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FM receivers for mono and stereo on a single chip

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Most households now possess a number of radio receivers. This wide availability of a once-expensive technical product has come about only through a phenomenal fall in the cost of manufacture. This in turn is the result of greatly increased production runs — now into millions — but more than anything it is due to new technological developments. Every now and then a fundamental advance is made, a step towards other, and cheaper, methods of achieving the required functions in the receiver. One such earlier step was the change from thermionic valves to transistors. In the new technology, costs are gradually being brought down by greater efficiency and economies of scale in production, but in the long run this downward trend in production costs must stop and another fundamental step forward becomes necessary. The article below describes such a new step: the integration of a complete FM mono receiver on a single chip. The number of trimming points — an important cost factor — has been reduced to one: the externally connected oscillator inductor. The other external components are a few capacitors and resistors, which do not have to meet any particularly arduous requirements. The new chip (TDA 7010T) is so small that extreme miniaturization is possible — and some wrist-watches on the market now contain a built-in FM receiver based on this chip. But even in this new technology further improvements are on the way; the successor, TDA 7020T, already in pilot production, will work on lower battery voltages and requires fewer external components. The article also presents a fundamental solution of the next step forward: a completely integrated FM stereo receiver.

Introduction

Although the first attempts to use integrated circuits in radio receivers to reduce the cost were made more than 15 years ago, most receivers built today still consist of a printed-circuit board with manually inserted discrete transistors, diodes, resistors, capacitors and inductors; the inductors generally have to be aligned. Integrated circuits are now frequently included in high-performance receivers to make tuning easier, through additional features such as noise muting, deviation muting (off-tune signal suppression), station preset, frequency display, frequency synthesis, touch control and remote control. But the basic circuit used in most of the less-expensive portables and clock radios has remained essentially unchanged for ten years or more.

This gives some idea of the difficulties encountered in designing a successful substitute for inductors that will combine high Q-factors with high signal-to-noise ratios at high operating frequencies. In principle, gyrators [1] can be used instead of inductors. However, with increasing operating frequencies these circuits become less attractive because of the limitations of dynamic range and Q-factor and because of their complexity and power dissipation.

Instead of designing an integrated radio receiver in which the inductors in the resonant circuits are replaced by active RC circuits, it may be possible to adopt other approaches that are less handicapped by the limitations of IC technology. Various systems have been proposed in which fewer inductors are necessary

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[1] B. D. H. Tellegen, The gyrator, a new electric network element, Philips Res. Rep. 3, 81-101, 1948.

for the receiver unit [2]-[7]. Compared with existing superheterodyne receivers the proposed receiver circuits use a significantly lower intermediate frequency, so that the intermediate-frequency bandpass filters may be replaced by *RC* filters that can be partly or completely integrated.

Depending on the choice of intermediate frequency, two classes of problems are encountered in FM receivers. With an intermediate frequency of zero [4]-[6] image-reception problems in superheterodyne receivers are eliminated. With these receiver systems, however, the amplitude of the i.f. signal cannot be limited before demodulation. This results in a 3-dB lower signal-to-noise ratio in the i.f. stage and in less suppression of unwanted amplitude modulation of the demodulated signal.

This problem is not encountered in a receiver system that has an intermediate frequency of 140 kHz [7], where a limiter/amplifier is used before the demodulator to suppress AM noise and peak distortions of the r.f. signal. With this system, however, there is an image response at a spacing of 280 kHz, almost coinciding with the centre frequency of a neighbouring channel at a spacing of 300 kHz. Because of this there are also two tuning positions for each station (quite apart from the double response on the outer slopes of the demodulator characteristic, which is dealt with below).

To be acceptable to the customer an integrated FM receiver should at least avoid the shortcomings of the proposed systems with low intermediate frequency. But if the shortcomings of today's superheterodyne receivers could also be eliminated, such an integrated receiver, in addition to being cheaper to produce, would have advantages for the user as well.

Compared with AM receivers, FM receivers have one fundamental weakness, which is that each station has at least three tuning positions. *Fig. 1* shows the amplitude of the demodulated audio signal of a typical superheterodyne portable receiver as a function of the tuning frequency, with the r.f. voltage V_{ant} at the antenna input as the parameter; the modulation is a constant 1000-Hz tone. Besides the correct response at the centre of fig. 1, there are two spurious responses, characterized by a reduced signal-to-noise ratio and increased harmonic distortion of the audio signal. The demodulation takes place on the sides of the demodulator curve, which are referred to here as 'noise slopes', because of the high noise level (see *fig. 2*). The position and relative intensity of the spurious responses depend on the antenna input voltage and the selectivity of the receiver. They are separated from the range for correct tuning by minima in the output voltage. It might seem as if these minima would make it easy to identify the spurious responses. However, this is not the case in the far more complex situation found in practice, where the modulation is not constant and sinusoidal as in fig. 1, but strongly time-dependent and often with significant overlap between the spurious responses of adjacent channels. These effects combine to give rather unsatisfactory tuning behaviour, especially for simple tuning by ear.

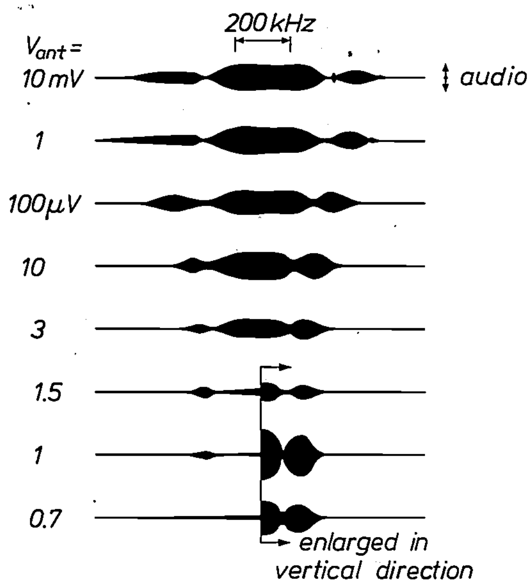

Fig. 1. The tuning behaviour of a conventional mono receiver at different antenna input voltages V_{ant}. The demodulated audio signal (vertical), here a 1000-Hz tone (*audio*), is shown as a function of the tuning frequency (horizontal). The audio signal has a maximum at the correct tuning position, and two secondary maxima to the left and right (on the outer slopes of the demodulator characteristic; see fig. 2). This makes it difficult to tune by ear. In the integrated FM receiver the two spurious responses are suppressed (see fig. 10).

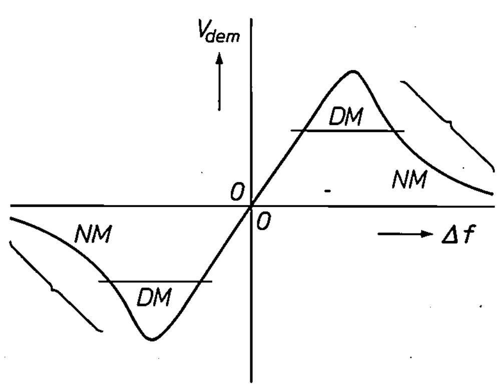

Fig. 2. Demodulator characteristic. This gives the relation between the demodulator output voltage V_{dem} and the frequency deviation Δf. The central part of the curve has the desired linear relation between the two. On detuning from the centre frequency, demodulation can also take place on the slopes *NS*; this is accompanied by distortion and noise. In our monolithic FM receiver these spurious responses are suppressed by a special circuit that reacts to frequency deviations (*DM*, 'deviation muting') or to excessive noise (*NM*, 'noise muting').

In the high-performance class of FM receivers squelch systems like noise muting [8] and deviation muting [9] are built in to suppress the spurious responses, and tuning meters are used to locate the correct tuning position. These features help to make tuning very much easier. Although cost reduction was the main objective in the integrated FM receiver presented here, it seemed a good idea to include these or similar features in the system. This had to be done, however, without unduly increasing the complexity, the power consumption or the number of external components and pins of the integrated circuit.

Fig. 3. A wrist-watch with a built-in FM receiver IC. The headphone lead also acts as the antenna.

Table I. Data for the integrated FM receivers TDA 7000/7010T and TDA 7020T.

	TDA 7000/ 7010T	TDA 7020T	
Supply voltage	4.5	3	V
Supply current	8	5.7	mA
Frequency range	1.5-110	0.5-110	MHz
Sensitivity (amplitude limiting −3 dB, input impedance 75 Ω, signal mute not operating)	1.5	1.5	μV
Max. input voltage (total harmonic distortion 10%, $\Delta f = \pm 75$ kHz, input impedance 75 Ω)	200	200	mV
Audio output voltage Minimum load	90 22000	100 35	mV Ω

The integrated mono receiver circuit described in this article is now being marketed under the type designation TDA 7000 (or TDA 7010T) [10]. An improved design has recently been brought out under the type designation TDA 7020T. Some data are given in *Table I*. *Fig. 3* shows a wrist-watch, now commercially available, which contains an FM mono receiver built around this integrated circuit.

A question that arises, of course, is whether it would also be possible to build an integrated FM stereo receiver. This is a much more complicated exercise in view of the large bandwidth (53 kHz) of the stereo multiplex signal compared with the bandwidth of the mono signal (15 kHz). Nevertheless, it does seem feasible, and a complete integrated circuit for a stereo receiver is now being studied. A block diagram of the design is presented at the end of this article.

[2] D. K. Weaver, Jr., A third method of generation and detection of single-sideband signals, Proc. IRE 44, 1703-1705, 1956.
[3] J. P. Costas, Synchronous communications, Proc. IRE 44, 1713-1718, 1956.
[4] J. G. Williford, U.S. patent No. 3 568 067.
[5] I. Vance, British patent No. 1 530 602.
[6] I. A. W. Vance, An integrated circuit VHF radio receiver, Proc. Int. Conf. on Land mobile radio, Bailrigg 1979, pp. 193-204.
[7] G. G. Gassmann, Ein neues Empfangsprinzip für FM-Empfänger mit integrierter Schaltung, Radio Mentor 32, 512-518, 1966 (in German).
[8] J. Craft, U.S. patent No. 3 714 583.
[9] I. Fukushima *et al.*, U.S. patent No. 3 851 263.
[10] W. H. A. van Dooremolen and M. Hufschmidt, A complete f.m. radio on a chip, Electron. Components & Appl. 5, 159-170, 1983.

Basic features of the FM mono receiver

The integrated FM mono receiver described here operates with an intermediate frequency of 70 kHz. This value would lead to unacceptable harmonic distortion at a frequency swing of ± 75 kHz, which is the maximum value allowed in FM broadcasting. This is evident, since there is no room in the intermediate-frequency band for a deviation of −75 kHz, which would extend into the range of 'negative frequencies'. The frequency scale is then 'folded' around 0 Hz (see fig. 13), resulting in unacceptable harmonic distortion. For this reason 'frequency feedback' is used in our system, to reduce the maximum frequency swing of the intermediate frequency to ±15 kHz. This frequency feedback is obtained by using the demodulated audio signal to control the frequency of the local oscillator in our superheterodyne receiver. This is done in such a way that the oscillator frequency 'travels some of the way' with the frequency deviation of the transmitter. If the transmitter has a frequency deviation of ± 75 kHz, then the oscillator is given a deviation of ± 60 kHz, resulting in an i.f. signal with a frequency deviation of only ± 15 kHz.

With our choice of intermediate frequency, image reception occurs at a spacing of 140 kHz, i.e. at the edge of the received channel. Background noise in this part of the frequency band is equal in strength to the background noise of the received signal. Thus, with a given input signal and a given bandwidth, the signal-to-noise ratio at the output of the mixer stage is in principle 3 dB lower than in systems with complete suppression of the image response. This entails an increase in the noise figure, but the increase is partly compensated by the reduced i.f. bandwidth [11], and partly by a lossless coupling between the r.f. amplifier and the mixer stage.

An advantage of using an intermediate frequency not equal to zero is that amplitude modulation in the received signal can be effectively suppressed. A high-gain limiter/amplifier is included before the demodulator. This gives good AM suppression and automatic volume control (AVC), even for weak input signals.

The circuit also includes 'deviation muting', a system that suppresses the audio signal if the tuning is incorrect or if the input signals are comparable with the input noise. It is based on the correlation between the i.f. signal and a delayed and inverted version of the i.f. signal. A qualitative description of the system is given in *fig. 4*. The signal *IF'* is derived from the amplitude-limited intermediate-frequency signal *IF* by delaying this signal by one half of its period at correct tuning, and inverting the delayed signal. This means that for correct tuning the two signals are identical (fig. 4*a*), giving high correlation. In this situation the demodulated audio signal is applied to the audio output. If the tuning is incorrect (fig. 4*b*), one half of the period of the signal *IF* no longer corresponds to the delay between the two signals *IF* and *IF'*. In this situation the correlation between the two signals is small or negative, and the demodulated audio signal is not applied to the output. In this way the spurious responses for large input signals are suppressed.

If the input signal is comparable in level to the input noise, the two signals *IF* and *IF'* are as shown in fig. 4*c*. Because of the low Q of the i.f. filter (about 0.7) the intervals between the successive zero crossings of the i.f. signal fluctuate considerably, again giving a low correlation between the two signals and hence muting of the demodulated audio signal. In this way the skirts of the spurious responses at large input signals (top of fig. 1) and the entire spurious responses for small input signals (bottom of fig. 1) are suppressed. Fig. 2 illustrates what is meant here by 'skirts'.

This muting system that comes into operation when correlation is low combines the characteristics of signal muting operated by noise (in which the noise on the envelope of the i.f. signal is detected [8]), and signal muting operated by incorrect tuning (which depends on the automatic frequency control [9]). The correlation system can be used in our receiver because the compressed frequency swing is never so large as to disturb the correlation seriously. In addition, advantage is taken of the delay network that is part of the demodulation circuit; this gives a delay of a quarter of a period. At the low intermediate frequency it is not necessary to use a tuned circuit either for this delay or for the additional quarter-period delay also required

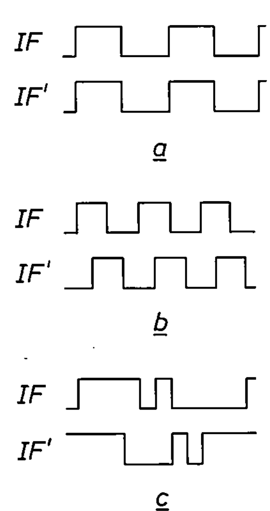

Fig. 4. Correlation measurement for signal muting. The i.f. signal is delayed by half a period, inverted (*IF'*) and compared with the original signal (*IF*). For correct tuning there is maximum correlation (*a*), for incorrect tuning less or no correlation (*b*), in the presence of noise alone there is no correlation (*c*). In cases *b* and *c* the signal is not supplied to the audio output.

for the muting, so that the entire signal-muting system requires few extra components and takes very little extra current. In our system the muting threshold is about five times as low as the muting thresholds of the most advanced systems based on detection of the noise on the envelope of the i.f. signal. It can therefore be used in portable receivers, where many of the input signals are only slightly above the input noise level.

A noise generator is used in combination with the muting system to give an audible tuning indication in the absence of a tuning meter.

The integrated FM mono receiver

Fig. 5 shows a block diagram of the integrated circuit with its three essential connections to the outside world: the antenna input, the audio output and the connection to the circuit used for tuning to the desired station.

The antenna is connected to the input of the broadband amplifier A_1, which amplifies throughout the entire FM band. The output of A_1 is connected to the r.f. mixer M_1, which performs the conversion to the intermediate frequency. The signal gain from antenna input to mixer output is about 26 dB.

The i.f. signal is filtered by a fourth-order active lowpass filter (LP_1) to suppress signals outside the selected channel [12]. The response of this filter is shown in *fig. 6*. The limiter/amplifier LA_1, with a gain of more than 90 dB, limits the i.f. signal, which has a high dynamic range at the output of the mixer, to a constant amplitude. To obtain the high gain of 90 dB at low current, LA_1 operates into a high impedance, and a buffer amplifier A_2 is added; this has a sufficiently low output impedance to reduce unwanted crosstalk between multipliers M_2 and M_3.

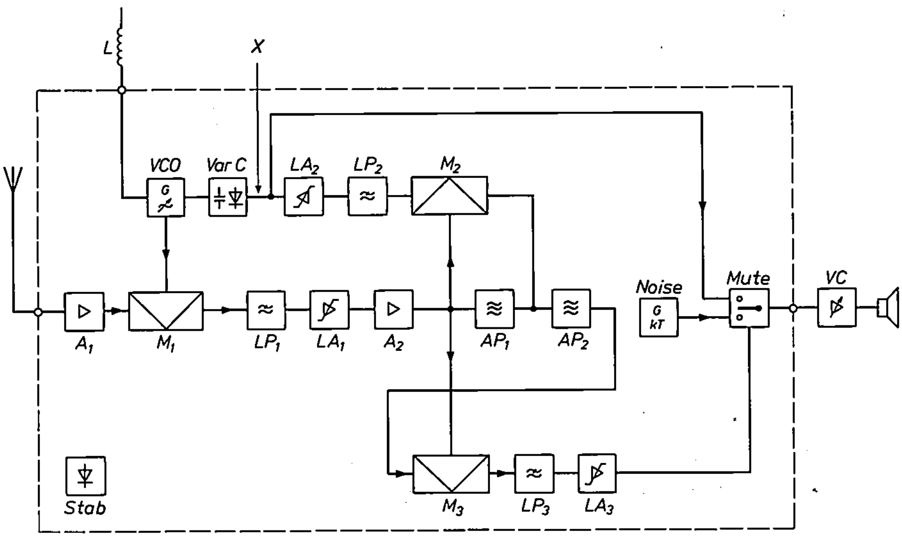

Fig. 5. Block diagram of integrated FM mono receiver. $A_{1,2}$ amplifier. $M_{1,2,3}$ multiplier circuit (mixer). $LP_{1,2,3}$ lowpass filter. $LA_{1,2,3}$ limiter/amplifier. $AP_{1,2}$ allpass network (gives 90° phase shift). *VarC* varactor. *VCO* voltage-controlled oscillator. *Noise* noise generator. *Mute* signal/noise switch. *Stab* stabilized direct voltage supply. *L* inductor (tunes *VCO* to the FM band). *X* point where the frequency-locked loop is regarded as open in the treatment of open-loop behaviour. *VC* volume control.

Demodulation and limitation of the frequency swing are achieved by converting the frequency into a voltage, which is then used to correct the voltage-controlled oscillator *VCO*. Associated with this oscillator are an integrated varactor diode *VarC* (a variable-capacitance diode) and the external resonant circuit

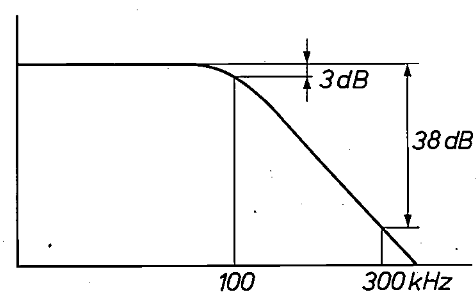

Fig. 6. Response of the lowpass filter LP_1. This fourth-order filter passes the band from 0 Hz to 100 kHz and acts as i.f. band filter in the integrated FM mono receiver, in which it determines the selectivity. A channel at a spacing of 300 kHz is attenuated by 38 dB.

[11] L. H. Enloe, Decreasing the threshold in FM by frequency feedback, Proc. IRE **50**, 18-30, 1962.

[12] A Sallen and Key configuration is used for this filter because it gives the best compromise between selectivity, current taken and signal-to-noise ratio. See R. P. Sallen and E. L. Key, IRE Trans. **CT-2**, 74-85, 1955.

mentioned above. The frequency is converted into a voltage by means of the multiplier M_2, in which the i.f. signal is multiplied by a version of the same signal shifted in phase by 90°. The phase shift takes place in the allpass filter AP_1. The combination of M_2 and AP_1 is called an FM quadrature detector (FMQD).

A mathematical description of quadrature demodulation is given on page 179, with a calculation of the signal delay in the quadrature demodulator.

The lowpass filter LP_2 is included in the frequency-locked loop thus formed. In addition to the audio frequency signal mentioned above, a d.c. signal also appears in the loop when the i.f. frequency does not have a value such that the delay in AP_1 is exactly a quarter-period of the carrier. This direct voltage provides automatic frequency control (AFC) by acting on the voltage-controlled oscillator *VCO*. The loop also contains a low-gain limiter/amplifier LA_2 to control the locking range of the AFC. Audio signals are extracted after this limiter/amplifier. The limiter does not distort the audio signals because as soon as it starts to operate the AFC is switched off; the receiver is then no longer tuned to the transmitter and the signal is suppressed.

The system that suppresses the signal in the absence of correlation (the correlation muting system) consists of two identical allpass filters AP_1 and AP_2, mixer M_3, lowpass filter LP_3 and limiter/amplifier LA_3. The correlation between the limited i.f. signal at the input of AP_1 and its delayed and inverted version at the output of AP_2 takes place in mixer M_3. The amplified and limited output signal of the correlator is used as a muting signal to suppress the audio signal for spurious responses.

Fig. 7 shows some of the control voltages in the integrated FM receiver as they would appear if the frequency-locked loop were opened at X in fig. 5; the responses shown are purely qualitative and relate to a particular level of the antenna signal. In fig. 7*a* and *b* the output voltages of demodulator M_2 and low-gain limiter/amplifier LA_2 are shown as a function of the difference between the antenna-signal frequency $f_{\rm ant}$ and the oscillator-signal frequency $f_{\rm osc}$. With the loop closed the control makes this difference equal to the centre frequency $f_{\rm c}$, giving correct tuning.

The output voltages of the correlator M_3 and of limiter/amplifier LA_3 are shown in fig. 7*c* and *d* as a function of $f_{\rm ant} - f_{\rm osc}$. There are two bands of $f_{\rm ant} - f_{\rm osc}$ where the signal is not suppressed but is applied to the output. One is centred around $f_{\rm c}$, the point of correct tuning. The other band with no muting is defined by $-f_2 < f_{\rm ant} - f_{\rm osc} < -f_1$. This means that a

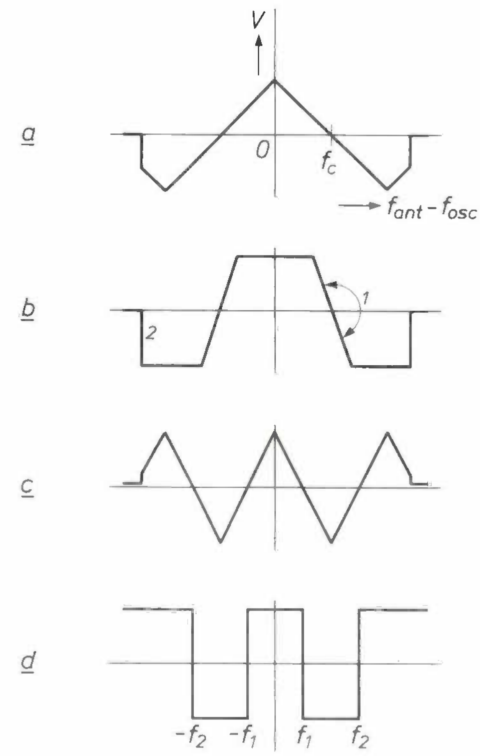

Fig. 7. Signal voltages V for frequency control and signal suppression (muting). Schematic representation of open-loop behaviour. *a*) Output voltage of mixer M_2 as a function of the difference $f_{\rm ant} - f_{\rm osc}$ between received frequency and oscillator frequency. $f_{\rm c}$ is the desired intermediate frequency to which $f_{\rm ant} - f_{\rm osc}$ is tuned in the closed loop. *b*) The same voltage after limiting and amplification in LA_2. Stable tuning positions are found on paths *1* and *2*, because if the value of $f_{\rm ant} - f_{\rm osc}$ is too high the control voltage is negative, so that $f_{\rm ant} - f_{\rm osc}$ is reduced again; the tuning position on path *2* (the slope) is one of the spurious responses. *c*) Output voltage of mixer M_3, used for deviation muting. *d*) The same voltage after limiting and amplification in LA_3. When the voltage is high, the output signal is suppressed (muted), thus eliminating the spurious respons on path *2*. The voltage is low in the desired frequency range between f_1 and f_2, and also at the image frequencies between $-f_2$ and $-f_1$; however, stable tuning is impossible at the image frequencies, because of the frequency control.

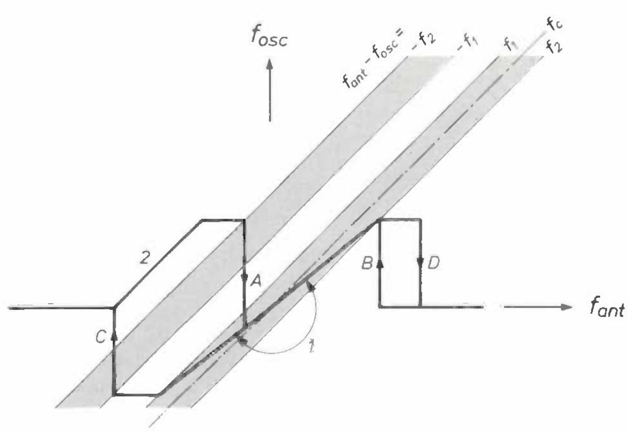

Fig. 8. Frequency control and signal muting with closed loop. (See fig. 7.) $f_{\rm c}$ desired intermediate frequency. *1* frequency trajectory at the correct tuning position; the frequency point moves to and fro along this path as a result of the frequency modulation. Because of the frequency feedback this line is not horizontal but sloping, and $f_{\rm osc}$ also varies. *Shaded area:* intermediate frequency bands where the audio signal is supplied to the receiver output; everywhere else the signal muting is operative. *2* second stable tuning position, but the signal is not supplied to the audio output. *A*, *B* frequency locking. *C*, *D* frequency release.

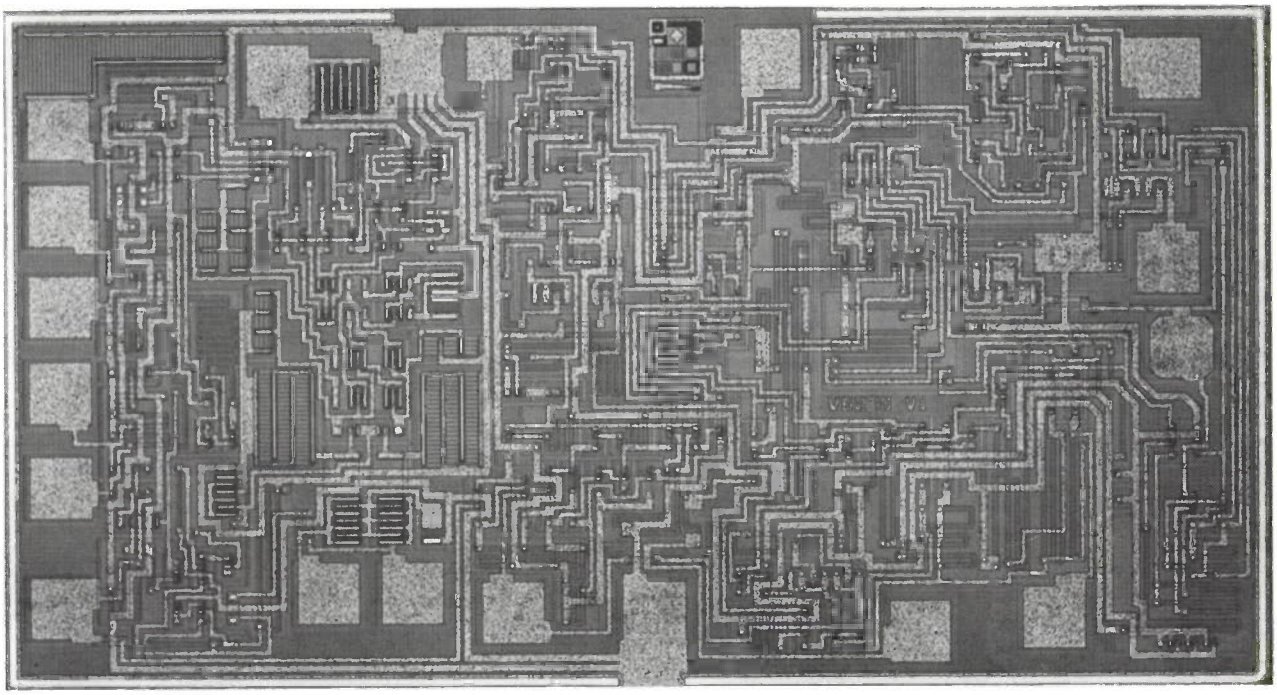

Fig. 9. Photomicrograph of the integrated FM mono receiver circuit TDA7000/7010T. Chip area 3.5 mm².

double tuning response is possible with an open loop. The second (spurious) response is at the image frequency.

When it is closed the frequency-locked loop gives positive feedback at the image frequency; a deviation from the image frequency generates a control voltage whose polarity is such that the deviation increases. Double tuning is therefore impossible with the loop closed.

This is illustrated in detail in *fig. 8*, which shows the paths taken by the control point as it moves towards the point of correct tuning in the plane f_{ant}, f_{osc}. The shaded areas indicate that there is no muting. The figure relates to the same antenna signal as in fig. 7.

With correct tuning (path *1*) the demodulated audio signal — which is a linear function of f_{ant} and also of f_{osc} — is not muted but is applied to the audio output (see fig. 7*d*). On path *2*, the noise slope of the demodulator characteristic, the frequency feedback is negative, so that a stable tuning position is also possible here. This path, however, lies in a region where signal muting is operative.

If the deviation from the point of correct tuning is too great the frequency-locked loop loses its hold on the oscillator frequency (fig. 8 *C*, *D*). As the point of correct tuning is approached, the transmitter eventually becomes 'locked' in the loop (fig. 8 *A*, *B*). Both events are accompanied by a sudden jump in the loop voltage, which would be audible in the audio signal if no countermeasures were taken. These transients in the audio signal are suppressed in two different ways.

The lock operation *B* and the loss of lock *D* take place in a region where $f_{ant} - f_{osc}$ is greater than f_2. In this region the signal muting is permanently in operation. The situation is different with loss of lock indicated by *C*; here the signal muting is not operative during the complete transition. The transition starts in a region where the mute is operative, then passes the region $-f_2 < f_{ant} - f_{osc} < -f_1$, where the signal is not muted and ends up in a region where the signal is muted again. To make this transition inaudible the time constant of the lowpass filter LP_3 (see fig. 5) has been made such that the input voltage of LA_3 remains positive during the short time interval necessary for crossing the region $-f_2 < f_{ant} - f_{osc} < -f_1$.

In a similar way the time constant of LP_3 also determines what can be heard during the locking operation *A*, where a transition takes place from the noise generator to the demodulated audio signal. Here the control point passes through four different regions with alternating states of the muting signal. A proper choice of the time constant ensures a smooth acoustical transition from the region $f_{ant} - f_{osc} < -f_2$, where the signal from the noise generator is passed to the output, to the region $f_{ant} - f_{osc} > f_1$ where the desired transmitter is heard at the output.

Measurements on the FM mono receiver

Fig. 9 shows a photograph of the integrated FM mono receiver circuit on a single chip. All the functions indicated in the block diagram in fig. 5 have been optimized for chip area, power consumption and supply voltage variations. The current taken by the IC is 8 mA at 4.5 V, but it operates at any supply voltage between 3 V and 18 V. The chip area is 4.5 mm^2 and the number of bonding pads is 18.

Measurements have been made on an experimental receiver containing this IC. Apart from the integrated circuit, the receiver consists of a resonant circuit used for tuning to the desired station, a number of ceramic capacitors and a resistor. The resonant circuit is tuned by a varactor. No trimming is required apart from the adjustment of the direct-voltage range for tuning the resonant circuit through the FM band. Because of the low intermediate frequency i.f. trimming is unnecessary: the tolerance on the values of the most critical of the fixed capacitors is about 100 times greater than the tolerance on the LC products of the resonant circuits used in a conventional receiver.

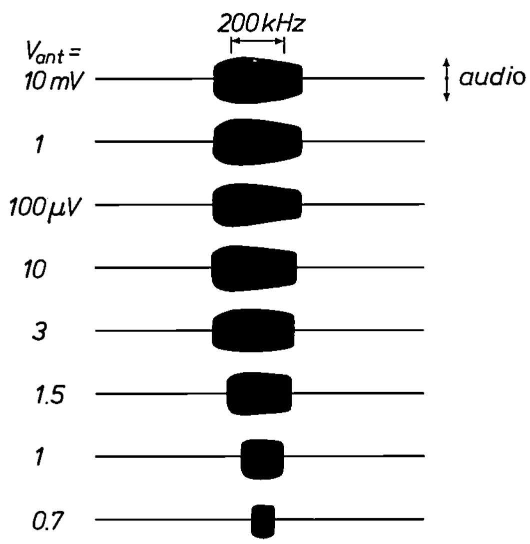

Fig. 10. Tuning behaviour of the integrated FM mono receiver (see also fig. 1).

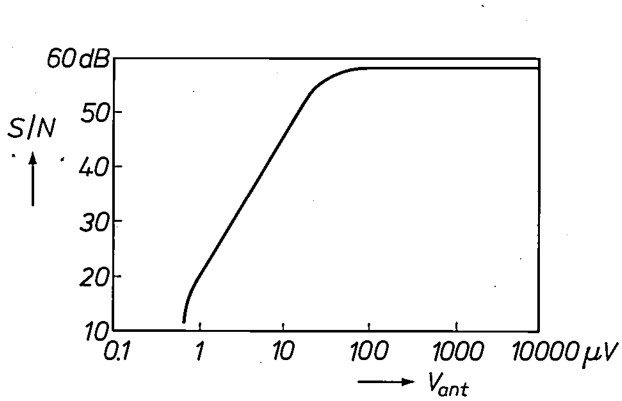

Fig. 11. Signal-to-noise ratio S/N of the demodulated signal (frequency 1 kHz, carrier frequency swing ± 22.5 kHz) as a function of the antenna voltage V_{ant} (across an input impedance of 40 Ω). At 1.5 μV the signal-to-noise ratio is 26 dB (this is the 'quieting sensitivity' as defined in the CCIR standard).

Fig. 10 shows the tuning behaviour of the experimental receiver. There are three improvements as compared with the tuning of a typical portable receiver, as shown in fig. 1:

- No spurious responses, at large or small input voltages. This is the result of the correlation muting system.
- Large range of correct tuning, even at small input voltages. This is achieved with the AFC. This function is illustrated by fig. 8, where a large variation in f_{ant} — which is equivalent to a large variation in the tuning frequency — is reduced to a small variation in the intermediate frequency $f_{ant} - f_{osc}$.
- No degradation of the audio signal for small antenna signals. This is due to the high gain of the limiter/amplifier LA_1 before the demodulator, which limits the i.f. signal even at input voltages lower than 1 μV. In conventional portable receivers, limiting at such low voltages results in stability problems because higher harmonics of the limited i.f. signal are radiated to the antenna. With the low intermediate frequency used in the design discussed here, it was possible to eliminate this radiation problem by careful layout of the integrated circuit.

In *fig. 11* the signal-to-noise ratio of the demodulated audio signal is shown as a function of the antenna input voltage; the frequency swing of the received signal is ± 22.5 kHz and the modulation frequency is 1 kHz. At 1.5 μV across an input impedance of 40 Ω the signal-to-noise ratio is 26 dB ('quieting sensitivity' as defined in the CCIR standard). The muting threshold is 0.7 μV, which practically coincides with the threshold of FM reception; below this level the signal-to-noise ratio decreases rapidly with decreasing input voltage. This means that with the correlation-muting system only audio signals of unacceptable reception quality are suppressed.

The fourth-order active RC filter LP_1 (figs. 5 and 6) has internal resistors and external capacitors. It provides 38 dB of attenuation for a neighbouring channel at a spacing of 300 kHz. The total harmonic distortion of the demodulated audio signal is defined by the characteristic curve of the integrated varactor diode. It has been measured as 1.8% at an input voltage of 1 mV and a frequency swing of ± 75 kHz. A better figure can be obtained by increasing the area of the integrated varactor.

Design of an integrated FM stereo receiver

Stability of frequency-locked loop

The bandwidth of the FM stereo signal that modulates the carrier is much larger than that of the FM mono signal (see *fig. 12*). As laid down in the international standards, the stereo signal, often called the stereo multiplex signal, consists of a baseband signal equal to the sum of the signals in the left and right channels (about 0-15 kHz), a pilot tone (19 kHz) and a subcarrier (38 kHz), modulated in amplitude by the difference between the left and right signals. This AM subchannel covers a band of 23 kHz to 53 kHz; the subcarrier is suppressed at the transmitter end and is reconstructed in the receiver from the pilot tone by frequency doubling [13].

The large bandwidth of the FM stereo multiplex signal (0-53 kHz) would lead to difficulties in a receiver like the one shown in fig. 5. The problem is connected with the stability of the frequency-locked loop, which

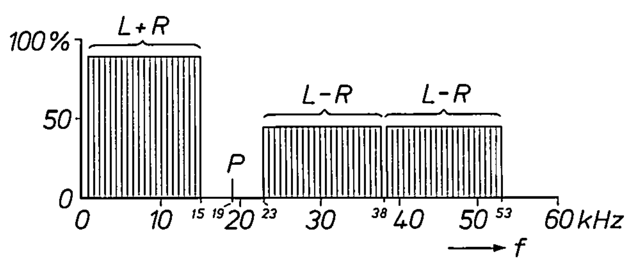

Fig. 12. Frequency spectrum of the stereo multiplex signal. The figure shows: from about 0 Hz to 15 kHz the stereo sum signal $L+R$, at 19 kHz the pilot tone P, from 23 kHz to 53 kHz the stereo difference signal $L-R$, amplitude-modulating a carrier at 38 kHz; this carrier is not transmitted. The vertical scale gives the contributions of each component to the frequency swing.

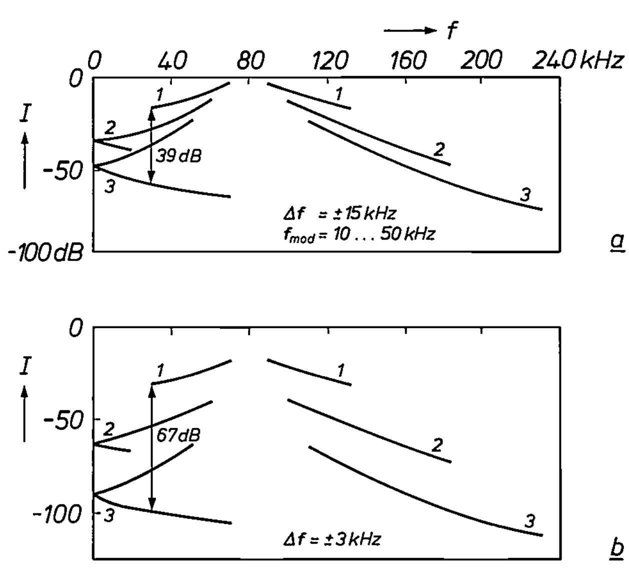

Fig. 13. Sideband 'folding'. The sideband level I for an FM signal is plotted against the frequency f (after conversion to an intermediate-frequency band centred on 80 kHz). The reference level of 0 dB corresponds to the unmodulated carrier. The diagram shows the sidebands of first, second and third order (*1, 2, 3*) for modulation by a sinusoidal tone at a frequency f_{mod} of 10 kHz to 50 kHz. The lower sidebands of second and third order extend further than the i.f. band permits; they are 'folded' around 0 Hz, causing harmonic distortion on demodulation. *a*) Weakly compressed or uncompressed frequency swing $\Delta f = \pm 15$ kHz. *b*) Highly compressed frequency swing $\Delta f = \pm 3$ kHz. The sidebands *2, 3* of second and third order are much weaker and the harmonic distortion is therefore less.

is responsible for compression of the frequency swing. For reception of the stereo multiplex signal, frequency-swing compression is even more necessary than for the mono signal, because of the larger bandwidth. The improvement resulting from frequency-swing compression is illustrated in *fig. 13*, where the level of the sidebands of an FM signal with a swing $\Delta f = \pm 15$ kHz (little or no compression) is compared with the level at $\Delta f = \pm 3$ kHz (high compression). A frequency swing of $\pm$ 15 kHz is the contribution from the encoded stereo difference signal if one of the channels is fully modulated and the other is quiescent. Because of the low intermediate frequency (80 kHz) assumed in fig. 13, the lower sidebands are 'folded' around the limit of 0 Hz, giving rise to harmonic distortion. Since the sidebands of higher order (*2, 3*) are disproportionately weaker when the swing is smaller, the resulting harmonic distortion is less.

The frequency-locked loop is a negative feedback system, which is only stable when the phase shift is less than 180° at the frequencies where the loop gain is greater than or equal to one. If the phase shift in this frequency range is 180°, the loop will oscillate spontaneously.

Now a phase shift is unavoidable in the filters used in this control loop: these are the i.f. filter (LP_1 in fig. 5), which determines the selectivity, the FM detector (AP_1 and M_2) and the loop filter (LP_2) that limits the bandwidth of the loop. To obtain a quantitative idea of the way in which the selectivity, the bandwidth of the closed frequency-locked loop and the transfer characteristic are affected by the stability condition, it is useful to have a simplified model of the frequency control system in the FM receiver. Let us first take another look at the mono receiver as shown in fig. 5.

Frequency control of mono receiver

Fig. 14 shows the model, including the transmitter. This is represented by the voltage-controlled oscillator VCO_1, to which the mono information V_i is supplied (there is no multiplex signal). VCO_1 delivers a frequency-modulated carrier. In the analysis that follows, the frequency of the unmodulated carrier is not relevant; we are more concerned with the frequency-deviation f_i. We shall therefore consider this alone in the following and merely state that VCO_1 delivers a signal f_i. This frequency deviation f_i is proportional to V_i.

All non-essential subfunctions in the receiver, such as the amplifiers and the signal muting, have been omitted. The oscillator VCO_2 is controlled by the

[13] N. van Hurck, F. L. H. M. Stumpers and M. Weeda, Stereophonic radio broadcasting, I. Systems and circuits, Philips Tech. Rev. 26, 327-339, 1965.

voltage V_o, which is produced by the control system, which gives it a frequency deviation f_o. The mixer M_1 produces the difference signal $f_i - f_o$; this passes through the i.f. filter LP_1, which determines the selectivity, and goes to the frequency demodulator AP_1/M_2, whose output signal $(V_i - V_o)_{del}$ is delayed. This signal passes through the loop filter LP_2 (bandwidth B) and its output signal is the control voltage V_o mentioned above.

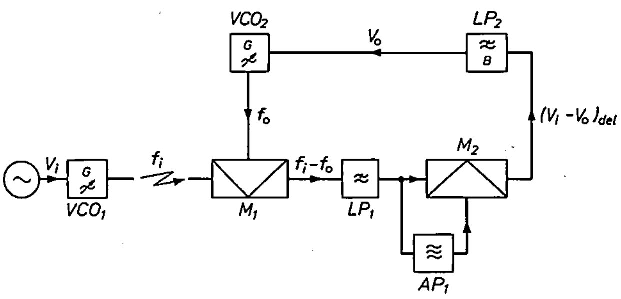

Fig. 14. Model of the frequency-locking system in the FM mono receiver. V_i mono signal. VCO_1 transmitter, frequency-modulated by V_i. f_i frequency deviation at receiver input. M_1 mixer. LP_1 i.f. filter. M_2 mixer for demodulation. AP_1 allpass filter (delay network) giving a 90° phase shift for the intermediate frequency. LP_2 loop filter. V_o control voltage. VCO_2 voltage-controlled oscillator. f_o deviation of the oscillator frequency. $(V_i - V_o)_{del}$ signal resulting from demodulation of $f_i - f_o$ and delayed by LP_1 and AP_1.

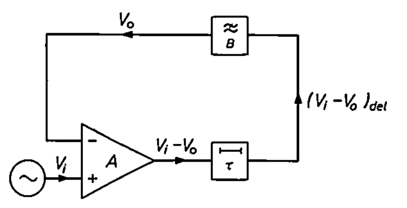

Fig. 15. Simplified version of fig. 14. The conversions from voltage to frequency deviation and frequency deviation to voltage have been omitted. The gain A is the product of the conversion factors of VCO_2 and AP_1/M_2.

The delay of $(V_i - V_o)_{del}$ is partly caused by the group delay of LP_1 and partly by the delay of the i.f. signal in AP_1. Both delays are added together to form a single delay τ.

When the voltage-frequency conversion in VCO_1 and VCO_2 is combined with the frequency-voltage conversion in demodulator AP_1/M_2, the result is a particularly simple model of an FM transmitter and FM receiver with a frequency-locked loop (*fig. 15*). The transmitter now consists only of the voltage source V_i, and the receiver consists of a differential amplifier of gain A, a delay element with delay τ and a loop filter of bandwidth B. The gain A is the product of the conversion gains of VCO_2 and AP_1/M_2.

The transfer function from transmitter to receiver is determined by these three quantities A, τ and B. The swing compression is $1/(1 + A)$. The part of the delay τ arising in LP_1 is connected with the selectivity, and the part arising in AP_1 is connected with the intermediate frequency. Finally, B is a quantity that determines the stability of the frequency-locked loop.

The contribution of the filter LP_1 to the delay τ can be determined quantitatively with the aid of *fig. 16*. The curves *Bu* represent the amplitude characteristic and the group delay of the filter as designed for the integrated FM mono receiver; it is a fourth-order Butterworth filter with a cut-off frequency of 100 kHz. It can be seen in fig. 16*b* that the group delay in the

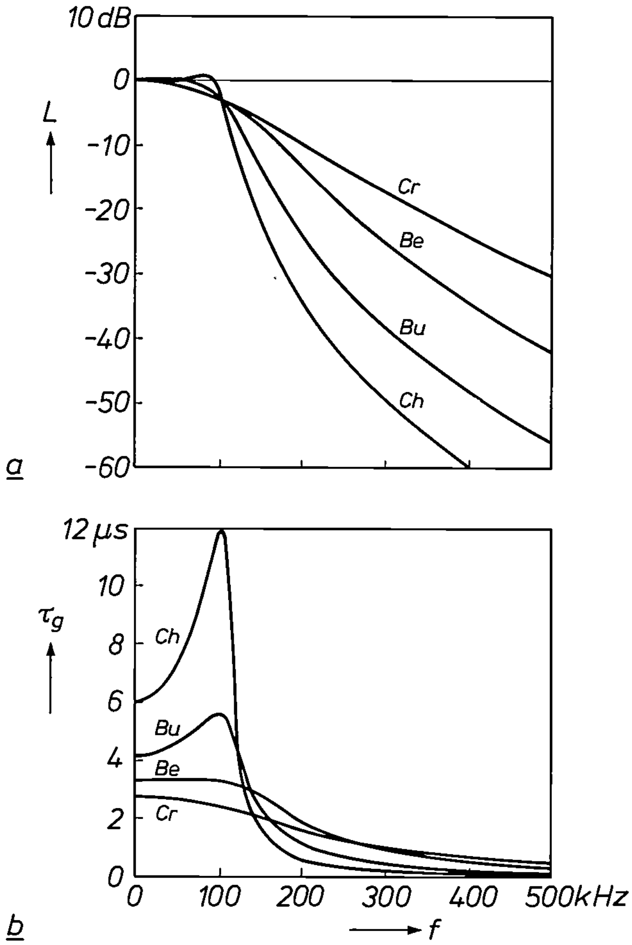

Fig. 16. Amplitude level L and group delay τ_g of the output signal as a function of frequency f for a fourth-order lowpass filter with a cut-off frequency of 100 kHz. *Cr* critically damped filter. *Be* Bessel filter. *Bu* Butterworth filter. *Ch* Chebyshev filter. The selectivity of the filters increases in this order, but at the same time the group delay in the passband increases.

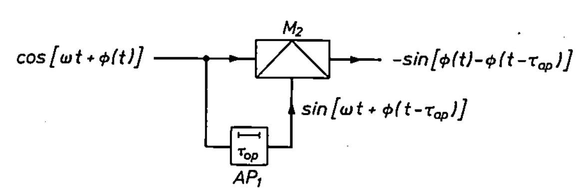

Fig. 17. Signal delay in the demodulator circuit AP_1/M_2. The delay τ_{ap} in the allpass filter AP_1 is a quarter of a period at the intermediate frequency. Because of the symmetry of the expression for the output signal a delay of $\frac{1}{2}\tau_{ap}$ can be assigned to it.

passband is reasonably constant and that LP_1 may be replaced by a frequency-independent delay element.

A further delay occurs in the frequency demodulator, which consists of the allpass filter AP_1 and the multiplier M_2 (*fig. 17*). The output signal of M_2 is the product of the i.f. signal and a delayed version of it obtained from AP_1; the delay in AP_1 corresponds to a phase shift of 90° at the nominal intermediate frequency. The symmetry of the contributions from the two input signals to the output signal suggests that the delay of the output signal is the average of the delays of the two input signals, i.e. half the group delay of AP_1.

This may be substantiated by a mathematical description of quadrature modulation. The product of the two input signals to M_2 (see fig. 17), taking τ_{ap} as the delay in AP_1, is:

$$\cos\{\omega t + \phi(t)\}\sin\{\omega t + \phi(t-\tau_{ap})\}$$
$$= \tfrac{1}{2}\sin\{2\omega t + \phi(t) + \phi(t-\tau_{ap})\}$$
$$- \tfrac{1}{2}\sin\{\phi(t) - \phi(t-\tau_{ap})\}.$$

The first term is at twice the intermediate frequency and is suppressed by the loop filter. The second term represents the low-frequency output signal; the variation of the phase $\phi(t)$ with time contains the audio information, since $d\phi/dt$ is the instantaneous value of the frequency deviation and hence of the audio signal. We can expand $\phi(t)$ and $\phi(t-\tau_{ap})$ about the time $t-\frac{1}{2}\tau_{ap}$ in a Taylor series and then subtract:

$$\phi(t) = \phi + \frac{\tau_{ap}}{2}\frac{d\phi}{dt} + \frac{\tau_{ap}^2}{8}\frac{d^2\phi}{dt^2} + \frac{\tau_{ap}^3}{48}\frac{d^3\phi}{dt^3} + \ldots$$
$$\phi(t-\tau_{ap}) = \phi - \frac{\tau_{ap}}{2}\frac{d\phi}{dt} + \frac{\tau_{ap}^2}{8}\frac{d^2\phi}{dt^2} - \frac{\tau_{ap}^3}{48}\frac{d^3\phi}{dt^3} + \ldots$$
$$\phi(t) - \phi(t-\tau_{ap}) = \tau_{ap}\frac{d\phi}{dt} + \frac{\tau_{ap}^3}{24}\frac{d^3\phi}{dt^3} + \ldots$$

All the terms on the right-hand side have the argument $t-\frac{1}{2}\tau_{ap}$. The terms of third and higher order are small, so that, to a good approximation,

$$\phi(t) - \phi(t-\tau_{ap}) \approx \tau_{ap}\frac{d\phi(t-\frac{1}{2}\tau_{ap})}{dt}.$$

The output signal thus carries the information of the time $t-\frac{1}{2}\tau_{ap}$, so that the delay in the demodulator circuit amounts to half the delay in the allpass filter AP_1.

Now that we know the total delay τ in the frequency-locked loop in fig. 15, we can calculate the effect of the bandwidth B of the loop filter on the transfer function V_o/V_i. *Fig. 18* gives three examples for bandwidths of 2 kHz, 3 kHz and 5 kHz. As might be expected, the −3-dB bandwidth of the closed loop increases and at the same time the stability decreases with increasing bandwidth B of the loop filter. Although a −3-dB bandwidth of 53 kHz is obtained at B = 5 kHz, there is a peak of 8 dB because of the reduced stability, and this is unacceptable for stereo reception. The bandwidth of 53 kHz is necessary, however, if the entire stereo multiplex signal is to contribute to the frequency-swing compression. The peaking can of course be reduced by making τ smaller, but this necessitates a less selective i.f. filter (see fig. 16, curves *Be* or *Cr*); in many situations there would then be unacceptable interference from strong adjacent stations.

There is therefore a dilemma between undistorted stereo reception on the one hand and sufficient selectivity on the other. We have succeeded in finding a solution, which will be explained here with the aid of a simplified model of a stereo transmitter and receiver.

Frequency control of a stereo receiver

The essence of our solution to the stability problem caused by the large bandwidth of the stereo multiplex signal is that the stereo difference signal that modulates the suppressed 38-kHz subcarrier does not pass through the loop filter. Instead it is first demodulated and passed through a lowpass loop filter of its own. The output is then remodulated, again producing an AM signal with suppressed carrier at 38 kHz. This in turn is added to the stereo sum signal, to provide oscillator control.

A simplified model of the system of frequency control in the combination of stereo transmitter and stereo receiver is illustrated in *fig. 19*. Here the signal from the stereo transmitter consists only of a subcarrier at 38 kHz, modulated in amplitude by V_i, the difference between the right and the left channels. The stereo receiver consists for the most part of the same components as the greatly simplified mono receiver in fig. 14: VCO_2, M_1, LP_1 and AP_1/M_2. Here, however, the loop filter LP_2 is replaced by mixer M_3, lowpass filter LP_3 and mixer M_4. The large-signal input to mixer M_4 is a 38-kHz subcarrier in phase with the subcarrier of the transmitter. The large-signal input to mixer M_3 is a delayed version of the subcarrier; the delay τ is equal to the delay in LP_1 and AP_1/M_2.

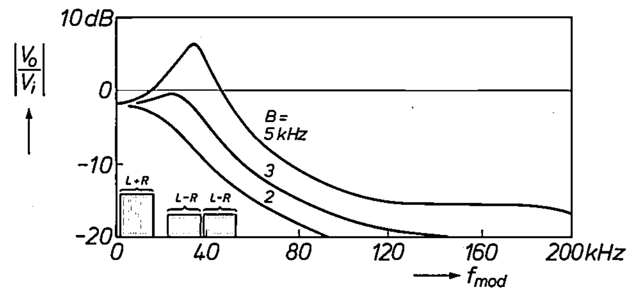

Fig. 18. Transfer function V_o/V_i in the closed frequency-locked loop; f_{mod} is the modulation frequency. The bandwidth B of the loop filter must be 5 kHz for all modulation frequencies up to 53 kHz of the stereo multiplex signal in fig. 12 to be passed, so that they can contribute to the swing compression. However, there is then an undesired peak of 8 dB, a sign of reduced stability.

The operation of the model is as follows. In M_0 the difference signal V_i modulates the subcarrier to produce V_i'; this then modulates VCO_1 to give the frequency deviation f_i. In a similar way the subcarrier is modulated in M_4 by the difference signal V_o obtained from the demodulation, resulting in a signal V_o', which agreement between model and experiment up to modulation frequencies of about 30 kHz with an open loop, and to about 80 kHz with the loop closed. The highest frequency in the stereo difference signal is 15 kHz, which can be processed without difficulty by a receiver of this type.

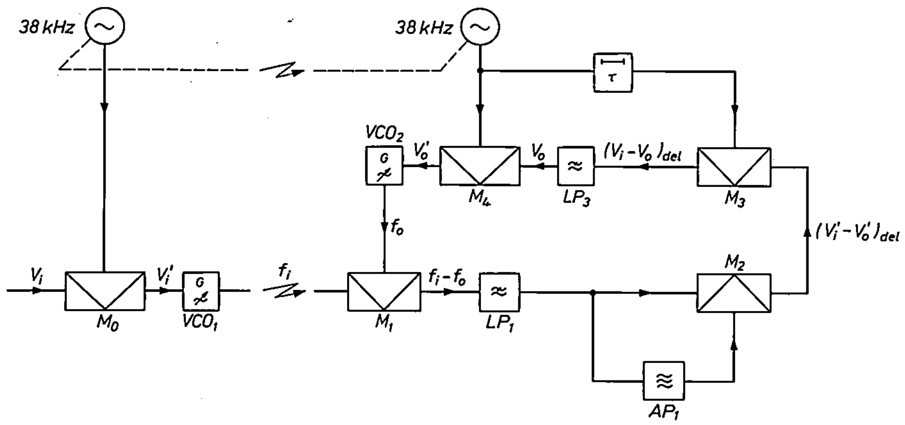

Fig. 19. Model of the frequency-control system in the FM stereo receiver. It is assumed that only a stereo difference signal V_i is transmitted; this is denoted as V_i' after modulation of the 38-kHz subcarrier. The model contains all the elements of the mono receiver in fig. 14. In the control loop now, however, demodulation of the 38-kHz subcarrier takes place in M_3 before the loop filter and remodulation to 38 kHz takes place in M_4 after the loop filter. In this model the 38-kHz generators in transmitter and receiver are assumed to be in phase; the signal delay τ in LP_1 and AP_1/M_2 makes it necessary to give the 38-kHz carrier the same delay to ensure correct phase in the demodulation in M_3.

then modulates VCO_2; the result is f_o. The output signal of mixer M_1 is the difference $f_i - f_o$. After filtering (in LP_1) and first demodulation (AP_1/M_2) this signal gives a delayed version $(V_i' - V_o')_{del}$ of the difference between the control signal V_i' of VCO_1 and V_o' of VCO_2. After a second demodulation in M_3, including compensation for the phase error introduced by LP_1 and AP_1/M_2, a delayed version $(V_i - V_o)_{del}$ of the difference between the difference signals V_i in the transmitter and V_o in the receiver is obtained. After filtering in loop filter LP_3 the output signal V_o is obtained in the same way as in the mono receiver.

Modulation by the 38-kHz subcarrier in M_0 and M_4 and corresponding inverse demodulation in M_3 take place without delay. The only delays in the receiver are the group delay of LP_1 and the delay in the demodulator AP_1/M_2. If the sum of these delays is equal to τ, and if the bandwidth of the loop filter LP_3 is equal to B, the transfer function V_o/V_i in the stereo receiver is the same as the transfer function in the highly simplified model of a mono receiver in fig. 15. The stability problem in the stereo receiver is therefore essentially the same as for the mono receiver and can therefore be solved with selectivity maintained.

Fig. 20 shows the calculated and measured transfer function V_o/V_i of a stereo receiver as in fig. 19 with open loop (OL) and closed loop (CL). There is good

A real stereo receiver must of course deal with the sum signal as well as the stereo difference signal. This means that the receiver must contain a loop as shown in fig. 14 in parallel with the loop shown in fig. 19.

In a practical receiver the subcarrier locking is obtained by means of a 19-kHz pilot tone, which has a delay τ with respect to the transmitter signal after demodulation. The regenerated 38-kHz subcarrier thus has the correct phase for the demodulation in M_3, but this phase must be advanced by an amount

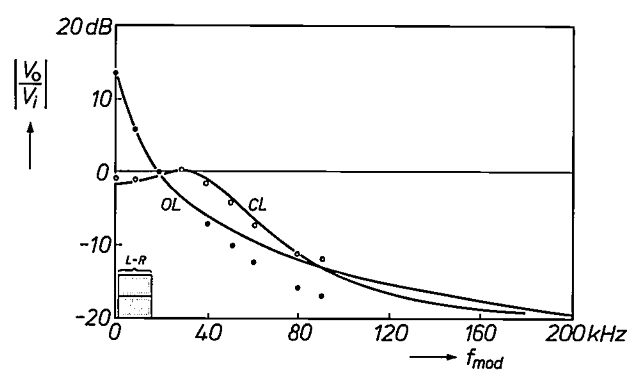

Fig. 20. Transfer function V_o/V_i in the open frequency-locked loop (OL) and in the closed loop (CL) in the stereo receiver; f_{mod} is the modulation frequency. The points indicate values calculated from a model like the one of fig. 19; the curves were measured for an experimental circuit. The stereo difference signal $L-R$ that the filter must pass has a bandwidth of 15 kHz.

$2\pi \times 38\,000 \times \tau$ for the remodulation in M_4. If, as assumed above, the i.f. filter LP_1 is a fourth-order Butterworth filter and the intermediate frequency is 70 kHz, the phase correction required at 38 kHz can even amount to about 90°. Without this phase correction the stereo difference signal modulated at 38 kHz would not contribute to the swing compression.

Block diagram of an integrated FM stereo receiver

The block diagram of the FM stereo receiver is given in *fig. 21*. The input stages are the same as in the FM mono receiver; the main additions are the more elaborate loop-filter network *LoopFi*, the subcarrier regenerator *SubcGen* and the matrix network *Matr*, in which the signals for the left and right channels are recovered from the sum and difference signals.

The stereo multiplex signal that appears at the output of the FM quadrature detector *FMQD* takes various paths in the loop-filter network. The lowpass filter LP_2 with a cut-off frequency of 5 kHz passes the baseband, but with an attenuation increasing by 6 dB/octave above 5 kHz (first-order filter). The multiplier circuit M_3 receives the 38-kHz subcarrier from the subcarrier regenerator and multiplies it by the multiplex signal. This results in demodulation of the difference signal, which is then limited in bandwidth in the first-order lowpass filter LP_3 (cut-off frequency also 5 kHz); after band limiting it is used for remodulating the amplitude of the subcarrier (multiplier M_4, amplitude modulation with suppressed carrier again). In this way a band filter centred on 38 kHz is obtained. In fact there is a temporary transformation from 38 kHz to 0 Hz, so that a simple lowpass filter is sufficient for limiting the passband.

A filtered multiplex signal is produced in an adder circuit and is returned via limiter/amplifier LA_2 to the

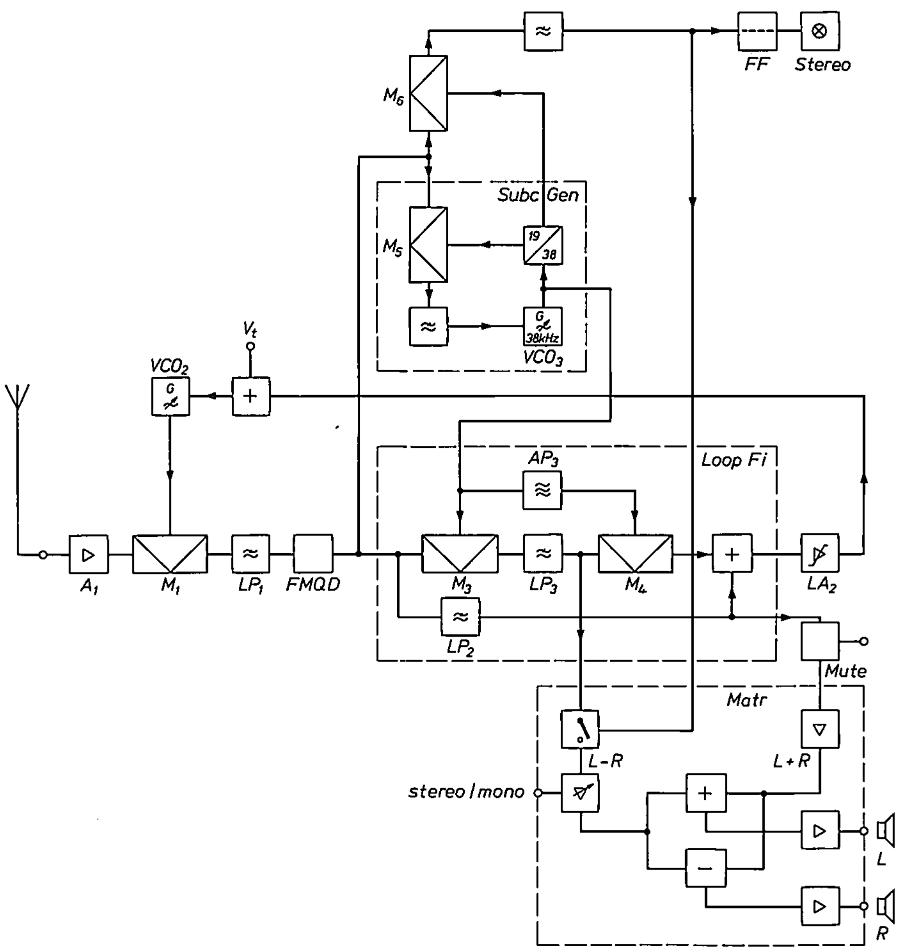

Fig. 21. Block diagram of an FM stereo receiver for integration. Many of the components are the same as in the mono receiver (see fig. 4). *FMQD* FM quadrature detector. *LoopFi* loop filter. *SubcGen* subcarrier generator. *Matr* matrix network (in which the left and right signals L and R are recovered from $L + R$ and $L - R$). *FF* bistable circuit. *Stereo* stereo indicator. *Mute* muting for off-tune signals.

voltage-controlled oscillator VCO_2 to give the required swing compression. The external tuning voltage V_t is added to the control signal.

The allpass filter AP_3 compensates for the delay of the 38-kHz carrier in the i.f. filter and the FM detector. As explained earlier, the phase of the subcarrier is advanced here by an amount τ, so that in M_4 the stereo difference signal modulates a carrier of the same phase as the subcarrier in the input signal of M_1.

The regeneration of the subcarrier for demodulating and remodulating the stereo difference signal in the loop-filter circuit *LoopFi* takes place in the stereo subcarrier regenerator *SubcGen*. This contains a phase-locked loop in which the 19-kHz pilot tone is multiplied by a 19-kHz signal, obtained by frequency division from a 38-kHz voltage-controlled oscillator (VCO_3). Any phase deviations cause a d.c. component in the product signal; this d.c. component is passed by a lowpass filter and is used to control the tuning of the oscillator.

The presence of the 19-kHz pilot tone is detected with the aid of multiplier M_6. Its output signal passes through a lowpass filter and controls a bistable circuit *FF*, which in turn switches on a stereo indicator; the same output signal operates the mono/stereo switch.

The demodulated stereo difference signal goes via this switch to a variable-gain amplifier and then to the matrix network, in which the signals for the left and right audio channels are derived from the difference and sum signals by addition and subtraction. The variable-gain amplifier makes it possible to vary the stereophonic effect (*stereo/mono*) in the reproduction. For a weak signal in noise it is better to attenuate the stereo difference signal; this reduces the stereo effect but improves the signal-to-noise ratio. On the other hand the stereo effect can be accentuated by amplifying the stereo difference signal more than proportionately, so that the sound sources seem to be further apart than the loudspeakers.

Summary. The article describes the integration of a complete FM mono receiver on a single chip. The external components are a single inductor (for the oscillator) and some capacitors and resistors. The use of a very low intermediate frequency (70 kHz) enables the i.f. bandpass filters to be replaced by lowpass filters. Harmonic distortion in such a limited i.f. band is avoided by compressing the frequency swing to ± 15 kHz. The IC is currently being marketed as type TDA 7000/7010T; an improved version, TDA 7020T, which operates with a lower battery voltage, is in pilot production. A monolithic FM stereo receiver is now being studied. In this receiver two frequency bands are selected from the stereo multiplex signal for oscillator control, since a frequency-locked loop with the full stereo multiplex bandwidth is not stable because of the group delay in a sufficiently selective i.f. filter.