatively cheap and plentiful will continue to give them a significant cost advantage over tantalum.

The diversity of tasks, channel frequencies, and types of modulation to which communications transmitters must nowadays be adaptable makes linearity of the power amplifier stages more important than ever. This article describes a linear power amplifier capable of delivering 400 W peak envelope power in the 1.6 to 30 MHz band with intermodulation distortion less than −26 dB. BLW96 planar interdigitated transistors are used in the output stage.

Wideband linear amplifiers in h.f. communications

J. LING

The h.f. communication portion of the radio spectrum covers the range 1.6 to 30 MHz and is utilised for a wide variety of services including short-wave broadcasting and point-to-point communications such as ship-to-ship and ship-to-shore, and for aircraft and national diplomatic traffic. Both speech (telephony) and data (telegraphy) are carried and a variety of modulation techniques are used; for example, amplitude modulation (a.m.), keyed carrier (A1), frequency shift keyed (f.s.k.), and single-sideband (s.s.b.). As the h.f. spectrum is fully utilised, especially at its lower end, it is of great importance that channel bandwidths are minimised and that the spectrum is used efficiently. The class of modulation must therefore be carefully selected to suit the type of traffic (Ref.1).

Modern h.f. communications equipment is often designed to transmit and receive more than one class of modulated signal and may even be required to operate with different types of modulation on each sideband, say speech on one and data on the other. In addition, because the quality of reception on some h.f. links varies considerably as a result of propagation conditions, it is desirable for transmitter/receiver systems to be able to switch rapidly from a poor-quality channel to a better one and to have sufficient flexibility to operate anywhere in the band. Equipment versatility is therefore essential.

The use of a wideband linear power amplifier in the final stages of a transmitter (Fig.1) is well established. Although all types of modulated signal do not necessarily require good linearity (f.m. and c.w., for example), it is still advantageous to use a linear amplifier for all modulation systems because of the complexity inherent in modifying the biasing and drive levels when switching from one type of modulated signal to another. In the case of single-sideband operation, the linearity requirement is determined by the need to amplify the transmitted signal and at the same time keep intermodulation distortion to a minimum. With low intermodulation distortion, it is practical to use the bandwidth occupied by one a.m. signal for two independent non-interfering s.s.b. signals of equivalent bandwidth.

Although tubes are still used in the power stages of larger transmitters (> 1 kW), transistors are now widely used in smaller equipments. Such transistors, although not capable of producing powers comparable with specialised transmitting tubes, can supply more than adequate power for many applications. They also have inherent wideband capability and can operate from low-voltage supplies such as vehicle batteries. Furthermore, with the increasing trend towards modular amplifier design (Fig.2), transistors can now be combined to produce total powers well in excess of 1 kW. One of the most advanced transistors of this type is the BLW96, which is capable of supplying up to 200 W p.e.p. with intermodulation distortion better than −30 dB.

In this article some of the factors governing the design of wideband linear amplifiers based on high-power h.f. transistors are outlined. As an example, a practical design using two BLW96 transistors in class AB, operating from a 50 V supply, and delivering up to 400 W p.e.p. with intermodulation distortion better than −26 dB, is described.

R.F. POWER TRANSISTORS

Silicon r.f. power transistors (such as those shown in Fig.3) represent a particularly specialised form of

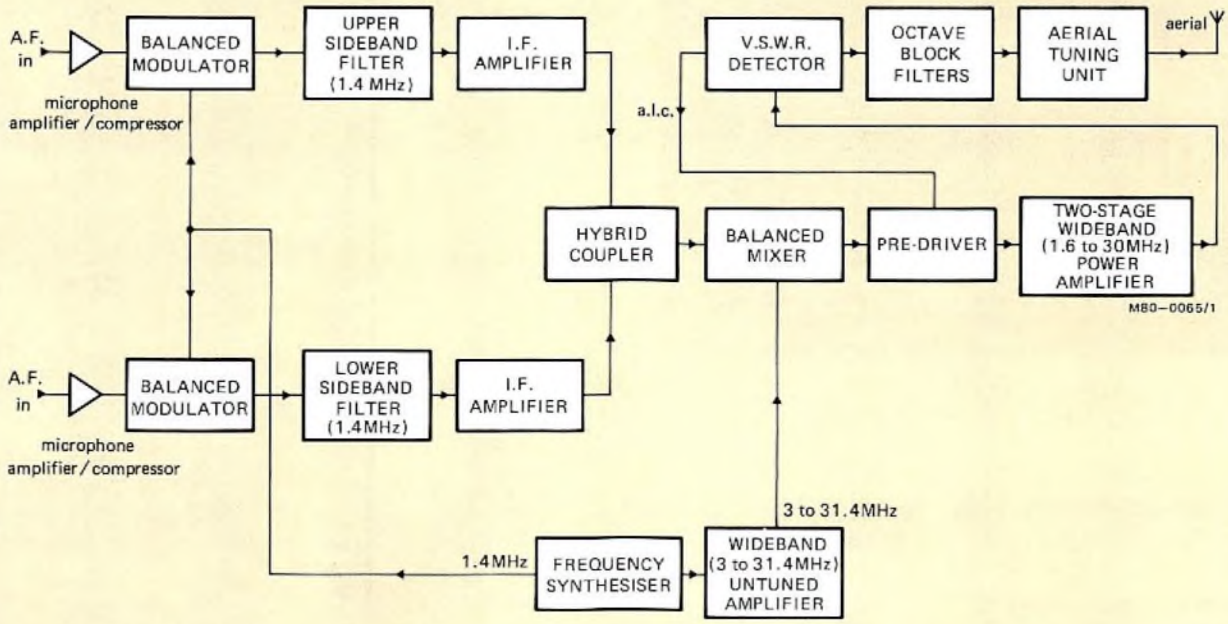


Fig.1 Typical independent sideband transmitter layout

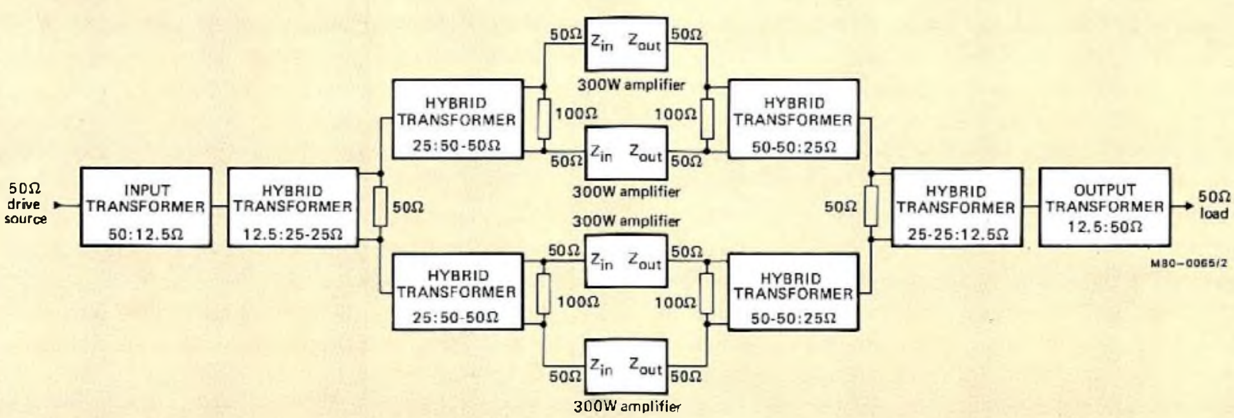


Fig.2 Modular construction of power amplifiers

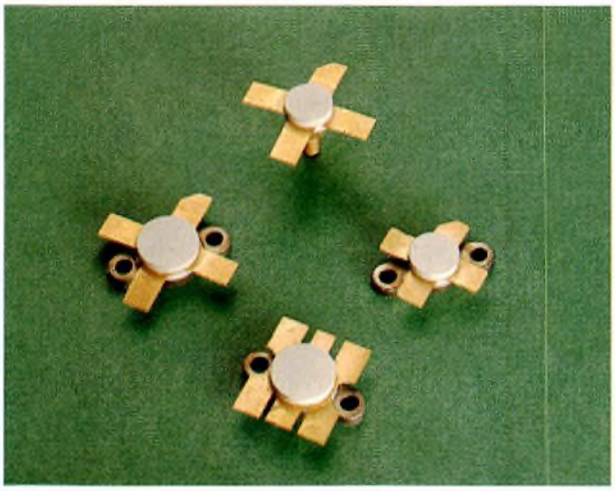


Fig.3 R.F. power transistors

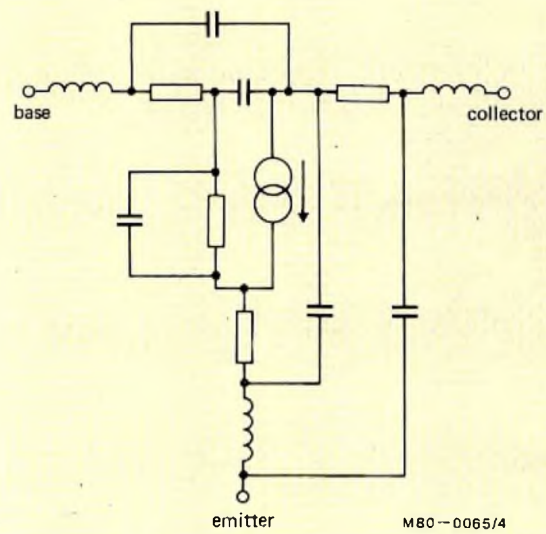


Fig.4 Large-signal equivalent circuit of r.f. power transistor (see Ref.2)

discrete semiconductor. For use in wideband linear amplifiers, the requirements of such transistors are:

- low thermal resistance;
- high cut-off frequency, ideally independent of collector current, ensuring good linearity up to high powers;
- high reliability under all modulation conditions.

To retain power handling capability at r.f., these transistors are constructed using a planar geometry in which the ratio of emitter periphery to emitter area is made as large as possible, consistent with practical photolithographic limits. The geometry is also determined by the need to conduct heat away from the chip efficiently and to maintain an even temperature over the entire area of the chip. For these and other reasons, an r.f. power transistor chip is diffused in the form of a large number of discrete separately-fed transistor elements, each with its own diffused ballast resistor. This prevents current crowding and consequent breakdown. The chip is generally mounted on a beryllia disc, providing electrical insulation from the header with an acceptably low thermal resistance and allowing the device to be used in a common emitter configuration. To ensure full r.f. performance with the device in circuit, parasitic inductances are minimised by the use of multiple internal bonding wires and low-inductance radial strips as external connectors. Two or more emitter connections are provided, again to limit parasitic resistance and inductance to earth.

An equivalent circuit of an r.f. power transistor is shown in Fig.4. This is similar to a conventional small-signal model but is applicable to large-signal class B or class AB operation. With the aid of computer analysis, such a circuit is used in amplifier design to predict the input impedance and gain when operating at the required power level into the chosen load impedance, (Refs. 2 and 3).

BLW96 POWER TRANSISTORS

The BLW96 is an advanced h.f./v.h.f. npn power transistor for use in high-power industrial and military transmitting equipment operating from nominal 50 V supplies. The device is generally used in conjunction with the BLW50F driver transistor and gives up to 200 W output power, allowing a practical 1 kW transmitter output stage to be built with only eight BLW96 transistors. It also has a very low thermal resistance and a high and constant f_T ($\approx$240 MHz). To ensure reliability in service, the BLW96 utilises well-proven header technology; in particular, special attention has been paid to the prevention of thermal fatigue which can be a problem in intermittent h.f. power operation.

The BLW96 comprises two chips divided into a total of 64 base areas, each with 18 associated emitter regions and individual ballast resistors (Fig.5). Current is thus distributed between 1152 transistor elements. By these means, a high r.f. power capability, deliberately conservatively quoted as 200 W, is achieved. The BLW96 has a non-saturable gain curve up to a minimum of 100 MHz (Fig.6). An amplifier using two BLW96 transistors driven by two BLW50F driver transistors is described later. First, however, some general points on the design of wideband linear amplifiers are considered.

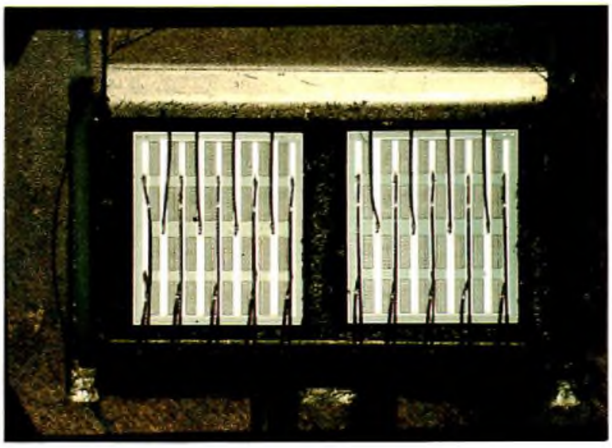


Fig.5 Unencapsulated BLW96 transistor

CIRCUIT DESIGN CONSIDERATIONS

In designing a wideband h.f. linear amplifier, there are three main requirements to be considered: power, bandwidth, and linearity.

Designing for power

Probably the most demanding aspect of design is that of achieving high-power capability in a compact equipment which must operate at high ambient temperatures. In obtaining the maximum possible power, there are two areas for consideration: thermal design and electrical design.

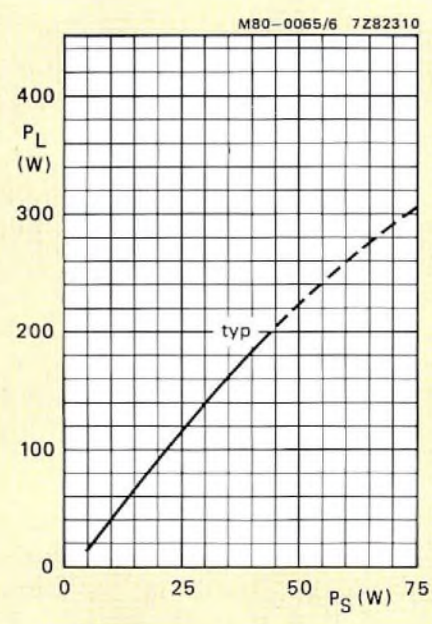


Fig.6 Load power as a function of drive power for BLW96. (V_{CE} = 50 V, f = 108 MHz, T_h = 25°C)

Thermally, it is necessary to reconcile the permitted maximum junction temperature of the power transistor with the worst-case ambient temperature at which the equipment is required to operate while at the same time observing the d.c. and r.f. SOAR ratings.

Equation 1 relates the power dissipated in the transistor P_c to temperature and thermal resistance.

$$P_c = \frac{T_j - T_a}{\theta_{j\text{-}mb} + \theta_{mb\text{-}h} + \theta_{h\text{-}a}} \tag{1}$$

where

P_c = total power dissipated in the transistor

T_j = transistor junction temperature

T_a = ambient temperature

$\theta_{j\text{-}mb}$ = thermal resistance between junction and the transistor encapsulation

$\theta_{mb\text{-}h}$ = thermal resistance between encapsulation and heatsink

$\theta_{h\text{-}a}$ = thermal resistance between heatsink and ambient.

In Eq.1, the designer must observe the upper limit on T_j set by the transistor manufacturer; T_a will be set by the equipment specification. The thermal resistance θ_{j-mb} is also fixed for the particular transistor used. However, θ_{mb-h} is partly under the control of the designer, depending as it does on the heatsink chosen and the assembly techniques used. There remain P_c and θ_{h-a} and a compromise between these two is generally necessary.

The total power P_c is determined by the operating r.f. output power of the transistor and the efficiency of the circuit. Efficiency is itself determined by the degree of linearity that can be tolerated which in turn sets the class of amplifier operation (A, AB, or B). It is also influenced by load variations (due, for example to the effects of a mismatched aerial). For a resistive load, the maximum r.f. power output is achieved by optimising the r.f. voltages and currents in the transistors, by correctly matching the transistor outputs to the load, and by using low-loss passive components and careful circuit construction.

For example, in a simple push-pull amplifier operating from a supply voltage V and driven to a sinusoidal peak current of I per transistor, the available r.m.s. power output from each transistor is given by:

$$P = \left(\frac{V - V_k}{4}\right) I \tag{2}$$

where V_k is the r.f. knee voltage of the transistor. For maximum power transfer to a specified load:

$$\frac{V - V_k}{I} = \frac{n^2 R}{4}, \qquad (3)$$

where n is the turns ratio of the matching transformer and R is the load resistance.

These equations are used for calculating the power available from the amplifier under specified conditions; for example, 50 V h.t., a 50 Ω load, and a simple integral turns ratio. They also indicate the peak r.f. collector current which each transistor must be capable of passing.

Designing for bandwidth

To achieve the required bandwidth, the circuit techniques used are generally similar to those used for lower frequency wideband power amplifiers with two notable exceptions.

First, to obtain a level frequency response, passive compensating networks are often used, rather than negative feedback. It is not always possible to use the latter technique because of the phase distortion inherent in power transistors operating at high frequencies.

Second, it is seldom practicable to set the operating conditions of the transistor (that is, the peak voltage and peak current) at levels where the transistor is matched perfectly to the load resistance. Impedance transformation is therefore necessary and, because the frequency range is so broad, the use of transformers wound on ferrite cores is the only practical approach. Furthermore, because the input impedance of an h.f. transistor is very low (≈ 1 Ω) and is complex and variable over the band, it is also necessary to use a wideband wound transformer together with a compensating network in the transistor input circuit. The design of these wideband wound transformers is in fact critical and one of the major considerations in wideband linear amplifier construction: more information on impedance transformation is given later.

Designing for linearity

In wideband h.f. power amplifiers, linearity is mainly influenced by amplitude distortion and the consequent generation of harmonics and intermodulation products. Amplifiers operating in class AB or class B generate relatively high harmonic powers: −10 to −16 dB is typical. Even-order harmonics can be reduced to about −40 dB by careful circuit design, in particular by accurate balancing of the push-pull stage, but little can be done to reduce the levels of odd-order harmonics and in practice these must be removed by filters at the output of the transmitter.

The most troublesome form of distortion is that due to odd-order intermodulation products, particularly third-order and fifth-order. These occur close to the carrier frequency and cannot be removed by filtering. They cause distortion on the channel in use as well as interference on neighbouring channels. The only way to minimise odd-order products is to bias the power transistors into a relatively linear operating region. In practice, depending on the levels of distortion acceptable, the power transistors are biased in class AB or, if thermal considerations and ratings allow, in class A. The driver stage is nearly always operated in class A and is designed to have about −6 dB better distortion performance than the overall amplifier requirement.

It is often noted that gain and intermodulation levels are not constant with frequency, and at certain points in the band, changes of level may occur, accompanied by corresponding changes in amplifier efficiency. These effects may be attributable to reflected harmonics which are not terminated either by the restricted bandwidth of the matching transformers or by the filters. It may be possible to improve performance by the use of crossover filters which divert harmonic power into discrete harmonic loads; this technique, however, is still in its infancy.

CIRCUIT CONFIGURATION

A schematic diagram of a two-stage wideband h.f. linear amplifier is shown in Fig.7. There are three major design aspects.

1) selection of transistor bias conditions;
2) design of transformers and baluns;
3) selection of passive components for the gain compensation networks.

Class AB push-pull operation for the output stage is assumed as this gives the best compromise in terms of linearity, dissipation, and cost. Operating in push-pull has the added advantage of minimising even-order distortion products and giving a relatively high output impedance, ensuring maximum power transfer to the load. As mentioned above, the driver stage generally operates in class A.

Transistor operating conditions

The performance of h.f. power transistors in linear amplifiers is limited both by thermal and electrical considerations. In a class AB amplifier, the most severe thermal conditions occur when the amplifier is operating with c.w. or f.m. signals at the maximum permissible heatsink temperature. It may then not be

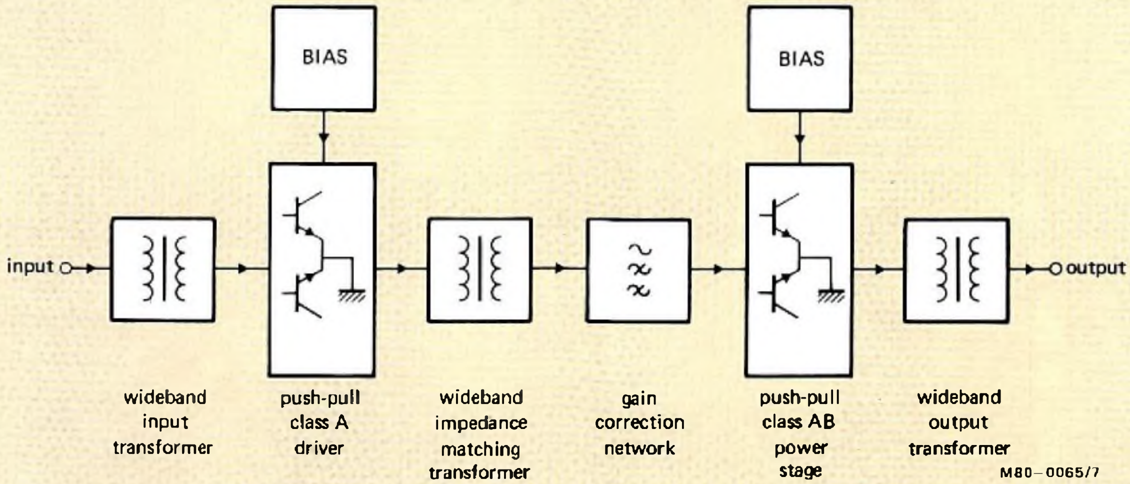


Fig.7 Block diagram of h.f. wideband linear amplifier

possible to drive the transistors at the published maximum power output, and downrating will be essential. Further downrating will be necessary if the amplifier is to operate into a mismatched load. With low duty-cycle s.s.b. or a.m. speech modulation, however, it is unlikely that the transistors will be thermally limited. Under such conditions, device failure would most likely be caused by the collector r.f. voltage limits being exceeded by operation into a mismatched load. Peak level detection circuits controlling the driver stage are often used to maintain the collector r.f. voltages within safe limits.

If it is necessary for the amplifier to operate with various degrees of mismatch, including open- and short-circuited aerials, then further protection must be provided. It is usual practice to current-limit the d.c. collector input to the maximum level permitted when the amplifier is correctly loaded and to provide a thermal cut-out which trips if the maximum heatsink temperature is exceeded.

Under severe environmental conditions, the above considerations may make it necessary to operate the power transistors at power levels between one half and one third of the theoretical maximum, especially when working near or at the maximum heatsink temperature. Fortunately, downrating of a class AB amplifier usually results in reduced intermodulation distortion.

The bias circuit for the power transistors must be capable of maintaining the necessary bias over the complete drive cycle and operating temperature range. To achieve this, the bias supply must be of very low impedance, be capable of supplying a current of $I_{c\,max}/h_{FE}$ to each transistor, and be temperature compensated. ($I_{c\,max}$ is the peak collector current under r.f. drive conditions and h_{FE} is the minimum specified gain at this current.) Temperature compensation is necessary because the quiescent current in bipolar devices is very temperature dependent. Recommended values of quiescent current for class AB operation are given in the published data. The quiescent current can be controlled, not perfectly, but sufficiently well to prevent thermal runaway and degradation of the intermodulation distortion performance, by the use of a thermally-tracking bias unit mounted on the heatsink as close as possible to the power transistors.

Impedance transformation

As shown in Fig.7, three impedance matching transformers are required in a typical two-stage wideband amplifier. The output transformer matches the calculated collector-to-collector resistance of the power stage to the resistive load, usually 50 Ω, while the input transformer matches the source resistance, again often 50 Ω, to the input resistance of the driver stage. The third transformer transforms the output resistance of the driver to the input of the power stage and in this case accurate power matching is particularly important. If an optimum power match is not achieved, the intermodulation distortion performance of the driver stage may be seriously degraded. The power capability of the driver stage would then have to be increased.

The impedance matching transformers must be designed and constructed to ensure low losses and accurate impedance transformation (Refs.4 and 5). Two basic types are used in wideband linear amplifiers: conventional, with monofilar windings, and transmission

Fig.8 Broadband transformers

line, with windings of miniature coaxial cable or twisted-pairs of single strand wire (Fig.8).

Conventional transformers

A wideband transformer can be wound in a conventional way using a core and separate primary and secondary windings. Tight coupling must be maintained between the windings to ensure accurate impedance transformation and low reflected power losses. The core material must maintain its permeability and have low losses over the whole band; a suitable grade of ferrite is 4C6.

In theory, an almost infinite range of turns ratios is possible, but this is far from the case in practice. For example, the use of fractional turns is not consistent with the need to keep connecting leads as short as possible. However, impedance ratios defined by m^2/n^2, where m and n are small integers, are achievable.

Transmission line transformers

In this type of transformer, the primary and secondary windings are combined to form one or more transmission lines (Ref.6). Transmission line transformers (such as the one shown in Fig.9) offer greatly improved h.f. performance because the self-capacitance and stray inductance of the windings, which set the h.f. limits of conventional transformers, are absorbed in the characteristics of the transmission lines. The range of transformed impedance achievable is, however, somewhat limited. Practical ratios are 1:1, 4:1, 9:1 and 16:1; apart from these it is difficult to obtain transmission lines of suitable impedance for the realisation of a required transformer. The transformers are generally wound on ferrite cores to minimise the length of line; 4C6 is a suitable grade for the h.f. band. However, the core material is not critical in transmission line transformers as the core is not used for coupling but only to increase the inductance of the windings at the low frequency end of the band.

For a transmission line transformer of impedance ratio n^2:1, the characteristic impedance of the line Z_0 should be chosen so that:

$$Z_0 = \sqrt{(R_L R_r)}, \qquad (6)$$

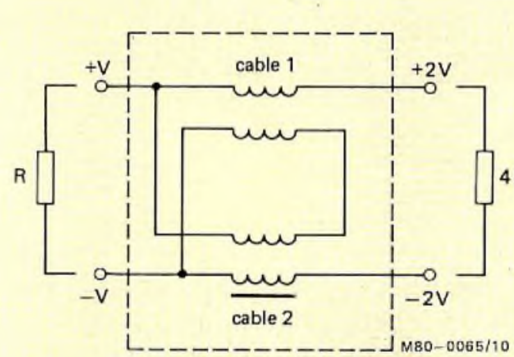


Fig.9 4:1 transmission line transformer

where R_L is the load resistance and R_T is the transformed resistance. Therefore:

$$Z_o = nR_L \quad (7)$$

As mentioned earlier, only a limited number of values of Z_o are practical with commercially available coaxial lines, and the use of versions with other values would be prohibitively expensive. The use of transmission line transformers is therefore limited to specific powers and supply voltages. For example, with a 50 Ω system operating from a 28 V supply at 100 W (assuming a peak collector swing of 25 V), 4:1 impedance matching would be required. This can be achieved using a 2:1 turns ratio transformer constructed with 25 Ω transmission line.

Passive components and circuit layout

The gain and input impedance of the power transistors vary considerably over the four octaves of the band. A passive network (Fig.10) is therefore included in the base circuits to compensate for the fall-off in gain with frequency. However, the design resistance of this network can be increased compared with the input resistance of the power transistors alone, and therefore the construction of the required matching transformer is eased.

The compensating networks are usually designed from data computed from equivalent circuit parameters. Alternatively, the input impedance of the transistors can be measured using tuned test amplifiers. After tuning the test amplifier with the transistor in circuit, an impedance meter or network analyser is substituted for the transistor and the impedance conjugate can be measured directly.

The passive components used in the compensating networks and elsewhere in the circuit can be standard types. However, types with low losses and low parasitic impedances must be chosen. These components are often used in parallel combinations to minimise parasitics still further and to handle the currents and powers required.

Because of the high circulating currents and low impedances in transistor r.f. power amplifiers, circuit layout and earth current paths are critical. Earth current paths should be as short as possible and should not interact with input and output networks.

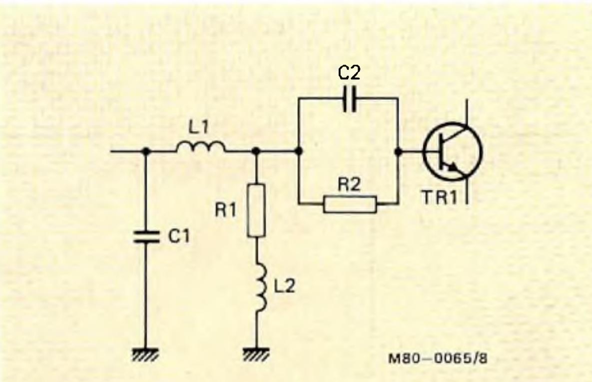


Fig.10 Passive compensating network

PERFORMANCE MEASUREMENT

Because a wideband untuned class AB amplifier has a relatively high harmonic output (typically −13 dB third harmonic), the measurement of peak envelope power can be subject to error depending on the method used. A typical measurement system is shown in Fig.11. The power meter measures the mean r.f. power from the amplifier, including all harmonics. With a single-tone drive signal, the p.e.p. and c.w. levels are equal. With the two-tone signal used for intermodulation distortion measurements, the p.e.p. output is twice the mean power indicated by the meter. However, with high levels of intermodulation distortion, this factor-of-two relationship no longer holds true.

With low harmonic levels, the output may be sufficiently undistorted to allow the p.e.p. to be measured directly by means of an oscilloscope connected across the load (CCIR method). Errors of up to 25% can occur with this method, however, if harmonic levels are significant.

Intermodulation products are measured using a two-tone drive signal and a spectrum analyser. The level of intermodulation distortion is normally quoted with respect to one tone of the two-tone signal; if quoted with respect to p.e.p., the intermodulation level will be 6 dB lower.

400 W P.E.P. WIDEBAND H.F. LINEAR AMPLIFIER

The circuit shown in Fig.12 has been developed to illustrate the performance capability of BLW96 transistors in a wideband linear h.f. amplifier. The two BLW96 transistors in the power stage operate in class AB and are driven by two BLW50F transistors in class A; the complete circuit operates from a +50 V supply and is water cooled. The amplifier can deliver up to 400 W c.w. into a wideband 50 Ω resistive load or, when operated with s.s.b. modulation, 400 W p.e.p. with intermodulation distortion better than −26 dB and an overall gain of 27 ± 2 dB over the 1.6 to 30 MHz band. When operated at 300 W p.e.p., intermodulation distortion falls to −30 dB, even with a reduced supply voltage (45 V).

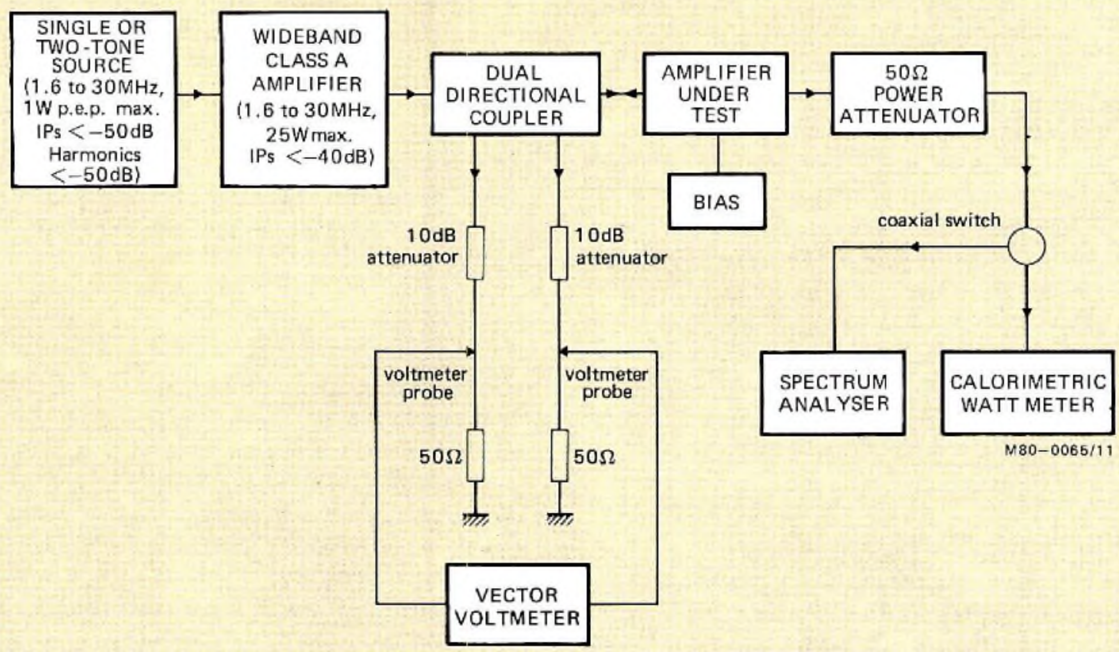


Fig.11 Typical performance measurement set-up

Design

In designing the amplifier, gain and input impedance data for both the BLW96 and BLW50F transistors, operating at the required power levels, are first calculated from equivalent circuit parameters. Single-stage two-transistor circuits are then constructed and the calculated values and bias conditions are checked. The performances of the individual stages are summarised in Table 1.

When the two stages are linked together as a two-stage amplifier, the total dissipation is dominated by the power levels and efficiency of the output stage. Table 2

TABLE 1

Performance data for driver and power stages

	Driver	Power stage
Bias		
V_{CE}(V)	44	50
I_C	2 x 1A	2 x 100 mA (zero signal)
Single-tone 1.6 to 30 MHz		
P_L(W)	25	400
Gain (dB):		
mid-band	15.8	13.4
min	15.7	13.4
max	16.4	15.8
Input v.s.w.r.:		
mid-band	1.35 : 1	1.1 : 1
max	1.36 : 1	1.45 : 1

TABLE 2

Efficiency measurements on single-stage BLW96 test circuit

C.W. signals 1.6 to 30 MHZ

P_L (W)	Best collector efficiency (%)	Worst collector efficiency (%)
400	65.5	47.5

Taking the worst case and assuming equal sharing and zero circuit losses, each transistor would dissipate 220 W.

Two-tone signals 1.6 to 30 MHz

P_L(p.e.p.) (W)	Best collector efficiency (%)	Worst collector efficiency (%)
400	50	37.7
300	44	32

Taking the worst case and assuming equal sharing and zero circuit losses, each transistor would dissipate 170 W.

CORRECTION The diagram below should be substituted for the one printed on page 219, E.C. & A. Vol.3, No.4.

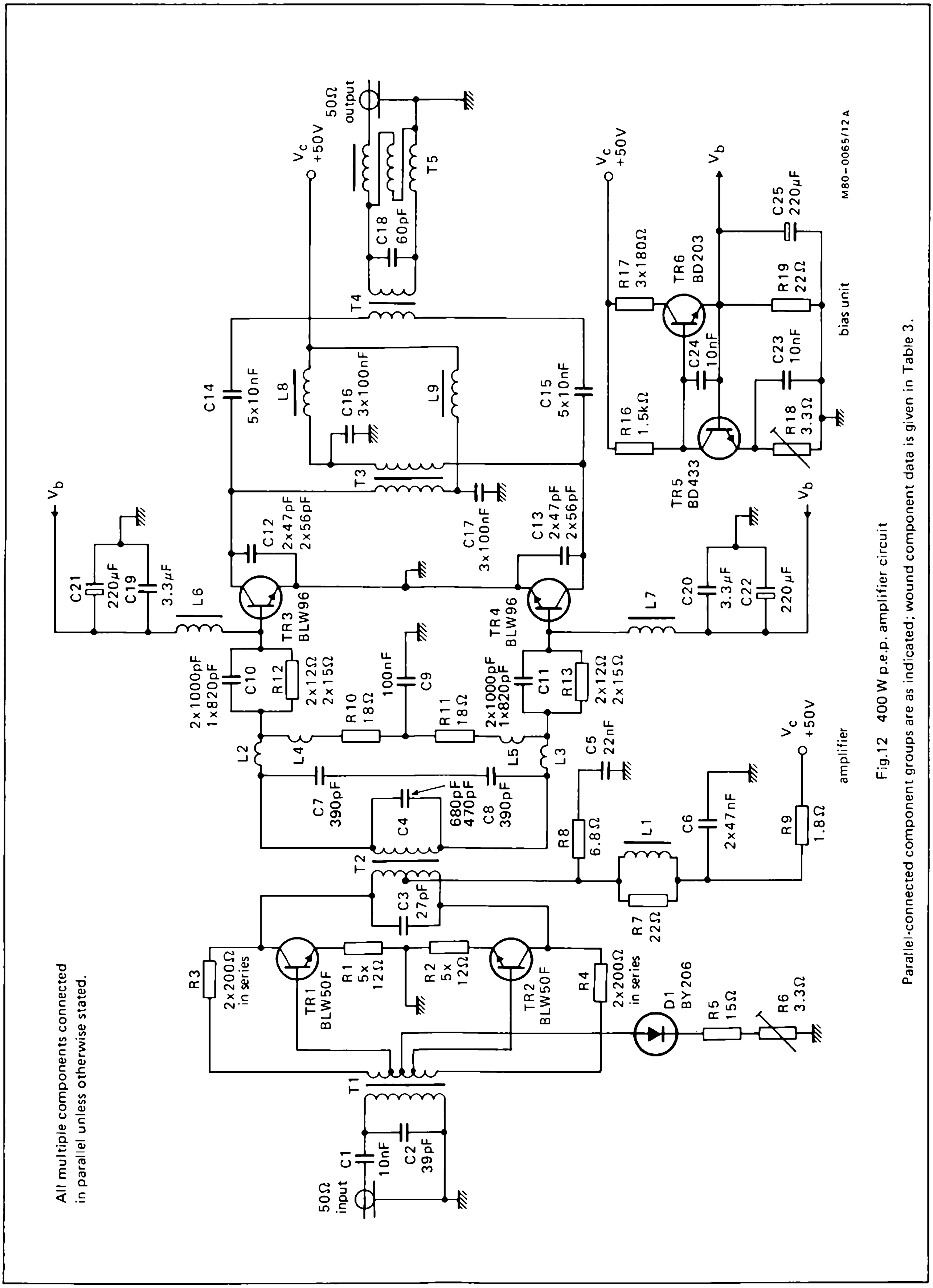


Fig.12 400 W p.e.p. amplifier circuit

Parallel-connected component groups are as indicated; wound component data is given in Table 3.

indicates best- and worst-case levels of collector efficiency achieved in a single-stage BLW96 test circuit operating between 1.6 and 30 MHz. Under worst-case (c.w.) conditions, it is seen that each BLW96 will dissipate 220 W.

The published r.f. thermal resistance of the BLW96 is 0.65°C/W (junction to heatsink) and the maximum permissible junction temperature is 200°C. Therefore, at 220 W dissipation, the heatsink temperature must not exceed 57°C. However, this assumes equal sharing and zero circuit losses and therefore in the two-stage amplifier in Fig.12, water cooling is used.

400 W p.e.p. two-stage circuit

The circuit in Fig.12 is constructed in accordance with the general principles and practice outlined earlier.

The driver stage is coupled directly to the power stage by means of a 100:5.5 Ω impedance matching transformer T2. The gain compensation network connected between the bases of TR3 and TR4 ensures that the output powers from each transistor are equal at all points in the band. In the case of the driver stage, negative feedback, both series and shunt, is used for compensation since the load on this stage is relatively constant.

The input transformer T1 is conventionally wound on a twin-hole ferrite bead, with the secondary tapped to allow connection to the bases of TR1 and TR2 and to the shunt feedback resistors R3 and R4. The centre-tap is connected to a bias temperature compensating network D1, R5, and R6. Interstage transformer T2 transforms the collector-collector impedance of the driver stage (100 Ω) to the input impedance of the power stage (5.5 Ω). This transformer is again conventionally wound with parallel windings on two stacked twin-hole ferrite beads to minimise leakage inductance. Bias for the driver is applied to the centre tap of T2, via an anti-parasitic network.

The main output transformer T4 transforms the collector-collector impedance of the output stage (9.6 Ω) to 50 Ω. It is conventionally wound on two stacked toroidal cores, with parallel windings. Transformer T3 supplies h.t. to the power transistors while at the same time offering a low impedance to in-phase transistor currents. It is wound with transmission line on a ferrite rod and is in effect a wideband centre-tapped choke. The output balun T5 is a transmission line type wound on two stacked toroids. All the wound components are provided with capacitive compensation for leakage reactance (C1, C2, etc.). Table 3 gives full winding details.

Bias is provided by a thermally-tracking circuit, also shown in Fig.12, in which TR5 acts as a temperature sensor. The circuit can supply up to 800 mA, depending on the value of R17 in the collector load of the BD203. In practice, R17 is made up of three 17 W wirewound resistors in parallel, mounted separately on stand-off insulators on the heatsink. The complete bias unit is thermally-mounted on the heatsink as close as possible to the power transistors.

Performance

Figures 13 to 17 indicate the overall performance of the complete amplifier. In Fig.13, gain and input v.s.w.r. under single-tone drive are shown as a function of frequency, with a 50 V supply and a load power of 400 W. Minimum gain and v.s.w.r. are seen to be 25 dB and 2:1 respectively. Figures 14 to 16 show third-order and fifth-

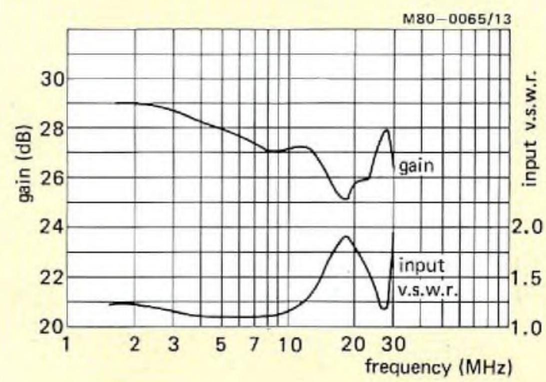


Fig.13 Gain and input v.s.w.r. as a function of frequency

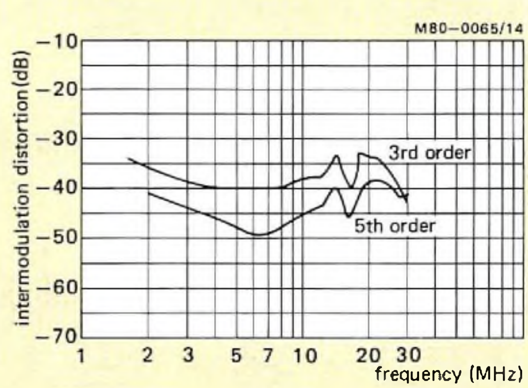


Fig.14 Intermodulation performance at 200 W p.e.p. output and 50 V supply

TABLE 3

Wound components (Fig.12)

L_1, L_6, L_7	2.5 turns, 0.55 mm tinned copper on 6-hole grade 3B ferrite bead type 4312 020 31500
L_8, L_9	3 parallel loops through 6-hole grade 3B ferrite bead type 4312 020 31500
L_2, L_3	13.9 nH
L_4, L_5	21 nH
T_1	1:1.5 turns ratio. Wound on 2-hole grade 4B1 ferrite bead type 4312 020 31520. Primary: 4 turns, 2 x 0.45 mm en. cu. in parallel. Secondary: 6 turns, 2 x 0.45 mm en. cu. in parallel, tapped at centre and at 2 x 1 turns from centre.
T_2	4.5:1 turns ratio. Comprises two transformers with primary and secondary windings in parallel, each wound on 2-hole grade 4B1 ferrite bead type 4312 020 31520 Primary: 9 turns, 0.45 mm en. cu. centre tapped. Secondary: 2 turns, 2 x 0.45 mm en. cu. in parallel.
T_3	4 turns, 1 mm twisted en. cu., wound on 50 mm of grade 4A10 ferrite aerial rod (or equivalent).
T_4	2.33:1 turns ratio. Comprises two transformers with primary and secondary windings in parallel, each wound on grade 4C6 ferrite toroids, 36 x 23 x 15 mm, type 4322 020 97200. Primary: 6 turns, 8 mm wide copper tape. Secondary: 14 turns, 4 x 0.5 mm en. cu. in parallel. Windings separated by 0.25 mm thick p.t.f.e. tape.
T_5	Output balun. Wound on two grade 4C6 toroids, 36 x 23 x 15 mm, type 4322 020 97200. 8 turns, 50 Ω coaxial cable with p.t.f.e. insulation and 4 mm external diameter, and 8 turns 1 mm en. cu. for the balancing winding.

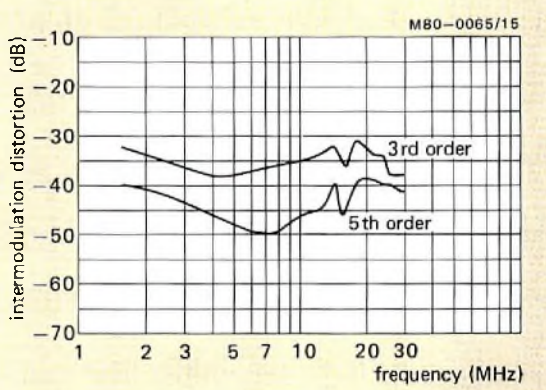


Fig.15 Intermodulation performance at 300 W p.e.p. output and 50 V supply

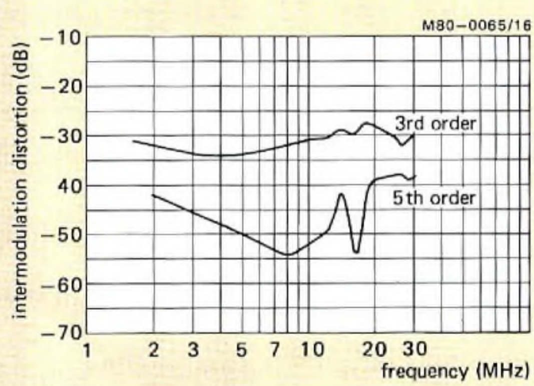


Fig.16 Intermodulation performance at 400 W p.e.p. output and 50 V supply

order intermodulation products under two-tone drive conditions with a 50 V supply and load powers of 200 W, 300 W, and 400 W p.e.p. It is seen that at 400 W, third-order intermodulation products are better than −26 dB and at 300 W, better than −30 dB. Fig.17 shows intermodulation performance at 300 W p.e.p. but with a 45 V supply; even in this case, intermodulation distortion is better than −30 dB.

Table 4 indicates the amplitudes of harmonics measured relative to the fundamental with the amplifier driven at 400 W c.w. Fig.18 shows a laboratory prototype of the amplifier.

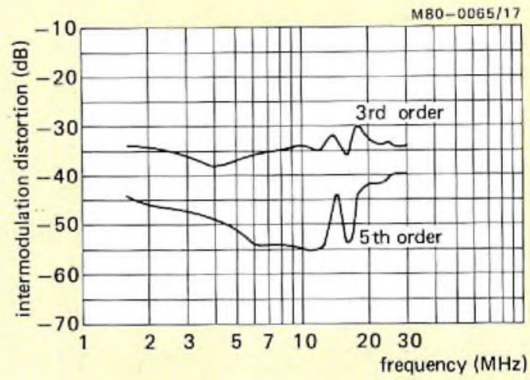


Fig.17 Intermodulation performance at 300 W p.e.p. output and 45 V supply

ACKNOWLEDGEMENT

The author would like to thank H. Benischek of the Elcoma R.F. Power Transistor Development Laboratory for his contribution to this article. He would also like to acknowledge the work of A.H.Hilbers of the Elcoma Central Application Laboratory.

REFERENCES

1. BETTS, J.A., 'High frequency communications', English Universities Press, 1967.
2. HILBERS, A.H., 'Large-signal behaviour of r.f. power transistors. Part 1: Analysis of the equivalent circuit,' Electronic Applications Bulletin, Vol.31, No.3, 1972, pp. 135 to 150.
3. MULDER, J., 'Large-signal behaviour of r.f. power transistors. Part 2: Computer program', Electronic Applications Bulletin, Vol. 31, No. 4, 1972, pp. 218 to 233.
4. HILBERS, A.H., 'High-frequency wideband power transformers', Electronic Applications Bulletin, Vol. 30, No.2, 1970, pp. 64 to 73.
5. HILBERS, A.H., 'Design of high-frequency wideband power transformers', Electronic Applications Bulletin, Vol.32, No.1, 1973, pp. 44 to 48.
6. ROTHOLZ, E., 'Transmission-line transformers', IEEE Transactions on Microwave Theory and Techniques, Vol. MTT-29, No.4, April 1981, pp. 327 to 331.

TABLE 4

Harmonic generation

f_1 MHz	f_2 dB	f_3 dB	f_4 dB	f_5 dB	f_6 dB	f_7 dB	f_8 dB	f_9 dB	f_{10} dB
1.6	−46	−19	−56	−34	−48	−48	−50	−45	−59
3.5	−45	−19	−50	−33	−58	−44	−56	−45	−60
7	−54	−18	−50	−29	−48	−40	–	–	–
10	−48	−17	−45	−32	−55	−50	–	–	–
14	−43	−16	−50	−44	–	–	–	–	–
20	−34	−25	–	–	–	–	–	–	–
28	−40	−45	–	–	–	–	–	–	–

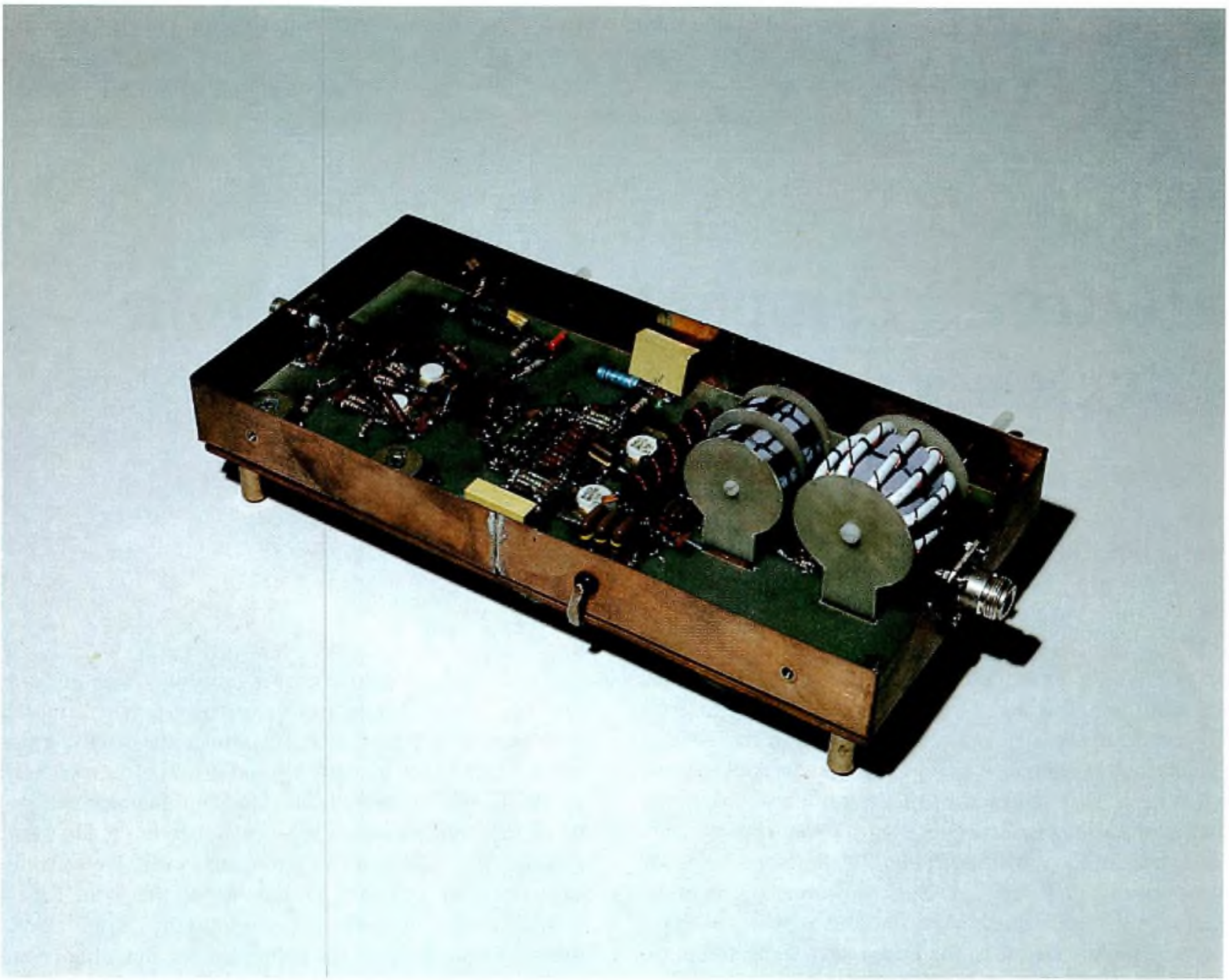


Fig.18 Laboratory prototype of 400 W p.e.p. wideband linear amplifier

While the styling of telephone subscriber sets has evolved over the years, not much has changed under the skin: in most cases metallic contacts, magnetic circuits, and carbon granules still do the work. Now the advance of microelectronics is making dramatic changes not only in how the work is done but also in the number and nature of the functions performed. Integrated circuits are now available for pulse and tone dialling, automatic redialling, storing a repertory of frequently called numbers, and even for ringing individual changes on the bell.

Electronic components for telephone subscriber sets

L. v. d. MEEBERG and H. J. M. OTTEN

Apart from the introduction of the dial in the 1920s and the addition of a few electronic components for line-current stabilisation and surge protection in the 1950s, the subscriber telephone set has scarcely changed electrically in a hundred years. Even the carbon microphone that first enabled voice signals to be transmitted over practical distances without amplification has held its place. During the past ten years, however, advances in microcircuit technology have resulted in the development of microprocessors, memories, and dedicated integrated circuits that are now beginning to replace the traditional components of the subscriber set, to the benefit of both its performance and appearance.

For example, integrated circuits for interrupted current-loop dialling (decadic or pulse) and for dual-tone multi-frequency (DTMF) dialling allow the dial to be replaced by a pushbutton keyboard. Also, by replacing the carbon microphone with an active type (electret or electro-dynamic) and using an anti-sidetone circuit, the speech and transmission functions can be performed by an integrated circuit. The same integrated circuit can also contain the previously mentioned dialling functions. Even the bell will eventually be replaced by an electronically driven transducer with a variety of ringing tones or chimes. Microcircuits also allow the incorporation of a host of new features such as automatic redialling, repertory dialling, automatic emergency call dialling, dialled number display, tariff-unit metering, hands-free use, remote wireless handset, and possibly a limited amount of data handling.

Another trend is toward digital telephone equipment and transmission systems. It is envisaged that the first digital telephone systems will be introduced within the next five years.

The long lifespan of existing subscriber sets, and the fact that their manufacture is often automated, means that several years must elapse whilst the new components prove their operational and economic superiority to the telephone authorities. Many of the new components will therefore have to work alongside the old ones, and the order and speed of approach to the all-electronic subscriber set will not necessarily be the same in all countries. We have therefore produced, or are developing, the wide range of microcircuits and other electronic components listed in Table 1 for telephones of the present and of the future. This article shows how the new components replace the old ones, and explains how our integrated circuits have been partitioned so that they can satisfy the varied requirements of the telephony market.

Incorporating such features as DTMF dialling, automatic redialling, and provision for hands-free use, this microprocessor-controlled telephone typifies the trend toward increased convenience, expanded facilities, and elimination of electromechanical components that is setting the pace for tomorrow telephones

(Photo courtesy of MBLE International, Brussels.)

TABLE 1
Dedicated components for telephone subscriber sets

	type	pulse generation	tone generation	line interface	speech function	dialling ease	ringing	protection	status
Bipolar ICs									
DTMF dialling circuits	TDA1077		●	●					production
	TEA1021		●	●					samples
	TEA1043		●	●					samples
microphone preamplifier	TCA980			●	●				production
speech/transmission circuit	N204*			●	●				samples
single-chip DTMF telephone	RO084*		●	●	●				development
CMOS ICs									
pulse dialling circuits	PCD3320	●				●			production
	PCD3321	●				●			production
	PCD3323	●				●			production
DTMF generator with redial; microcomputer peripheral DTMF generators	PCD3310 family		●			●			development
dedicated microcomputer dialler	PCD3330	●	●			●			development
serial 1K RAM	PCD8571					●			development
LCD driver	PCE2111					●			production
multi-tone ringer							●		development
Other components									
voltage-dependent resistor	8222 298 12141*							●	samples
zener bridge	BZW10 series			●				●	production
MOSFETs		●		●			●	●	development
loudspeakers	AD2071/3071				●		●		production
special transducer for ringer							●		development
liquid-crystal display	LC 7020 160					●			samples
display module	M 7020 160					●			samples

* development type number.

DIALLING CIRCUITS

Pulse dialling circuits

Integrated circuits PCD3320 and PC3323 (development type numbers MH320 and MH323) for interrupted current-loop dialling were described in Ref. 1. These have now been augmented by the PCD3321, which uses the same die and provides many of the same features as the 28-pin PCD3323 but in an 18-pin package. Table 2 compares the three circuits.

Using a keyboard it is possible to dial the digits faster than they may be transmitted. All three circuits therefore incorporate a RAM that can store up to 23 digits for transmission at the required rate. The number last dialled can be redialled by pressing a single button on the keyboard.

DTMF dialling circuits

The TDA1077 bipolar circuit for DTMF dialling described in Ref. 2 is succeeded by the TEA1021, which has lower minimum supply current: 8 mA typical, 10 mA maximum. Despite simpler peripheral circuitry, it meets the CEPT distortion requirements and, unlike the TDA1077, does not require a 15 mH choke to satisfy the −80 dBm requirement for spurious output above 50 kHz.

Another bipolar DTMF circuit is the TEA1043, which incorporates a mute output, but at the cost of slightly higher distortion: a spurious frequency of 12.2 kHz at −52 dBm and one of 36.6 kHz at −66 dBm.

Some CMOS circuits for DTMF are also in development: the PCD3310 family. Being intended for use with an electronic speech/transmission circuit, these do not include an output stage or line adaptation.

One member of the family is designed for keyboard control and incorporates such features as last-number redial, 70 ms minimum tone and pause duration, and register recall ('flash' in Great Britain, 'hold' in the U.S.A.).

Others are stripped versions minus the memory and timing functions, intended for use with a microcomputer. One of these has an I^2C bus interface.

SPEECH/TRANSMISSION CIRCUITS

N204 is the development type number of a speech/transmission circuit that includes the following features:

- receiver amplifier and balanced microphone amplifier with gain setting and adaptive gain control responsive to line current
- gain control adaptable for use with two feeding-bridge resistances and various exchange voltages
- anti-sidetone circuit
- muting input for interrupted current-loop or DTMF dialling
- DTMF input with externally adjustable sensitivity
- supply output for interrupted current-loop or DTMF dialling circuit, microcomputer, electret microphone, and/or additional circuits such as RAMs and display drivers.

Another speech/transmission circuit in development is the RO084, which incorporates a DTMF generator but does not have adaptive gain control. It can operate in conjunction with a single-contact keyboard, but the keyboard inputs can also be connected to a microcomputer if extended dialling facilities are required. Target specifications of the N204 and RO084 are summarised in Table 3.

Table 4 summarises the specification of the TCA980G, a separate amplifier for use with electrodynamic and magnetodynamic microphones.

TABLE 2
Comparison of pulse dialling circuit facilities

circuit	PCD3320	PCD3321	PCD3323
number of pins	18	18	28
dialling pulse			
frequency, f_{DP} (Hz)	10	10/16/20	10/16/20
mark/space ratio	3:2	3:2 or 2:1	3:2 or 2:1
interdigit pause ($T_{DP} = 1/f_{DP}$)	$8T_{DP}$	$8T_{DP}$	$8T_{DP}$ or $9T_{DP}$
duration of line-power interruption to cause operation in stand-by	$>1.6T_{DP}$	$>1.6T_{DP}$	$>1.6T_{DP}$ or $>3.2T_{DP}$
access pause duration	N/A	$32T_{DP}$	$32T_{DP}$ or $64T_{DP}$

TABLE 3
Abridged target specification for transmission circuits

	N204	RO084	
supply:			
line current	10-140	10-140	mA
corresponding nominal d.c. voltage	4-7	4.8-7	V
adjustable d.c. voltage	yes	no	
impedance:			
nominal impedance	600	600	Ω
balance return loss against 600Ω	⩾20	⩾20	dB
adjustable impedance	yes	yes	
microphone amplifier:			
nominal gain	45	50±2	dB
manual adustability	+8 to −8	0 to −15	dB
maximum output voltage (d = 10%)	+6	+6	dBm
symmetric input impedance	2 × 2	2 × 2	kΩ
noise level on line	⩽−70	⩽−70	dBmp
mute input	yes	internal	
telephone amplifier:			
nominal gain (including sidetone circuit)	−4	0±2	dB
manual adjustability	+6 to −6	0 to −10	dB
maximum output voltage (d = 10%)		800	mV_{rms}
telephone impedance range		300-400	Ω
mute input	yes	internal	
gain control:			
control range, microphone + telephone amplifier	6	–	dB
feeding bridge correction	400 or 800	–	Ω
exchange voltage correction range	24 to 60	–	V
gain control disable switch	yes	–	
DTMF section:			
crystal frequency	–	3.58 or 4.78	MHz
lower tone level (adjustable)	–	−11 to −6	dBm
level inaccuracy in fixed application	–	±2	dB
pre-emphasis of higher tones	–	2±1	dB
keypad type	–	single/double	
spurious signal level	–	CEPT	
encapsulation	18-pin DIL	24-pin DIL	

TABLE 4
TCA980G microphone amplifier

line current range, I	10−100 mA
corresponding d.c. voltage	4.5−7 V
nominal voltage gain at I = 10 mA	48.3 dB
output impedance	150 Ω
maximum r.m.s. output voltage (d = 10%)	1 V
package	9-pin SIL (metal can optional)

RINGER CIRCUIT

Many telephone authorities are dissatisfied with the timbre of the conventional bell which contains too many high-frequency components inaudible to the elderly. They have therefore specified several tones that an electronic ringer must be able to generate, preferably in agreeable sequences or chimes. Such a ringer should also incorporate means for varying the tone sequence so that when telephones are installed close together it is easy to tell which one is ringing. Although the ringer function could be combined with the speech/transmission function, it is preferable to keep it separate: more than one ringer is often wanted with one subscriber set.

Figure 1 shows an electronic ringer that meets these requirements. The integrated circuit can generate one of

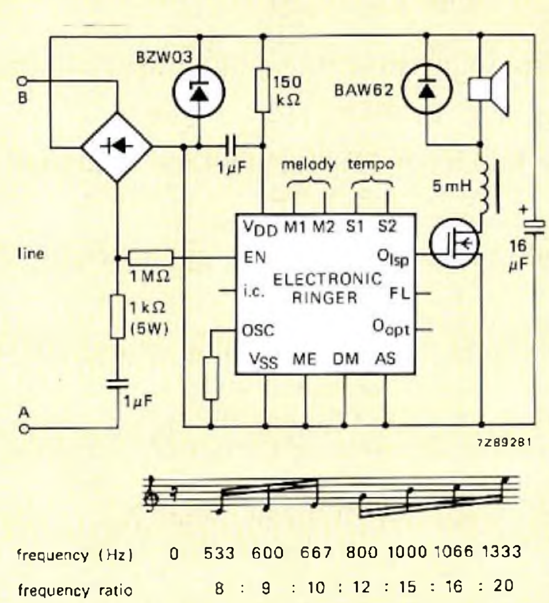


Fig.1 Electronic ringer, with notation showing the mask-programmable tones

four mask-programmable tone sequences consisting of a number of time intervals, each of which can be occupied by one of seven tones –533 Hz, 600 Hz, 667 Hz, 800 Hz, 1000 Hz, 1066 Hz, 1333 Hz – or a pause. The tones and their harmonic relation have been selected to meet the requirements of most European telephone authorities and to provide the best possible aureal perception. The tempo of the selected sequence can be varied in four steps. For maximum flexibility the required tone sequence and tempo are selected by hard-wiring or switch-programming four address pins. This means that if, for example, a two-tone ringer is required, tone sequences are stored which differ only in the number of time intervals allotted to each tone or pause, and two of the address lines are used to select the required repetition rate. An optional automatic swell increases the output level in two increments of about 10 dB for the first and second repetition of the tone sequence during ringing.

An important feature of the ringer is the enable circuit. The tone-sequence generator reacts to the ringer voltage only if it is more than 15 V r.m.s. and its frequency is between 13.33 Hz and 30 Hz or between 20 Hz and 60 Hz (pin-selectable).

A rugged 50Ω loudspeaker with characteristics that remain stable in the long term converts the output from the integrated circuit into an audible signal. To avoid the necessity of using an output transformer, the loudspeaker is driven by a discrete MOSFET operating in class D. The integrated circuit also has an output for driving a visual ringing indicator. A diode bridge rectifies the ringer voltage to supply the IC: a zener diode protects the IC against overvoltage.

CIRCUITS FOR FEATURE-PHONES

To supplement our PCF8500 microcomputer range we are developing a telephony-dedicated CMOS microcomputer, the PCD3330, which is able to accept data from all known types of pushbutton dialling units. It can store a number of up to 18 digits for direct dialling, redialling or extended redialling, and can store a repertory of up to ten numbers. The repertory can be expanded to 100 numbers by connecting CMOS RAMs type PCD8571 to the two-wire I^2C bus of the PCD3330. The facilities of the microcomputer-controlled telephone can also be extended by connecting additional circuits such as display drivers PCE2111 to the I^2C bus. The PCD3330 has four muting outputs and outputs for both interrupted current-loop and DTMF dialling. A delayed reset which operates after a line power break is built in. An access pause system may also be included. Some characteristics of the PCD3330 are:

- very low current consumption
- 8-bit CPU, 2K-byte ROM, 200-byte RAM
- adapted 8500 instruction set
- two I/O ports specially designed for keyboard scanning
- output port with high current drive capability
- DTMF output
- serial I/O port (I^2C bus)
- additional interrupt (any key down)
- oscillator for 3.58 MHz crystal.

The software can be adapted to all telephone authority specifications with respect to dialling and programming procedures, and to various keyboard expansion possibilities.

The first version, the PCD3331, is designed to function in accordance with the German postal specification FeAp82. It is controlled by:

- standard function keys 0–9, A–D, # and *
- four special function keys (programming, redial, register recall, and access pause)
- an optional 10-key repertory dialler extension for ten on-chip 16-digit numbers
- four diode option switches for selection of mark/space ratio in interrupted current-loop dialling mode, selection of power-failure reset delay, normal/direct call mode, and dialling mode.

It has an on-chip program for a 16-digit LCD controlled by two PCE2111 display drivers connected to the I^2C bus. A module containing the 16-digit LCD and two PCE2111s is available from Videlec under the type number M 70 20 160.

PROTECTION COMPONENTS

The solid-state circuits in the subscriber set must be supplied at the correct polarity, whichever way round the lines are connected. They must also be protected against overvoltages such as might be caused either by high-voltage transients due to induced current surges on the lines (secondary lightning effects) or by inadvertent connection of the lines to the mains. The BZW10 described in Ref. 4 performs both functions. During normal operation its bridge-connected diodes ensure the correct polarity; and in the event of current surges, they absorb most of the excess energy. In case of accidental connection to the mains, the diodes limit the voltage applied to the subscriber-set circuits to a safe level. The BZW10-12 and BZW10-15 are for sets with DTMF dialling. A new zener diode, the BZW03, is to become available for protection of the electronic ringer circuit.

Pulse-dialling sets can be protected by a passive component with a non-linear current/voltage characteristic (e.g. our voltage-dependent resistor 8222 298 12141).

ARCHITECTURE OF ELECTRONIC SUBSCRIBER SETS

The first step in converting subscriber sets to electronic operation will nearly always be replacement of the rotary dial by a pushbutton keyboard operating in conjunction with either a pulse generator for interrupted current-loop dialling or a tone generator for DTMF dialling. This effectively divides electronic telephone production into two main streams, one for sets with pulse dialling and one for sets with tone dialling. Subsequent steps are replacement of the carbon microphone by an active transducer such as an electret or electrodynamic microphone, and replacement of the transformer hybrid by an integrated speech/transmission circuit. The sequence continues with the incorporation of features such as repertory dialling, last-number redial, extended redial, dialled number display, and tariff-unit metering. There will also be some sets capable of either pulse or DTMF dialling.

The ringer is a completely separate function and can therefore be replaced by electronics at any stage.

Telephones for pulse dialling

Figure 2 shows the architecture of three pushbutton subscriber sets for interrupted current-loop dialling using integrated circuits PCD3320/21/23.

Figure 2(a) shows an insert unit to replace the dial in a conventional set having a transformer hybrid; the muting relay inhibits the speech function during dialling.

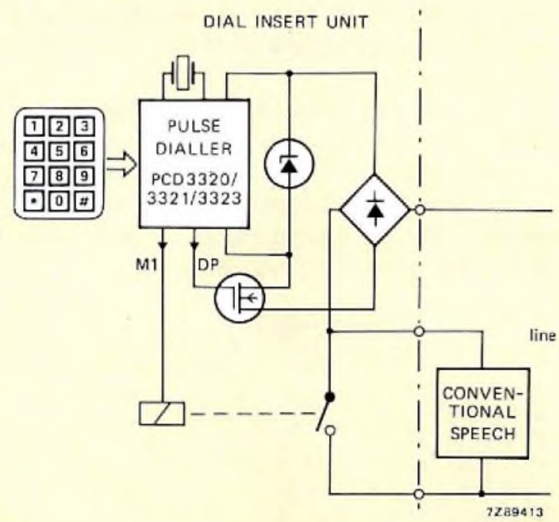
Figure 2(b) shows a parallel circuit which is advantageous when a common contact is not required but a conventional speech part is. The line current flows either through the speech part or through a dummy load and is interrupted by the M3 output of the dialling IC.

In Fig.2(c) the dialling IC operates in conjunction with an N204 speech/transmission IC. The latter works with either an electret or an electrodynamic microphone and has a special input for muting.

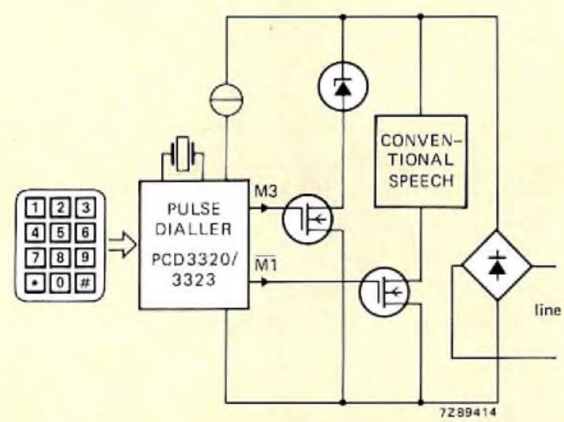
Telephones for DTMF dialling

Figure 3 shows the architecture of three pushbutton sets for DTMF dialling.

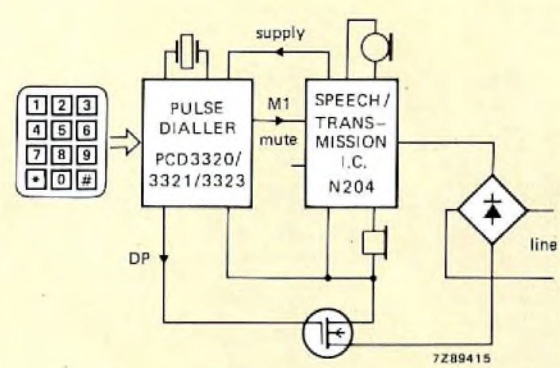
Figure 3(a) shows a DTMF insert to replace a rotary dial. This requires a DTMF generator with output stage and voltage stabiliser such as the TDA1077 or its improved successor the TEA1021, both of which require a common contact to switch between speech and dialling.



(a) Serial-mode insert with conventional speech circuit



(b) Parallel-mode with conventional speech circuit

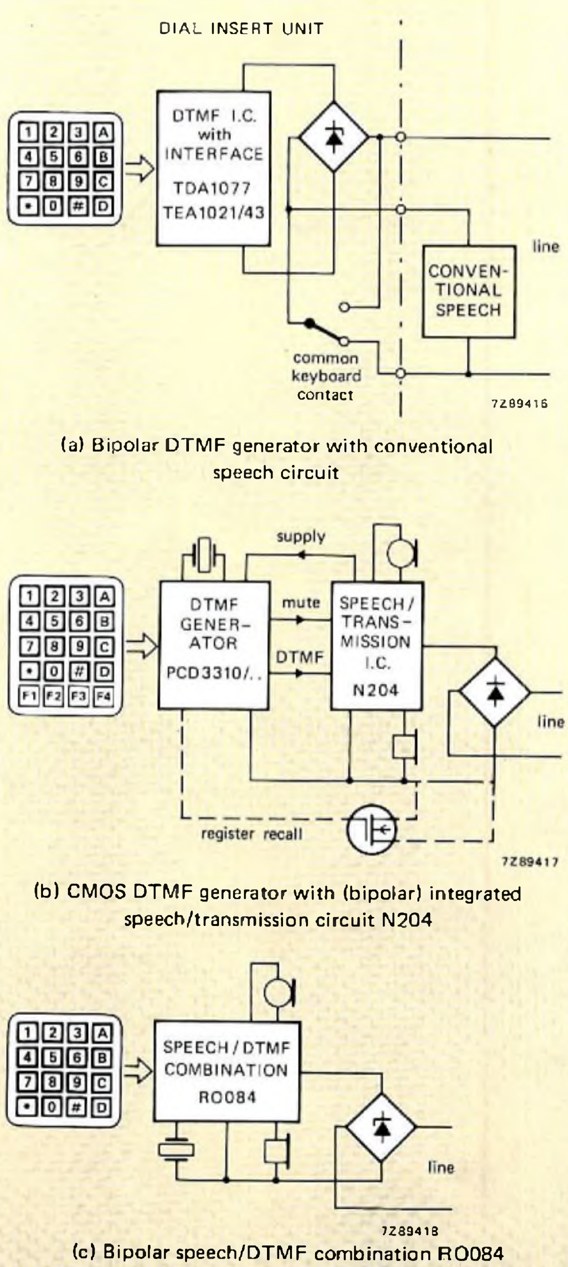


(c) Combination with integrated speech/transmission circuit N204

Fig.2 Subscriber-set architecture for interrupted current-loop dialling

With the TEA1043, the performance of which is comparable to that of the TEA1021, the common contact can be eliminated in favour of discrete semiconductors for muting.

In Fig.3(b) the CMOS DTMF generator PCD3310, without output stage but with some features such as redial and calibrated tone bursts, operates in conjunction with the speech/transmission circuit N204 which incorporates a voltage stabiliser and an audio output stage for both speech and DTMF signals.

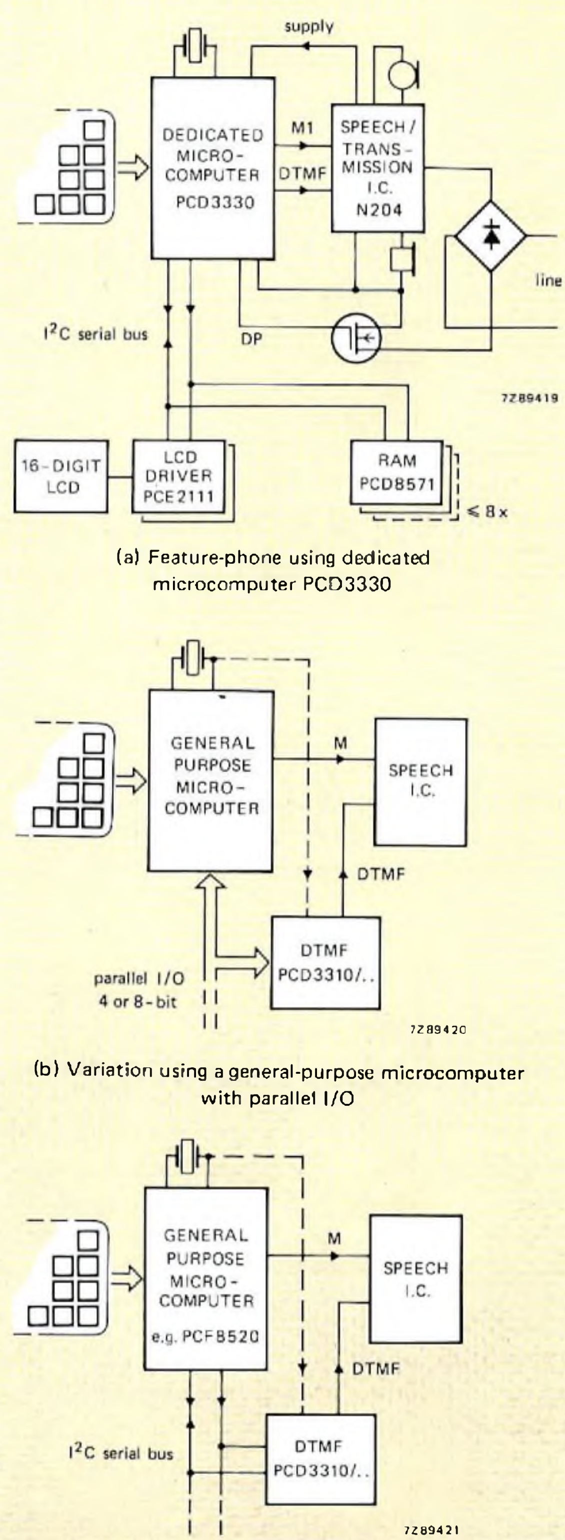


(a) Bipolar DTMF generator with conventional speech circuit

(b) CMOS DTMF generator with (bipolar) integrated speech/transmission circuit N204

(c) Bipolar speech/DTMF combination RO084

Fig.3 Subscriber-set architecture for dual-tone multi-frequency (DTMF) dialling



(a) Feature-phone using dedicated microcomputer PCD3330

(b) Variation using a general-purpose microcomputer with parallel I/O

(c) Variation using a general-purpose microcomputer with I²C bus

Fig.4 Subscriber-set architecture with extended features

Telephones with extended features

Figure 4(a) shows a subscriber set with additional features such as last-number redial, extended redial, repertory dialling, register recall, dialled-number display, and tariff-unit metering. The PCD3330 dedicated microcomputer used is suitable for either interrupted current-loop or DTMF dialling. Its on-chip memory capacity of ten 16-digit-plus-two-access-pause numbers can be augmented by connecting up to eight CMOS RAMs type PCD8571 to the I^2C bus. The two display drivers type PCE2111 can drive a 16-digit 7-segment LCD to display the number called and the tariff units metered.

Figure 4(b) shows a variation using a general-purpose microcomputer with parallel I/O in conjunction with a member of the PCD3310 family of CMOS DTMF generators.

If one of the general-purpose microcomputers of the 8500 series is preferred (e.g. the PCF8500), the serial I/O in conjunction with the I^2C bus, can lead to simpler interconnection. Another member of the PCD3310 family, a DTMF generator with I^2C interface only, can then be used (see Fig.4(c)).

REFERENCES

1. FASSER, J. and HODEMAEKERS, A. M. 1980. 'Interrupted current-loop dialling for pushbutton telephones' Electronic Components and Applications, Vol. 2, No. 3, pp 175-189.
2. GEBOERS, J. J. A., JANSEN, D. J. G. and KAIRE, J. C. 1978. 'Two-tone telephone dialling' Electronic Components and Applications, Vol. 1, No. 1, pp 29-41.
3. MOELANDS, A. P. M. 1980. 'Serial I/O with the MAB8400 series microcomputers' Electronic Components and Applications, Vol. 3, No. 1, pp 38-46.
4. 1979. 'BZW10 transient suppression bridge for telephones' Electronic Components and Applications, Vol. 1, No. 2, pp 125-127.

Neither a thyristor nor a transistor, the GTO has practical advantages over both as a power switching device, provided it is properly used. This article describes four GTO drive circuits for applications ranging from domestic applicances to television line-deflection circuits and switched-mode power supplies.

Basic GTO drive circuits

F.J. BURGUM

The development of power electronic switching techniques has received a strong impetus from the introduction of the gate turn-off switch (GTO). A three-terminal pnpn device, the GTO is similar in construction and performance to the conventional thyristor, and like the thyristor it can be turned on by a positive current injected into the gate. However, the GTO is a more versatile device since it can be turned off by drawing current from the gate. This is achieved by applying a voltage of between −5 and −10 V directly between gate and cathode. Thus while the GTO has the high blocking voltage and high overcurrent capability of the thyristor, these qualities are combined with the ease of gate drive and fast switching associated with the bipolar transistor. An introduction to the GTO is given in Ref.1, and a detailed guide to the use of the device data is given in Ref.2.

The full performance of any device can only be fully realised if it is incorporated into a correctly designed circuit. This can be a particular problem with new devices, such as the GTO, since they may represent a significant departure from the circuit designer's previous experience. This article describes four GTO drive circuits which will facilitate correct operation of the device over a wide range of applications.

The appropriate circuit for a given application is largely dependent upon whether or not isolation is needed, and on the required range of duty cycle and switching frequency. Control signal isolation is often essential in professional switched-mode power supplies (SMPS) and motor speed control systems, although it is not usually required in self-contained domestic appliances where low cost is of paramount importance. In SMPS, the duty cycle is typically in the range 5 to 50% at a switching frequency of the order of tens of kHz, while in a.c. motor control circuits, the duty cycle must range from less than 1% to more than 99% at frequencies which are often as low as tens of hertz.

NON-ISOLATED DRIVE CIRCUITS

The application of the GTO is still in its infancy and the drive circuits are the subject of rapid development and improvement. Reference 2 gives two examples of non-isolated drive circuits, and the two circuits which are described in this section have been developed from these earlier types. In the case of the first circuit to be described, the improvements are relatively minor; for completeness, however, the circuit is described in full.

Simple direct gate drive

A non-isolated gate drive circuit is shown in Fig.1. Combining simplicity and low cost, this circuit has a limited range of duty cycle and the range of switchable current is determined by the gate supply voltage. The main applications of this circuit are in television SMPS, horizontal deflection circuits, and simple series-resonant power supplies (SRPS).

The GTO is turned on by gate current supplied by TR1. Resistor R1 limits the initial level of gate current, and while TR1 is conducting charge builds up on capacitor C1. The maximum voltage to which C1 is charged is limited to about 10 V by the supply voltage and the values of R1 and R2. When C1 has charged, the GTO gate current will have fallen to a low level determined by R2.

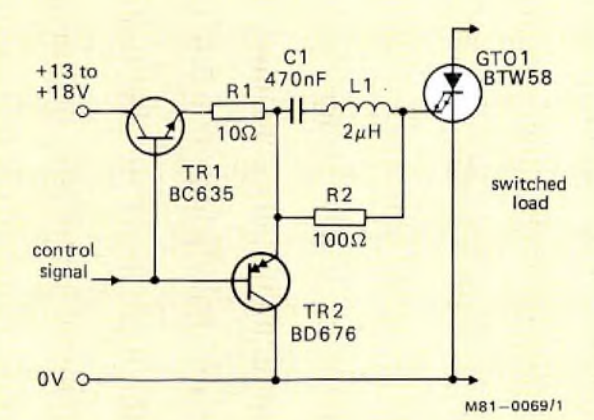
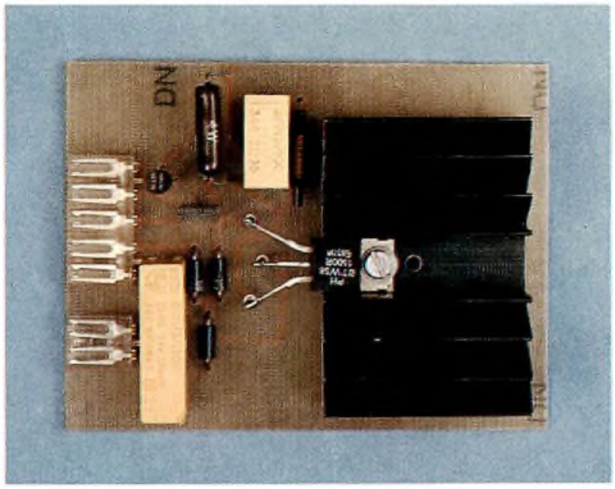


Fig.1 – Simple direct drive circuit



Laboratory model of circuit shown in Fig. 1

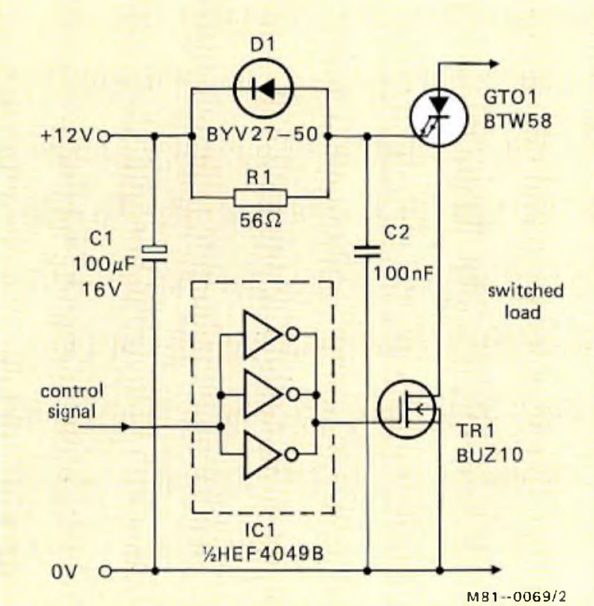
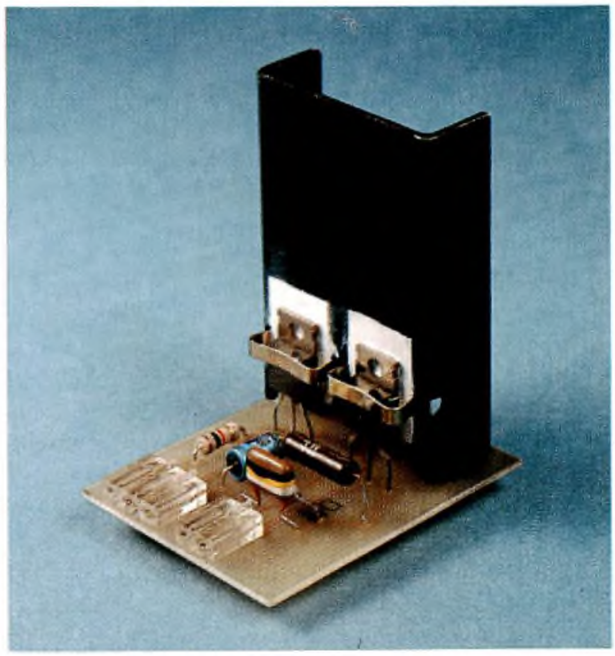


Fig.2 – MOS-compatible cathode drive circuit



Laboratory model of circuit shown in Fig. 2

Turn-off is initiated when the darlington transistor TR2 is turned on. The voltage on C1 is then applied directly between the GTO gate and cathode. Charge is therefore extracted from the gate, and the GTO is turned off. The capacitance of C1 is such that the charge extracted from the gate at turn-off reduces the voltage across C1 by one or two volts only.

The limitation on the range of duty cycle associated with this circuit arises from the need for the GTO conduction period to be long enough to charge C1 via TR1 and R1. This is particularly important during the first few switching cycles when C1 is receiving its initial charge.

This circuit can turn off a peak anode current of 6 A, subject to a maximum rate of rise of applied anode voltage of 500 V/μs and a peak anode voltage of 1500 V. The minimum on-time of the GTO is 5 μs.

MOS-compatible cathode drive

The circuit shown in Fig.2 controls the GTO by switching the cathode current with a POWERMOS transistor. The benefits resulting from this arrangement include a MOS-compatible control input, good switching characteristics, and a controllable GTO anode current of more

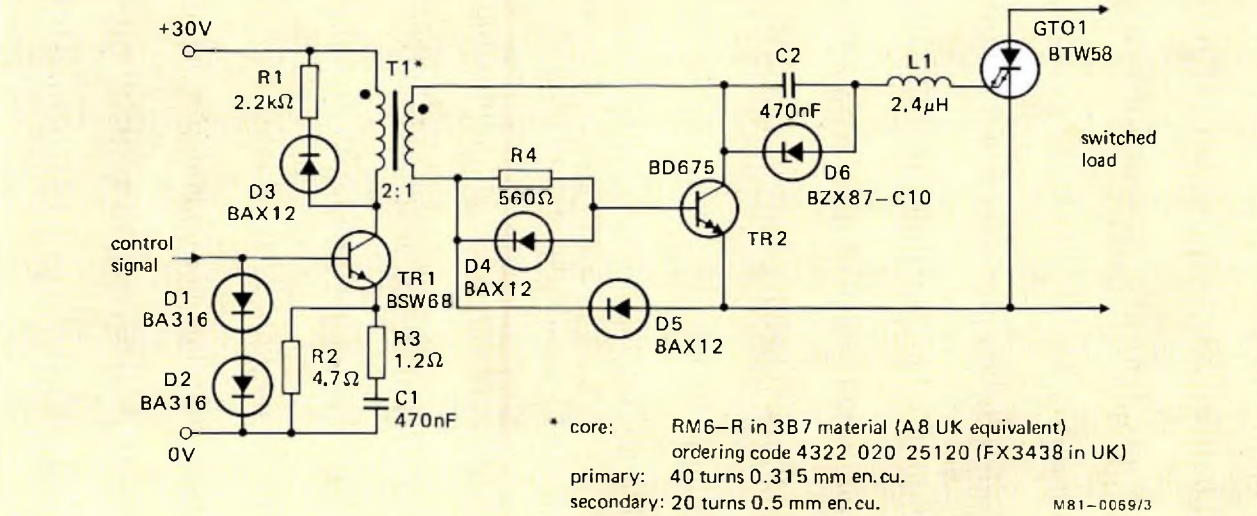


Fig.3 – Isolated SMPS and SRPS drive circuit

than 10 A. The circuit can be used for pulsed switching over a very wide range of duty cycle, and the switching frequency extends down to d.c. The principal application area for this circuit is the control of d.c. motors in large domestic appliances.

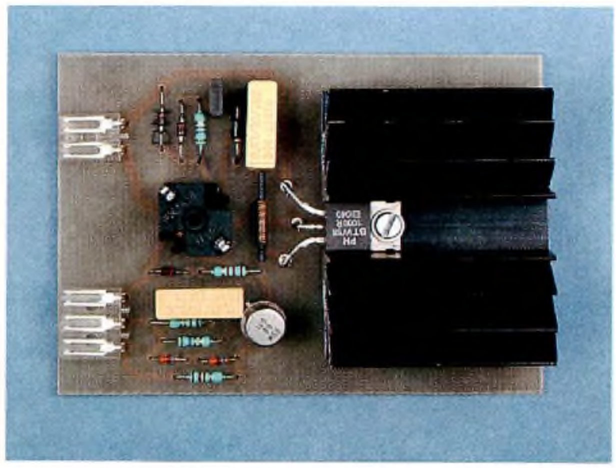
With TR1 conducting, the GTO is turned on by discharging current into the gate from C2. When C2 has discharged, the gate current is maintained via R1. Supplying the initial gate current by discharging C2 ensures that the GTO is forced rapidly into full conduction, even with a high rate of rise of anode current. When TR1 is turned off, the GTO cathode current is very quickly diverted into the gate. Initially this gate current flows into C2, and when C2 is charged to 12 V the current flows into C1 via D1. The energy extracted from the gate of the GTO at turn-off is thus stored in C2, and is used to turn the GTO on in the next switching cycle.

This circuit can turn off a peak anode current of 12A. The rate of rise of applied anode voltage should be limited to 500 V/μs, and the peak anode voltage should not exceed 1500 V. There is no limitation on the range of duty cycle.

ISOLATED DRIVE CIRCUITS

SMPS and SRPS drive circuit

In the drive circuit shown in Fig.3, the control signal is isolated by a simple pulse transformer. This circuit is restricted to applications in which the switching frequency is high enough to allow the use of a small transformer with a low leakage inductance. The minimum duty cycle is determined by the time required to charge C2 in the GTO gate circuit, while the maximum duty cycle is limited by the time needed to demagnetise the transformer core. Isolated SMPS and SRPS are the principal application areas, where the duty cycle is typically in the range 5 to 50% and the switching frequency is 10 to 40 kHz.



Laboratory model of circuit shown in Fig. 3

The GTO is switched on when TR1 is brought into conduction. With TR1 conducting, the current in the primary of transformer T1 is controlled by the voltage on the base of TR1 and the network R2, R3, and C1 in the emitter circuit of TR1. This current has a high initial value of about 500 mA, falling to 125 mA with a time-constant of 600 ns. The turns ratio of T1 is 2:1, giving an initial gate current of about 1 A falling to 250 mA. The GTO gate current flows via C2, L1, the gate-cathode junction, and D5. While current is flowing into the gate,

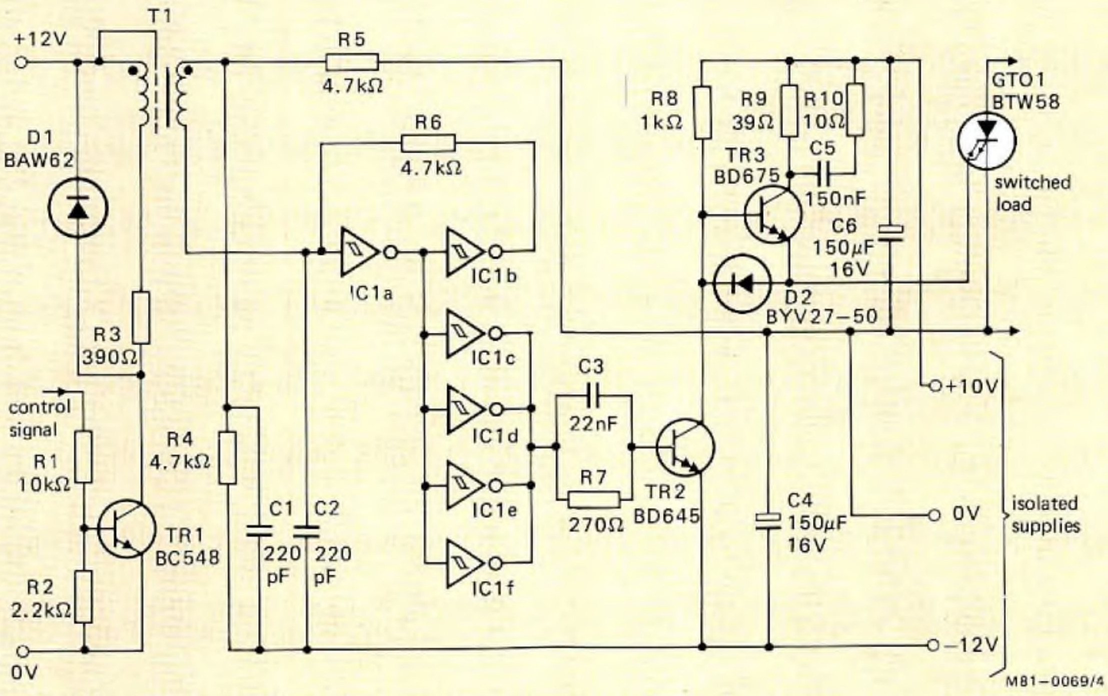
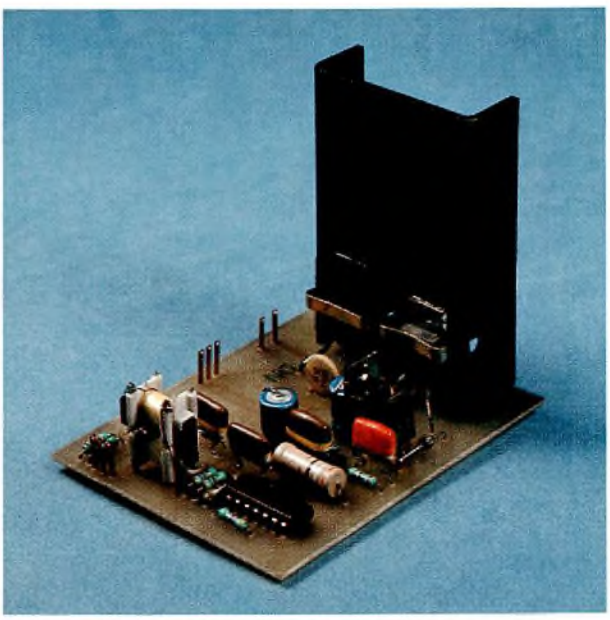


Fig.4 — Isolated drive circuit with a wide switching frequency range



Laboratory model of circuit shown in Fig. 4

TR2 has a negative base-emitter voltage and is thus kept in the off-state. The voltage across C2 is limited to 10 V by the voltage regulator diode D6.

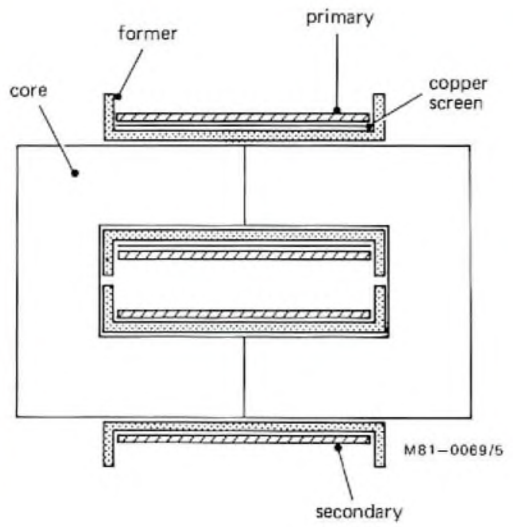
To turn off the GTO, transistor TR1 is turned off. The energy stored in T1 causes the voltage across the secondary winding to reverse. This voltage drives TR2 into conduction, applying the voltage on C2 across the GTO gate-cathode junction. Current flows out of the gate to discharge C2, turning off the GTO.

By selecting a capacitor of suitable size, the drop in voltage across C2 at turn-off is limited to only one or two volts. The inductance L1 in the gate circuit prolongs the period during which the GTO gate-cathode junction is in reverse avalanche breakdown, and also ensures that TR2 continues to conduct until the GTO has completely turned off. Diode D4 is included to ensure a rapid turn-off for TR2 when TR1 is turned on. Without this diode a portion of the initial peak GTO gate current would be short-circuited by TR2.

With this circuit the controllable peak GTO current is 5 A, subject to a maximum rate of rise of anode voltage of 800 V/μs and a peak anode voltage of 800 V. The duty cycle range is 5 to 50%.

Drive circuit with wide switching frequency range

In the circuit shown in Fig.4, the control signal is isolated by a small pulse transformer T1 with minimum inter-winding capacitance and low inductance. The construction of T1 is indicated in Fig.5. In addition to the power supply for the pulse transformer switching transistor



core: 2 off U10/8/3 in 3C8 material (A16 UK equivalent) ordering code 3122 134 91160 (FX3676 in UK)
primary: 27 turns of 0.25 mm en.cu.
secondary: 27 turns of 0.25 mm en.cu.
copper screen under primary winding

Fig.5 – Isolating transformer for the circuit of Fig.4

TR1, this circuit requires isolated supplies to turn the GTO on and off. However, the circuit was designed to be used in a three-phase a.c. motor drive inverter, and in such an application the cost of providing these extra supplies is relatively low in comparison with the cost of the six switching devices. The design of a suitable power supply is described in the next section.

The inductance of T1 is made low, so that the primary circuit L/R time-constant is about 1 µs. The transformer secondary voltage is thus a differentiated version of the input control signal. The two LOCMOS gates IC1a and IC1b act as a combined Schmitt trigger and memory circuit, and so the output of IC1a is a restored square wave corresponding to the input control signal. The output of IC1a, buffered by IC1c to IC1f, controls the darlington transistor TR2. This transistor is used for turning the GTO off, extracting current via D2 into the smoothing capacitor C4, connected to the –12 V supply. The inductance in the loop formed by the GTO gate-cathode junction, D2, TR2, and C4 should be kept as low as possible, preferably less than 1 µH.

When TR2 is turned off, TR3 conducts and the GTO is turned on by positive gate current defined by the network R9, C5, and R10. The allowable minimum GTO on- and off-periods have been made sufficiently low for the circuit to be suitable for pulse-width modulated a.c. motor speed control. There is no lower limit to the switching frequency.

This circuit can turn off a peak anode current of 6 A, the minimum pulse width is 2 µs, and the switching frequency range is d.c. to 5 kHz. The rate of rise of applied voltage across T1 should not exceed 2 kV/µs.

Multiple-output isolated power supply

Fig.6 shows a simple and economical power supply, suitable for generating the multiple isolated supplies required when using the circuit of Fig.4 in the inverter of an a.c. motor speed control system. The oscillator IC1 switches TR1 on and off at about 60 kHz. Transformer T1 has a 1:3 turns ratio, giving a secondary voltage of about 65 V peak-to-peak. Transformers T2, T3, and T4 step this down to about 22 V peak-to-peak for the three upper GTOs in the inverter bridge, and transformer T5 provides 22 V peak-to-peak for the three lower GTOs in the bridge. When TR1 is conducting, diodes D5 to D10 conduct, charging the capacitors connected to the positive supply in the GTO drive circuits (C6 in Fig.4). When TR1 is turned off, the energy stored in the cores of transformers T1 to T5 charges (via diodes D11 to D16) the capacitors connected to the negative supply in the GTO drive circuits (C4 in Fig.4). Voltage regulator diodes D17, D18, D19, and D20 limit the negative outputs to –12 V.

ACKNOWLEDGEMENT

The first three circuits covered in this article were designed respectively by H.W. Evers of CAB Eindhoven, D.R. Hyde of MAL Mitcham and E.B.G. Nijhof of CAB Eindhoven. Their contribution is gratefully acknowledged.

REFERENCES

1. BURGUM, F.J., NIJHOF, E.B.G., and WOODWORTH, A., 'Gate turn-off switch', Electronic Components and Applications, Vol.2, No.4, August 1980, pp.194 to 201.
2. WOODWORTH, A., 'Understanding GTO data as an aid to circuit design', Electronic Components and Applications, Vol.3, No.3, pp.159 to 166.

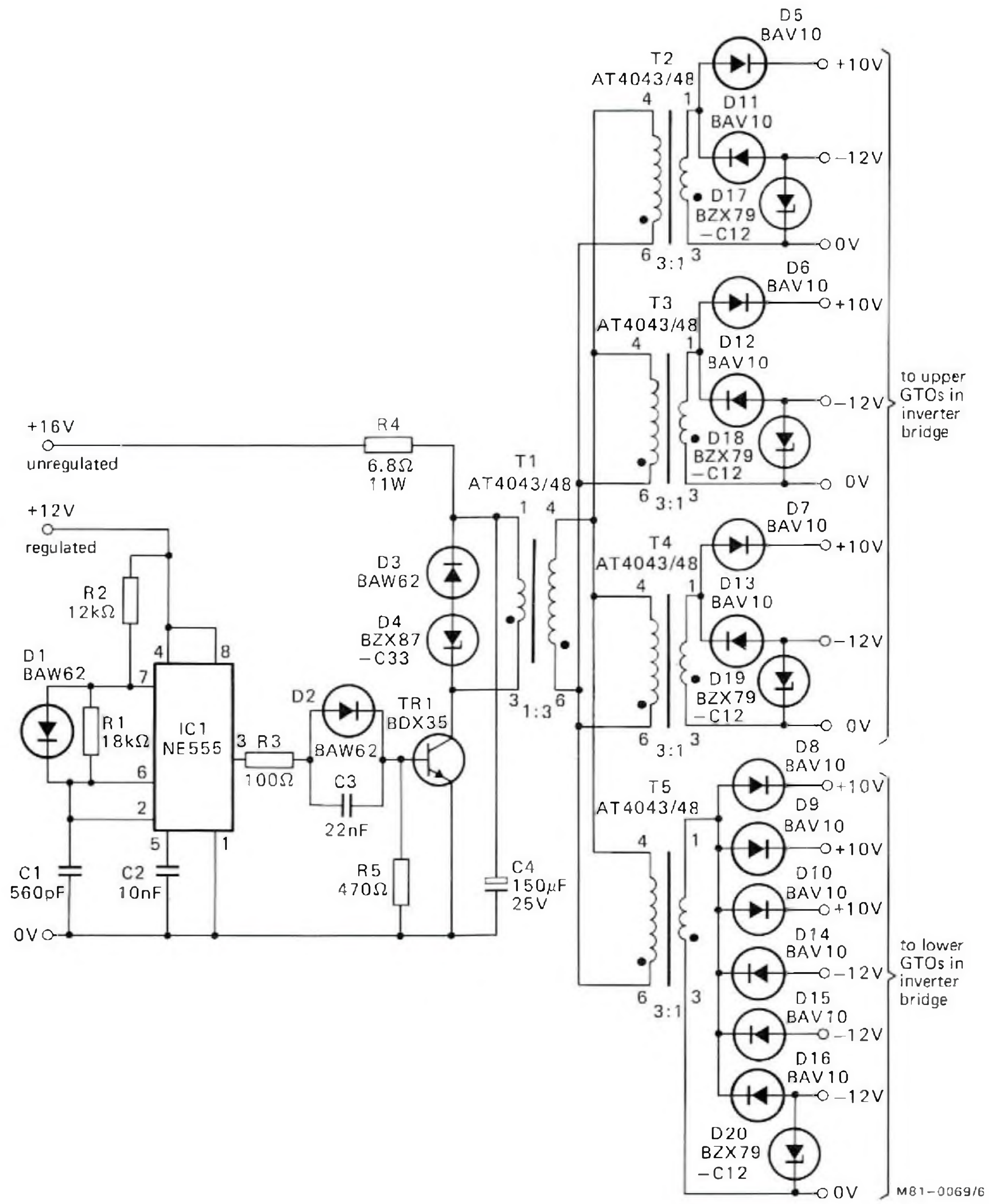


Fig.6 — Multiple-output isolated power supply

Closed-circuit tv cameras for surveillance often have to work under conditions that range from full sunlight to starlight: an eight-order-of-magnitude variation that no camera tube can handle alone. This article describes a practical combination that can. It consists of a Newvicon camera tube with a microchannel-plate image intensifier to adapt it to starlight, and an auto-iris objective lens with neutral density centre-spotting to safeguard the image intensifier in sunlight.

Lens-coupled intensifier unit for low-light-level CCTV cameras

The most widely used sensors for low-light-level tv cameras are

- silicon multiple-diode vidicons (silicon vidicons)
- intensified silicon-intensified-target vidicons (ISVs)
- fibre-optic or lens-coupled image intensifier/camera tube combinations.

Important factors to be considered when choosing the best sensor for a given application are the range of illuminance in which the camera must operate, the spectral content of the scene illumination, the resolution required, and the size, weight and cost of the camera.

Silicon vidicon. The simplest sensor is the silicon vidicon whose high sensitivity, compared with that of a conventional vidicon, is mainly due to its good infrared response. However, when used with visible light alone, it performs only slightly better than a conventional vidicon. Thus, the silicon vidicon can only be used down to a scene illuminance of about 10 lux (twilight). Sensors for use in lower illuminance must incorporate some form of light intensification.

Diode intensifier/vidicon combination. The simplest arrangement is a vidicon (or one of its derivatives) fibre-optically coupled to a single-stage diode intensifier. This combination performs well at an illuminance of 0.1 lux (full moonlight), making it suitable for some surveillance applications in towns where backscatter of light from the atmosphere provides this level of illuminance. The use of a cascade intensifier can extend the operating range down to a scene illuminance of about 1 mlux. Another sensor that can be used at this light level is the ISV.

ISV. The ISV comprises, in one evacuated envelope, a single-diode image intensifier (less screen) coupled to a vidicon with a silicon intensified target that is sensitive to photoelectrons. The ISV can be operated down to the photoelectron noise limit. A disadvantage of the ISV is that it requires a dedicated camera design and, furthermore, failure of either the intensifier part or camera tube part means replacing the whole tube.

A disadvantage common to all the combinations just described is that they use first-generation intensifiers

Using an XX1500TV image intensifier in front of the camera tube enables CCTV cameras to operate down to starlight scene illumination. Top left: standard Video 40 camera; bottom right: the same camera with intensifier unit which are intolerant of bright light sources and rapid changes in scene illumination. Highlight blooming makes these intensifiers unsuitable for most urban surveillance applications.

Microchannel-plate intensifier/Newvicon combination. The combination of second-generation intensifier and Newvicon can also operate down to the photoelectron noise limit. In addition, it has the advantage of using an intensifier that is small and light and that has intrinsically good highlight suppression.

FIBRE-OPTIC OR LENS COUPLING?

There are three basic ways to couple an intensifier to a camera tube:

- direct fibre-optic coupling
- earthed-plane fibre-optic coupler
- lens coupling.

Although fibre-optic (FO) coupling has higher light-coupling efficiency, lens coupling has decisive practical advantages for an add-on unit.

Direct FO coupling

Here, the FO output window of an image intensifier without an integral power supply is mounted in optical contact with the FO faceplate of the camera tube.

Disadvantages

- Requires a special camera tube with FO faceplate.
- Requires special mechanical mounting of the intensifier and camera tube to ensure optical contact of the fibre optics.
- Requires a separate non-standard power supply for the intensifier. Design of the supply is more complicated for a second-generation intensifier than for a first-generation one. To prevent spurious signals in the camera-tube output, the screen electrode potential must be ripple-free; this requires complex earthing arrangements.
- Requires a non-standard intensifier with flying leads.

Earthed-plane FO coupler

The way round the power supply and earthing problems associated with direct FO coupling is to use an earthed-plane FO coupler. The coupler is a fibre-optic disc having on one face a transparent conductive film that is connected to a conductive coating around the disc circumference. The coupler is placed between the intensifier screen and the camera-tube faceplate with its conductive face in contact with the intensifier screen. The coating is connected to the main camera-tube earth and to a can screening the intensifier.

The coupler enables a standard intensifier with integral power supply to be used, but the other disadvantages of FO coupling remain and new problems appear.

Disadvantages

- Still requires a special camera tube with FO faceplate.
- Still requires special mechanical mounting of the intensifier and camera tube to ensure optical contact of the fibre optics.
- Additional light-coupling loss occurs between the intensifier and camera tube.

Lens coupling

Advantages

- Existing cameras can be used with an add-on unit.
- Uses a basically standard image intensifier and standard camera tube (i.e. not FO).
- Field replacement of either the intensifier or the camera tube is possible.
- System is not restricted to a 1:1 aspect ratio, so smaller camera tubes can be used.

Disadvantages

- Lower light-coupling efficiency than with FO coupling.
- Longer optics; however, the possibility of using smaller camera tubes compensates.

CAMERA WITH XX1500TV ADD-ON UNIT

As Figure 1 shows, the elements of the camera are:

Objective. This should be any 1-inch format camera tube lens having a C-mount, the type of lens depending on the field of view and f-number required. For high scene illumination, the lens must be fitted with an iris stop and light attenuating filters to ensure that the illumination of the XX1500TV photocathode does not exceed the maximum continuous rating of 1 mlux.

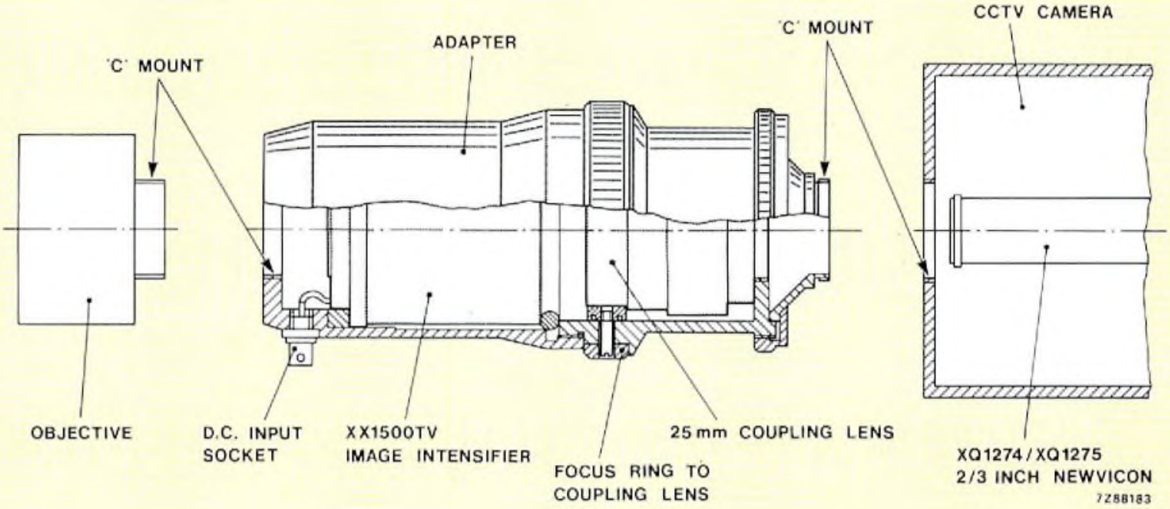


Fig.1 Components of a low-light-level CCTV camera incorporating an XX1500TV image intensifier

XX1500TV image intensifier. The XX1500TV is a low-distortion self-focusing microchannel-plate intensifier with an integral power supply having automatic brightness control (ABC). Minimum gain of the XX1500TV is 65 000 and the intensifier has intrinsically good high-light suppression and bright-source protection. Figure 2 shows the transfer characteristic. Table 1 shows the specification.

The intensifier requires a low-impedance d.c. voltage supply (2–3.6 V) which can be derived from the television camera through a suitable regulator, or from a separate source, e.g. two 1.5 V dry cells. Power consumption is less than 80 mW. A filter may be required to prevent any interference on the low voltage side of the intensifier power supply reaching the video amplifier of the camera, see Fig.3. The leads from the filter to the intensifier should be as short as possible. The add-on unit should be earthed.

There is a rectangular mask on the FO input window to reduce veiling glare caused by light entering the intensifier outside the field of view of the system. The use of the mask enables the light level at which the intensifier ABC operates to be increased.

Coupling lens. The coupling lens focuses the image on the FO screen of the XX1500TV onto the faceplate of the camera tube. In the add-on unit described here, a Fujinon 25 mm, f/1.4 lens was used.

Camera tube. The camera tube is a 2/3-inch Newvicon XQ1274 (magnetic focusing) or XQ1275 (electrostatic focusing). The Newvicon is chosen because of its high sensitivity and low lag. Figure 4 shows the light transfer characteristic of the Newvicon. Table 2 shows the specification.

The add-on unit can be adapted for use with 1-inch camera tubes.

TABLE 1
Brief specification of the XX1500TV

photocathode	S25	
white light sensitivity	350	μA/lm
sensitivity at λ = 800 nm	35	mA/W
sensitivity at λ = 850 nm	25	mA/W
gain	70 000	
input illuminance for onset of ABC	0.6	mlx
modulation transfer factor		
2.5 cycles/mm	min. 90%	
7.5 cycles/mm	min. 65%	
16 cycles/mm	min. 30%	
limiting resolution	36 (~390 TV lines)	lp/mm
useful photocathode area	10.8 mm × 14.4 mm	

TABLE 2
Brief specification of the XQ1274 and XQ1275

dark current at 25 °C	5	nA
signal current for a faceplate illuminance of 1 lux, colour temp. 2856 K	260	nA
resolution at picture centre	650	TV lines
max. spectral response	750	nm

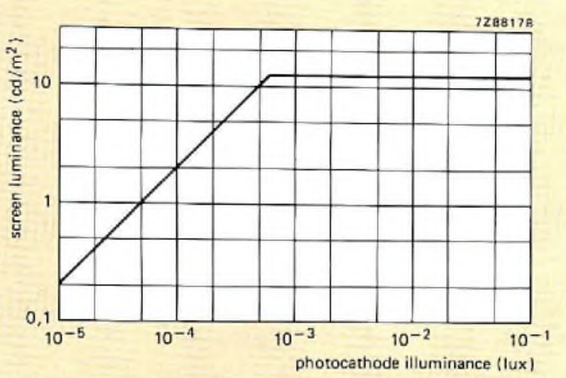
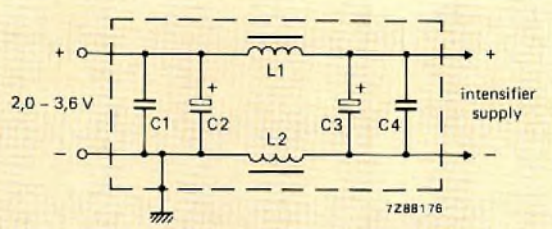


Fig.2 Transfer characteristic of the XX1500TV



C1, C4	47 nF ceramic
C2, C3	220 µF, 6.3 V
L1, L2	1 mH

Fig.3 Suggested decoupling filter for the intensifier power supply

PERFORMANCE

Newvicon signal current as a function of scene illuminance

The relation between the illuminance at the intensifier input window, intensifier screen luminance and the illuminance of the Newvicon faceplate as functions of scene illuminance and reflectivity are given in the Appendix, Eq.(1) to (4). For a camera with the optical parameters shown in Table 3, the illuminance E_i of the input window of the intensifier in lux is

$E_i = 5.9 \times 10^{-2} E_s r$ from Eq.(1), see Appendix

where

E_s is the illuminance of the observed scene in lux, and

r is the average reflectivity of the scene.

The intensifier screen luminance L_s in cd/m² for an intensifier with a gain of 70 000 is

$L_s = 2.2 \times 10^4 E_i$ from Eq.(3)

From Eq.(4), the illuminance of the Newvicon faceplate E_N in lux is

$E_N = 2.0 \times 10^3 E_i$.

Because the image seen by the Newvicon appears on the green P20 phosphor screen of the intensifier, the effective sensitivity of the Newvicon is only 0.24 times the value published for light from a 2856 K source. Hence, the effective faceplate illuminance $E_{N(eff)}$ is $4.8 \times 10^2 E_i$.

TABLE 3
Optical parameters of LLLTV camera with XX1500TV and XQ1274

objective		
relative aperture f_o	1.9	optional values
transmission coefficient T_o	0.85	
coupling lens		values for the adapter
relative aperture f_c	1.4	
transmission coefficient T_c	0.62	
magnification between intensifier screen and Newvicon faceplate	0.65	

For a typical Newvicon whose signal current is 260 nA for 1 lux faceplate illuminance from a 2856 K source, the signal current I_s (also in nA) in the intensified camera is

$$I_s = 260\, E_{N(eff)}$$
$$\approx 1.3 \times 10^5 E_i$$
$$\approx 7.5 \times 10^3 E_s r.$$

Figure 4 shows the Newvicon signal current as functions of input illuminance and scene illuminance times scene reflectivity for the XQ1274 Newvicon alone, and for the camera with XX1500TV as an intensifier for the Newvicon.

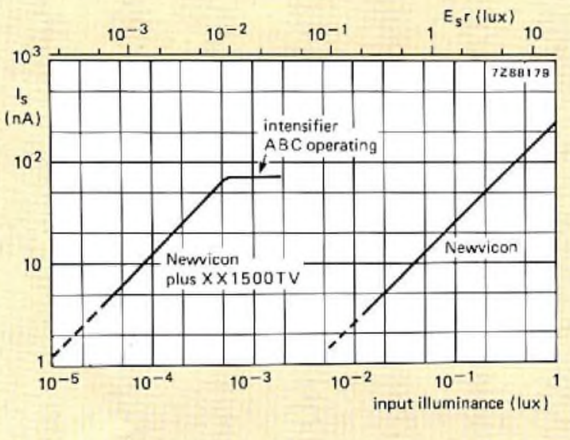


Fig.4 Signal current of the XQ1274 Newvicon, I_s, as functions of input illuminance and scene illuminance times scene reflectivity, $E_s r$, for the bare Newvicon and for one coupled to an XX1500TV intensifier by means of the adapter. Intensifier gain 70 000; $f_c/1.4$; $T_c = 0.62$; $m = 0.65$; $f_o/1.9$; $T_o = 0.85$

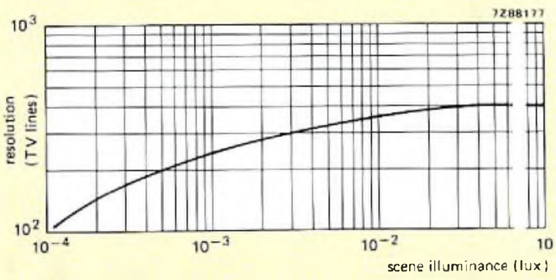


Fig.5 Resolution of a CCTV camera with XQ1274 Newvicon fitted with an XX1500TV intensifier and adapter as a function of scene illuminance. Intensifier gain 70 000; objective Tamron 75 mm, f/1.9; coupling lens Fujinon 25 mm, f/1.4; target USAF 1951 resolution target array with 100% contrast; scene reflectivity 100%

The intensifier ABC operates when $E_i > 0.54$ mlux. Increased illuminance will produce no significant increase in signal current, which at 0.54 mlux illuminance is about 70 nA, see Fig.4. At this illuminance, camera resolution has almost reached its maximum of 400 TV lines (Fig.5).

Picture quality depends on the performance of the camera video amplifier. However, a signal current of 30 nA is usually sufficient for a good picture; one as low as 4 nA is still adequate for surveillance applications and requires a scene illuminance times scene reflectivity of only 0.6 mlux. This means that the camera can operate down to at least 1 mlux scene illuminance (starlight). Use of a faster objective, for example, one with f/0.95 could (assuming the same system transmission) improve the performance four times.

Light attenuation at high illuminance

Although an input illuminance greater than 0.54 mlux will not increase the signal current, it may reduce the operational life of the intensifier. When used in surveillance, a CCTV system with image intensifier is likely to be used 24 hours a day. If no artificial illumination is available, the camera must be able to operate in a scene illuminance that changes by some 8 orders of magnitude: from 10^5 lux in bright sunlight, down to 10^{-3} lux in starlight.

For optimum performance, maximum intensifier life, and avoidance of the 'chicken-wire' effect, the continuous illuminance of the intensifier input window should be restricted to a maximum of 1 mlux. For bright sunlight, this requires an attenuation of 10^8 which can be realised with an objective lens having a T-number of 3600 (T-number equals f-number divided by the square root of transmission). Auto-iris lenses with neutral density spotting at the centre of the lens can be used to achieve this attenuation. When using an auto-iris lens, ensure that iris control starts at a light level below that which causes the intensifier ABC to operate.

APPENDIX

The illuminance of the input window of the intensifier E_i, in lux, is

$$E_i = E_s r T_o / 4f_o^2 \qquad (1)$$

where

- E_s is the illuminance of the observed scene in lux
- r is the average reflectivity of the scene
- T_o is the transmission of the objective
- f_o is the relative aperture of the objective.

The illuminance of the faceplate of the Newvicon E_N, in lux, is

$$E_N = \pi L_s T_c / 4f_c^2 (m + 1)^2 \qquad (2)$$

where

- L_s is the intensifier screen luminance in cd/m^2
- T_c is the transmission of the coupling lens
- f_c is the relative aperture of the coupling lens
- m is the magnification of the image between the intensifier screen and the Newvicon faceplate.

The gain G of the intensifier is:

$$G = \pi L_s / E_i \qquad (3)$$

From Eq.(2) and (3),

$$E_N = G T_c E_i / 4f_c^2 (m + 1)^2 \qquad (4)$$

Simplified synchronous power pack with diode-split transformer for colour TV

A switched-mode power supply which is synchronised with the line scan in a colour television receiver is able to generate all of the d.c. supply voltages, including the e.h.t., and provide drive for the horizontal deflection circuit. Such a Synchronised Power Pack (SPP) is well-known as an economical, compact and efficient mains-isolated voltage source. It also allows the bulky conventional line output transformer to be reduced to a small loading coil or choke. We have now refined the SPP, extended its facilities, and developed it into a new colour television supply system known as the Simplified Synchronous Power Pack (S^2P^2). Although this system provides the same mains-isolated power supplies as the SPP, it is more compact and has fewer components. Furthermore, the scan voltage for the horizontal deflection coils can be derived directly from the S^2P^2, thereby entirely eliminating the conventional horizontal deflection circuit.

The main advantages of the S^2P^2 are:

- All of the supplies for the receiver, including the e.h.t., are derived from the S^2P^2 transformer.
- The e.h.t. source impedance is low.
- The power consumption is less than 40 W. This has been achieved by reducing the switching losses and eliminating the double-conversion method of obtaining the e.h.t. and scan voltages.
- Reduced circuit board area, mainly due to the replacement of the conventional horizontal output transformer and elimination of horizontal deflection circuitry.
- The S^2P^2 output transformer provides mains isolation which conforms to IEC65 standards.

SPP PRINCIPLE

Before describing the operating principles of S^2P^2, it is useful to review the operating principles of the basic SPP shown in Fig.1.

When S_1 is closed, the rectified mains input V_{in} is connected across L_1 and the primary winding of T_1. The current through L_1 therefore increases linearly according to $dI/dt = V/L$, and energy ($\frac{1}{2}LI^2$) is stored in L_1. When S_1 opens, the polarity of the voltage across L_1 reverses and the stored energy is transferred, via D_1, to a charge on C_1. By varying the duty factor (δ) of the switch, the amount of energy stored in L_1 can be controlled and the voltage across C_1 will conform to the expression:

$$V_{C1} = \delta V_{in}.$$

The circuit C_1, T_1, C_f, D_3 and S_1 operates like that of a horizontal timebase circuit in which C_1 maintains a constant potential across the primary winding of T_1 during the 'scan' period, and negative-going 'flyback' pulses appear across the primary winding of T_1 during the 'flyback' period. Rectification of a secondary winding voltage from T_1 during the 'scan' period provides mains-isolated low-voltage high-current supplies for the receiver circuits. Rectification of a separate secondary winding voltage from T_1 during the 'flyback' period provides mains-isolated high-voltage low-current e.h.t. and focus supplies.

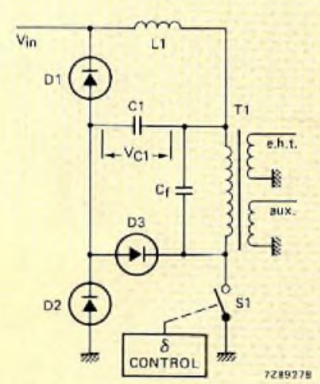


Fig.1 Basic SPP circuit

S^2P^2 PRINCIPLE

Extension of the SPP principle allows the value of C_f to be selected so that the drive voltage for the horizontal deflection coils can be derived from a secondary winding on T_1 without having to increase the maximum voltage rating of the switch. The basic circuit is given in Fig.2. In this circuit, the voltage across C_1 conforms to the expression

$$V_{C1} = \frac{\delta V_{in}\, m}{1 - \delta(1 - m)}$$

This expression is valid for $m < 1$, $m = 1$ (conventional SPP) and $m > 1$. With correct choice of value ($m \approx 0.5$), the circuit can operate with a mains input voltage within the range of at least 165 V to 265 V. Interchanging the V_{in} and D_1 connections to L_1 gives a reciprocal value for m. This allows the circuit to operate at lower mains input voltage (e.g. 110 V).

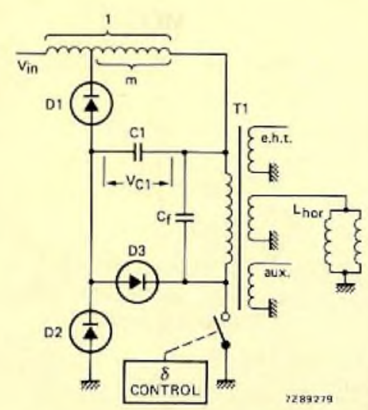


Fig. 2 Basic S^2P^2 circuit

S^2P^2 TRANSFORMER

Transformer DT2076/80 shown in Fig.3 is specifically designed for the S^2P^2 circuit in the economy colour television receiver shown in block form in Fig.4. A brief specification for the transformer is as follows:

Primary voltage	120 V
Secondary voltages	
horizontal deflection coil drive	110 V and 120 V
e.h.t.	23 kV
maximum focus voltage	33% of e.h.t.
vertical deflection stage	26 V
small-signal circuits	16 V (suitable for 12 V stab.)
picture tube heater	6.3 V r.m.s.
video amplifiers	200 V

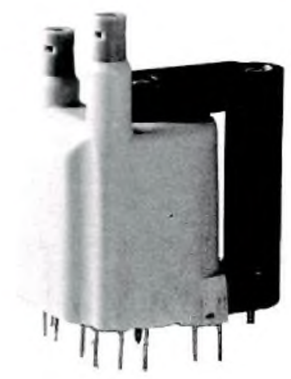


Fig.3 S^2P^2 output transformer

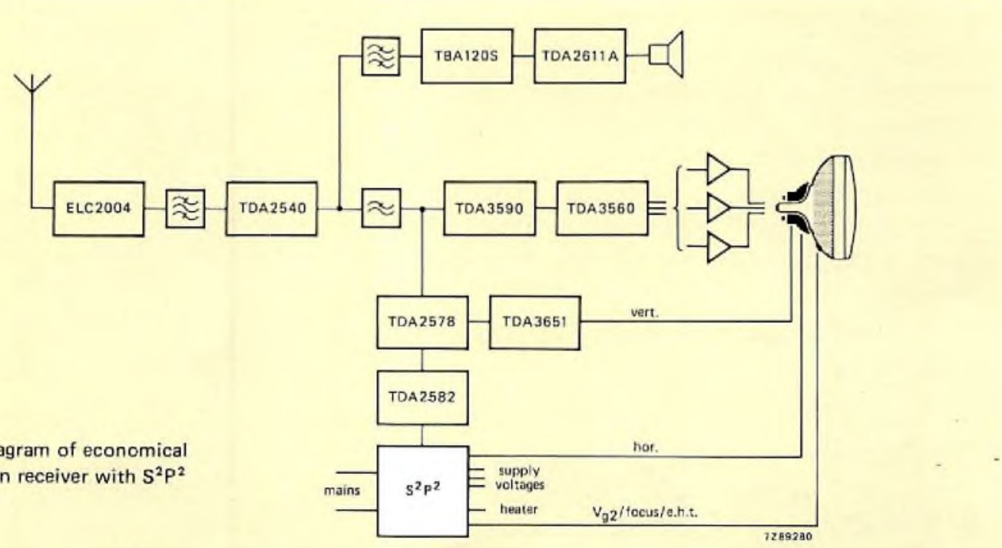


Fig.4 Block diagram of economical colour television receiver with S^2P^2

Concluding the series on motor-speed control that has appeared in previous issues, this article discusses the design of a switched-mode power supply for all the stabilised voltages required by the described PWM system.

Auxiliary power supply for a.c. motor speed control system

F.J. BURGUM and L.E. JANSSON

The design and operating principles of our Pulse-Width Modulation (PWM) speed control system for a.c. motors have been extensively described in the series of articles listed in Refs.1 to 6. This article, concluding the series, is devoted to the switched-mode power supply which provides the various isolated low-voltage d.c. outputs required for the operation of the PWM system. The detailed design of the PWM system will vary with differing user requirements, and these variations in design will be reflected in changes to the specification of the power supply outputs. The power supply described in this article should therefore be seen as an example illustrating the general principles rather than a universally applicable design.

PWM DRIVE FOR A.C. MOTORS

A block diagram of the PWM a.c. motor speed control system is shown in Fig.1. The three-phase mains input is connected to the bridge rectifier via an interference filter. An input filter smooths the rectified mains, and the output from the input smoothing capacitors provides the d.c. supply for the inverter. The output from the inverter consists of three-phase sinewave-weighted pulse-width-modulated waveforms at the selected carrier frequency; this gives sinusoidal motor currents with low harmonic content. The waveforms are synthesised by a purpose-designed LSI circuit, type HEF4752V. Control of the system is achieved by the analogue control inputs (under user control) and three internal feedback signals.

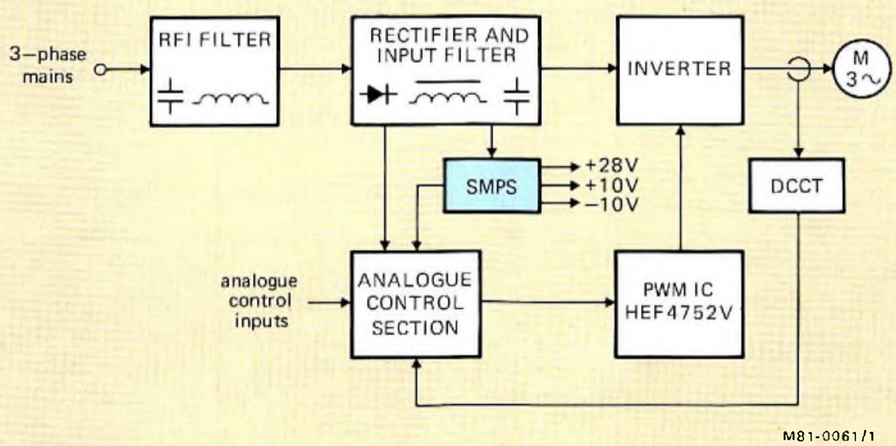


Fig.1 – PWM motor drive system

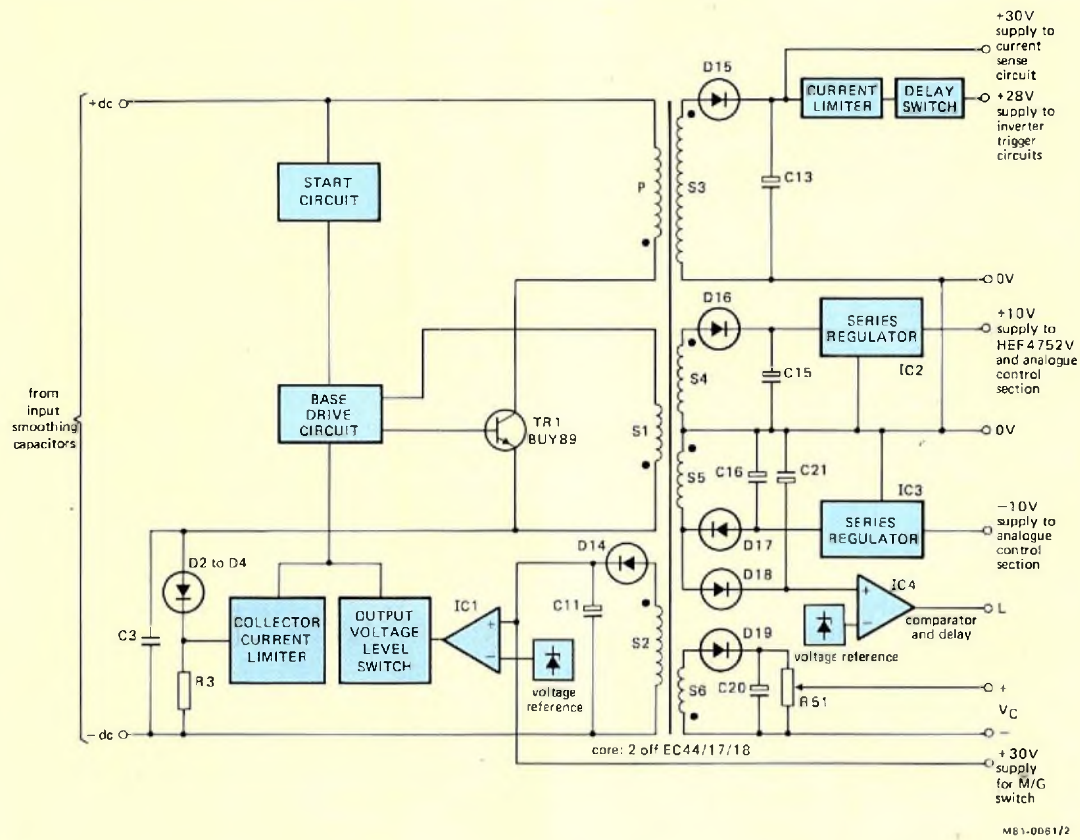


Fig.2 — Block diagram of switched-mode power supply
(component numbers correspond to identical components in Fig.3)

These feedback signals provide information on the voltage across the input smoothing capacitors, the level of motor current, and indicate whether the motor is in the motor mode or generator mode. A full description of the control process is given in Ref.4.

When the motor is operating in the generator mode, the energy produced is stored in the input smoothing capacitors, and the voltage across these capacitors will thus vary considerably. Since this voltage provides the input to the power supply, tolerance of a wide input voltage range is an essential requirement of the power supply design. Meeting this requirement provides an incidental advantage in that the power supply will then operate without modification from a battery, or from a single-phase or three-phase mains input.

POWER SUPPLY SPECIFICATION

The power supply is required to provide the following seven outputs.

1) +28 V for the inverter trigger circuits
2) +10 V for the LOCMOS IC (HEF4752V) and the analogue control section
3) −10 V for the analogue control section
4) +30 V for the motor current sense circuit
5) +30 V for the motor mode/generator mode indicator switch ($\overline{M}$/G)
6) +10 V logic signal (L) to enable/disable the complete drive system at switch-on and switch-off
7) isolated signal V_C proportional to the d.c. voltage across the input smoothing capacitors

The circuit must generate these outputs from d.c. input voltages in the range 200 to 800 V.

POWER SUPPLY CIRCUIT DESIGN

The switched-mode power supply is a self-oscillating flyback converter using a voltage feedback winding to regulate variations in outputs 1 to 6 (see above specification) caused by variations in the high-voltage d.c. input (200 to 800 V). In addition, integrated series regulators limit variations in the +10 and −10 V supplies arising from changes in load current. The power switching element is the new BUY89 high-voltage transistor, and the transformer is wound on two EC44/17/18 ferrite cores. A block diagram of the circuit is shown in Fig.2, and a complete circuit diagram is shown in Fig.3.

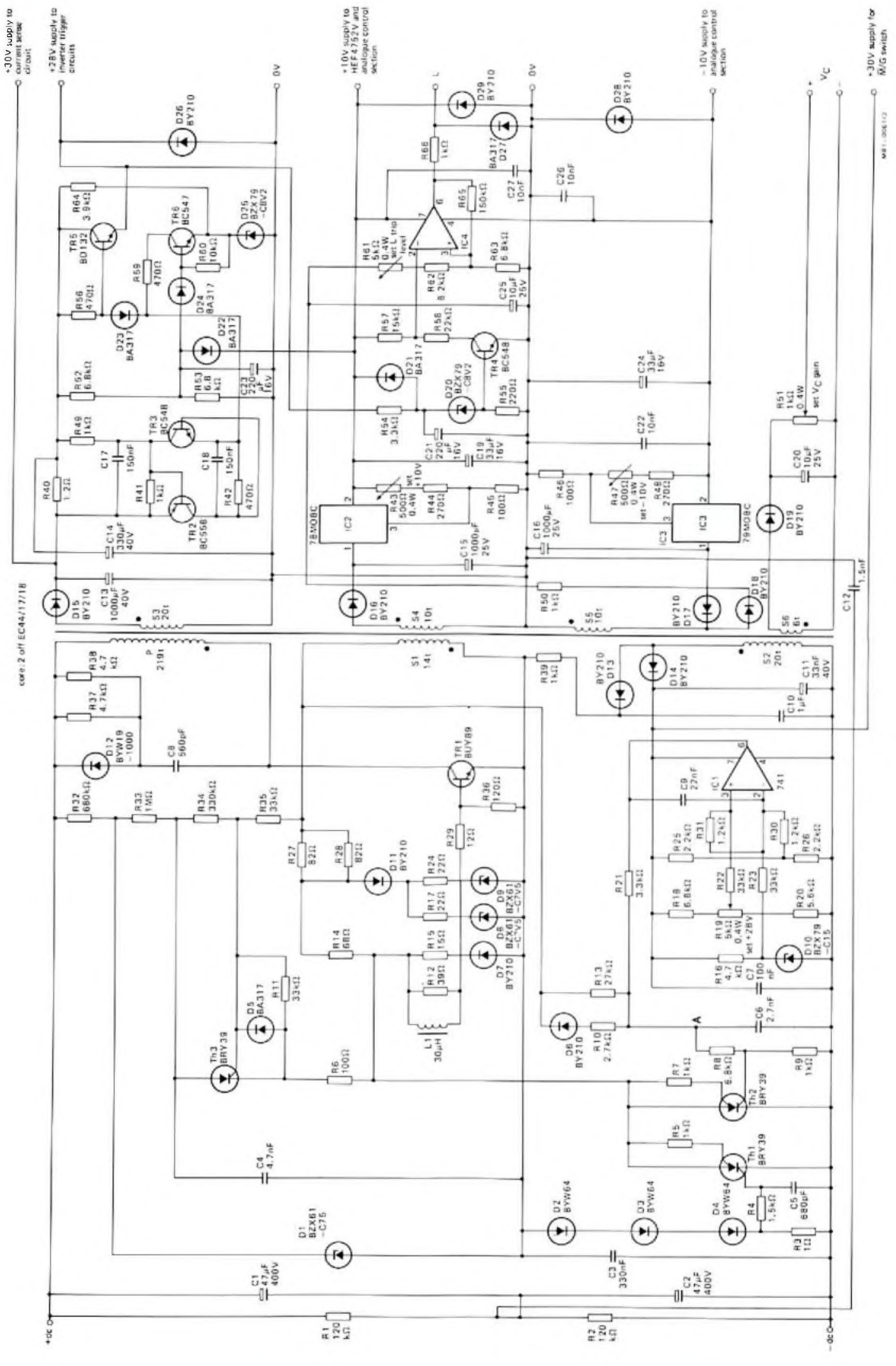


Fig.3 – Switched-mode power supply circuit diagram

Start circuit

With the power supply connected to the input smoothing capacitors, capacitor C4 (Fig.3) is charged via R32 and R33. The gate voltage of Th3 rises until the device conducts at which point C4 is discharged through Th3, R6, L1, R12, R29, and the base of TR1. This turns TR1 on, and the base drive to maintain conduction is provided by S1.

Switching action

Although the switching element TR1 is initially driven into conduction by the start circuit, the basic switching action of the power supply is achieved by switching TR1 off via Th2 in the voltage stabilisation circuit, and by switching TR1 on by the action of S1.

With TR1 in conduction, capacitor C6 is charged via R13. The potential at the point A (Fig.3) will rise until Th2 is switched on, which in turn switches off the base drive of TR1.

With TR1 off, the collector voltage of TR1 rises to about 340 V above the supply voltage, and remains at this level until the energy stored in the transformer is transferred to the output. A voltage oscillation in the primary then causes the collector voltage to fall below the supply voltage. The e.m.f. in S1 reverses, providing the base drive to switch TR1 back into conduction.

Voltage stabilisation

The voltage stabilisation circuit compensates for changes in the d.c. output level by varying the mark/space ratio of the switched waveform. The variations in mark/space ratio are effected by raising and lowering the current flowing into point A via R21.

Changes in the d.c. output level are sensed by the voltage feedback winding S2. When the d.c. output rises above the desired level, the output of IC1 goes positive, the potential at point A rises, reducing the time to reach the trip-on level of Th2. The mark/space ratio of the chopped waveform is thus reduced and the d.c. output voltage falls. Conversely, if the d.c. output voltage falls below the desired level, the output of IC1 goes negative, the potential at point A falls, and the mark/space ratio of the chopped waveform is increased.

Collector current limiter

The collector current limiter protects TR1 against excessive collector current. The action of the current limiter is determined by the voltage across R3, which will rise as the current through TR1 rises. Above a certain voltage Th1 switches on, thereby switching off the base drive of TR1.

Output stabilisation and protection

While the voltage stabilisation circuit described above provides excellent stability against variations of input voltage, it does not give the same protection against output voltage variations that result from changes in load current. Integrated series regulators (IC2 and IC3) are used to protect the +10 and −10 V supplies against variations of this kind.

The +10, −10, and +28 V outputs are protected against indefinite short-circuits. The +10 and −10 V outputs use fold-back regulators, and the +28 V output is switched off under overload conditions by a current limiter. The full specification of the +10, −10, and 28 V outputs is given in Table 1.

Switch-on and switch-off

At switch-on the voltage across the input smoothing capacitors rises with a time-constant determined by the d.c. link charging circuit (see Ref.4); a value of 66 ms is typical. To ensure safe operation of the PWM system, the trigger circuit supply voltage is inhibited until the input voltage is sufficiently high. Conversely, at switch-off the trigger circuit supply is disabled before the input voltage falls below a preset minimum.

TABLE 1

Detailed specification for +10, −10, and +28 V outputs

Nominal output V	Maximum current mA	Output variation with output current varied from zero to maximum	Output variation with input voltage varied from 200 to 800 V d.c. mV	Peak-to-peak ripple mV
+10	110	<100 mV	< 50	<100
−10	140	<100 mV	< 50	<100
+28	500	24 to 32 V	<100	<150

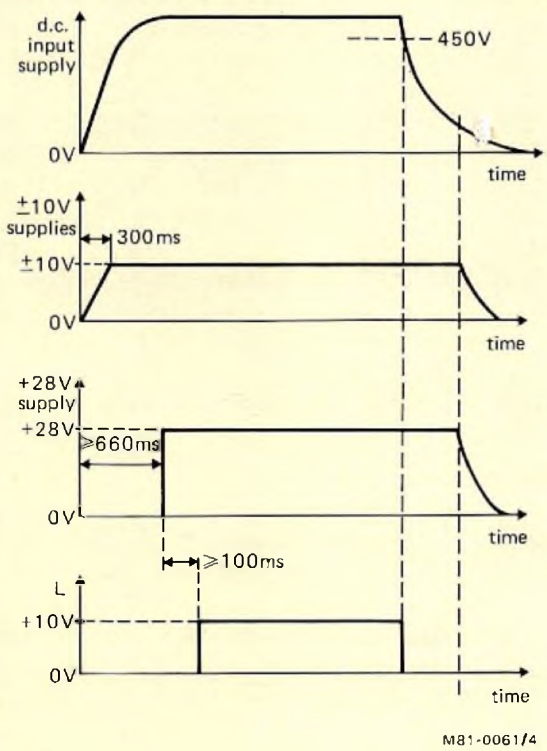


Fig.4 — Switch-on and switch-off waveforms

The various voltage waveforms at switch-on and switch-off are shown in Fig.4. During the initial switch-on period, the +10 and −10 V supplies are established after approximately 300 ms. The +28 V output is disabled by a delay switch for approximately 660 ms. The control signal L is generated using the −10 V transformer winding in the forward converter mode. The output from this winding is proportional to the d.c. input voltage, and is compared with a reference voltage in a comparator and delay circuit. Consequently, the output from the comparator and delay (the signal L) is inhibited for a time that depends on the magnitude of the d.c. input voltage. The L signal is used to enable the complete drive system. Exactly how this is done will depend on the design of the analogue control section. At switch-off L goes LOW when the d.c. input voltage falls below 450 V.

Transformer

Details of the transformer design are listed in Table 2. The transformer is wound on two EC44/17/18 ferrite cores, ordering code 8213 140 25320*. The gap spacer thickness is 0.34 mm.

COMPLETE A.C. MOTOR SPEED CONTROL SYSTEM

This article concludes the description of our PWM speed control system and completes the series listed as Refs. 1 to 6. However, the described system is subject to continuing development, and individual articles in the series may not always reflect the most recent innovations. Requests for information on these should be addressed to the editors.

*UK ordering code FX3838

TABLE 2

Transformer for switched-mode power supply

Circuit reference Fig.3	Total number of turns	Turns per layer	Number of layers	Copper diameter mm	D.C. output voltage
P	219	73	3	0.16	–
S1	14	14	1	0.63	–
S2	20	20	1	0.63	+30 V (internal)
S3	20	20	1	0.63	+30 V (external)
S4 } bifilar	10	20	½	0.63	+14 V (+10 V regulated)
S5 } bifilar	10	20	½	0.63	−14 V (−10 V regulated)
S6	6	6	1	0.63	$\frac{\text{d.c. input voltage}}{36.5}$

REFERENCES

1. HOULDSWORTH, J.A. and ROSINK, W.B., 'Introduction to PWM speed control system for 3-phase AC motors', Electronic Components and Applications, Vol.2, No.2, February 1980, pp.66 to 79.
2. BURGUM, F. and NIJHOF, E.B.G., 'Inverter circuit for PWM motor speed control system', Electronic Components and Applications, Vol.2, No.3, May 1980, pp.130 to 142.
3. STARR, B.G. and van LOON, J.C.F., 'LSI circuit for AC motor speed control', Electronic Components and Applications, Vol.2, No.4, August 1980, pp.219 to 229.
4. ROSINK, W.B., 'Analogue control system for an AC motor with PWM variable speed drive', Electronic Components and Applications, Vol.3, No.1, November 1980, pp.6 to 15.
5. HOULDSWORTH, J.A., 'Purpose-designed ferrite toroids for isolated current measurement in power electronic equipment', Electronic Components and Applications, Vol.3, No.2, February 1981, pp.101 to 109.
6. HOULDSWORTH, J.A. and SCHMICKL, H. 'Electrolytic capacitors for industrial applications', Electronic Components and Applications, Vol.3, No.3, May 1980, pp.167 to 180.

Quality line

The fundamental definition of Quality is fitness for use: fitness, not only initially, but during the required lifetime of a product. Thus, ultimately, the only true measure of the quality of components is the success with which they can be incorporated in equipment, and the reliability of that equipment in service.

Both aspects of quality have a large influence on equipment-manufacturing costs. If a large proportion of finished equipments contain faults – high fall-off rate – the cost of correction will be high. If automatic assembly methods are used, the cost of correction – a labour-intensive activity – will be disproportionately high. Field servicing – call rate – is also expensive; moreover, frequent failures damage the OEM's reputation.

There is a limit to the quality that can be achieved by component-manufacturer and OEM working in isolation. Continual contact is necessary if the component-manufacturer's test and optimisation criteria are to be relevant to the applicaton, and if the OEM is fully to appreciate the capabilities of the components. Furthermore, analysis by the manufacturer of components that fail to meet the application requirements in any way can provide both parties with the data essential to improving the quality achieved where it really matters – in use.

HIGH-VOLTAGE POWER TRANSISTORS CONSISTENTLY RELIABLE

Analysis of endurance testing of high-voltage power transistors shows that their reliability approaches that of their low-voltage counterparts. The testing was part of routine Quality-Control Acceptance Testing carried out between 1977 and 1979. It includes life tests of 168 h, and 168 h to 1000 h duration, with extended tests of up to 10 000 h duration. Testing is carried out under Absolute-Maximum Rating conditions.

The 168 h tests indicate the level of early failures. After 168 h, the devices have entered the constant-failure period; longer-term tests check the indications of the 1000 h tests.

Our range of high-voltage power transistors has been developed to cover all the major areas of application: tv and VDU deflection, electronic ignition, switched-mode power supplies and electronic motor control. Rigorous quality control and the most advanced fabrication techniques achieve the high, built-in quality essential to the meeting of modern requirements.

During the period 1977 to 1979, over 10 million device-hours routine testing was carried out on these devices. This includes over 12 million device-cycles of thermal-fatigue (power-cycling) testing. The table gives the cumulative results from these tests.

Derating from Absolute-Maximum Rating conditions to normal operating conditions indicates that failure rates of the order of 10^{-6}/h can be expected. This is comparable to the failure rates experienced with conventional, low-voltage power transistors.

Early-failure precentages and failure rates of high-voltage power transistors obtained from endurance tests carried out from 1977 to 1979 as part of routine Quality-Control evaluation

Note: all data corrected to a confidence level of 60%.

device type	early failures	failure rates (1000 h tests)	failure rates (>1000 h tests)
BD232	0.21%	6.53×10^{-6}/h	22.8×10^{-6}/h
BU208A	0.40%	4.60×10^{-6}/h	4.0×10^{-6}/h
BU426	0.95%	4.4×10^{-6}/h	7.8×10^{-6}/h
BUX80	0.19%	12.5×10^{-6}/h	3.8×10^{-6}/h
BUX82	0.75%	14.7×10^{-6}/h	3.4×10^{-6}/h
BUX84	0.31%	5.94×10^{-6}/h	* –
BUX86	1.01%	19.6×10^{-6}/h	21.7×10^{-6}/h

* These tests not performed.

More detailed information about these tests is available as a separate Technical Publication.

Research news

S.P.I. - the ultimate radio tuning aid

In recent years, microcircuits have allowed considerable improvement of the tuning characteristics and operating facilities in domestic and car radios. Fully electronic tuning with frequency synthesisers, digital storage of the frequencies of favourite programmes, and digital frequency indication are now commonplace. Even facilities such as automatic search tuning and microcomputer control are now incorporated in some top-class radios. All of these tuning systems have one drawback however; the listener must still initially locate the desired programme manually. With an f.m. radio, this can be a very tedious task since there are so many closely-spaced, often duplicated transmissions, no names on the tuning scale, a not very accurate manual tuning system and a host of four-digit frequencies to remember. Community aerial systems add extra aggravation because they often change the frequency of the received programmes before they reach the listener's radio. Car radio users experience particular difficulty as they traverse the boundaries of transmitter service areas of travel through mountainous regions. They are then distracted from their driving each time the signal fades by having to retune their radios to search for a transmission that allows continued listening to the desired programme.

We have now overcome all these problems by developing a Station/Programme Identification system (S.P.I.). With this system, f.m. radio transmissions include a subcarrier digitally modulated with information regarding the type of programme material being transmitted and the name of the radio station. The S.P.I. data is decoded in the receiver to provide an alphanumerical display of the station name and a programme type code. For example, HILV4 CM6 would indicate that the radio is tuned to the sixth classical music programme of the day from station Hilversum 4 in The Netherlands. S.P.I. data is already being continuously transmitted in The Netherlands and test transmissions are in progress in Austria, France, Germany and Switzerland.

The transmitted S.P.I. data can of course be extended to provide many automatic radio tuning facilities and functions in addition to the identification of the station and programme type. A few examples are:

- switching on a tape recorder to record a chosen type of programme material
- switching on the radio at predetermined times
- tuning to a programme selected by using a light-pen to read a bar code printed in a programme guide or newspaper
- tuning to a chosen type of programme such as classical music or news reports
- setting separate sound levels for speech and music
- transmission of 'radio text' for weather forecasts, traffic conditions, temperature, etc.

S.P.I. DATA TRANSMISSION STANDARDS

Since the S.P.I. subcarrier must not cause audible interference by beating with the 19 kHz stereo pilot tone, its frequency must be carefully chosen. Gaps are available in the f.m. mpx frequency spectrum at 15 kHz – 19 kHz, 19 kHz – 23 kHz and above 53 kHz. Experiments have shown that the two most suitable frequencies are $7/8 \times 19\,\text{kHz} = 16.625\,\text{kHz}$, and $9/8 \times 19\,\text{kHz} = 21.375\,\text{kHz}$. The present test transmissions use a low-level 16.625 kHz subcarrier which contributes only 250 Hz (0.3%) to the total carrier frequency deviation caused by the f.m. mpx signal (75 kHz max.).

The digital S.P.I. data consists of up to 19 alphanumerical characters, each represented by a 6-bit binary code. These 114 bits are preceded by a 14-bit burst for synchronisation, making a maximum total S.P.I. message length of 128 bits.

The bit transmission rate must be fast enough to allow quick read-out and prevent flicker of the display in the

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

receiver and to allow correct decoding of the S.P.I. message during search tuning. On the other hand, it must not be so fast that it causes excessive frequency deviation of the subcarrier. Taking these considerations into account, a repetition rate of about five messages (640 bits) per second is suitable. If the bit transmission rate is linked to the frequency of the reference carrier used in the S.P.I. decoder in the receiver (sub-multiple of 19 kHz), very reliable bit synchronisation is obtained. A bit transmission rate of 19 kHz/32 = 593.75 bits/s has therefore been chosen.

The digital S.P.I. data stream is modulated onto the 16.625 kHz subcarrier using two-phase phase-shift keying (PSK). Figure 1 shows that, with this method of modulation, the subcarrier phase is unchanged to transmit a binary 0, and is shifted 180° to transmit a binary 1.

S.P.I. DATA DECODER

To achieve a high signal-to-noise ratio and thereby minimise the probability of error in reception of the PSK S.P.I. data, coherent detection is used in the S.P.I. decoder. A decoder arrangement which has proved to give good results during the test period is shown in Fig.2.

In Fig.2 the 16.625 kHz PSK S.P.I. signal and the 19 kHz pilot tone are separated from the demodulated stereo multiplex signal by a bandpass filter. The 19 kHz pilot tone is then recovered in a PLL and mixed with the 16.625 kHz signal so that the S.P.I. signal frequency is converted to 19 kHz – 16.625 kHz = 2.375 kHz. This signal is then passed through a coherent detector, the 2.375 kHz reference carrier for which is derived by dividing the 19 kHz pilot tone frequency by eight. For coherent detection however, the reference carrier must be of the same phase as the transmitted S.P.I. subcarrier. The phase of the reference carrier is therefore corrected in a voltage controlled phase shift network controlled by a signal derived by detecting the phase difference between the reference carrier and the incoming S.P.I. signal. Since the modulated 2.375 kHz S.P.I. signal does not contain a discrete frequency component at the subcarrier frequency (as shown in Fig.1, its phase is shifted 180° for each binary 1 transmitted), it cannot be compared with the 2.375 kHz reference carrier to derive the reference carrier phase correction signal. The modulated 2.375 kHz S.P.I. signal is therefore passed through a square-law amplifier, the output from which is the same form as a full-wave rectified version of the modulated waveform in Fig.1(b). In other words, it has a fixed phase relationship to the transmitted S.P.I. subcarrier, and a ripple frequency of 2 x 2.375 kHz = 4.75 kHz as shown in Fig.1(c). To derive the reference carrier phase control signal, the output from the square-law amplifier is clipped and phase compared with the reference carrier which has also been frequency doubled to 4.75 kHz.

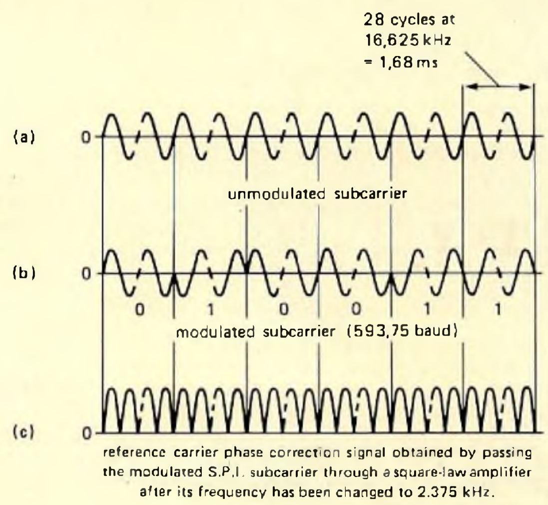


Fig.1 The two-phase PSK modulated subcarrier for transmitting S.P.I. data

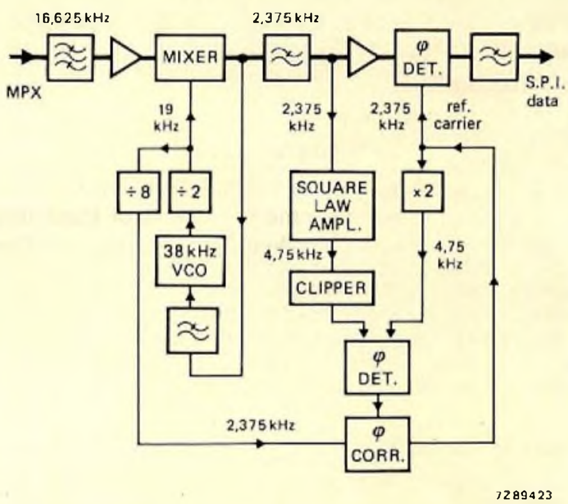


Fig.2 Block diagram of an S.P.I. decoder

After decoding, the S.P.I. signal is passed through a low-pass filter (integrator) with a time-constant of one S.P.I. data bit period. The output from the low-pass filter at the end of a bit period is a logic HIGH level when an antiphase subcarrier burst is received (binary 1 transmitted), and is zero when an in-phase subcarrier burst is received (binary 0 transmitted).

FURTHER INFORMATION

The S.P.I. system is described in more detail in Philips Technical Review, Volume 39, 1980, no. 8.

Abstracts

A.M. stereo – a new dimension for car radios

Proposals for a.m. stereo transmission in the U.S.A. are expected to result in adoption of a standard in which the L – R signal and a 5 Hz stereo pilot tone are phase modulated onto the a.m. carrier. Simple modification, requiring few additional components, enables a receiver previously described in EC&A, Vol. 3, No. 2, to decode such transmissions with good channel separation and to provide the usual stereo indication function.

Advances in solid-aluminium electrolytic capacitors

New etching technology roughly halves the size of solid-aluminium electrolytic capacitors, at no cost to the reliability, long life, or resistance to temperature extremes for which they are noted. Capacitors based on the new technology rival the smallness of tantalum electrolytics, and prospective further improvements promise to extend them into the range of capacitance-voltage product that has hitherto been the exclusive domain of wet electrolytics.

Wideband linear amplifiers in HF communications

R.F. power transistors are now widely used in the linear power stages of transmitters delivering up to about 1 kW output power, and devices capable of supplying up to 200 W p.e.p. with intermodulation distortion better than −30 dB are now available. This article summarises the factors governing the design of wideband linear amplifiers based on high-power r.f. transistors, and as an example describes a push-pull amplifier using only two BLW96 power transistors driven by two BLW50F transistors. The complete amplifier delivers up to 400 W p.e.p. with intermodulation distortion better than −26 dB.

Electronic components for telephone subscriber sets

Integrated circuits for interrupted current-loop or dual-tone multi-frequency dialling allow the traditional dial to be replaced by a keyboard. If the carbon microphone is replaced by an electret or electrodynamic microphone, the speech and transmission functions can also be performed by an IC. IC-controlled electronic ringers, offering a choice of distinctive tones or chimes, can replace the bell. Microcircuits also allow the incorporation of such features as automatic redialling, repertory dialling, dialled number display, tariff-unit metering, hands-free use, remote wireless handset, and the possibility of some data handling.

Basic GTO drive circuits

The development of electronic power switching techniques has received a strong impetus from the introduction of the gate turn-off switch (GTO). This article describes four GTO drive circuits which will facilitate the correct operation of the GTO over a wide range of applications. Isolated and non-isolated applications are covered, together with variations in the required range of duty cycle and switching frequency.

Auxiliary power supply for a.c. motor speed control system

This article, the last in a series, describes an auxiliary switched-mode power supply which provides the various low-voltage d.c. outputs required for the operation of our PWM speed control system for a.c. motors. The power switching element is the new BUY89 high-voltage transistor, and the transformer is wound on two EC44/17/18 ferrite cores.

AM-Stereo – eine neue Dimension im Autoradio

Vorschläge für AM-Stereo-Übertragungen in den USA lassen die Annahme einer Übertragungsnorm erwarten, bei der das (L – R)-Signal und ein 5 Hz Stereo-Pilotton den AM-Träger frequenzmodulieren. Es ist möglich, einen in dieser Zeitschrift, Band 3, Nr. 2 beschriebenen Empfänger durch wenige Bauelemente so zu erweitern, dass derartige Stereosignale mit guter Kanaltrennung decodiert werden und die übliche Stereoidentifikation durchgeführt wird.

Fortschritte bei Al-Kondensatoren mit festem Elektrolyten

Durch eine neue Ätztechnik gelang es, die Grösse von Kondensatoren mit festem Elektrolyten auf etwa die Hälfte herabzusetzen, ohne Einschränkung der Zuverlässigkeit, der Lebensdauer und der Temperaturbelastbarkeit. Die nach der neuen Technik hergestellten Kondensatoren können in ihrer Grösse mit Tantal-Elektrolytkondensatoren konkurrieren und dürften mit den noch zu erwartenden Verbesserungen ein Kapazitäts-Spannungsprodukt erreichen, das bist jetzt ausschliesslich den Kondensatoren mit nassem Elektrolyten vorbehalten war.

Lineare Breitbandverstärker für HF-Fernmeldeübertragungen

HF-Leistungstransistoren werden heute allgemein in linearen Leistungsverstärkerstufen von Sendern bis zu etwa 1 kW Ausgangsleistung eingesetzt, und derartige Verstärker können noch Leistungen bis zu 200 W (PEP) bei Intermodulationsabständen besser als 30 dB abgeben. In diesem Artikel werden die Faktoren behandelt, die den Entwurf von linearen Breitbandverstärkern mit HF-Leistungstransistoren bestimmen, und als Beispiel wird ein Gegentakt-Verstärker beschrieben, der zwei Leistungstransistoren BLW96 in der Endstufe und zwei Transistoren BLW50F in der Treiberstufe enthält. Der komplette Verstärker liefert eine Leistung bis zu 400 W (PEP) bei einem Intermodulationsabstand besser als 26 dB.

Elektronische Bauelemente für Telefonapparate

Integrierte Schaltungen für Impulswahlverfahren (IWV) oder Mehrfrequenzwahlverfahren (MFV) ermöglichen, die traditionelle Wählscheibe durch ein Tastenfeld zu ersetzen. Wenn statt des Kohlemikrofons Elektret- oder elektrodynamische Mikrofone verwendet werden, kann weiterhin eine integrierte Schaltung die Funktion der Sprechschaltung übernehmen. Elektronische Wecker, die die Klingel ersetzen und von integrierten Schaltungen gesteuert werden, erlauben die Wahl von Tönen mit unterschiedlichen charakteristischen Klangeindrücken, Mikroschaltkreise gestatten schliesslich die Einführung einer Reihe von Erweiterungen und Bedienungserleichterungen wie Anrufwiederholung, Kurzwahl, Rufnummeranzeige, Zählung der Gebühreneinheiten, Einführung von Freisprecheinrichtungen und drahtlosen Handapparaten sowie die Möglichkeit von Datenübertragungen.

Ansteuerschaltungen für GTO-Thyristoren

Die Entwicklung auf dem Gebiet des elektronischen Schaltens von Leistungen hat mit der Einführung von GTO-Thyristoren starke Impulse erhalten. Dieser Artikel beschreibt vier GTO-Ansteuerschaltungen, die den korrekten Einsatz von GTO-Thyristoren über einen weiten Anwendungsbereich erleichtern. Es werden Anwendungen mit und ohne galvanische Trennung sowie mit verschiedenen Tastverhältnissen und Schaltfrequenzen behandelt.

Stromversorgung zum Betrieb eines Systems zur Drehzahlregelung von Induktionsmotoren

Dieser Artikel, der letzte einer Serie, beschreibt ein zusätzliches Schaltnetzteil, welches die verschiedenen Versorgungsgleichspannungen liefert, die für unser System zur Drehzahlregelung von Induktionsmotoren mittels Impulsbreiten-Modulation benötigt werden. Als Leistungsschalter wird der neue Hochvolt-Transistor BUY89 verwendet; als Transformatorkern dienen zwei Ferrite-Kerne EC44/17/18.

Stéréo a.m. – une nouvelle dimension en autoradios

On s'attend à ce qu'aux Etats-Unis des propositions relatives aux émissions en stéréo a.m. aboutissent à l'adoption d'une norme suivant laquelle le signal G-D et une tonalité pilote stéréo 5 Hz seraient modulés en phase sur la porteuse a.m. Une modification simple, demandant peu de composants additionnels, permet à un récepteur précédemment décrit dans EC&A, Vol. 3, No. 2, de décoder de telles émissions avec une bonne séparation des canaux et de fournir la fonction d'indication stéréo habituelle.

Progres dans les condensateurs électrolytiques en aluminium solide

La nouvelle technologie de l'attaque des métaux permet de réduire de moitié les dimensions des condensateurs électrolytiques en aluminium solide, sans affecter la stabilité, la longévité et la résistance aux températures élevées qui font leur réputation. Les condensateurs utilisant cette nouvelle technologie rivalisent avec les condensateurs électrolytiques au tantale de moindres dimensions.

Amplificateurs linéaires à large bande en communications à haute fréquence

Actuellement, les transistors de puissance haute fréquence sont largement utilisés dans les étages de puissance linéaires d'émetteurs dont la puissance de sortie atteint environ 1 kW. Il existe des dispositifs capables de fournir une puissance de crête atteignant 200 W avec une distorsion d'intermodulation meilleure que −30 dB. L'article résume les facteurs qui régissent la conception des amplificateurs linéaires large bande utilisant des transistors haute fréquence de grande puissance et décrit à titre d'exemple un amplificateur push-pull qui n'utilise que deux transistors de puissance BLW96 pilotés par deux transistors BLW50F.

Composants électroniques pour postes téléphoniques d'abonnés

Le cadran classique peut être remplacé par un clavier, grâce à des circuits intégrés pour numérotation multi-fréquence à boucle de courant interrompu ou tonalité double. Les fonctions vocales et de transmission peuvent également être assurées par un circuit intégré à condition de remplacer le microphone à charbon par un microphone électrodynamique ou à électret. La sonnerie peut être remplacée par des timbres électroniques commandés par circuits intégrés, offrant un choix de tonalités ou carillons différants. Les microcircuits rendent également possible l'incorporation de fonctions telles que la renumérotation automatique, la composition automatique de numéros en mémoire, l'affichage du numéro composé, le comptage des unités de tarification, l'emploi mains libres, le combiné sans fil, et, le traitement de données.

Circuits pilotes pour commutateurs à grille de contrôle (GTO)

L'introduction du commutateur à grille de contrôle (GTO) a donné une forte impulsion au développement des techniques de commutation électronique de puissance. Cet article décrit quatre circuits pilotes pour commutateurs à grille de contrôle qui faciliteront le fonctionnement correct de ceux-ci dans une large gamme d'applications. Des applications avec et sans isolement sont décrites, en même temps que sont précisées les gammes correspondantes de coefficient d'utilisation et de fréquence de commutation.

Alimentation auxiliaire pour variateurs de vitesse de moteurs à courant alternatif

Cet article, le dernier d'une série, décrit une alimentation auxiliaire à découpage, qui fournit les divers courants continus basse tension nécessaires au fonctionnement de notre variateur de vitesse à modulation par impulsions de largeur variable pour moteurs à courant alternatif. L'élément de commutation de puissance est le nouveau transistor haute tension BUY89 et le transformateur est bobiné sur deux noyaux de ferrite EC44/17/18.

A.M. estéreo – una nueva dimensión en auto-radio

En U.S.A. se aguardan sugerencias para hallar una norma en la transmisión a.m. estéreo, en la cual la señal L−R y un tono piloto estéreo de 5 Hz estén en fase moduladas sobre la portadora de a.m. Una simple modificación, con pocos componentes adicionales, permite que el receptor descrito en EC&A, Vol. 3, Núm. 2, decodifique estas transmisiones con una separación de canales y proporcione la función de indicación estéreo usual.

Adelantos en condensadores electrolíticos de alumino sólido

Una nueva tecnología de grabado al aguafuerte ha disminuido a aproximadamente la mitad el tamaño de los condensadores electrolíticos de alumino sólido, sin que por ello sufrieran la fiabilidad, larga duración o resistencia a temperaturas extremas, por lo que son notorios. Los condensadores basados en la nueva tecnología compiten con la pequeñez de electrolíticos de tantalio, y probables adelantos futuros prometen extenderlos a la serie de productos de tensión capacitiva que hasta ahora ha sido del dominio exclusivo de electrolíticos húmedos.

Amplificadores lineales de banda ancha en comunicaciones de r.f.

Los transistores de potencia de r.f. que se emplean frecuentemente en las etapas lineales de potencia de emisores que entregan una potencia de hasta 1 kW. Actualmente existen dispositivos capaces de proporcionar hasta 200 W p.e.p., con una distorsión de intermodulación mejor que −30 dB. Este artículo resume los factores que gobiernan el diseño de amplificadores lineales de banda ancha basados en transistores de potencia de r.f. de potencia elevada, y como ejemplo describe un amplificador cascodo que utiliza solamente dos transistores de potencia BLW96, excitados mediante dos transistores BLW50F. El amplificador completo entrega hasta 400 W p.e.p., con una distorsión de intermodulación mejor que −26 dB.

Componentes electrónicos para equipos teléfonicos de abonado

Los circuitos integrados para petición de línea por lazo de corriente interrumpida o bién para tono dual multifrecuencia, permiten sustituir el marcador tradicional por un teclado. Si el micrófono de carbón es reemplazado por un electret o micrófono electrodinámico, las funciones de conversación y transmisión también pueden ser realizadas por un circuito integrado. Los timbres electrónicos controlados por un circuito integrado, ofrecen una elección de distintos tonos o ritmos y pueden reemplazar el timbre clásico. Los microcircuitos también permiten la incorporación de características tales como marcador automático, repertorio de marcado, visualización del número marcado, medida de la tarifa, uso con las manos libres, auriculares sin hilos y la posibilidad de tratamiento de datos.

Circuitos básicos de excitación del GTO

El desarrollo de técnicas electrónicas de conmutación de potencia han presentado un fuerte ímpetu desde la aparición del tiristor bloqueable por puerta (GTO). Este artículo describe cuatro circuitos de excitación del GTO que facilitarán el funcionamiento correcto del GTO en un amplio margen de aplicaciones. Se tratan aplicaciones aisladas y no aisladas, junto con las variaciones del margen requerido del ciclo de trabajo y frecuencia de conmutación.

Fuente de alimentación auxiliar para un sistema de control de velocidad de motor de c.a.

Este artículo, último de una serie, describe una fuente de alimentación conmutada auxiliar que proporciona las diversas salidas de c.c. de baja tensión requeridas para el funcionamiento de nuestro sistema de control de velocidad y para motores de c.c. El elemento de conmutación de potencia es el nuevo transistor de alta tensión BUY89, y el transformador está bobinado en dos núcleos de ferrita EC44/17/18.

Authors

Jim Ling joined Mullard in 1937 and worked initially on the development of receiving tubes. After service with the Royal Air Force from 1940 as a radio mechanic he returned to Mullard in 1946 to work on cathode ray tubes. Since 1950, as an application engineer, he has been concerned with various aspects of transmitter design and with r.f. heating. In recent years he has been engaged in the development of linear h.f. amplifiers.

Frank Burgum who gained his M.Sc. in solid-state physics at Chelsea College, University of London, in 1971, joined the instrumentation and control group of Mullard application laboratories in 1973. He has worked on switched-mode power supplies and speed control of industrial three-phase motors, with a special interest in computer-aided design, and is now a member of the Systems Application Centre for Power at Mitcham.

Leo van de Meeberg was born in Rotterdam in 1941 and joined Philips in 1959 as a loudspeaker designer. He left the company in 1969 to study electrical engineering at the University of Technology, Eindhoven. After graduating he joined the Central Application Laboratory of the Elcoma Division where he worked on integrated circuits for telephony and, since 1980, has been head of the telephony group.

Evert H. L. J. Dekker took his doctorate in solid-state chemistry at the University of Technology, Eindhoven, in 1975. He then joined Philips' Research Laboratories, where he worked on magnetic bubble circuit and memory design. Since 1979 he has been responsible for solid electrolytic capacitor development for the Electronic Components and Materials Division at Zwolle.

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

L. E. Jansson joined Mullard Research Laboratories, Redhill, in 1954 to work on applications of high-frequency transistors. In 1958 he transferred to Southampton to develop h.f. amplifiers and radio receivers. Since 1970 he has been at Mullard Application Laboratories, Mitcham, where, as member of the Systems Application Centre for Power, he is now mainly concerned with switched-mode power supply.

Helfried Schmickl studied mechanical and electrical engineering and took his doctorate at the University of Technology, Graz, Austria, in 1970. After working on capacitor development for Siemens G.m.b.H., Deutschlandberg, he joined Philips in 1973 and since 1977 has been in charge of electrolytic capacitor development, first at Klagenfurt and, since 1979, as development chief at Zwolle. He is a reader in television engineering at the University of Technology, Graz.

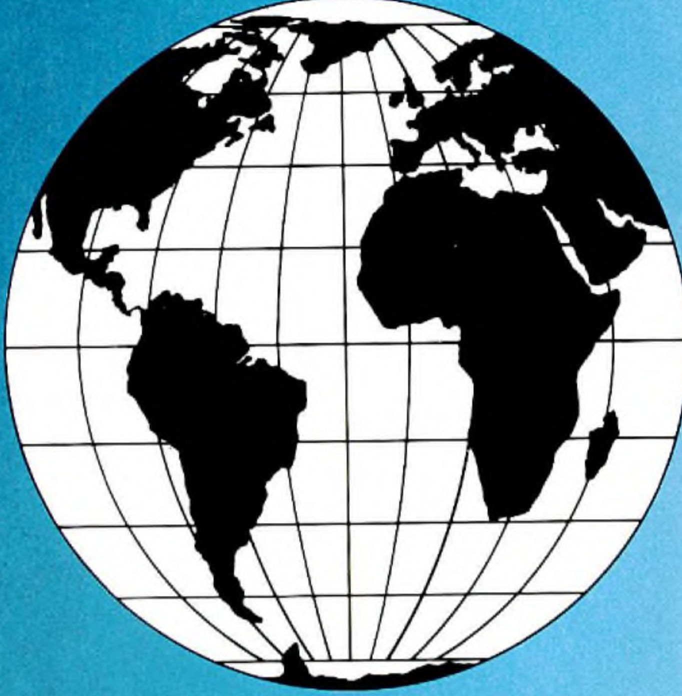
Winfried Jansen was born in Bonn, Germany, in 1947 and received his degree in electrical engineering from the University of Technology, Aachen, in 1975. Upon graduation, he joined Valvo in Hamburg as a designer in the consumer linear integrated circuit group. His primary work has been in f.m. receiver and interference absorption circuits, a field in which he has several patents pending.

Willy Kanow was born in Berlin in 1939 and graduated in communications technology at Gauss Academy of Engineering, Berlin, in 1964. From 1965 till 1979 he was with Loewe in Berlin, where he was head of the hi-fi radio laboratory. Since then he has been head of the radio group of Valvo Application Laboratory in Hamburg.

Electronic components and materials

for professional, industrial and consumer uses

from the world-wide Philips Group of Companies



Argentina: FAPESA, Av. Crovara 2550, Tablada, Prov. de BUENOS AIRES, Tel. 652-7438/7478.
Australia: PHILIPS INDUSTRIES HOLDINGS LTD., Elcoma Division, 67 Mars Road, LANE COVE, 2066, N.S.W., Tel. 427 08 88.
Austria: ÖSTERREICHISCHE PHILIPS BAUELEMENTE Industrie G.m.b.H., Triester Str. 64, A-1101 WIEN, Tel. 62 91 11.
Belgium: M.B.L.E., 7, rue du Pavillon, B-1030 BRUXELLES, Tel. (02) 242 7400.
Brazil: IBRAPE, Caixa Postal 7383, Av. Brigadeiro Faria Lima, 1735 SAO PAULO, SP, Tel. (011) 211-2600.
Canada: PHILIPS ELECTRONICS LTD., Electron Devices Div., 601 Milner Ave., SCARBOROUGH, Ontario, M1B 1M8, Tel. 292-5161.
Chile: PHILIPS CHILENA S.A., Av. Santa Maria 0760, SANTIAGO, Tel. 39-40 01.
Colombia: SADAPE S.A., P.O. Box 9805, Calle 13, No. 51 + 39, BOGOTA D.E. 1., Tel. 600 600.
Denmark: MINIWATT A/S, Emdrupvej 115A, DK-2400 KØBENHAVN NV., Tel. (01) 69 16 22.
Finland: OY PHILIPS AB, Elcoma Division, Kaivokatu 8, SF-00100 HELSINKI 10, Tel. 1 72 71.
France: R.T.C. LA RADIOTECHNIQUE-COMPELEC, 130 Avenue Ledru Rollin, F-75540 PARIS 11, Tel. 355-44-99.
Germany: VALVO, UB Bauelemente der Philips G.m.b.H., Valvo Haus, Burchardstrasse 19, D-2 HAMBURG 1, Tel. (040) 3296-1.
Greece: PHILIPS S.A. HELLENIQUE, Elcoma Division, 52, Av. Syngrou, ATHENS, Tel. 915 311.
Hong Kong: PHILIPS HONG KONG LTD., Elcoma Div., 15/F Philips Ind. Bldg., 24-28 Kung Yip St., KWAI CHUNG, Tel. 12-24 51 21.
India: PEICO ELECTRONICS & ELECTRICALS LTD., Ramon House, 169 Backbay Reclamation, BOMBAY 400020, Tel. 295144.
Indonesia: P.T. PHILIPS-RALIN ELECTRONICS, Elcoma Div., Panim Bank Building, 2nd Fl., Jl. Jend. Sudirman, P.O. Box 223, JAKARTA, Tel. 716 131.
Ireland: PHILIPS ELECTRICAL (IRELAND) LTD., Newstead, Clonskeagh, DUBLIN 14, Tel. 69 33 55.
Italy: PHILIPS S.p.A., Sezione Elcoma, Piazza IV Novembre 3, I-20124 MILANO, Tel. 2-6994.
Japan: NIHON PHILIPS CORP., Shuwa Shinagawa Bldg., 26-33 Takanawa 3-chome, Minato-ku, TOKYO (108), Tel. 448-5611.
(IC Products) SIGNETICS JAPAN, LTD, TOKYO, Tel. (03)230-1521.
Korea: PHILIPS ELECTRONICS (KOREA) LTD., Elcoma Div., Philips House, 260-199 Itaewon-dong, Yongsan-ku, C.P.O. Box 3680, SEOUL, Tel. 794-4202.
Malaysia: PHILIPS MALAYSIA SDN. BERHAD, Lot 2, Jalan 222, Section 14, Petaling Jaya, P.O.B. 2163, KUALA LUMPUR, Selangor, Tel. 77 44 11.
Mexico: ELECTRONICA S.A. de C.V., Varsovia No. 36, MEXICO 6, D.F., Tel. 533-11-80.
Netherlands: PHILIPS NEDERLAND B.V., Afd. Elonco, Boschdijk 525, 5600 PB EINDHOVEN, Tel. (040) 79 33 33.
New Zealand: PHILIPS ELECTRICAL IND. LTD., Elcoma Division, 2 Wagener Place, St. Lukes, AUCKLAND, Tel. 894-160.
Norway: NORSK A/S PHILIPS, Electronica, Sørkedalsveien 6, OSLO 3, Tel. 46 38 90.
Peru: CADESA, Rocca de Vergallo 247, LIMA 17, Tel. 62 85 99.
Philippines: PHILIPS INDUSTRIAL DEV. INC., 2246 Pasong Tamo, P.O. Box 911, Makati Comm. Centre, MAKATI-RIZAL 3116, Tel. 86-89-51 to 59.
Portugal: PHILIPS PORTUGESA S.A.R.L., Av. Eng. Duharte Pacheco 6, LISBOA 1, Tel. 68 31 21.
Singapore: PHILIPS PROJECT DEV. (Singapore) PTE LTD., Elcoma Div., Lorong 1, Toa Payoh, SINGAPORE 1231, Tel. 25 38 811.
South Africa: EDAC (Pty.) Ltd., 3rd Floor Rainer House, Upper Railway Rd. & Ove St., New Doornfontein, JOHANNESBURG 2001, Tel. 614-2362/9.
Spain: COPRESA S.A., Balmes 22, BARCELONA 7, Tel. 301 63 12.
Sweden: A.B. ELCOMA, Lidingövägen 50, S-115 84 STOCKHOLM 27, Tel. 08/67 97 80.
Switzerland: PHILIPS A.G., Elcoma Dept., Allmendstrasse 140-142, CH-8027 ZÜRICH, Tel. 01/43 22 11.
Taiwan: PHILIPS TAIWAN LTD., 3rd Fl., San Min Building, 57-1, Chung Shan N. Rd, Section 2, P.O. Box 22978, TAIPEI, Tel. (02)-5631717.
Thailand: PHILIPS ELECTRICAL CO. OF THAILAND LTD., 283 Silom Road, P.O. Box 961, BANGKOK, Tel. 233-6330-9.
Turkey: TÜRK PHILIPS TICARET A.S., EMET Department, Inonu Cad. No. 78-80, ISTANBUL, Tel. 43 59 10.
United Kingdom: MULLARD LTD., Mullard House, Torrington Place, LONDON WC1E 7HD, Tel. 01-580 6633.
United States: (Active devices & Materials) AMPEREX SALES CORP., Providence Pike, SLATERSVILLE, R.I. 02876, Tel. (401) 762-9000.
(Passive devices) MEPCO/ELECTRA INC., Columbia Rd., MORRISTOWN, N.J. 07960, Tel. (201) 539-2000.
(IC Products) SIGNETICS CORPORATION, 811 East Arques Avenue, SUNNYVALE, California 94086, Tel. (408) 739-7700.
Uruguay: LUZILECTRON S.A., Avda Rondeau 1576, piso 5, MONTEVIDEO, Tel. 91 43 21.
Venezuela: IND. VENEZOLANAS PHILIPS S.A., Elcoma Dept., A. Ppal de los Ruices, Edif. Centro Colgate, CARACAS, Tel. 36 05 11.
For all other countries apply to: PHILIPS INDUSTRIES, Electronic Components and Materials Division, Marketing Communications, Building BA, 5600 MD EINDHOVEN, THE NETHERLANDS, Telex 35000, Tel. (040) 723142

A20

Printed in The Netherlands

9398 702 70011